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Preface

Broadcast television reception uses the radio spectrum. In the days before digital TV, spots, hash, snowstorms, colour and vision distortion and occasionally complete loss of picture were all symptoms of the same cause – electromagnetic interference. Interference still occurs, but its manifestation on digital pictures is different – pixellation, “blocking” in parts, and complete freezing of the picture are the modern curse, and might perhaps be regarded as more annoying than the analogue effects.

It is irritating for the viewer when the picture flickers or is wiped out during a crucial programme, just as it is irritating for a music lover who has carefully taped an important radio broadcast only to find that the quiet passages are ruined by the intrusion of the neighbour’s electric drill. It is far more critical when the emergency services are unable to communicate within a city centre because their radio signals are obscured by the electromagnetic “smog” emitted by thousands of computer terminals in the buildings around them.

The coexistence of all kinds of radio services, which use the electromagnetic spectrum to convey information, with technical processes and products from which electromagnetic energy is an undesirable by-product, creates the problem of what is known as “electromagnetic compatibility” (EMC). The solution is a compromise: radio services must allow for a certain degree of interference, but interfering emissions may not exceed a certain level, which normally involves measures to limit or suppress the interference energy. There is an economic trade-off inherent in this compromise. A lower level of interference would mean that less powerful transmitters were necessary, but the suppression costs would be higher. Alternatively, accept high power transmitters – with the attendant inefficient spectrum usage – in return for lower suppression costs. This economic balance has been tested over the past decades with the establishment of various standards for allowable levels of interference.

The problems of EMC are not limited to interference with radio services. Electronic equipment of all kinds is susceptible to malfunctions caused by external interference. This phenomenon is becoming more noticeable for two reasons: the greater pervasiveness and interaction of electronic products in all aspects of daily life, and the relatively worse immunity of modern equipment using plastic cases and microprocessors. Susceptibility to interference is now an issue for many kinds of electronic device, especially those whose continued correct operation is vital for safety or economic reasons. Automotive and aviation control systems are examples of the former category, banking and telecommunication networks are examples of the latter.

Recognizing the need for EMC protection measures and at the same time to eliminate the protectionist barriers to trade throughout the European Union, the European Commission adopted in 1989 a Directive “on the approximation of the laws of the Member States relating to electromagnetic compatibility”, otherwise known as the EMC Directive. It was revised, and a new edition was published at the end of 2004. It is discussed in detail in Chapter 2 of this book.

Every company that manufactures or imports electrical or electronic products should have in place measures that will enable its products to comply with the Directive. This means that an awareness of EMC must penetrate every part of the enterprise. EMC is undoubtedly affected by the design of the product, and the design and development group is where the awareness normally starts. But it also depends on the way an individual product is put together, so it affects the production department; by the way it is installed, so it affects the installation and service technicians, and the user documentation; it needs to be assured for each unit, so it affects the test department; it impacts the product's marketing strategy and sales literature, so it affects the sales and marketing departments; and it ultimately affects the viability and liabilities of the company, so it must be understood by the senior management.

There are various means of implanting and cultivating this awareness. An EMC training course would be a good starting point. You could bring in consultants to handle every aspect of the EMC compliance process, but for many products this would be expensive and cumbersome and would not necessarily result in improved awareness and expertise within the company where it was really needed. It would also be possible to send every appropriate member of staff on a training course. This would certainly raise awareness but it may not prove so effective in the long run, since EMC techniques also need to be practised to be properly understood.

A typical compromise is to nominate one person, or a group if the resources are available, to act as the centre of EMC expertise for the company. His, her or its responsibility should be to implement the requirements of the EMC Directive and any other EMC specifications to which the company may need to work. In the long term, it should also be to make the EMC centre redundant: to imbue a knowledge of EMC principles into each operating division so that they are a natural part of the functioning of that division. This, though, takes years of continuous oversight and education. Meanwhile, the tasks include:

- reviewing each new product design throughout the development and prototyping stages for adherence to EMC principles, and advising on design changes where necessary;
- devising and implementing an EMC test and control plan for each product;
- supervising pre-compliance and compliance tests both in house and in liaison with external test houses;
- maintaining an intimate knowledge of the EMC standards and legislation that apply to the company's products;
- liaising with marketing, sales, production, test, installation and servicing departments to ensure that their strategies are consistent with EMC requirements.

This gives an indication of the breadth of scope of the EMC engineer's job. It is comparable to that of the quality department, and indeed can sometimes be incorporated within that department.

Preface to the fourth edition

This fourth edition comes out 14 years after the first. In that time the EMC Directive has become fully functional and the vast majority of manufacturing companies have become familiar with it. But the EMC world has not stood still: the Directive has been substantially revamped, more new product standards have been published, new test methods have become established and much has been learned to improve old tests. Although Maxwell's laws haven't changed, there is more understanding of how best to apply them to maximize the compatibility of individual products. The onward and upward march of clock speeds and the shrinking of product, package and interconnect dimensions has continued. And so, even if you were familiar with earlier editions, you will find quite a lot of new material in this one.

This book is intended to help the work of the company's EMC centre. It seems to be serving its purpose: I have been pleasantly surprised by how widely it has been recommended. It can be used as a reference for the EMC engineer, as background reading for designers and technicians new to the subject, or as part of the armoury of the development group tackling a new project. It is structured into three parts. The first part (Chapters 1–5) discusses the European legislative framework now erected to encompass EMC and its near relative, the radio and telecom performance requirements. These chapters are mainly non-technical in nature. Chapter 1 introduces the subject of interference, and Chapter 2 goes on to discuss the provisions of the EMC Directive and the means of achieving compliance with it. Chapter 3 reviews the R&TTE Directive. Chapter 4 details the standards-making structure and describes the various standards that are now in existence and which are relevant for compliance with the Directive. Chapter 5 covers requirements for those sectors that are only peripherally affected by the R&TTED and EMCD: automotive, military, aerospace and rail transport.

Both design engineers and project managers need to have a feel for the tests to which their products will be submitted. As well as witnessing tests carried out by third party test houses, it is likely that many will be doing pre-compliance and full compliance tests themselves. The second part of the book looks at these areas. Chapter 6 covers the test methods for RF emissions that are laid down in the standards and which have to be followed both in-house and by external test houses. Chapter 7 does the same for the immunity tests: RF immunity, ESD and transient immunity. Chapter 8 considers the low frequency techniques, both mains input harmonics and flicker emissions, and immunity to magnetic field and voltage dips and interruptions. Finally, the tests do not happen by themselves: they must be planned, and Chapter 9 looks at this sadly neglected aspect of EMC compliance.

The third part of the book discusses techniques for achieving an acceptable EMC performance at minimum extra cost, at the design stage. It is usually possible to add screening and suppression components to an existing design to enable it to meet EMC standards. This brute force method is expensive, time-consuming and inefficient. Far better is to design to the appropriate principles from the start, so that the product has a good chance of achieving compliance first time, or if it doesn't then modifications are made easy to implement.

Chapter 10 covers the basic principles involved in coupling electromagnetic interference from a source to a victim. Chapter 11 looks at the techniques which can be applied before resorting to the more traditional methods of screening and suppression: attention to equipment and PCB layout and grounding, and Chapter 12 discusses choice of circuit configuration, components and software features. Chapter 13 carries on to detail the accepted "special" EMC techniques, which include cable configuration and termination, and filtering methods and components. Shielding theory and practice – the two are not always related – are covered in Chapter 14. Many products are used within systems, and so product designers need an appreciation of system-related EMC, which is encapsulated in Chapter 15. Chapter 16 discusses EMC management and control principles and, finally, a series of appendices gather together some reference information, along with a few case studies and a comment on CAD.

Much of the book grew out of course notes that were prepared for seminars on Design and Test for EMC, and I am grateful to those designers who attended these seminars and stimulated me to continually improve and hone the presentation. Many people have helped with its progress. I would particularly like to acknowledge the work of Prof. Andy Marvin and Dr John Dawson and their colleagues at York University, as well as that of Dr Jasper Goedbloed and Prof. Piet van der Laan. I have had a long and fruitful relationship with Schaffner EMC and TÜV Product Service, and am continually grateful especially to Ken Webb, Pete Dorey, David Riley, John Dearing, Ray Hughes and Nick Smith. I must also thank Dr Alastair Duffy, and particularly my consultant colleagues, Dave Imeson, Keith Armstrong and Phil Carter. As always the responsibility for this book remains the author's alone. I hope you find it useful.

Tim Williams
September 2006

Chapter 1

Introduction

1.1 What is EMC?

Electromagnetic interference (EMI) is a serious and increasing form of environmental pollution. Its effects range from minor annoyances due to crackles on broadcast reception, to potentially fatal accidents due to corruption of safety-critical control systems. Various forms of EMI may cause electrical and electronic malfunctions, can prevent the proper use of the radio frequency spectrum, can ignite flammable or other hazardous atmospheres, and may even have a direct effect on human tissue. As electronic systems penetrate more deeply into all aspects of society, so both the potential for interference effects and the potential for serious EMI-induced incidents increases.

Electromagnetic *compatibility* (EMC), then, is the absence of effects due to EMI. The definition of EMC, as it appears in the International Electrotechnical Vocabulary [152], is:

> The ability of a device, equipment or system to function satisfactorily in its electromagnetic environment without introducing intolerable electromagnetic disturbance to anything in that environment.

Some reported examples of electromagnetic *in*compatibility are:

- in Norfolk, various makes of car would "go crazy" when they passed a particular air defence radar installation – dashboard indicators dropping to zero or maximum, lights and engines cutting out;
- on one type of car, the central door locking and electric sunroof would operate when the car's mobile transmitter was used;
- new electronic push-button telephones installed near the Brookmans Park medium wave transmitter in North London were constantly afflicted with BBC radio programmes;
- interference to aeronautical safety communications at a US airport was traced to an electronic cash register a mile away;
- the instrument panel of a well-known airliner was said to carry the warning "ignore all instruments while transmitting HF";
- electronic point-of-sale units used in shoe, clothing and optician shops (where thick carpets and nylon-coated assistants were common) would experience lock up, false data and uncontrolled drawer openings;
- when a piezo-electric cigarette lighter was lit near the cabinet of a car park barrier control box, the radiated pulse caused the barrier to open and drivers were able to park free of charge;

- lowering the pantographs of electric locomotives at British Rail's Liverpool Street station interfered with newly installed signalling control equipment, causing the signals to "fail safe" to red;
- a digital TV set-top box initiated an air–sea rescue operation in Portsmouth harbour by creating an emission on the distress frequency;
- two Navy warships nearly collided when the radar transmissions of the frigate HMAS Anzac disabled the steering of the minehunter HMAS Huon, Huon passing ahead of Anzac "at close range".

Many other examples have been collected over the years; the "Banana Skins" column in the *EMC Journal*, collated by Keith Armstrong, is a fruitful source, and the EMC group of the former UK Radiocommunications Agency commissioned an EMC Awareness web page introducing the subject [203], which also contains a number of examples. Here are a few issues in more detail.

1.1.1 Portable electronic devices in aircraft

Mobile cellular telephones are rapidly establishing themselves, through their sheer proliferation, as a serious EMC threat. Passengers boarding civil airliners are now familiar with the announcement that the use of such devices is not permitted on board. They may be less familiar with why this is regarded as necessary. The IFALPA International Quarterly Review reported 97 EMI-related events due to passenger "carry-on" electronic devices since 1983. To quote the Review:

> ... By 1990, the number of people boarding aeroplanes with electronic devices had grown significantly and the low-voltage operation of modern aircraft digital electronics were potentially more susceptible to EMI.
>
> A look at the data during the last ten years indicates that the most likely time to experience EMI emissions is during cruise flight. This may be misleading, however. During the last three years, 43% of the reported events occurred in cruise flight while an almost equal percentage of events occurred in the climb and approach phases.
>
> Of particular note: during the last three years the number of events relating to computers, compact disc players, and phones has dramatically increased and these devices have been found to more likely cause interference with systems which control the flight of the aircraft.
>
> Recognising an apparent instrument or autopilot malfunction to be EMI related may be difficult or impossible in many situations. In some reported events the aircraft was off course but indications in the cockpit displayed on course. Air traffic controllers had to bring the course deviations to the attention of the crews. It is believed that there are EMI events happening that are not recognised as related to EMI and therefore not reported.

Particular points noted by the Review were that:

- events are on the rise;
- all phases of flight are exposed (not just cruise);
- many devices may cause EMI (phones, computers, CD players, video cameras, stereos);
- often there will be more than one device on a flight;
- passengers will turn on a device even after being told to turn it off[†];
- passengers will conceal usage of some devices (phones, computers);
- passengers will turn devices on just after take-off and just prior to landing;
- phones are a critical problem;

- specific device type and location should be recorded and reported by the crew;
- when the emitting EMI device is shut off, the aircraft systems return to normal operation (in the case of positioning errors a course change may be necessary);
- flight attendants should be briefed to recognize possible EMI devices.

In 2000, the Civil Aviation Authority carried out tests on two aircraft parked at Gatwick, which reinforces the ban on the use of mobile phones while the engine is running [64]. The tests revealed that interference levels varied with relatively small changes in the phone's location, and that the number of passengers on the flight could affect the level, since they absorbed some of the signal. Further testing has been done since, publicly reported in [198], which showed that at the GSM mobile frequencies it was possible to create the following interference effects:

- compass froze or overshot actual magnetic bearing;
- instability of indicators;
- digital VOR (VHF Omnidirectional Ranging, an aeronautical navigation aid using the VHF spectrum) navigation bearing display errors up to 5 degrees;
- VOR navigation To/From indicator reversal;
- VOR and ILS (Instrument Landing System, an aeronautical navigation aid using the VHF spectrum) course deviation indicator errors with and without a failure flag;
- reduced sensitivity of the ILS Localiser receiver;
- background noise on audio outputs.

Nevertheless, there is considerable public pressure to allow use of cellphones on board, and the fact that more often than not interference *doesn't* actually create problems has led to the perception that there *isn't* a problem. To deal with both of these issues, some airlines are trialling a system which actually allows cellphones to be used via a pico-cell base station on the aircraft, communicating via satellite with the ground networks. The EMC implication of this is that because the base station is very local, the phones are able to transmit at their minimum power, thus (hopefully) eliminating EMC interactions. Even so, there will almost certainly be restrictions on use, only above 10,000 feet and not during takeoff and landing.

1.1.2 Interference to medical devices

Another critical area with potentially life-threatening consequences is the EMC of electronic medical devices. A 1995 review article [122] described three incidents in detail and listed more than 100 EMI problems that were reported to the US Food & Drug Administration between 1979 and 1993. It states bluntly that:

> EMI-related performance degradation in electronic medical devices has resulted in deaths, serious injuries, and the administration of inappropriate and possibly life-threatening treatment.

† Especially if they regard their need for personal communication as more important than a mere request from the crew. [64] reports that an aircraft carrying a German foreign minister had to make an emergency landing "after key cockpit equipment cut out". It was claimed that mobile phone transmissions could be the only explanation and it was said that, "despite repeated requests from the crew, there were still a number of journalists and foreign office personnel using their phones".

The detailed case studies were as follows:

- apnea monitors: the essential function of an apnea monitor is to sound an alarm when breathing stops; the devices are used in hospitals and frequently prescribed for home use in the case of infants who either have exhibited or are at risk of experiencing prolonged apnea. After numerous reports of unexplained failure on the part of apnea monitors to alarm even upon death, their susceptibility to radiated RF was evaluated by the US Center for Devices and Radiological Health (CDRH). Most commercial apnea monitors were found to erroneously detect respiration when exposed to relatively low field strengths, a situation that could result in failure to alarm during apnea. Most monitors were found to be susceptible above 1V/m; one particular model was susceptible to pulsed fields above 0.05V/m.
- anaesthetic gas monitor: the CDRH received several reports of erroneous displays and latch-up of an anaesthetic gas monitor during surgery. None of the reports mentioned EMI as a possible cause. FDA investigators found that the manufacturer had a list of 13 complaint sites, and his own investigations revealed that interference from certain types of electrosurgery units disrupted the communication link between the monitor and a central mass spectrometer, causing the monitor to fail to display the concentration of anaesthetic gas in the operating room during surgery.
- powered wheelchairs: a QA manager at a large wheelchair manufacturer had received reports of powered wheelchairs spontaneously driving off kerbs or piers when police or fire vehicles, harbour patrol boats, or CB or amateur radios were in the vicinity. Though CDRH databases showed reports of unintended motion – in several cases involving serious injury – none of these incidents had been attributed to EMI. When CDRH investigated the EMI susceptibility of the motion controllers on various makes of powered wheelchairs and scooters, they discovered susceptibilities in the range of 5 to 15V/m. At the lower end of the range, the electric brakes would release, which could result in rolling if the chair happened to be stopped on an incline; as the field strength at a susceptible frequency was increased, the wheels would actually begin turning, with the speed being a function of field strength.

Another issue is the effect on hearing aids:

> The problem of interference to hearing aids has been known for some time. Digital mobile phones use a form of radio transmission called Time Division Multiple Access (TDMA), which works by switching the radio frequency carrier rapidly on and off. If a hearing aid user is close to a digital mobile telephone, this switching of the radio frequency carrier may be picked up on the circuitry of the hearing aid. Where interference occurs, this results in a buzzing noise which varies from very faint to maximum volume of the aid... [A specialist standards panel] has determined that, although digital mobile telephones are being looked at as the source of likely interference, all radio systems using TDMA or similar transmissions are likely to cause some interference.
>
> *BSI News December 1993*

These are all examples of the lack of a product's "fitness for purpose": that is, to operate correctly and safely in its intended environment, which includes the electromagnetic environment. There are clear safety implications in the reports.

1.1.2.1 Hospital and emergency service radio management

Many types of hospital equipment are susceptible to RF radiation from hand-portable mobile radio transmitters – diagnostic equipment such as ECGs, EEGs, pulse oximeters and other physiological monitoring equipment; and therapeutic equipment such as infusion pumps, ventilators and defibrillators. Physiological (patient-coupled) monitoring equipment is very sensitive and hence very susceptible, although for every device type, some models consistently perform better than average (they exhibit good EMC design). The type of modulation employed by the mobile transmitter can be significant. For example, an external pacemaker withstood a GSM signal (modulated at 217Hz) at 30V/m field strength, but TETRA modulation (17Hz) caused interference at 3V/m.

This is of particular concern for ambulances, which in Europe are mandated to use the TETRA system for emergency communications, but which also carry an array of patient-coupled instrumentation for life support purposes. This has led to the UK's Medicines and Healthcare Products Regulatory Agency (MHRA) recommending as follows [199]:

- the use of portable handsets and cellphones inside ambulances should be restricted;
- special precautions are needed if a patient with an external pacemaker is being transported;
- displaying warning notices, providing staff training, and relocating parking bays are possible actions if risks of interference prove unacceptable when emergency vehicles are parked immediately outside patient treatment areas;
- caution should be exercised when treating patients with medical devices at the scene of an accident if an emergency vehicle is nearby;
- mobile data terminals should be subjected to any restrictions which are locally applied to cellphones.

Various studies have tested medical devices and recommend that a distance of 1 to 1.5m be maintained between typical hand-portable transmitters and medical equipment. The MHRA tested 178 different models of medical device using a wide range of radio handsets. Overall, in 23% of tests medical devices suffered electromagnetic interference (EMI) from handsets. 43% of these interference incidents would have had a direct impact on patient care, and were rated as serious. Only 4% exhibited effects with cellphones at 1m distance, although at that distance emergency and security handsets had much greater effects [200].

The difficulty with controlling the use of radio communications in hospitals and other medical situations is well illustrated in the MHRA's guidance document [201], which itself refers to an ISO technical report on the subject [170]:

> Overly-restrictive policies may act as obstacles to beneficial technology and may not address the growing need for personal communication of patients, visitors and the workforce. At the other extreme, unmanaged use of mobile communications can place patients at risk.

The guidance stresses the need for an effective policy for healthcare providers to manage the use of the radio frequency spectrum in their own sites. This includes considering areas where medical devices will not be affected and therefore no restrictions apply, and other areas where authorized staff can use communication devices authorized by the hospital. Incidents should be reported when a medical device is suspected to have suffered electromagnetic interference.

1.1.2.2 Diathermy and electrosurgery

As well as radio communications, medical diathermy and electrosurgery are well known as a source of significant interference problems that most surgeons simply learn to cope with. Medical diathermy (tissue heating) used for physiotherapy typically operates at 27MHz with RF powers up to 400W, although modern pulsed diathermy uses average RF powers around 40W; but these levels are more than enough to interfere with many kinds of electro-medical equipment, particularly monitors.

1.1.3 Thermostats

Thermostats and other automatic switching contacts of all sorts are a major source of noise complaints, particularly when they are faulty. The former UK Radiocommunications Agency dealt with many cases of interference caused by thermostats or the radio-suppression components fitted to them. In about 90% of these cases, the interference is attributable to thermostats in gas boilers. It seems that, as these operate in a heat-stressed environment, they are prone to more rapid deterioration than other domestic thermostats such as room thermostats, cylinder thermostats and diverter switching valves. Sometimes the offending thermostat is found in the house that is suffering the interference, although there have been cases where the source of the interference has been found some distance away.

New domestic appliances are required to pass tests for "discontinuous disturbance" emissions (the current harmonized standard is EN 55014-1), but this does not guarantee that such products will remain noise-free after many years of operation. The limits for RF emissions are related in a complex way to the repetition rate and duration of the automatic switching event.

An example is the interference signal generated from a boiler gas control valve and its associated thermostat switching from stand-by to ON and vice versa. The low power single-phase arc causes a short burst of radiation. When the thermostat is malfunctioning this burst of radiation can be heard as a rough rasping noise which typically lasts for a few seconds but may last for 20 seconds or more. It repeats typically every 10 minutes but in some cases, a faulty thermostat may arc several times per minute. This kind of interference, which is intermittent in nature, is mostly noticed in relation to the reception of analogue TV signals at 470 to 850 MHz and sometimes on FM radio at 88–108 MHz.

Replacing the faulty thermostat will normally resolve the problem, but a better solution is to fit suppression to all such switching contacts. This prevents the arc forming at the instant of switching and if properly designed has the side effect of lengthening the contact life, but the added cost is usually viewed unfavourably by manufacturers.

1.1.4 The quacking duck

In a lighter vein, probably the least critical EMC problem this author has encountered is the case of the quacking duck: there is a toy for the under-fives, which is a fluffy duck with a speech synthesizer which is programmed to quack various nursery rhyme tunes. It does this when a certain spot (hiding a sensor) on the duck is pressed, and it shouldn't do it otherwise. Whilst it was in its Christmas wrapping in our house, which is not electrically noisy, it was silent. But when it was taken to our daughter's house and left in the kitchen on top of the fridge, next to the microwave oven, the Christmas present quacked apparently at random and with noone going near it. Some disconcerting moments arose before it was eventually explained to the family that this was just another case of bad EMC and that they shouldn't start to doubt their sanity!

1.2 Compatibility between and within systems

The threat of EMI is controlled by adopting the practices of electromagnetic compatibility (EMC), as defined earlier. The concept of EMC has two complementary aspects:

- it describes the ability of electrical and electronic systems to operate without interfering with other systems;
- it also describes the ability of such systems to operate as intended within a specified electromagnetic environment.

Thus it is closely related to the environment within which the system operates. Effective EMC requires that the system is designed, manufactured and tested with regard to its predicted operational electromagnetic environment: that is, the totality of electromagnetic phenomena existing at its location. Although the term "electromagnetic" tends to suggest an emphasis on high frequency field-related phenomena, in practice the definition of EMC encompasses all frequencies and coupling paths, from DC through mains supply frequencies to radio frequencies and microwaves. And "phenomena" is not restricted to radio-based phenomena but also transient events and power-related disturbances.

1.2.1 Intra-system EMC

There are two approaches to EMC. In one case the nature of the installation determines the approach. EMC is especially problematic when several electronic or electrical systems are packed into a very compact installation, such as on board aircraft, ships, satellites or other vehicles. In these cases susceptible systems may be located very close to powerful emitters and special precautions are needed to maintain compatibility. To do this cost-effectively calls for a detailed knowledge of both the installation circumstances and the characteristics of the emitters and their potential victims. Military, aerospace and vehicle EMC specifications have evolved to meet this need and are well established in their particular industry sectors.

This, then, can be characterized as an int*ra*-system approach: the EMC interactions occur between parts of the overall system, the whole of which is amenable to characterization. It may not be necessary or desirable to draw a boundary around individual products in the system, but rather to consider how they affect or are affected by other parts of the same system. Mitigation measures can be applied as easily, and sometimes more easily, at the system level as at the equipment level.

1.2.2 Inter-system EMC

The second approach assumes that the system will operate in an environment which is electromagnetically benign within certain limits, and that its proximity to other sensitive equipment will also be controlled within limits. This approach is appropriate for most electronics used in homes, offices and industry, and similar environments. So for example, most of the time a personal computer will not be operated in the vicinity of a high power radar transmitter, nor will it be put right next to a mobile radio receiving antenna. This allows a very broad set of limits to be placed on both the permissible emissions from a device and on the levels of disturbance within which the device should reasonably be expected to continue operating. These limits are directly related to the class of environment – domestic, commercial, industrial, etc. – for which the device is marketed. The limits and the methods of demonstrating that they have been met form the basis for a set of standards, some aimed at emissions and some at immunity, for the EMC performance of any given product *in isolation*. This makes it an int*er*-system approach rather than an int*ra*-system approach, and means that a necessary part of the process is defining the boundary of the product – easy for typical commercial electronic devices, harder when it comes to installations.

Note that compliance with such standards will not guarantee electromagnetic compatibility under all conditions. Rather, it establishes a probability (hopefully very high) that equipment will not cause interference nor be susceptible to it when operated under *typical* conditions. There will inevitably be some special circumstances under which proper EMC will not be attained – such as operating a computer within the near field of a powerful transmitter – and extra protection measures must be accepted.

1.2.3 When intra-system meets inter-system

Difficulty arises when these two approaches are confused one with the other, or at the interface where they meet. This can happen when commercial equipment is used in other environments, for instance on vehicles or in aircraft, and we get the issues referred to earlier, for instance with passenger electronic devices; the PED might be compliant with its normal requirements but that isn't necessarily relevant to its use in these different surroundings. Military projects might require commercial-off-the-shelf (COTS) products to be procured, but their EMC performance requirements are substantially mismatched to military needs. Grounding and bonding techniques which are necessary and appropriate for intra-system requirements can be misapplied to attempt to meet the EMC Directive.

From the product designer's point of view, many of the necessary techniques are similar or common to both approaches, but there are instances where they diverge and so we need to be clear about which approach is being considered in any given case.

1.3 The scope of EMC

1.3.1 Malfunction of control systems

Solid state and especially processor-based control systems have taken over many functions which were earlier the preserve of electromechanical or analogue equipment such as relay logic or proportional controllers. Rather than being hard-wired to perform a particular task, programmable electronic systems rely on a digital bus-linked architecture in which many signals are multiplexed onto a single hardware bus under software control. Not only is such a structure more susceptible to interference, because of the low level of energy needed to induce a change of state, but the effects of the interference are impossible to predict; a random pulse may or may not corrupt the operation depending on its timing with respect to the internal clock, the data that is being transferred and the program's execution state. Continuous interference may have no effect as long as it remains below the logic threshold, but when it increases further the processor operation will be completely disrupted. With increasing functional complexity comes the likelihood of system failure in complex and unexpected failure modes.

Clearly the consequences of interference to control systems will depend on the value of the process that is being controlled. In some cases disruption of control may be no more than a nuisance, in others it may be economically damaging or even life threatening. The level of effort that is put into assuring compatibility will depend on the expected consequences of failure.

1.3.1.1 Phenomena

Electromagnetic phenomena which can be expected to interfere with control systems are:

- supply voltage interruptions, dips, surges and fluctuations;
- fast transient overvoltages (spikes and surges) on supply, signal and control lines;
- radio frequency fields, both pulsed (radar) and continuous, coupled directly into the equipment or onto its connected cables;
- electrostatic discharge (ESD) from a charged object or person;
- low frequency magnetic or electric fields.

The stress levels that can occur for each of these phenomena depend largely on the local environment, with industrial areas expected to be more stressful than, say, residential or commercial. Special environments such as military, automotive and aerospace can be expected to have even higher levels of some phenomena. The levels are reflected in the standards for immunity testing, which are covered in later chapters.

Note that we are not directly concerned with the phenomenon of component damage due to ESD, which is mainly a problem of electronic production. Once the components are assembled into a unit they are protected from such damage unless the design is particularly lax. But an ESD transient can corrupt the operation of a microprocessor or clocked circuit just as a transient coupled into the supply or signal ports can, without actually damaging any components (although this may also occur), and this is properly an EMC phenomenon.

1.3.1.2 Software

Malfunctions due to faulty software may often be confused with those due to EMI. Especially with real-time systems, transient coincidences of external conditions with critical software execution states can cause operational failure which is difficult or impossible to replicate, and the fault may survive development testing to remain latent for years in fielded equipment. The symptoms – system crashes, incorrect operation or faulty data – can be identical to those induced by EMI. In fact you may only be able to distinguish faulty software from poor EMC by characterizing the environment in which the system is installed.

1.3.2 Immunity of data and programme processing

A very large class of EMC problems comes not with the impact of disturbance on electronic controls, but on various types of system which process audio, video material or data – telecomms and consumer applications being the main areas. Interference here manifests as corruption of data or degradation of the audio or video programme that is being enjoyed.

The phenomena which cause the effects are the same as those discussed above in section 1.3.1.1. In the worst case these can cause complete loss of function; as before, in digitally-based systems this is usually a result of transient interference interrupting the processor program control. But before that level is reached, lower levels of transients and continuous radio frequency disturbances can affect the analogue sections of the equipment and cause annoying effects on the programme material itself, particularly demodulation effects which appear as distortion of the video picture or extraneous audible noise on an audio programme. The buzzing noise that occurs when you place a mobile cellphone next to your radio or TV set, which warns you that it is about to ring, or when you are engaged on a call, is an example of this phenomenon. Other examples which were more prevalent in the days before RF immunity requirements became well established are the pickup of taxi mobile transmitters on your home hi-fi, or the susceptibility of early electronic telephones to the programmes of high power local broadcasting transmitters.

Because of the heavy reliance of developed societies on data communications, the impact of interference on telecommunications networks is a critical issue and the network equipment itself must be capable of withstanding the full range of phenomena at quite substantial levels; for instance the lightning surge immunity of exchange equipment and mobile phone base stations, both at the equipment and the system level, is carefully specified by the network operators.

1.3.3 Interference with radio reception

Bona fide users of the radio spectrum have a right to expect their use not to be affected by the operation of equipment which is nothing to do with them. Typically, received signal strengths of wanted signals vary from less than a microvolt to more than a millivolt, at the receiver input. If an interfering signal is present on the same channel as the wanted signal then the wanted signal will be obliterated if the interference is of a similar or greater amplitude. The acceptable level of co-channel interference (the "protection factor") is determined by the wanted programme content and by the nature of the interference. Continuous interference on a high fidelity broadcast signal would be unacceptable at very low levels, whereas a communications channel carrying compressed voice signals can tolerate relatively high levels of impulsive or transient interference.

Analogue sound and TV broadcasting are being replaced by digital broadcasting like Digital Radio Mondial (DRM) which is intended to replace the AM radio in the MF and HF bands, Digital Audio Broadcasting (DAB or T-DAB) operated in the VHF and UHF bands, and Digital Video Broadcasting Terrestrial (DVB-T) operated in the UHF bands. These digital radio services require lower RF protection ratios (17dB for DRM, 20dB for DVB-T and 28dB for DAB) than radio services with analogue modulation (which need protection ratios of about 27dB for AM, about 48dB for FM and about 58dB for TV). On the other hand, the transition between no interference and unacceptable interference is narrower than for analogue modulation: digital systems do not fail gracefully.

1.3.3.1 Setting the limits

Radiated interference, whether intentional or not, decreases in strength with distance from the source. For radiated fields in free space, the decrease is inversely proportional to the distance provided that the measurement is made in the far field (see section 10.1.4.2 for a discussion of near and far fields). As ground irregularity and clutter increase, the fields will be further reduced because of shadowing, absorption, scattering, divergence and defocussing of the diffracted waves. Annex D of CISPR 11

[161] suggests that for distances greater than 30m over the frequency range 30 to 300MHz, the median field strength varies as $1/d^n$ where n varies from 1.3 for open country to 2.8 for heavily built-up urban areas (n = 1 for ideal, far field, free-space propagation). An average value of n = 2.2 can be taken for approximate estimations; thus, increasing the separation by ten times would give a drop in interfering signal strength of 44dB. In the near field – for a frequency of 30MHz the transition distance is 1.6m, extending further away proportionally as the frequency decreases – the factor n depends also on the detailed characteristics of the emitting source and cannot be generalized, but remains theoretically in the range 2 to 3.

Below 30MHz the dominant method of coupling out of the interfering equipment is via its connected cables, and therefore the radiated field limits are translated into equivalent voltage or current levels that, when present on the cables, correspond to a similar level of threat to HF and MF reception. Electric or magnetic field coupling to nearby antennas in this frequency range is in most cases of minor importance compared with conduction coupling to the mains input of the affected receiver.

Limits for unintentional emissions are based on the acceptable interfering field strength that is present at the receiver – that is, the minimum wanted signal strength for a particular service modified by the protection ratio – when a nominal distance separates it from the emitter. This will not protect the reception of very weak wanted signals nor will it protect against the close proximity of an interfering source, but it will cover the majority of interference cases and this approach is taken in all those standards for emission limits that have been published for commercial equipment by CISPR (see Chapter 4). CISPR 16-4-4 gives an account of how such limits are derived, including the statistical basis for the probability of interference occurring.

Historically, the rationale for particular limits has relied upon statistics of interference complaints. A draft amendment to CISPR 16-4-4 [168] points out that with the advent of digital communications this is becoming harder to rely on:

> If a digital mobile phone or a wireless LAN receiver cannot receive the signal from the nearest base station or access point because of an unwanted emission from a nearby equipment, the user will never suspect this equipment and will even not consider the possibility of an interference to occur. He will assume that the coverage of the network is poor and will move to another place to make his call or to get his connection. Furthermore, as these systems are generally frequency agile, if one channel is interfered with, the system will choose another channel, but if all other channels are occupied, then the phone will indicate that the network is busy, and once again, the user will think the network capacity is not large enough to accommodate his call, but he will never suspect an EMC problem.
>
> Generally for analogue systems, one can hear the interference. With digital and mobile systems, interference is much less noticeable† (muting in audio reception, or frozen images on the TV set for DVB). In addition, modern digital modulations implement complex escape mechanisms (data error correction, frequency agile systems...) so that the system can already be permanently affected from an EMC point of view before an interference case actually is detected.
>
> The evolutions detailed above – generalisation of mobile use of radio receivers and move from analogue to digital radio services – will not reduce the number of interference situations, but continues to decrease the probability of getting significant numbers of interference complaints indicating an existing EMC problem.

1.3.3.2 Malfunction versus spectrum protection

It should be clear from the foregoing discussion that RF emission limits are not determined by the need to guard against malfunction of equipment which is not itself a

† It's questionable that the interference is less "noticeable"; on the other hand it is harder to attribute the noticeable effects unequivocally to interference.

radio receiver. As discussed in the last section, malfunction requires fairly high energy levels – RF field strengths in the region of 1–10 volts per metre for example. Protection of the spectrum for radio use is needed at much lower levels, of the order of 10–100 microvolts per metre – ten to a hundred thousand times lower. RF incompatibility between two pieces of equipment neither of which intentionally uses the radio spectrum is very rare. Normally, equipment immunity is required from the local fields of intentional radio transmitters, and unintentional emissions must be limited to protect the operation of nearby intentional radio receivers. The two principal EMC aspects of emissions and immunity therefore address two different issues.

1.3.3.3 Free radiation frequencies

Certain types of equipment, collectively known as industrial, scientific and medical (ISM) equipment, generate high levels of RF energy but use it for purposes other than communication. Medical diathermy and RF heating apparatus are examples. To place blanket emissions limits on this equipment would be unrealistic. In fact, the International Telecommunications Union (ITU) has designated a number of frequencies specifically for this purpose, and equipment using only these frequencies (colloquially known as the "free radiation" frequencies) is not subject to emission restrictions. Table 1.1 lists these frequencies.

Centre frequency, MHz	Frequency range, MHz	
6.780	6.765 – 6.795	*
13.560	13.553 – 13.567	
27.120	26.957 – 27.283	
40.680	40.66 – 40.70	
433.920	433.05 – 434.79	*
2,450	2,400 – 2,500	
5,800	5,725 – 5,875	
24,125	24,000 – 24,250	
61,250	61,000 – 61,500	*
122,500	122,000 – 123,000	*
245,000	244,000 – 246,000	*

* maximum radiation limit under consideration

The frequency range 902 – 928MHz is also allowed in Region 2 only (the Americas, Canada and Greenland).

Table 1.1 ITU designated industrial, scientific and medical free radiation frequencies (Source: CISPR 11 [161])

1.3.3.4 Co-channel interference, spurious emissions and blocking

A further problem with radio communications, often regarded as an EMC issue although it will not be treated in this book, is the problem of co-channel interference from unwanted transmissions. This is caused when two radio systems are authorized to use the same frequency on the basis that there is sufficient distance between the systems, but abnormal propagation conditions increase the signal strengths to the point at which interference is noticeable. This is essentially an issue of spectrum utilization.

A transmitted signal may also overload the input stages of a nearby receiver which is tuned to a different frequency and cause desensitization or distortion of the wanted signal. Transmitter outputs themselves will have spurious frequency components present as well as the authorized frequency, and transmitter type approval has to set limits on these spurious levels. This is more properly viewed as a spectrum management concern and is sometimes called “mutual interference”.

Although this book doesn’t consider the subject further, it is important particularly for interactions between systems on the same platform, such as telecommunication installations sharing one mast, or multiple transmitter/receiver installations in one vehicle. Issues which have to be addressed are:

- spurious and harmonic emissions from the transmitters: emissions of any frequency which is not intended;
- correct frequency and power control of transmitters;
- spurious responses of the receivers: reception of any frequency which is not intended;
- adjacent channel blocking of receivers: unintended deafness induced by loud signals on nearby channels;
- sidelobe antenna patterns of both transmitters and receivers;
- intermodulation effects in transmitters, receivers and nearby structures, which create different frequencies from the interaction of existing ones.

1.3.4 Disturbances of the mains supply

Mains electricity suffers a variety of disturbing effects during its distribution. These may be caused by sources in the supply network or by other users, or by other loads within the same installation. A pure, uninterrupted supply would not be cost-effective; the balance between the cost of the supply and its quality is determined by national regulatory requirements, tempered by the experience of the supply utilities. Typical disturbances (see also section 10.3.5) are:

- *medium-term voltage variations*: the distribution network has a finite source impedance and varying loads will affect the terminal voltage. Not including voltage drops within the customer’s premises, an allowance of ±10% on the nominal voltage will cover normal variations in Europe.
- *voltage fluctuations*: short-term (sub-second) fluctuations with quite small amplitudes are annoyingly perceptible as flicker on electric lighting, though they are comfortably ignored by electronic power supply circuits. Generation of flicker by high power load switching is subject to regulatory control.
- *voltage dips and interruptions*: faults on power distribution systems cause up to 100% voltage drops but are cleared quickly and automatically by protection devices, and throughout the rest of the system the voltage immediately recovers. Most consumers therefore see a short voltage dip. The frequency of occurrence of such dips depends on location and seasonal factors.
- *waveform distortion*: at source, the AC mains is generated as a pure sine wave but the reactive impedance of the distribution network together with the harmonic currents drawn by non-linear loads causes voltage distortion. Power converters and electronic power supplies are important contributors to non-linear loading. Harmonic distortion may actually be worse at points

remote from the non-linear load because of resonances in the network components. Not only must non-linear harmonic currents be limited but equipment should be capable of operating with up to 10% total harmonic distortion in the supply waveform.

- *transients and surges*: switching operations generate transients of a few hundred volts as a result of current interruption in an inductive circuit. These transients normally occur in bursts and have risetimes of no more than a few nanoseconds, although the finite bandwidth of the distribution network will quickly attenuate all but local sources. Rarer high amplitude spikes in excess of 2kV may be observed due to fault conditions. Even higher voltage surges due to lightning strikes occur, most frequently on exposed overhead line distribution systems in rural areas.

All these sources of disturbance can cause malfunction in systems and equipment that do not have adequate immunity.

Mains signalling

A further source of incompatibility arises from the use of the mains distribution network as a telecommunications medium, or mains signalling (MS). MS superimposes signals on the mains in the frequency band from 3kHz to 150kHz and is used both by the supply industry itself and by consumers. Unfortunately this is also the frequency band in which electronic power converters – not just switch-mode power supplies, but variable speed motor drives, induction heaters, fluorescent lamp inverters and similar products – operate to their best efficiency. There are at present very few pan-European standards which regulate conducted emissions on the mains below 150kHz, although EN 50065-1 sets the frequency allocations and output and interference limits for MS equipment itself. Overall, compatibility problems between MS systems and such power conversion equipment can be expected to increase.

1.3.5 Power line telecoms

Power Line Telecommunication (PLT or PLC, Power Line Communications, or Broadband over Power Line, BPL, in the USA) is a means of transmitting broadband data over the installed base of mains electricity supply cables. It can be used in two ways:

- Access to the home, to deliver the data connection from the service provider;
- Networking within the home, for data interconnection between mains-connected devices.

The two applications use different frequency ranges: low frequency (1.6–10MHz) for access, and high frequency (10–30MHz) for in-home, as specified for Europe in ETSI TS 101 867. The generally accepted power level for adequate operation of a PLT system is –50 to –40dBm/Hz. Measured in a 9kHz bandwidth, as is the normal method for interference measurements at these frequencies, this implies a power level of around –10 to 0dBm, which across the differential 100 ohm resistance of the power network is 100–110dBμV (0.1–0.32V). This compares with the allowed levels for conducted emissions in the domestic environment, with which most if not all electronic product designers are familiar, of 60dBμV in a comparable frequency range – one hundred times lower.

1.3.5.1 The politics of PLT

Because it can deliver domestic broadband access as an alternative to other providers such as cable and telephone companies, PLT is viewed favourably by regulators on the grounds of extending competition. The local loop, or the “last mile” (delivery of the broadband data finally into the home or office), appears as a bottleneck in the process of liberalizing the competitive environment for the telecommunications infrastructure, particularly in breaking the perceived stranglehold of the “incumbents” (pre-existing telecom providers). Hence any technology which promises to unblock this bottleneck is encouraged by the European authorities. PLT is clearly such a technology.

But meanwhile, some European nations have implemented regulations which would allow them to control it if there was any threat of interference becoming widespread. In Germany, the standard NB30 put down radiated emissions limits in the 1.6–30MHz range. In the UK, the Radiocommunications Agency standard MPT1570 was also published, though it covered a lower frequency range. Naturally, this put a brake on PLT activity in these countries, since investors were wary of supporting systems which might quickly turn out to be illegal, and it also meant that there were differences in approach across the European Union.

In 2001 the European Commission placed a mandate on the standard bodies ETSI and CENELEC (mandate M/313) to create a standard for the EMC of Telecommunications Networks. This was addressed by a Joint Working Group of the two bodies but the difficulties involved, particularly that of finding agreement on a set of limits for radiated emissions from the network which would satisfy all participants, meant that the work on it eventually stalled. The Commission subsequently issued a Recommendation in April 2005 [191] which includes the following wording:

> 2. ... Member States should remove any unjustified regulatory obstacles, in particular from utility companies, on the deployment of broadband powerline communications systems and the provision of electronic communications services over such systems.
>
> 3. Until standards to be used for gaining presumption of conformity for powerline communications systems have been harmonised under Directive 89/336/EEC, Member States should consider as compliant with that Directive a powerline communications system which is:
>
> – made up of equipment compliant with the Directive and used for its intended purpose,
>
> – installed and operated according to good engineering practices designed to meet the essential requirements of the Directive.
>
> The documentation on good engineering practices should be held at the disposal of the relevant national authorities for inspection purposes as long as the system is in operation.

We will meet the phrase “good engineering practices” elsewhere (section 2.2.5.4) in the context of fixed installations.

1.3.5.2 The use of the HF spectrum

The slice of spectrum from about 1 to 30MHz (MF and HF) is unique in that it can support long distance communication, and so it is important to broadcasters and many other users. Sky-wave propagation in the HF bands enables an international broadcaster to reach a target country without having a transmitter within the target area. This has political consequences, since it means that an audience can be reached without the co-operation of that country's authorities – which cannot be said for other kinds of access, including any kind of internet delivery. The BBC's World Service, for instance, is broadcast on several HF frequencies and is heard by many people in countries that have no free media of their own.

As well as broadcasting, aeronautical and marine communications use the HF band for long-distance communication when the mobile station is out of reach of ground-based VHF stations, which is a large proportion of their journeys.

1.3.5.3 PLT's interference capability

Interference from PLT systems stands outside the general regime of interference control, which as we will see applies limits to the amount of noise injected onto the mains supply. The principal emissions are deliberately injected onto the supply wiring, rather than accidentally as is the case with other sources such as fluorescent light inverters or computer power supplies. From access-PLT systems, the interference will be largely continuous rather than intermittent, and will potentially affect all households being supplied from a substation in a PLT-active zone, whether they are a subscriber or not. Even in-home systems could interfere with other parties connected to the same service entrance.

The interference signal will stretch broadband across the whole of the spectrum occupied by the modem's output, and within a given region of spectrum it will be impossible to tune it out. In the quiescent state some systems will create a pulsing type of signal which may or may not be subjectively less annoying than the continuous noise which occurs when the system is actually passing data. Some systems may use low-frequency carriers such that a continuous audible tone is present across the spectrum.

Dependence on quality of wiring

The mains supply wiring both to and within a domestic house was never intended to carry high frequencies. At some frequencies the signal may be transmitted with little loss, but at others the attenuation can be severe, and this characteristic can change with time as users plug various appliances into the mains supply. This means that in order to work at all, the amplitude of the signal must be high enough to ride over any interference already present on the network, and must adapt to time-dependent changes in this interference and the network attenuation. Current-generation PLT systems are designed to do this.

A critical parameter which determines the amount of radiation that the mains wiring creates is the “Longitudinal Conversion Loss” (LCL) of the cable, discussed in section 13.1.9.1. Simply put, this is the ratio between the signal level which appears across the wires, intentionally, due to the desired data transmission, and which to a first order does not radiate; and the signal level in common mode – all wires together – which represents the leakiness of the cable and which contributes the lion's share of the radiation. Data cables which carry broadband signals, of which Ethernet is the most typical example, are very tightly specified for good LCL, which ensures that the RF leakage from the data signal is kept low. This is not the case for mains wiring. The most important aspect of cable design which affects LCL is the physical balance of the two wires that make up the cable. Each conductor must be tightly coupled to the other so that they cancel each other’s interaction with the environment. Data cables are tightly twisted in a controlled way to achieve this. Not only is mains wiring not controlled in this way, it is commonly installed in direct contravention of this principle.

1.3.5.4 Is PLT the same as other interferers?

PLT supporters propose that there should be parity (at least) between the emissions compliance requirements that a PLT system has to meet, and those applied to other devices such as information technology, lighting, or household appliances. CISPR conducted limits, it is said, have been adequate to protect the HF spectrum so far and therefore any system limits should be no more onerous than levels derived from these. This argument overlooks a number of important points:

- a victim won't be able to get away from PLT interference. When a whole street or a whole building is wired for PLT, it will be pervasive and re-positioning the victim will not work. CISPR limits assume that mitigation by separation from a local interferer is possible;
- PLT is always on. CISPR limits incorporate a relaxation which takes into account the probability of non-coincidence in time of source and victim. For PLT, this relaxation must be ignored;
- EMC engineers know that most products which comply with CISPR conducted limits do so with a good margin, often at least 20dB, in the frequency range above 2MHz. If CISPR limits do indeed protect HF reception, this factor should not be overlooked, since such a margin will not be enjoyed by a PLT product.

In fact, PLT modems are unable to operate anywhere near the mains conducted emissions limits in force in CISPR at the moment.

1.3.5.5 Cumulative effects

The threat to victim receivers in close proximity to the PLT system is not the only one that concerns radio administrations. If PLT were to be widely implemented within any country, the total radiated power available would be sufficient to increase the radio noise floor at distances remote from the source, potentially even in other countries. If, say, an entire city was to be wired for PLT, this could form an aggregate transmitter whose RF energy was reflected from the ionosphere and illuminated a continent. In addition, an aircraft flying over such a city might find that its ability to receive HF signals was curtailed. The UK's Civil Aviation Authority has expressed its concern that "aeronautical services are under threat from cabled telecommunications services." Established HF propagation models exist for this phenomenon and a number of studies have been carried out to try and model the possible outcome. The problem with any such study is that for the time being it must remain theoretical, since it's impossible to validate the radiation (as opposed to propagation) models used for prediction until there are sufficient installed systems to be statistically acceptable; but by then the roll out may be so advanced that it will be impossible to stop it.

1.3.5.6 Compliance status of PLT devices

The EC's Recommendation on PLT quoted above refers to a system being "made up of equipment compliant with the Directive". Here is the nub of the question: how can PLT modems be made compliant with the EMC Directive? It is the case that some PLT modems are already on the market in Europe and are CE Marked, which means that their manufacturers believe that they meet the essential requirements of the EMC Directive. But there are no standards specifically for such devices and for now, no such device could actually meet the usual standards. In fact, existing products appear to use the Technical Construction File route (under the first EMC Directive, see section 2.1.3) which implies that a Competent Body has issued a certificate of compliance with the essential requirements. It is difficult to know how this has been justified.

1.3.5.7 Opening the floodgates

It must be assumed that the mains supply already carries noise from other apparatus which may approach the usual conducted emission limits, even if everything connected is in full compliance with the EMC Directive. For PLT to operate at all, its signals must be greater than this minimum noise level, almost by definition; and so it must breach these limits, again by definition. This is indeed so, by several tens of dB, and if it were not, PLT could not operate. Yet all other mains-connected equipment, such as ITE, medical products, household appliances, lighting and so forth, is subject to the standard limits.

What is to prevent the manufacturers of such equipment, which after all forms the vast bulk of products placed on the market within the EU, from demanding to know why PLT has received such special treatment? If PLT can flagrantly flout the limits, they would say, so can we. But of course, were they to do that, it would open the floodgates to an uncontrolled escalation of interference on the mains wires. More bluntly, it would drive a horse and cart through the principles of interference control established over decades.

This exposes a contradiction at the core of the case for PLT. It can only operate if it is indeed granted special status to apply RF disturbances to the mains lines. It must, in fact, be regarded as a special case in the context of the EMC Directive. It cannot possibly comply with the requirement not to generate an electromagnetic disturbance exceeding "a level allowing radio and telecommunications equipment and other apparatus to operate as intended" because, since the limits are set to achieve this requirement, it must itself exceed those limits and therefore breach the requirement.

In fact, it seems at the time of writing that PLT is losing its attractiveness both to the regulators and to its hopeful operators, mainly because of the regulatory EMC problems rehearsed above. Probably, it will be relegated to a curiosity in years to come, superseded by less polluting, more reliable and more effective wired and wireless ways of delivering broadband internet access.

1.3.6 Other EMC issues

The issues discussed above are those which directly affect product design to meet commercial EMC requirements, but there are some other aspects which should be mentioned briefly.

1.3.6.1 EEDs and flammable atmospheres

The first is the hazard of ignition of flammable atmospheres in petrochemical plant, or the detonation of electro-explosive devices in places such as quarries, due to incident RF energy. A strong electromagnetic field will induce currents in large metal structures which behave as receiving antennas. A spark will occur if two such structures are in intermittent contact or are separated. If flammable vapour is present at the location of the spark, and if the spark has sufficient energy, the vapour will be ignited. Different vapours have different minimum ignition energies, hydrogen/air being the most sensitive. The energy present in the spark depends on the field strength, and hence on the distance from the transmitter, and on the antenna efficiency of the metal structure. BS 6656 [173] discusses the nature of the hazard and presents guidelines for its mitigation.

Similarly, electro-explosive devices (EEDs) are typically connected to their source of power for detonation by a long wire, which can behave as an antenna. Currents induced in it by a nearby transmitter could cause the charges to explode prematurely if the field was strong enough. As with ignition of flammable atmospheres, the risk of premature detonation depends on the separation distance from the transmitter and the efficiency of the receiving wire. EEDs can if necessary be filtered to reduce their susceptibility to RF energy. BS 6657 [174] discusses the hazard to EEDs.

1.3.6.2 Data security

The second aspect of EMC is the security of confidential data. Low-level RF emissions from data-processing equipment may be modulated with the information that the equipment is carrying – for instance, the video signal that is fed to the screen of a VDU. These signals could be detected by third parties with sensitive equipment located outside a secure area and demodulated for their own purposes, thus compromising the security of the overall system. This threat is already well recognized by government agencies and specifications for emission control, under the Tempest scheme, have been established for many years. Commercial institutions, particularly in the finance sector, are now beginning to become aware of the problem.

1.3.6.3 Electromagnetic weapons

The idea that an intense broadband radiated pulse could be generated intentionally, and used to upset the operation of all potentially susceptible electronics within a certain range, is gaining credence. Because of the almost universal social reliance on electronic systems, an attack that simultaneously crashed many computer networks could indeed have substantial consequences. It is known that US and other military researchers are working on such technology, but we can also imagine less sophisticated devices being within reach of many other organizations or individuals.

The more sensationalist press, of course, has a field day with this idea – phrases such as "frying computer chips" are used with abandon. Realistically, the amount of energy needed to generate a wide-area pulse would be so enormous that only disruption, not damage, is at all likely. This is precisely the effect of a high altitude nuclear explosion, which generates a sub-nanosecond nuclear electromagnetic pulse (NEMP) that is disruptive over an area of hundreds of square kilometres. The idea that attracts military researchers now is to do this more discreetly. The limitation of any such weapon is its uncertainty. Unless you know exactly what kind of electronics you are attacking, and how well protected it is, it is hard to predict the damage that the weapon will cause. Equipment that is immune to a local electrostatic discharge (ESD, as described in these pages) is likely to have good immunity to electromagnetic warfare.

On the other hand, radio communications are easy to disrupt by jamming transmissions and this subject is well established in the military canon; there are military aircraft that are specifically designed to carry high power broadband jammers and it is not hard to imagine that they are used with some enthusiasm in "theatres of war" (it is said that one such aircraft suffered an engine shutdown when its jammers were turned on in earnest). Under such circumstances, the issue of the commercial application of radiated emission limits is of no more than academic interest.

1.3.7 The compatibility gap

The increasing susceptibility of electronic equipment to electromagnetic influences is being paralleled by an increasing pollution of the electromagnetic environment. Susceptibility is a function partly of the adoption of VLSI technology in the form of microprocessors, both to achieve new tasks and for tasks that were previously tackled by electromechanical or analogue means, and the accompanying reduction in the energy required of potentially disturbing factors. It is also a function of the increased penetration of radio communications, and the greater opportunities for interference to radio reception that result from the co-location of unintentional emitters and radio receivers.

At the same time more radio communications mean more transmitters and an increase in the average RF field strengths to which equipment is exposed. A study has been reported [30] which quantified this exposure for a single site at Baden, Switzerland, for one year; this found the background field strength in the shortwave band regularly approaching, and occasionally exceeding, levels of 1V/m. Also, the proliferation of digital electronics means an increase in low-level emissions which affect radio reception, aptly described as a form of electromagnetic "smog".

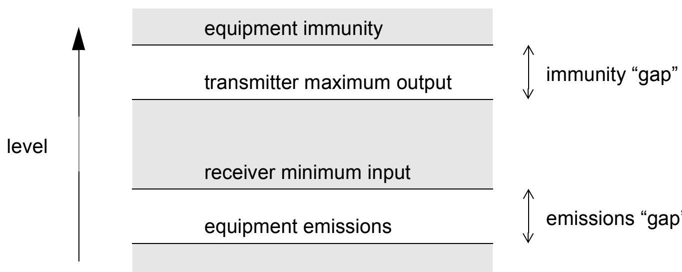


Figure 1.1 The EMC gap

These concepts can be graphically presented in the form of a narrowing electromagnetic compatibility gap, as in Figure 1.1. This "gap" is of course conceptual rather than absolute, and the phenomena defined as emissions and those defined as immunity do not mutually interact except in rare cases. But the maintenance of some artificially-defined gap between equipment immunity and radio transmissions on the one hand, and equipment emissions and radio reception on the other, is the purpose of the application of EMC standards, and is one result of the enforcement of the EMC Directive.

1.4 Electromagnetic fields and human health

This book is about EMC in electronic products, but one subject which any EMC engineer is regularly quizzed about is the related issue of the direct health hazards of electromagnetic fields. Apart from its application under the R&TTE Directive and relevance to radiated RF immunity tests at high levels, it's not a subject that will take up a lot of space in this book.

It is undoubtedly the case that high levels of electromagnetic energy can create effects in human tissue. This then implies that below some level the EM energy can be regarded as "safe" and above it as "hazardous", and that exposure to it should be controlled. The exposure levels to which most if not all regulatory limits adhere are laid down in the ICNIRP Guidelines [196].

1.4.1 ICNIRP basic restrictions

The International Commission on Non-Ionising Radiation Protection is a non-governmental expert body and its guidance document refers to the known short-term and immediate health effects. These are:

- stimulation of peripheral nerves and muscles;
- shocks and burns through touching conducting parts;
- tissue heating due to energy absorption.

For absorption of energy, EM fields can be divided into four frequency ranges:

- frequencies from about 100kHz to less than about 20MHz, at which absorption in the trunk falls rapidly with decreasing frequency, and significant absorption may occur in the neck and legs;
- frequencies in the range from about 20MHz to 300MHz, at which relatively high absorption can occur in the whole body, and to even higher values if partial body (e.g. head) resonances are considered;
- frequencies in the range from about 300MHz to several GHz, at which significant local, non-uniform absorption occurs;
- frequencies above about 10GHz, at which energy absorption occurs primarily at the body surface.

The ICNIRP guidelines' levels are set on the following basis:

> Available experimental evidence indicates that the exposure of resting humans for approximately 30 minutes to EMF producing a whole-body SAR of between 1 and 4 W kg^{-1} results in a body temperature increase of less than 1°C. Animal data indicate a threshold for behavioral responses in the same SAR range. Exposure to more intense fields, producing SAR values in excess of 4 W kg^{-1}, can overwhelm the thermoregulatory capacity of the body and produce harmful levels of tissue heating.
>
> Many laboratory studies with rodent and non-human primate models have demonstrated the broad range of tissue damage resulting from either partial-body or whole-body heating producing temperature rises in excess of 1–2°C. The sensitivity of various types of tissue to thermal damage varies widely, but the threshold for irreversible effects in even the most sensitive tissues is greater than 4 W kg^{-1} under normal environmental conditions. These data form the basis for an occupational exposure restriction of 0.4 W kg^{-1}, which provides a large margin of safety for other limiting conditions such as high ambient temperature, humidity, or level of physical activity.

SAR stands for Specific Absorption Rate, quoted in watts per kilogram, and is a measure which is given as a Basic Restriction in the ICNIRP guidelines. Depending on frequency, the physical quantities used to specify the basic restrictions on exposure to EMF are current density, SAR, and power density. Protection against adverse health effects requires that these basic restrictions are not exceeded.

1.4.1.1 Reference levels

"Reference levels" of exposure are provided for comparison with measured values of physical quantities (electric or magnetic field strength); compliance with all reference levels given in the ICNIRP guidelines will ensure compliance with basic restrictions.

In many circumstances actually measuring SAR is either not feasible (because it would imply invasive measurements in a human body) or would not be economically justified. In these cases the fields to which people are actually exposed can be measured directly and compared to the reference levels. It is important to note that these are not limits but levels above which you should investigate the exposure regime: provided that basic restrictions are met and adverse indirect effects can be excluded, these field strength values can be exceeded. Typical circumstances in which the reference levels are applied are in exposure of the public or workers to fields from base station transmitter installations (though not in evaluation of hand-held transmitters). For EMC test labs, the radiated RF immunity test also provides an opportunity to consider them (see section 7.1.3.5).

1.4.1.2 SAR measurement

Reference levels are not relevant for handheld transmitters since the coupling takes place in the near field of the transmitting antenna. For this situation, it is necessary to measure SAR directly using a phantom of a human head or body, filled with a tissue-simulating liquid (actually a mixture of sugar, salt and water) which gives the correct dielectric properties at the relevant frequency, and allows a specialized E-field probe to be positioned inside the head cavity. This method is now well established for mobile phones and is being extended to other types of radio transmitter which are used close to the body. It has some similarities to EMC measurements but also many more differences; relatively few generalist test labs can do it, and it is mostly carried out by the mobile phone manufacturers themselves.

1.4.2 Athermal effects

Effects at much lower levels of RF that may be experienced by some people but are not based on tissue heating (athermal effects) are still controversial and disputed, and are not covered by the ICNIRP guidance or indeed any other officially published and recognized documents. ICNIRP itself says "available data are insufficient to provide a basis for setting exposure restrictions" and this position has been maintained for some years, although numerous studies have been published which purport to show one effect or another on biological systems of various frequencies and modulation schemes, including those used by mobile phone transmissions. It seems that there are two problems with this body of research:

- much of the experimental data is open to criticism and has been found difficult or impossible to replicate;
- the implication for radio transmitting authorities, if low-level EM energy is ever conclusively shown to have serious health hazards, is such that many radio-based systems would be shut down overnight, and this is a nettle that, unsurprisingly, noone in authority has seen fit to grasp.

Nevertheless, there is a substantial body of public opinion which believes that such a low-level hazard exists, and agitates against the continued extension of the radio transmitting infrastructure, particularly as evidenced by mobile phone base station installations. It is largely in (perhaps misplaced) response to this public opinion that standards for measuring the RF emissions of virtually any electronic apparatus have been drafted. Since these are essentially safety standards they are only relevant in the context of the R&TTE Directive.

Chapter 2

The EMC Directive

2.1 History

The relaxed EMC regime that existed throughout most of Europe up until the early 90s was overturned with the adoption by the European Commission on 1st January 1992 of the EMC Directive, 89/336/EEC [177]. This was at the time widely regarded to be "the most comprehensive, complex and possibly contentious Directive ever to emanate from Brussels" [34]. It has eventually been superseded by its second edition, published at the very end of 2004. This chapter examines the provisions of the new Directive and how manufacturers will need to go about complying with it. But first, we have to understand the background to European Directives and the history of the first EMC Directive.

2.1.1 The New Approach Directives

Of the various aims of the creation of the Single European Market, the free movement of goods between European states† is fundamental. All Member States impose standards and obligations on the manufacture of goods in the interests of quality, safety, consumer protection and so forth. If there are detailed differences in procedures and requirements, these act as technical barriers to trade, fragmenting the European market and increasing costs because manufacturers have to modify their products for different national markets.

For many years the Commission tried to remove these barriers by proposing Directives which gave the detailed requirements that products had to satisfy before they could be freely marketed throughout the Community, but this proved difficult because of the detailed nature of each Directive and the need for unanimity before it could be adopted. In 1985 the Council of Ministers adopted a resolution setting out a "New Approach to Technical Harmonisation and Standards".

Under the "new approach", directives are limited to setting out the essential requirements which must be satisfied before products may be marketed anywhere within the EU. The technical detail is provided by standards drawn up by the European standards bodies CEN, CENELEC and ETSI. Compliance with these standards is expected to demonstrate compliance with the essential requirements of each Directive. All products covered by each Directive must meet its essential requirements, but all products which do comply, and are labelled as such, may be circulated freely within the Community; no Member State can refuse them entry on technical grounds. Decisions on new approach Directives are taken by qualified majority voting, eliminating the need for unanimity and speeding up the process of adoption.

A document (the "Blue Guide") was published in early 2000 by the European Commission setting out the way in which new approach Directives should be

† Appendix E lists the EU and EEA Member States.

implemented in a relatively harmonised fashion. This says that a new approach Directive contains the following elements [179]:

- the *scope* of the Directive;
- a statement of the *essential requirements*;
- the *methods of satisfying* the essential requirements;
- how *evidence of conformity* will be provided;
- what *transitional arrangements* may be allowed;
- a statement confirming entitlement to *free circulation*;
- a *safeguard procedure*, to allow Member States to require a product to be withdrawn from the market if it does not satisfy the essential requirements.

It is the responsibility of the European Commission to put forward to the Council of Ministers proposals for new Directives. Directorate-General III of the Commission has the overall responsibility for Directives in the electrical sector. The actual decision on whether or not to adopt a proposed Directive is taken by the Council of Ministers, by a qualified majority vote. Texts of Directives proposed or adopted are published in the *Official Journal of the European Union*. Consultation on draft Directives is typically carried out through European representative bodies and in working parties of governmental experts.

2.1.2 Background to the legislation

In the UK, previous legislation on EMC was limited in scope to radio communications. Section 10 of the Wireless Telegraphy Act 1949 enables regulations to be made for the purpose of controlling both radio and non-radio equipment which might interfere with radio communications. These regulations have taken the form of various Statutory Instruments (SIs), which cover interference emissions from spark ignition systems, electromedical apparatus, RF heating, household appliances, fluorescent lights and CB radio. The SIs invoked British Standards which are closely aligned with international and European standards. The power existed to make regulations regarding the immunity to interference of radio equipment but this was never used.

At the European level various Directives have been adopted over the years, again to control emissions from specific types of equipment. Directive 72/245 EEC, adopted in June 1972, regulates interference produced by spark ignition engines in motor vehicles. Directives 76/889 EEC and 76/890 EEC, amended by various other subsequent Directives, apply to interference from household appliances and portable tools, and fluorescent lamps and luminaires. These latter two were superseded and repealed by the first EMC Directive. Each Member State is required to implement the provisions of these Directives in its national legislation, as described above for the UK.

This previous legislation is not comparable in scope to the EMC Directive, which covers far more than just interference to radio equipment, and extends to include immunity as well as emissions.

2.1.3 The first EMC Directive

89/336/EEC became fully operational in 1996 after an extended transition period lasting four years. During the initial period, manufacturers had to follow a fairly steep learning curve to get to grips both with the legal requirements and the technical subject of EMC: many had never had to deal with the issue before. The European Commission clearly assumed that manufacturers would be able to assume responsibility for the compliance of their products from the start, but in fact for many this was a new subject and a substantial extra burden, and the implementation of the Directive was at best patchy for several years. A detailed guidance document [180] was published by the Commission in 1997 which cleared up many of the misunderstandings that had been created, and by the turn of the century the understanding of the Directive within electrical and electronics companies had largely settled down.

Development of enforcement followed a similar uneven course. The Directive itself did not mandate any particular approach to enforcement and so different Member States took different approaches, depending on their particular culture and attitude towards European-initiated laws. This led to a noticeable variation in the gradient of the playing field across Europe, somewhat along the lines of the "traffic lights" principle†. In the UK, the job of enforcement was given to the Trading Standards authorities, more at home with dealing with rogue traders and the safety of consumer products like toys; only a few Trading Standards Officers would consider themselves even slightly familiar with the issues of EMC, and this combined with under-resourcing has meant that enforcement of the Directive in the UK has been somewhat limited in scope.

In fact, a large part of the pressure forcing companies to comply has come not from the enforcement authorities, but from other customers. For instance, in the consumer sector most of the market for electrical goods is dominated by a few large national or international retailers, and these demand correct declarations of conformity as part of their conditions of trade. Many non-consumer products such as industrial instrumentation are sold not to the end user but to a systems integrator, and again these have found that they need to ask for EMC compliance from their suppliers as part of their own compliance stance.

The first EMC Directive had two alternative routes to compliance: application of and self certification to harmonised standards, which was used in the vast majority of cases; and generation and assessment of a Technical Construction File, which was meant to be used if the manufacturer didn't or couldn't use standards, and which involved a review by a Competent Body.

2.1.3.1 SLIM

In 1998 the EMC Directive was subject to a review under the SLIM (Simpler Legislation for the Internal Market) process. Ten governmental experts were charged with making recommendations for improvements and simplifications to the Directive. Their report looked at the following areas amongst others:

- functional safety – explicitly, not to be part of the EMC Directive;
- immunity requirements – to be retained, but clarified;
- standards – a strategic review panel to be set up;
- large machines and installations – not subject to conformity assessment;
- definitions of certain types of environment.

The SLIM recommendations have formed the foundation for a complete rewrite of the EMC Directive, but it would be wrong to suggest that everything the SLIM group recommended has been taken on board unaltered in the new edition. Indeed, the second edition has introduced some new aspects which need a whole new raft of interpretation.

† In Northern European countries, compliance with traffic lights is mandatory. In Central Europe, they are advisory. In Southern Europe, traffic lights are for decoration only.

2.1.3.2 Other Directives

Apart from the EMC Directive, other new approach Directives adopted which may affect some sectors of the electrical and electronic engineering industry include:

- Toy safety (88/378/EEC)
- Non-automatic weighing machines (90/384/EEC)
- Medical devices (93/42/EEC)
- Active implantable electromedical devices (90/385/EEC)
- Machinery safety (89/392/EEC)
- Gas appliances (90/396/EEC)
- Lifts (95/16/EC)
- Refrigeration appliances (96/57/EC)
- In vitro diagnostic medical devices (98/79/EC)
- Radio and telecommunications terminal equipment (99/5/EC) (covered in Chapter 3)

Some of these contain their own EMC requirements which either partially or completely supplant the EMCD, as mentioned in section 2.2.2.3. In addition to this list, there are two other Directives which are relevant although they are not strictly "new approach" Directives. These are the Low Voltage Directive (73/23/EEC)(LVD) and the Automotive EMC Directive (95/54/EC, now superseded by 2004/104/EC). The LVD is concerned with safety, not EMC, but as a result of the CE Marking Directive the CE Mark attests to conformity with this Directive as well as any other applicable new approach Directives.

The Automotive EMC Directive is not a CE Marking Directive. It applies to vehicles and their electrical/electronic sub-assemblies and requires a form of type approval, signified by the 'e' marking. It is discussed in section 5.1.1.

2.2 The second EMC Directive

The second edition of EMC Directive 2004/108/EC [183] was published on 31st December 2004†. The timescale for its adoption is as follows:

- Entry into force: 20th January 2005
- Transposition: Member States shall adopt and publish the provisions necessary to comply with this Directive by 20th January 2007. They shall apply those provisions as from 20th July 2007, and from that date equipment claiming compliance with 2004/108/EC can be put on the market.
- Directive 89/336/EEC is repealed as from 20th July 2007.
- Transitional provisions: Member States shall respect compliance with 89/336/EEC for equipment which was placed on the market before 20th July 2009.

† At the time of writing, the UK legislation and the EU Guidance for the second edition Directive had not been completed. The discussion in the rest of this chapter draws on draft legislation and guidance [192][193], and on consultation workshops that the UK DTI and EMCTLA sponsored in 2005 and 2006.

2.2.1 What changes?

The essential requirements have not changed except for detailed wording, particularly that immunity now requires operation "without unacceptable degradation of its intended use". The routes to compliance however have been changed. Whereas before, if you didn't "apply" harmonised standards you had to use the Technical Construction File route and pass your TCF by a Competent Body, now, the TCF route has been swept away. The overarching requirement is for the manufacturer to perform and document an "EMC assessment" of his product. This may still involve compliance with harmonised standards: "The correct application of all the relevant harmonised standards whose references have been published in the OJEU shall be equivalent to the carrying out of the electromagnetic compatibility assessment." But it is also made clear that "Compliance with a harmonised standard is not compulsory", and the "presumption of conformity" which compliance offers is limited to the scope of the harmonised standard(s) applied and the relevant essential requirements covered by such harmonised standard(s).

The real change here is that, if you don't apply correctly all the relevant harmonised standards, you can, indeed must, still carry out an EMC assessment; but there is no requirement to have this assessment reviewed by a third party such as a Competent Body. On the other hand, you may do so voluntarily, and the Directive allows for "Notified Bodies" to be appointed on pretty much identical terms to the old Directive's Competent Bodies, so that these organizations will most probably continue to offer their services under the new Directive. Their use, though, is entirely at the manufacturer's discretion.

The other major change is to the regime for fixed installations. Equipment that is regulated by the Directive now has two subsets: apparatus, and fixed installations. The change allows different regulatory provisions for each. "Apparatus" now covers virtually everything that was covered by the old Directive, including mobile installations and systems; such apparatus is subject to the full provisions of the Directive, but there is an explicit exemption now for benign apparatus. Fixed installations on the other hand are not subject to much of the conformity assessment procedures, but do have to follow other new requirements.

2.2.2 Scope, requirements and exceptions

2004/108/EC applies to apparatus which is liable to cause electromagnetic disturbance or which is itself liable to be affected by such disturbance. "Apparatus" is defined as any finished appliance or combination of appliances made commercially available as a single functional unit, and intended for the end user. Essentially, anything which is powered by electricity is covered, regardless of whether the power source is the public supply mains, a battery source or a specialized supply. Specific exceptions are discussed in section 2.2.2.3.

An electromagnetic disturbance is any electromagnetic phenomenon which may degrade performance, without regard to frequency or method of coupling. Thus radiated emissions as well as those conducted along cables, and immunity from EM fields, mains disturbances, conducted transients and RF, electrostatic discharge and lightning surges are all covered. *No* specific phenomena are *excluded* from the Directive's scope.

2.2.2.1 Essential requirements

The essential requirements of the Directive (Article 5 and Annex I) are that the apparatus shall be so designed and manufactured, having regard to the state of the art, that:

- the electromagnetic disturbance it generates does not exceed a level above which radio and telecommunications equipment and other equipment cannot operate as intended;
- it has a level of immunity to the electromagnetic disturbance to be expected in its intended use, which allows it to operate without unacceptable degradation of its intended use.

The intention is to protect the operation not only of other users' radio and telecommunications equipment but any equipment which might be susceptible to EM disturbances, such as information technology or control equipment. At the same time, all equipment must be able to function correctly in whatever electromagnetic environment – that is, all electromagnetic phenomena observable in a given location – it might reasonably be expected to occupy. Notwithstanding these requirements, any Member State has the right to apply special measures with regard to the taking into service of apparatus, to overcome existing or predicted EMC problems at a specific site or to protect the public telecommunications and safety services. Compliance with the essential requirements will be demonstrated via the performance and documentation of an "EMC assessment". This is discussed in section 2.3.

2.2.2.2 Sale and use of products

The Directive applies to all apparatus that is placed on the market or taken into service. The definitions of these two conditions do not appear within the text of the Directive but are the subject of several paragraphs in the 1997 Guidelines [180] and in the Blue Guide [179].

Placed on the market

The "market" means the market in any or all of the European Economic Area (EEA); products which are found to comply within one state are automatically deemed to comply within all others. "Placing on the market" means the *first* making available of the product within the EEA, so that the Directive covers only new products manufactured within the EEA, but both new and used products imported from a third country. Products sold second hand within the EEA are outside its scope. Where a product passes through a chain of distribution before reaching the final user, it is the passing of the product from the manufacturer into the distribution chain which constitutes placing on the market. If the product is manufactured in or imported into the EEA for subsequent export to a third country, it has not been placed on the market.

The Directive applies to each individual item of a product type regardless of when it was designed, and whether it is a one-off or high volume product. Thus items from a product line that was launched at any time before 1996 had to comply with the provisions of the first Directive after 1st January 1996. Put another way, there is no "grandfather clause" which exempts designs that were current before the Directive took effect. However, products already *in use* before 1st January 1996 do not have to comply retrospectively. The transitional provisions for the second EMCD are that you can use the methods of compliance of either, from 20th July 2007 up to 20th July 2009.

The picture becomes less clear for products which are modified before being re-sold. If used apparatus, which has been modified during its operational life, is placed on the EU market then the procedure will depend on the extent to which those modifications affect the EMC performance of the apparatus. If the apparatus has been maintained and also modified, but not to any significant degree in terms of EMC, then it is termed “reconditioned” and is not subject to the provisions of the Directive. If the apparatus has been modified more substantially then it is referred to as “upgraded” and the Directive does apply. The choice remains in the hands of whoever carries out the modifications and places the product on the market, who is expected to carry out a new EMC assessment – or at least to validate the conclusions of the original assessment.

If the manufacturer resides outside the EEA, then the responsibility for maintaining the certificate of conformity with the Directive rests with the person placing the product on the market for the first time within the EEA, i.e. the manufacturer’s authorized representative or the importer. Any person who produces a new finished product from already existing finished products, such as a system builder, is considered to be the manufacturer of the new finished product.

Taken into service

“Taking into service” means the first *use* of a product in the EEA by its final user. If the product is used without being placed on the market, if for example the manufacturer is also the end user, then the protection requirements of the Directive still apply. This means that sanctions are still available in each Member State to prevent the product from being used if it does not comply with the essential requirements or if it causes an actual or potential interference problem.

2.2.2.3 Exceptions

There are a few specific exceptions from the scope of the Directive. Self-built amateur radio apparatus is specifically excluded. Military equipment is excluded, at least in the UK, as a result of an exclusion clause in the Treaty of Amsterdam which established the European Union, but equipment which has a dual military/civil use will be covered when it is placed on the civilian market. Equipment which falls within the scope of the R&TTE Directive is specifically excluded, since this Directive refers back to the EMCD’s essential requirements for its EMC provisions. Another exclusion is aeronautical products, parts and appliances that are referred to in Regulation (EC) No. 1592/2002 on common rules in the field of civil aviation, i.e. equipment that will be used on aircraft. Equipment for display at trade fairs need not comply provided that it carries a warning notice, and provided that “adequate measures are taken to avoid electromagnetic disturbances”.

Benign equipment

The most widely drawn exclusion is for electromagnetically benign equipment. This is defined as equipment, the *inherent qualities* of which are such that:

(a) it is incapable of generating or contributing to electromagnetic emissions which exceed a level allowing radio and telecommunication equipment and other equipment to operate as intended; and

(b) it will operate without unacceptable degradation in the presence of electromagnetic disturbance normally consequent upon its intended use.

Both conditions must be met, but if they are, such equipment is completely excluded from the application of the Directive. As a general rule, the requirements are contingent on the nature of the apparatus: such that it cannot create power supply or RF disturbances, and does not include active electronic components which could be affected by such disturbances. Most electronic equipment could not be regarded as benign, since it wouldn't meet one or both of these criteria, but some simple electrical apparatus such as switches, fuses, induction motors and filament lamps may be.

The picture is slightly clouded by the EC's draft guidance, which allows as examples "Protection equipment which only produce transitory disturbances of short duration" and "High voltage types of equipment in which possible sources of disturbances are due only to localised insulation stresses which may be the result of the ageing process and are anyway under the control of other technical measures", even though such devices do create disturbances, which can affect radio reception.

Other Directives

The only other exclusions are for those types of apparatus which are subject to EMC requirements in other Directives or regulations, to the extent that those Directives specifically lay down, in whole or part, the essential EMC requirements. These are:

- 90/385/EEC active implantable medical devices;
- 98/79/EC in vitro medical devices;
- 93/42/EEC medical devices;
- 72/245/EEC EMC of vehicles as amended by 95/54/EC and 2004/104/EC, but see section 5.1.1.2 regarding non-safety-critical aftermarket equipment;
- 75/322/EEC EMC of agricultural or forestry tractors as amended by 2000/2/EC;
- 97/24/EC two- and three-wheel motor vehicles;
- 96/98/EC on marine equipment;
- 90/384/EEC non-automatic weighing instruments.

In addition, equipment covered by the Measuring Instruments Directive 2004/22/EC which is CE marked and "M" marked according to that Directive, is excluded in respect of the immunity of such equipment. This exclusion is somewhat more complex, since the MI Directive is voluntary for some instruments. For the MI and non-automatic weighing instruments Directives, the EMCD still applies as far as emissions requirements are concerned.

2.2.2.4 Components

The question of when does a "component" (which is not within the scope of the Directive) become "apparatus" (which is) has been problematical. The Commission's 1997 guidelines introduced the concept of the "direct function", that is, any function which fulfils the intended use specified by the manufacturer in the instructions to the end user. It is available without further adjustment or connections other than those which can be performed by a technically naive user. Any component without a direct function is clearly not apparatus and is therefore outside the scope of the Directive. Thus individual small parts such as ICs and resistors are definitely outside the Directive.

Unfortunately, the second edition of the Directive, in defining what is meant by a component, has completely ditched this distinction. The question has now become, is the device intended to be placed on the market for incorporation into an apparatus by an end user, or likely to follow this path even if not intended directly for the end user? If so, then it is apparatus and must follow the full procedure required by the Directive. If not, then such components must be intended for incorporation into apparatus by other manufacturers, who take on the responsibility for compliance of their final product.

A component may be complex provided that its only purpose is to be incorporated inside an apparatus, but the manufacturer of such a component must indicate to the equipment manufacturer how to use and incorporate it. The distinction is important for manufacturers of board-level products and other sub-assemblies that may appear to have their own function and are marketed separately, yet cannot be used separately from the apparatus in which they will be installed.

However, in the particular case of plug-in cards for personal computers, which are supplied by a third party for the user to insert, the situation has been clarified: although such boards clearly need a host computer to have any purpose, they are placed on the market for the final end user and therefore need to carry a CE mark. They will need to be tested in a "representative" host computer, and certified accordingly. This position has been carried through into the harmonised standards, particularly EN 55022, which includes specific advice on how to treat such modules.

For products which may be both supplied to OEMs for incorporation into other apparatus, and supplied to the end user – an example might be some types of industrial sensor – then the item becomes apparatus and needs separate certification. If the manufacturer can insist that the item is only ever sold to OEMs then it is a component. This distinction has been made by many suppliers to shrug off the responsibility of ensuring that their products are properly specified for EMC ("Oh no, the Directive doesn't apply to us, we make components"). But since their customers are demanding EMC performance specifications anyway, to help them meet their own responsibilities, this is not a sustainable position.

2.2.3 The CE mark and the paperwork

The manufacturer or his authorized representative is required to attest that the protection requirements of the Directive have been met. This requires two things:

- he issues a declaration of conformity, which must be kept available to the enforcement authority for ten years following the last date of manufacture (a subtle change from the first Directive, which referred to the last date of placing on the market);
- he affixes the CE mark to the apparatus, or if this is not possible, to its packaging, instructions or guarantee certificate, in that order of priority.

2.2.3.1 The CE Mark

A further Directive concerning the affixing and use of the CE mark was adopted in 1993 [182]. This Directive harmonised the provisions regarding CE marking among the various previous new approach Directives. The mark consists of the letters CE as shown in Figure 2.1. The mark should be at least 5mm in height and be affixed "visibly, legibly and indelibly" but its method of fixture is not otherwise specified. Affixing this mark indicates conformity not only with the EMC Directive but also with the requirements of any other Directives relevant to the product which provide for CE marking – for instance, an electrical toy with the CE mark indicates compliance both with the Toy Safety Directive and the EMC Directive. Many electrical products fall under the scope of the Low Voltage Directive and the CE mark also indicates compliance with this. But during the transition period of any such applicable Directive, the CE mark need not indicate compliance; those Directives which *are* complied with should be listed in the appropriate documentation, such as the declaration of conformity.

It's worth pointing out that the CE Mark is not in any sense a quality mark, since all

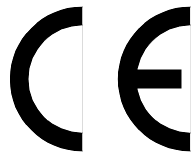


Figure 2.1 The CE mark

products that desire European market access must have it. Nor is it an approval mark, since in most cases there has been no third party oversight to check that it is valid. Its presence is largely meaningless in any technical sense; if you want to know anything about the performance of a product, you need to demand much more than simply that it carries the CE Mark.

2.2.3.2 The D of C

The EC Declaration of Conformity is required in every case. It must include the following components:

- a reference to the EMC Directive;
- a description of the apparatus to which it refers;
- the name and address of the manufacturer and, where applicable, the name and address of his authorized representative;
- a dated reference to the specifications under which conformity is declared;
- the date of the declaration;
- an identification of the signatory empowered to bind the manufacturer or his authorized representative.

It should also, perhaps self-evidently, certify that the apparatus to which it relates conforms with the protection requirements. You can write a single D of C in such a way that it covers all Directives that apply; you don't need a separate declaration for each.

Under the EMC Directive, an actual copy of the D of C is not required to accompany each item of equipment, although under the R&TTE Directive (see section 3.2.4) it is. Perhaps to forestall anticipated problems with some countries' enforcement officers who may be less than fully familiar with the distinction, many manufacturers take the safe route of including it in the documentation anyway. There is no requirement to have the CE mark on the D of C, but neither is it forbidden.

2.2.3.3 Description

The description of the apparatus should be straightforward; assuming the equipment has a type number, then reference to this type number (provided that supporting documentation is available) should be the starting point. The new EMCD is more stringent with regard to identification than the old one, requiring that equipment is identified by "type, batch, serial number or any other information" which would identify it. The description on the D of C should then encompass the range of such identifications so that there is an unambiguous correlation between the two.

Difficulties can arise when the type is subjected to revision or modification. At what stage do modifications or updates result in a new piece of equipment that would require re-certification? If the declaration of conformity refers to the Widget 3000 with software version 1.0 launched in 1996, does it continue to refer to the Widget 3000S of 2003 with version 3.2? The sensible approach would be to determine whether the modifications had affected the EMC performance and if so, re-issue the declaration for the new product; but this will require that you re-test the modifications, with the attendant cost penalties, or you exercise some engineering judgement as to whether a minor change will affect performance. No general guidance can be given on this point, but it should be clear that the breadth of the EMC requirements means that very few modifications will have absolutely no effect on a product's EMC performance.

2.2.3.4 Signatory

The empowered signatory will not necessarily be competent to judge the technicalities of what is being declared. Normally this will be one of the directors of the manufacturing or importing company, since the phrase used in the Directive is "empowered to bind" the manufacturer or his authorized representative. In small companies the technical director will probably be close enough to the product in question to understand the detail of its EMC performance, but in medium or large-scale enterprises the directors will expect to rely on the technical advice of their product development and manufacturing engineers and/or the EMC test and management personnel. Such companies will have to define clearly the levels of responsibility that exist for each person involved in making the declaration.

2.2.3.5 Specifications

In most cases the references to the specifications under which conformity is declared, will be a list of the harmonised standards – with dates, to distinguish which version has been used – that have been applied to the apparatus. If harmonised standards have been used only partially or not at all, a reference to the technical documentation (see section 2.3.1.3) where the detailed technical EMC assessment is given should be included as well as a reference to any identifiable non-harmonised standards or specifications that have been applied.

The reference to specifications raises the question of whether you have to *test* to these specifications. Three possibilities are apparent. Firstly, you may deem that the product intrinsically meets the requirements of the Directive and does not need testing. An example might be a simple linear unregulated stand-alone power supply which is below the power level at which harmonic currents are controlled. You may sometimes be able to convince your signatory that this is a competent engineering judgement, but many electronic products will not be able to follow this option.

Secondly, you may be able to make a declaration based on pre-existing test results. If for example you have already been conforming to existing non-harmonised standards, then you may be confident enough to state that the product will meet the appropriate harmonised standards without further testing, or with only partial testing.

The final option is to test fully to harmonised standards. For a sophisticated product this will be lengthy and expensive, and may involve some complex judgements as to what tests to apply. For new products though, testing will be essential. In fact, the new EMCD has subtly introduced a significant extra hurdle: it says (in the preamble)

> Compliance with a harmonised standard means conformity with its provisions and demonstration thereof by the methods the harmonised standard describes or refers to.

This appears to mean, in effect, that you must follow in every case the full compliance test method, in all its complexity and expense, whenever you are going to make a compliance declaration with a reference to a harmonised standard. The concept of pre-compliance testing, discussed in later chapters in this book and used successfully for many years in pursuance of the old Directive, has been closed off. However, what is almost certain to be the outcome of this is that pre-compliance testing will continue, supported by a brief statement in the technical documentation that harmonised standards have been applied “in part”, and justifying the effectiveness of such partial application. In essence, the manufacturer will make an assessment of the risk of not following harmonised standards in full versus the risk (in terms of cost to the company) of doing so.

A fourth option, of course, is not to test at all; just make the declaration, stick on the CE mark and hope that nobody ever notices. A reputable company, of course, won’t take this route, but the possibility of competitors doing so may be a factor in assessing your market position.

2.2.3.6 Other information

In keeping with the EC’s general desire to keep the user informed, the second edition EMCD makes new requirements for information to be supplied with each apparatus. As well as identification of the apparatus and the name and address of the manufacturer, there should be:

- information on any specific precautions that must be taken when the apparatus is assembled, installed, maintained or used in order to ensure that when put into service the apparatus complies with the protection requirements;
- a clear indication (easily understood by the intended user) of restriction of use for apparatus for which compliance with the protection requirements is not ensured in residential areas;
- information required to enable the apparatus to be used in accordance with its intended purpose.

It is assumed that if no information is given with the apparatus, then users can install, use and maintain it without any special considerations regarding its EMC aspects, and it will still comply with the protection requirements. This is naturally the preferred default position, but the possibility exists that you could, for instance, require the use of, say, particular cables or earthing regimes if these were necessary for compliance, provided that this was made clear to the user.

The residential area limitation has been a knotty problem for the Commission for some time. It has its origins in the text of a warning notice which appears in CISPR 22/EN 55022, the IT equipment emissions standard, referring to Class A equipment:

> Such equipment should not be restricted in its sale but the following warning shall be included in the instructions for use:
>
> **Warning**
>
> This is a class A product. In a domestic environment this product may cause radio interference in which case the user may be required to take adequate measures.

This text has caused headaches for the lawyers because, properly speaking, Class A equipment should never be marketed for or used in the domestic environment. The text is in the international (CISPR) version of the standard and other non-European countries, which have varying legal requirements, do not want to see it removed. The adopted compromise has been to delete the phrase “Such equipment should not be restricted in its sale" from the EN modified version of CISPR 22, and make it also a requirement in the new EMCD that "apparatus for which compliance with the protection requirements in residential areas is not ensured by the manufacturer has to be accompanied by a clear indication of this restriction of use".

2.2.4 Manufacturing quality assessment

The Directive covers every individual, physically existing finished product, but it would be impractical to test every item in series production fully for all the EMC characteristics that it must exhibit. The conformity assessment procedures for all the technical harmonization Directives are contained in Council Decision 90/683/EEC [181]. This document contains a range of modules which may be applied in the case of each specific Directive. However, the EMC Directive does not specifically refer to this Decision, and therefore conformity assessment requirements have been left somewhat open.

2.2.4.1 Production control

The second edition Directive requires in Annex II that "the manufacturer must take all measures necessary to ensure that the products are manufactured in accordance with the technical documentation ... and with the provisions of this Directive that apply to them". No specific means of determining what these measures might be are mentioned in either the Directive or the Council Decision.

CISPR sampling schemes

For many years before the adoption of the EMC Directive, the standards committee CISPR (see section 4.1.1.2) had recognized the need for some form of production quality testing, and had incorporated sampling schemes into the RF emission standards which form the basis of EN 55011, EN 55014 and EN 55022. The purpose of these schemes is to ensure that at least 80% of series production complies with the limits with an 80% confidence level, the so-called 80/80 rule. Practically, to comply fully with the 80/80 rule the manufacturer has to aim at about 95% of the products being in compliance with the specified limit.

The first scheme requires measurements of the actual emission levels from between 3 and 12 identical items, from which the mean and standard deviation are derived. The limit levels are then expressed in the form:

$$L \geq \overline{X} + k \cdot S_n \qquad (2.1)$$

where $\overline{X}$ is the arithmetic mean and S_n the standard deviation of the measured emission levels, and k is a constant derived from the non-central t-distribution between 2.04 and 1.2 depending on sample size

If the emission levels are similar between items (a low value of S_n) then a small margin below the limit is needed; if they are highly variable, then a large margin is needed. This sampling method can only be applied to emissions measurements and cannot be used for immunity.

A second scheme which is applicable to both emissions and immunity is based on recording test failures over a sample of units. Compliance is judged from the condition that the number of units with an immunity level below the specified limit, or that exceed the emissions limits, may not exceed c in a sample of size n, as per Table 2.1: this test is based on the binomial distribution and produces the same result as the first, in accordance with the 80/80 rule.

Table 2.1 Acceptable number of failures versus sample size n

n (sample size)	7	14	20	26	32
c (no. of failures)	0	1	2	3	4

As well as the above sampling schemes, published EN standards also allow a single test to be made on one item only, but then advise that subsequent tests are necessary from time to time on samples taken at random from production. The "one item only" dispensation is, of course, used in the vast majority of cases, although manufacturers with a high production volume and, especially, their own in-house test facilities, may well carry out some sample testing to reinforce and protect their compliance position.

Significantly, according to the CISPR standards the banning of sales is to occur only after tests have been carried out in accordance with one or other sampling scheme. This appears to put an onus on the enforcement authorities always to carry out testing of multiple samples in their market surveillance campaigns.

2.2.5 Fixed installations

The question of how the EMCD applies to installations has been an issue ever since the original Directive was under discussion:

> ... Finally, there is something of a problem concerning systems comprising assemblies of apparatus. In some cases, the EMC of a system can differ from the EMC of its constituent parts, and a system which is marketed as such requires separate certification under the Directive. To a large extent, the problems can be resolved by ensuring that the standards contain adequate practical test methods for systems, but some systems can only be tested "in situ" – ie long after the marketing stage. A similar problem applies to "one off" items of large capital equipment. We are trying to obtain clarification of the legal implications and will report back.
>
> *John Ketchell, DTI, 28th February 1989, in a circular to interested parties regarding progress on the negotiations prior to introducing the EMC Directive*

> The application of the EMC Directive to installations is, from the experience obtained over the last four years, a very controversial issue.
>
> *EC Guidelines on the application of the EMC Directive, July 1997, section 6.5.2.1*

Throughout the life of the first EMCD, systems and installations have been a bone of contention. So it is hardly a surprise that the biggest effect of the second edition is to substantially change the compliance regime for installations.

2.2.5.1 Changes in the second EMCD

The second EMC Directive has a separate section on fixed installations, defined as follows:

> Fixed installation means a particular combination of several types of apparatus and, where applicable, other devices, which are assembled, installed and intended to be used permanently at a predefined location

This distinguishes a fixed installation from a mobile installation, which has its own definition as "a combination of apparatus and, where applicable, other devices, intended to be moved and operated in a range of locations", and which is subject to the same compliance regime as other apparatus. A fixed installation has to be installed applying "good engineering practices" and respecting the information on the intended use of its components, with a view to meeting the protection requirements. Those good engineering practices are to be documented, and the documentation held by the person(s) responsible for as long as the fixed installation is in operation. However, fixed installations do not need to be CE marked or subject to a declaration of conformity. Also, apparatus that is intended for incorporation into a specified fixed installation does not have to comply with certain aspects of the Directive.

This represents a significant change in the way that fixed installations and the apparatus that goes into them are regulated. Not all of the implications are yet clear, but one is that operators of installations will have to designate a "responsible person" who has responsibility for the establishment of compliance of a fixed installation with the relevant essential requirements. This will be new to most if not all such operators. Another is that the vague term "good engineering practices" will have to be fleshed out; at present there are no harmonised standards for installations, but the Commission clearly envisages that some will have to be created.

The principal dilemma of applying the Directive to complete installations is that to make legally relevant tests is difficult, but the nature of EMC phenomena is such that to test only the constituent parts without reference to their interconnection is meaningless. Two main possibilities have been explored in TCFs under the first EMCD. The first is to ensure that the system is built out of individual items which are themselves compliant, and that the method of installation follows suppliers' instructions such that this compliance is not breached. The TCF would reference installation drawings and work instructions that ensure this. This approach is continued and expanded in the new regime.

The alternative has been to perform limited site testing once the installation has been assembled to show that it is compliant. Since this would be less comprehensive, it would need to be balanced by a greater amount of documentation in the form of a matrix defining the EMC threats and a rationale for the claim to compliance, including a justification for the tests that were done. The new regime does not mention testing in any specific way.

2.2.5.2 Scope

The definition shown above for a fixed installation distinguishes it from a large system (discussed below in section 2.2.6) in that it is not a single functional unit, that is, it is "a particular combination of several types of apparatus", and it is not placed on the market and cannot enjoy "free movement" around the EU: it is "assembled, installed and intended to be used permanently at a predefined location". The phrase "intended to be used permanently" has caused problems in interpretation; some people have thought that it meant being used 24 hours a day. Actually, the likely interpretation (still to be fully accepted at the time of writing) is related to duration. The requirement would be fulfilled if it were intended at the time of putting into service that the constituent parts were to be used in the defined location for their expected lifetime. If they were expected to be moved for use at another location, it would not.

It is recognized that modifications will be made to fixed installations throughout their operational life, and such modifications would not invalidate the original intent. Indeed, many such installations (such as telecommunications centres) are in a continuous state of modification, and the legislation cannot be drafted in such a way as to prevent this. But the implication of the requirements are that the EMC dimension of such modifications must always be taken into account and properly documented.

One problem that arises when considering fixed installations is what the boundaries of the installation are, and therefore what parts have to be considered – and indeed, what entity should take responsibility for it. For instance, in a campus with a network of power supplies and telecommunication lines, where is the physical "edge" of the installation? Sometimes the service entrance of the conductors can be easily located, but not always, and this certainly doesn't define the boundary for radiated phenomena. Knowing the boundary should enable the electromagnetic environment to be defined, but there will be many examples where the boundary is fuzzy at best. The question also introduces the idea of "nested" installations, that is, smaller installations within a larger one, with different boundaries and responsibilities. The DTI's draft guidance wryly points out that in the case of interference between installations in such a situation, "it is likely that only by further co-operation between the responsible persons will a solution be able to be found".

There is another consequence of the generalized definition given above, which if pursued to its logical end could have dramatic consequences. This is that there is no effective lower limit on the size or extent of the installation. It is clear, historically, that the definition was meant to encompass large premises such as telecommunication centres, factories, power stations or substations, and so forth. But actually, any location which has "a particular combination of several types of apparatus" that are intended only for that location would fall within the scope, and therefore need a responsible person and a documentation of good practice. It is possible that this definition could extend down to the level of individual shops, offices and even homes, if electrical or electronic equipment is installed in them which is specific to that one location. Such a possibility is not excluded by the EC's draft guidance, which suggests "The definition covers all installations from the smallest residential electrical installation through to national electrical and telephone networks, including all commercial and industrial installations."

2.2.5.3 The responsible person

Annex I of the EMCD requires that

> good engineering practices shall be documented and the documentation shall be held by the person(s) responsible at the disposal of the relevant national authorities

and Article 13.3 says

> Member States shall set out the necessary provisions for the identifying the person or persons responsible for the establishment of compliance of a fixed installation

In the UK, the DTI has drafted a definition of the responsible person as the person who, "by virtue of their ownership or control of the relevant fixed installation is able to determine that the configuration of the installation is such that when used it complies with the protection requirements". It is accepted that this may be a different entity according to circumstances. For example in some cases it could the site owner, in others the operator of the installation, in others the maintainer of the installation, and so on. Operators will need to identify the responsible person before the installation is taken into service, and often this will have to be determined contractually.

The implication of the DTI's definition above is that whoever the entity turns out to be for a given case, they will need to have a sufficient understanding of the EMC issues raised by the installation to be able to make the required judgement, as well as having sufficient seniority to be able to take on the required responsibility. As a matter of simple observation, such individuals, groups or departments rarely exist in typical organizations that administer installations, although there are exceptions: for instance large telecom organizations will usually have the technical capability in-house, as will groups in some sectors that already deal with installation EMC, such as the railway industry as discussed in section 5.4.

Consequently, we can envisage either of two outcomes:

- either organizations will take their responsibilities seriously, and appoint or employ an individual, a consultant or an in-house group to deal with the likely proliferation of documentation that the new regulations imply; or
- comprehension of the scope and implications of the new regulations will be so minimal, and their enforcement so weak, that little if anything will be done to bring installations into compliance with the new regulations.

2.2.5.4 Good engineering practices

There is no definition in the EMCD of what is meant by the phrase "good engineering practices", except that by implication they are capable of being documented. There are various good practices that have been established for electrical installations, for instance in the UK the IEE's Wiring Regulations, now enshrined in BS 7671; but these have nothing significant to say about EMC and certainly wouldn't be useful in complying with the essential requirements. It is an emerging consensus that what is meant is "good *EMC* engineering practices", but this leaves the field wide open for interpretation. To date, such interpretation has been highly generalized, pointing at best to a few IEC documents such as those in the IEC 61000-5 series and to the existence of some books on the subject, a canon to which this author amongst others has already contributed [22]. Much of the more detailed material in fact applies to military systems and would be irrelevant or hard to apply in the context of the EMC Directive.

Certainly, it seems very unlikely that any responsible person would be prosecuted for non-compliance on the basis that their documented engineering practices were actively bad for EMC; even if this were to be the case, actually proving it in a court of law would be expensive and uncertain in outcome at best. For instance, many system designers and installers use traditional engineering practices that they assume are good, such as single-point earthing and terminating cable screens at one end only using long lengths of wire, when in fact these are inimical to good EMC at the frequencies of interest for the EMCD's protection requirements. To prove the case, though, would require long, involved and probably inconclusive technical arguments that would not be compatible with an effective enforcement action. What is left is the requirement to "respect the information on the intended use of components", and this boils down to ensuring that if any such components include strictures on their installation and operation that have relevance for EMC, these should be observed and the fact that they have been observed should be documented.

The level of detail that is needed in the documentation will vary according to the complexity of the fixed installation. It should be sufficient to enable an enforcement authority to determine whether good practices have been followed. When interpreting this requirement, you should bear in mind that enforcement authorities (certainly in the UK) are not experts in EMC.

One point to note about the good practices that might be adopted is the distinction between intra-system and inter-system EMC, as already discussed in section 1.2. The EMCD is only concerned with interference between the fixed installation and the outside world, that is, *inter*-system EMC. So, although of great interest to the owner and operator of the installation, good practices for *intra*-system EMC, which are intended only to ensure mutual compatibility of the various parts of the installation, are of themselves not relevant for the EMCD.

2.2.5.5 Specific apparatus

Apparatus that is intended for incorporation into a specified (named) fixed installation – and not otherwise commercially available to an end user as a single functional unit – does not have to comply with:

- the essential requirements;
- the conformity assessment procedure;
- the technical documentation and EC declaration of conformity – but some technical documentation is required;
- CE marking.

The basis for this difference is that it is not necessary, and sometimes not possible, to carry out a conformity assessment for apparatus that is intended solely for a specific fixed installation, in isolation from that installation. This exemption cannot be used for apparatus which may become available in a general sense. Apparatus can only benefit from it if there is a direct link between the manufacturer of that specific apparatus and those responsible for the fixed installation for which that apparatus is intended. A relationship, probably contractual, between the provider and the customer is required. The information requirements remain, that is, the apparatus must be fully identified and include the name and address of its manufacturer; and it needs to document exactly which fixed installation it is intended for "and its EMC characteristics" and must indicate "the precautions to be taken for the incorporation of the apparatus into the fixed installation in order not to compromise the conformity of that installation".

The reference to EMC characteristics of the installation implies that the manufacturer of such apparatus must be familiar with the installation's electromagnetic environment. The level of detail needed, and the form that the information should take, is not given: it could mean as little as classifying the fixed installation as having a domestic, commercial, light-industrial, or industrial electromagnetic environment as per the generic standards; but this would hardly be enough to design a product that would not compromise the compliance of the installation. Similarly, indicating the precautions to be taken for EMC may simply mean listing well-known EM mitigation measures and letting the installer decide what to do; or it could mean giving the installer clear and detailed instructions for every aspect of the installation. Both of these issues are so novel, and the available guidance so limited, that for the present it is hard to see how an acceptable general practice could be developed to deal with the requirements.

The Commission's draft guidance envisages a number of instances where the exemption for specific apparatus might be used:

- specific apparatus made according to particular specifications and intended for a fixed installation or installations considered as equivalent by the manufacturer;
- apparatus designed according to a specification given by a customer, intended for a given fixed installation;
- apparatus derived from a generic model adapted to the specific need of the customer or to the specificity of any particular location, in a fixed installation;
- apparatus made in small series and delivered for incorporation into a well-defined type of installation, each item necessitating appropriate EMC adjustments at the final location.

2.2.6 Systems

A fixed installation cannot "enjoy free movement" within the EEA, in contrast to a system (or a moveable installation) which can. A typical system might be a personal computer workstation comprising the PC, monitor, keyboard, printer and any other peripherals. If the units were to be sold separately they would have to be tested and certified separately; if they were to be sold as a single package then they would have to be tested and certified as a package.

The definition unfortunately does not help system builders who will be "placing on the market" – i.e. supplying to their customer on contract – a single system, made up of separate items of apparatus but actually sold as one functional unit, not necessarily intended for a pre-defined location. Many industrial, commercial and public utility contracts fall into this category. According to the published interpretation, the overall assembly should be regarded as a system and therefore should comply as a package. As it stands at present, there are no standards which specifically cover large systems, i.e. ones for which testing on a test site is impractical, although some emissions standards do allow measurements in situ. These measurements are themselves questionable because of the difficulty of distinguishing external interference at the measurement position from that due to the installation, and because the variability of the physical installation conditions introduces reflections and standing waves which distort the measurement. There are no provisions for large systems in the immunity standards. Therefore the EMC assessment has to be performed in the absence of standards, but there is little guidance as to how to interpret the Directive's essential requirements in these cases.

2.2.7 Implementation, enforcement and sanctions

Member States cannot impede for EMC reasons the free circulation of apparatus covered by the Directive which meets its requirements when properly installed and maintained, and used for its intended purpose. They must presume that apparatus which bears the CE mark, and for which the manufacturer has performed and documented an EMC assessment, does in fact comply with the protection requirements unless there is evidence to the contrary.

On the other hand, Member States are required to ensure that equipment which is found not to comply is not placed on the market or taken into service, and to take appropriate measures to withdraw non-compliant apparatus from the market. Legislation which translates the Directive's requirements into national law in each Member State was required to be in place by 20th January 2007.

In the UK the enforcement regime includes the issue of "suspension notices" which prohibit the supply or use of specified equipment that the enforcement authorities believe does not comply with the EMC requirements. The notice may or may not have immediate effect, depending on the urgency of the situation; an appeal procedure allows persons on whom a notice is served to make representations for it to be revoked. Enforcement authorities can also apply to a court for forfeiture of apparatus, with its consequent destruction, modification or disposal, and officers of the enforcement authorities may be empowered to enter premises and inspect or seize apparatus and documents.

2.2.7.1 Offences

The EMC legislation does include criminal sanctions. But because of the difficulty of judging whether or not apparatus actually complies with the requirements, the UK legislators have not created an absolute criminal offence of supplying or using non-compliant equipment. Users and retailers cannot normally be expected to know whether or not the apparatus in question is non-compliant. Criminal offences on other fronts are necessary, for instance to guard against misuse of the CE mark or the provision of false or misleading information, and to penalize breaches of prohibition notices.

2.2.7.2 Practice

Two important questions are: how is enforcement operated in practice, and is the Directive enforced equally in all Member States (the so-called "level playing field"). These questions are directly related to the resources that national governments are prepared to devote to the task. The UK DTI has previously indicated that its enforcement efforts are complaint-driven. As well as investigating interference complaints arising from actual use of apparatus, it is open to complaints that apparatus does not conform to the Directive's requirements regardless of whether or not there is a problem in its use. A possible source of complaint will therefore be from companies testing samples of their competitors' equipment and, if they find that it does not comply, "shopping" them to the authorities. It is also likely, though, that such complaints will need to be backed by serious evidence of non-compliance before the authority will take them seriously.

On the other hand, the German authorities have stated that it is necessary to gain information from the market in the form of random spot checks in order to react to violations [115]. Germany already had a strong regime for the control of RF emissions before the first EMCD in the form of the mandatory VDE standards, and these were stricter than the EN standards which are now used to demonstrate compliance with the Directive. The Germans have been concerned that the Directive might dilute the effectiveness of their previous regime, and have therefore insisted that it is thoroughly enforced.

It is apparent that differences in enforcement practices within the various Member States work contrary to the stated intent of the Directive, which is to reduce technical barriers to trade. Article 10 of the Directive requires that "where a Member State ascertains that apparatus bearing the CE Marking does not comply with the [protection] requirements, it shall take all appropriate measures to withdraw the apparatus from the market, prohibit its placing on the market or restrict its free movement", and shall immediately inform the Commission of any such measure. If the Commission finds, after consultation, that the action is justified, it will inform all other Member States. The competent Member State shall then take appropriate action against the author of the attestation. Therefore any Member State can take immediate action to prohibit an offending apparatus from its own market, but sanctions against the company that put the apparatus on the market in another Member State are dependent on the deliberations of the Commission and on the enforcement practices of the latter Member State.

2.2.7.3 Interpretation

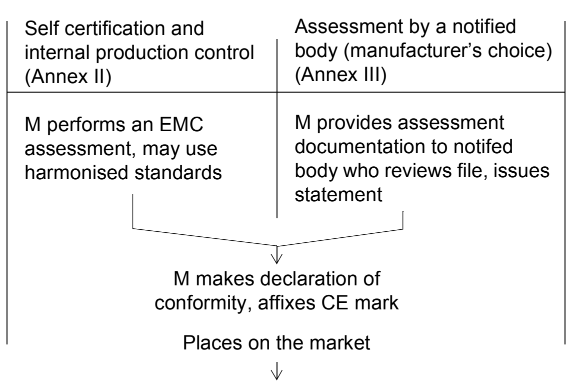
As has already been indicated, the Directive is so widely drawn that many of its provisions have to be the subject of interpretation. You might expect that this interpretation would be the function of the EC or of the national implementing authorities, but it has been remarkably difficult to obtain answers to detailed questions of interpretation from such authorities, especially so when there is a technical dimension to the question. The 1997 Guidelines gave considerable interpretive help, but do not cover many more abstruse technical issues.

The new Directive also has a guidance document, a draft of which [192] is referred to at various points throughout this chapter. The intent of this is that it should have been available in advance of Member States' transposition of the Directive into their own laws, so that such transposition took place more uniformly than was the case for the first Directive. Because of the delays and contention in creating the document, this has happened only imperfectly. Furthermore, it is intended as an internet-based document, partly for ease of access and linking to other sources of information, but also so that it can be readily updated if the need arises. This latter aim may not be seen entirely as an advantage by its users, who generally would like such guidance to be consistent rather than fluid, but the Commission is distinctly in favour of it.

In the UK, a group known as the EMC Test Laboratories Association (EMCTLA) was formed primarily to ensure a reasonable uniformity of approach to the assessment of technical construction files under the first EMCD by different competent bodies. Within that association, a working group has been set up which responds to queries regarding the implementation of the Directive and issues technical guidance notes which have a reasonably wide circulation, and which can form the basis for a uniform interpretation. These guidance notes are available via the EMCTLA's website; many are still relevant, despite the removal of the TCF route. The EMCTLA has been instrumental in founding a pan-European Association of Competent Bodies which is intended to spread this uniform approach within the EU.

2.3 Compliance of apparatus with the Directive

Of themselves, the essential protection requirements are too generalized to enable manufacturers to declare that their product has met them directly. So Annexes II and III of the Directive provide a more detailed route (Figure 2.2) for manufacturers to achieve compliance with them.



M: manufacturer, authorized representative within the EU, or person who places apparatus on the EU market

Figure 2.2 Routes to compliance

2.3.1 Self certification and internal production control

The route which is expected to be followed by most manufacturers is that given in Annex II, known as self certification with internal production control. This requires you as the manufacturer firstly to perform

> an electromagnetic compatibility assessment of the apparatus, on the basis of the relevant phenomena, with a view to meeting the protection requirements ... The electromagnetic compatibility assessment shall take into account all normal intended operating conditions. Where the apparatus is capable of taking different configurations, the electromagnetic compatibility assessment shall confirm whether the apparatus meets the protection requirements ... in all the possible configurations identified by the manufacturer as representative of its intended use.

On the assumption that the assessment does indeed confirm that the apparatus meets the protection requirements, you then have to

> draw up technical documentation providing evidence of the conformity of the apparatus with the essential requirements of this Directive, [and] hold the technical documentation at the disposal of the competent authorities for at least ten years after the date on which such apparatus was last manufactured.

This documentation requirement is no more than you would have done under the first edition of the Directive, since you would have kept your own CE Marking documentation on file for your own purposes, whether or not you ever involved a third party such as a Competent Body. In other words, there is little change so far from the first Directive. Having developed the documentation, you then make the Declaration of Conformity and prepare to retain that also for ten years. You are then free to apply the CE Mark and market the apparatus, ensuring that each product is manufactured in accordance with the technical documentation and in compliance with the essential requirements (hence, "internal production control").

2.3.1.1 The EMC assessment

We now have to interpret what is meant by the phrase "EMC assessment". Annex II.1 specifically refers to "on the basis of the relevant phenomena, with a view to meeting the protection requirements". It also makes clear that the assessment must cover all configurations and normal operating conditions. In the first edition of the Directive, either you had to apply harmonised standards or you had to generate a Technical Construction File and run it past a Competent Body. In this new regime, all the responsibility of what to do is in your hands. The Commission's view is, fairly clearly, that having let manufacturers in gently with the first EMCD, the time has come for them to grow up and deal with the EMC of their products on their own.

Use of harmonised standards in the assessment is not mandatory, but if they are not used or are used only in part, you then have to decide how you are going to show that the protection requirements have been met. This means that the assessment will have to look at least at:

- a description of the apparatus' operating conditions and its intended use;
- the environments in which the apparatus will be used;
- the electromagnetic phenomena covered and compatibility levels applied;
- the performance criteria of the apparatus;
- test plan, including techniques, emissions limits and imunity stress levels;
- EMC design and installation techniques used;

- the user's information requirements, including any restriction (or not) on use in residential areas, and any precautions on installation;
- definition of the test selection matrix for variants, using worst case criteria.

Clearly, in the absence of other guidance, this could be a substantial exercise in its own right. So at this point, the Directive refers you to standards.

2.3.1.2 Application of harmonised standards

Annex II.1 says

> The correct application of all the relevant harmonised standards whose references have been published in the Official Journal of the European Union shall be equivalent to the carrying out of the electromagnetic compatibility assessment.

Article 6.2 goes on to say

> The compliance of equipment with the relevant harmonised standards whose references have been published in the Official Journal of the European Union shall raise a presumption, on the part of the Member States, of conformity with the essential requirements referred to in Annex I to which such standards relate.

But

> This presumption of conformity is limited to the scope of the harmonised standard(s) applied and the relevant essential requirements covered by such harmonised standard(s).

Harmonised standards are those CENELEC, CEN or ETSI standards which have been announced in the *Official Journal of the European Union* (OJEU). In the UK these are published as BS EN standards, and each Member State has a similar system, administered by its own national standards body, for publishing them. They are not available from CENELEC itself, although ETSI ones can be obtained (for free) direct from ETSI. The most widely used ones are detailed in Chapter 4, and section 2.3.4 covers the use of standards.

The potential advantage of using standards in the EMC assessment from the manufacturer's point of view is that the assessment process is simplified. There is no mandatory requirement for testing by an independent test house. The only requirement is that the manufacturer makes a declaration of conformity (see section 2.2.3.2) which references the standards against which compliance is claimed. Of course the manufacturer will normally need to test the product to assure himself that it actually does meet the requirements of the standards, but this could be done in-house. Many firms will not have sufficient expertise or facilities in house to do this testing, and will therefore have no choice but to take the product to an independent test house. This is discussed further in sections 2.3.3 and 16.3. But the long-term aim ought to be to integrate the EMC design and test expertise within the rest of the development or quality department, and to decide which standards apply to the product range, so that the prospect of self certification for EMC is no more daunting than the responsibility of functionally testing a product before shipping it.

2.3.1.3 The technical documentation

The documentation needs to support the case that you have made using the EMC assessment for meeting the protection requirements for any particular product. One requirement for placing on the market in the draft UK regulations is that "the technical documentation has been prepared and is available".

Contents

Annex IV of the Directive specifies the general content of the documentation, whose overall purpose is "to enable the conformity of the apparatus with the essential requirements to be assessed". The content includes:

- a general description of the apparatus (one purpose of this is to allow an unambiguous link between the actual apparatus and the documentation);
- evidence of compliance with the harmonised standards, if any, applied in full or in part;
- where the manufacturer has not applied harmonised standards, or has applied them only in part, a description and explanation of the steps taken to meet the essential requirements of the Directive.

The third of these of course is the record of the EMC assessment described in 2.3.1.1. Notice that the second requirement is for "evidence" of compliance with standards; this may well not mean a full test report. What constitutes a minimum of "evidence" has yet to be ascertained.

As we have already mentioned, the products that are placed on the market must be "manufactured in accordance with the documentation". Since products change and develop over their lifetime, this means that the documentation should keep pace: as is necessary in any case, you will need to assess engineering changes for their EMC implications. Any that ripple through to affect the content of the technical documentation should result in appropriate changes to it.

Whilst the purpose of the documentation is clear enough, what is not clear is who will be doing the assessment of conformity with the essential requirements – in other words, who is the documentation being written for? Will it be an enforcement officer who has just a sketchy knowledge of EMC technicalities, or will it be an EMC consultant who is prepared and able to pick holes in even the densest material? Presumably, this will depend on the context of whatever enforcement action is being taken; so it is up to you as technical author to anticipate this context and set the technical level of the documentation appropriately.

2.3.2 The notified body

A substantial change between the first and second editions of the Directive is the removal of the old Technical Construction File route to compliance, and the concomitant disappearance of Competent Bodies. Instead, we now have Notified Bodies (NBs) and their use by the manufacturer is entirely voluntary.

Annex III of the Directive describes the use of Notified Bodies (Annex II is the self certification/internal production control procedure discussed above):

> 1. This procedure consists of applying Annex II, completed as follows:
>
> 2. The manufacturer or his authorised representative in the Community shall present the technical documentation to the notified body referred to in Article 12 and request the notified body for an assessment thereof. The manufacturer or his authorised representative in the Community shall specify to the notified body which aspects of the essential requirements must be assessed by the notified body.
>
> 3. The notified body shall review the technical documentation and assess whether the technical documentation properly demonstrates that the requirements of the Directive that it is to assess have been met. If the compliance of the apparatus is confirmed, the notified body shall issue a statement to the manufacturer or his authorised representative in the Community confirming the compliance of the apparatus. That statement shall be limited to those aspects of the essential

requirements which have been assessed by the notified body.

4. The manufacturer shall add the statement of the notified body to the technical documentation.

There are a number of features to point out here.

2.3.2.1 The primacy of the manufacturer

Not only is it entirely at the manufacturer's discretion whether or not to use an NB, but the manufacturer tells the NB what to do, not *vice versa* (Annex III.2). You may choose to ask an NB to assess the documentation as a whole, covering the totality of the essential requirements, or only in part. Of course, the statement that the NB issues will only cover those aspects that it has reviewed. So, for instance, you may be entirely comfortable with your compliance with emissions limits but you may want an opinion on how the product complies with the immunity requirements.

The NB only provides a "statement", and only if the compliance is confirmed. The statement has to be included with the technical documentation, but there is no indication that there is any responsibility on or ability of the NB to prevent the manufacturer placing the product on the market if it believes that compliance is not confirmed; that remains firmly the responsibility of the manufacturer. Manufacturers are free to choose any NB. There is no need to choose an NB located in the country where the apparatus is manufactured, nor in the country to which the apparatus will be shipped, marketed or taken into service. If the manufacturer has used the services of an NB for one of his products there is no obligation to use the same NB for another one, or for modifications of the originally assessed product.

A manufacturer may consult more than one NB simultaneously in respect of any apparatus: there is no prohibition on this, and the NB's opinion is not binding. In other words, it is apparently legitimate to "shop around", should you so desire, until you get a favourable opinion.

The Directive at Article 12 specifies how NBs are to be appointed. The requirements are similar to those for Competent Bodies under the first Directive, but sufficiently different that it is unclear as to whether all such existing Competent Bodies will choose to or be able to convert to NB status. An NB can only render services within its scope of designation which may be limited to certain categories of apparatus, certain essential requirements or certain other specific aspects. But for the time being there are no agreed criteria within the EU for the definition of "scope of designation"; different countries are likely to have different ideas.

2.3.2.2 When would a manufacturer use an NB?

At first sight, there seems to be little point in consulting an NB. It adds expense, but has no teeth and is not a requirement of the process.

A common use of the TCF route and the involvement of a Competent Body under the first edition Directive was to certify the compliance of a fixed installation. As we have seen, the requirements for installations have changed drastically, and Notified Bodies have no role in assessing the good practice documentation that is required, even though that would be one of their most useful functions, and one in which (as Competent Bodies) they actually have gained some experience. Their role is limited entirely to assessing the compliance of *apparatus*.

The NB cannot perform the product's EMC assessment or generate the technical documentation itself: in no way does it supplant the Annex II procedure, which remains firmly in the hands of the manufacturer. The NB is, in fact, required to be independent in the process of "preparing the reports and performing the verification function". (This may not stop you contracting one NB to create the documentation and another to assess it, if you are prepared to sign up to the result.)

The main driving force for involving an NB will be where the manufacturer requires reassurance by a third party to increase his confidence in the compliance process, or where the equipment is so complex that its assessment requires expert guidance. In most cases, this could be done by any competent EMC consultant; the use of an NB merely gives the imprimatur of appointment by the statutory authority. However, the NB statement can be seen as a shield against future enforcement action, and some manufacturers will be prepared to pay for such insurance. It may also be viewed in a few cases as an extra marketing tool.

2.3.3 Testing

Except in the case of benign products, which it is clear will intrinsically not cause interference or be susceptible to it, such as an electric fire or pocket torch, each manufacturer will need to submit products to some degree of EMC testing to be sure that they comply with the Directive. Later chapters consider EMC test methods in detail. To cover the eventual requirements of the standards, the scope of the tests will need to include mains harmonic, conducted and radiated RF emissions, plus immunity to RF, transients, electrostatic discharge and supply disturbances. A test facility to address all these phenomena at compliance level is beyond the budget of all but the largest companies. Not only are a screened room and/or an open area test site and all the test equipment needed, but also the staff to run the facility – which itself requires a level of skill, experience and competence not usually found in most development or test departments. A large company may have the product volume and available capital which allows investment (of the order of £1m) in an in-house facility of this nature, and there are several such companies throughout Europe who have taken this step. The EMCD places no external constraints on the operation of these in-house test facilities. The management options for, and approaches to, testing are discussed in section 16.3.

2.3.4 Using standards

Compliance with harmonised standards is "equivalent to" the performance of the EMC assessment (see 2.3.1.2). This mechanism depends on the availability of standards which can be applied to the product in question. The detail of the appropriate standards is covered in Chapter 4; this section will discuss their general availability and applicability.

Prior to the adoption of the first EMC Directive, the EMC standards regime had developed in a somewhat piecemeal fashion. The existing standards fell into a number of categories:

- RFI: intended to protect the radio spectrum from specific interference sources, such as information technology equipment, motor vehicle ignition, household appliances or fluorescent lights
- mains emissions: specifically harmonic currents and short-term variations, to protect the low-voltage power distribution network
- product- and industry-specific: to ensure the immunity from interference of particular types of product, such as process instrumentation or legal metrology, or to regulate emissions from equipment that will be used in a specific environment, such as marine equipment

These standards are not over-ridden by the Directive; those which have been harmonised by CENELEC may be applied to products within their scope and are regarded as adequate to demonstrate compliance, at least as far as their scope and coverage of the essential requirements allows.

2.3.4.1 The generic standards

In the early days of the EMC Directive, there were many industry sectors for which no product-specific standards had been developed. This was especially so for immunity, which was a new concept for many products. In order to fill this gap wherever possible, CENELEC gave a high priority to developing the Generic Standards. These are standards with a wide application, not related to any particular product or product family, and are intended to represent the essential requirements of the Directive. They are divided into two groups, one for immunity and one for emissions, each of which has separate parts for different environment classes (Table 2.2).

Table 2.2 The generic standards

Environment	Residential, commercial and light industry	Industrial
Emissions	EN 61000-6-3	EN 61000-6-4
Immunity	EN 61000-6-1	EN 61000-6-2

Where a relevant product-specific standard does exist, this takes precedence over the generic standard. It is quite common, though, for a particular product to need a variety of standards. For instance, a fire alarm may be covered by one product standard for mains harmonic emissions, another for immunity, and the generic standard for emissions. All these standards must be satisfied before compliance with the Directive can be claimed. Although a comprehensive range of product standards has been developed, other mixed combinations still occur and there will always be unusual products that “fall through the cracks”.

Environment classes

The distinction between environmental classes is based on the electromagnetic conditions that obtain in general throughout the specified environments [128]. The inclusion of the “light industrial” environment (workshops, laboratories and service centres) in class 1 has been the subject of some controversy, but studies have shown that there is no significant difference between the electromagnetic conditions at residential, commercial and light industrial locations. Equipment for the class 2 “industrial” environment is considered to be connected to a dedicated transformer or special power source, in contrast to the class 1 environment which is considered to be supplied from the public mains network.

2.3.4.2 Performance criteria

A particular problem with immunity is that the equipment under test may exhibit a wide variety of responses to the test stimulus. This can range from a complete lack of response, through a degradation in the accuracy of measured variables to total corruption of its operation. The same problem does not exist for emissions, where comparison with a defined test limit is possible. To account for this variety, the generic immunity standards include three generalized performance criteria for the purpose of evaluating test results. In the test report, you must include a functional description and a specific definition of performance criteria based on these, during or as a consequence of the EMC testing. The definitions of these criteria can be found in section 9.3. Most noteworthy is that the criteria are grounded on what performance the user may reasonably expect *or is told to expect*. In other words, if you specify a given performance loss during application of the immunity test and write this into the user documentation, then provided the equipment does not actually become unsafe as a result of the test, you have met the requirements of the generic standards.

2.3.4.3 Basic and product standards

The tests defined in the generic standards are based only on internationally approved, already existing standards. For each electromagnetic phenomenon the test procedure given by such a standard is referenced, and a single test level or limit is laid down. No new tests are defined in the body of any generic standard.

Those standards which are referenced in the generic standards, for example the various parts of EN 61000 along with some of the CISPR standards, are known as "basic" standards. This means that such standards are entirely devoted to aspects of EMC that will prove to be of general interest and use to all committees developing other standards – for instance, product-specific standards. Generally, a product-specific standard will take a form similar to the generic standard, with similar limits, but will be more specific as regards operational modes and configurations, and about performance criteria that are considered acceptable. It will refer to the basic standards for the test methods wherever possible.

2.3.4.4 Harmonisation

The discussion in this chapter has repeatedly referred to standards as being "harmonised". This has a special meaning under the EC's New Approach, which covers the EMC and R&TTE Directives as well as many others. The meaning and structure is comprehensively outlined in the Commission's "Blue Guide" [179], which covers the implementation of New Approach Directives. In summary:

> Harmonised standards are European standards, which are adopted by European standards organisations, prepared in accordance with the General Guidelines agreed between the Commission and the European standards organisations, and follow a mandate issued by the Commission after consultation with the Member States. European standards are technical specifications adopted by European standards organisations for repeated or continuous application, with which compliance is not compulsory. According to the internal rules of these organisations, European standards must be transposed at national level. This transposition means that the European standards in question must be made available as national standards in an identical way, and that all conflicting national standards must be withdrawn in a given period.
>
> Harmonised standards in the meaning of the New Approach are deemed to exist when the European standards organisations formally present to the Commission the European standards elaborated or identified in conformity with the mandate. Harmonised standards are not a specific category amongst European standards. The terminology used in New Approach directives is a legal qualification of technical specifications existing as European standards, but to which a special meaning has been given by these directives.

From this, you can see that international (IEC) standards by themselves cannot be harmonised; only their European editions can be. Harmonised standards provide a *presumption of conformity* with the essential requirements, once their reference has been published in the *Official Journal of the EU*. The objective of publishing the reference is to set the earliest date for the presumption of conformity to take effect, and the entry in the OJ includes this information. The OJ listing is updated at occasional intervals, and the most recent can be found via links on the EC's Europa website.

In fact, because of this system, publication in the OJ assumes a considerable importance especially when difficulties have arisen with the application of a particular standard, and it has been amended or revised to take these into account. You cannot legitimately use a new version of a standard until it has been harmonised by an OJ listing. In 2005 and 2006 a complex situation arose with the application of EN 55022 which was exacerbated by the Commission's failure to publish necessary updates in any reasonable timescale, and indeed to introduce errors when it did. It's probably fair to say that the resulting quagmire has severely tested many participants' faith in the compliance system as it now exists.

2.3.4.5 Presumption of conformity

The meaning of this term is important: it says that if you correctly apply all relevant harmonised standards, it is to be *presumed* that you comply with the Directive's essential requirements; but it doesn't *guarantee* that you do. There may be deficiencies in a standard even if you apply it correctly, which mean that it is possible for a product still to cause or suffer from interference, despite complying with its requirements. Some standards have known loopholes which still exist, although their existence has been well documented for many years. Since you are required to certify compliance with the essential requirements, there will always be some element of risk which is unavoidable. The best that can be said is that an enforcement authority is likely to have a harder time proving non-compliance if you can show correct application of the standards.

The presumption of conformity is limited to the scope of the standard, and to the relevant essential requirements that are covered by the standard, and to *compliance with* (not merely application of) the standard (see 2.3.1.2). Not many standards explicitly state which essential requirements they cover, although it is slowly becoming more common as standards are updated. Just because it is listed in the OJEU, with a title that applies to your apparatus, this does not mean it covers exactly what you want it to, wholly and completely. So you still need to carry out (and document) some analysis yourself, in order to tell whether you are using an adequate standard or group of standards, or whether you need to expand the EMC assessment to cover those aspects that the chosen standard doesn't. As we see later in Chapter 9, a test plan is still needed to interpret even a comprehensive standard in the light of your particular product, so you would have had to do some such analysis anyway.

There is a further question as to what is meant by "correct application" of a harmonised standard. As discussed in section 2.2.3.5, the Directive states that this means doing everything by the book, that is, in a manner fully compliant with the methods as given in the standard. You may be prepared to do this, or to get a test house to do it for you; the cost may be justified. But if it isn't, and you use only "pre-compliance" methods of test, then your EMC assessment should include an analysis of the limitations of such methods and how you have dealt with them, since you will then only have "partially" applied a harmonised standard.

In respect of the protection requirements, the EMC Directive, at recital 13, indicates that harmonised standards are understood to "reflect the generally acknowledged state of the art as regards EMC matters in the European Union". Where harmonised standards are not employed in full, the manufacturer should have regard to the state of the art in terms of the services to be protected, and the electromagnetic disturbances to which his apparatus may be subjected.

2.4 Action for compliance for a product manufacturer

The second edition EMC Directive is all about risk. The dimensions of the risk are either that you will take insufficient measures to prove compliance with the Directive, in terms of identifying, performing the EMC assessment and testing to the correct harmonised standards, and meeting their requirements; or, you will spend so much time and effort in doing this that your product will not make it to market on time or to budget.

With this in mind, the steps to take for a new product to achieve compliance with the EMC Directive and bear the CE mark can be detailed as follows.

A. Self certification

1. From the marketing specification, determine what type of product it will be and what environment it will be sold for use within, and hence which if any product-specific or product-family standards published in the OJEU apply to it. If your company only ever makes or imports products for one particular application then you will be able to use the same product standard(s) for all products.
2. If no product standards apply, or they don't cover all the essential requirements (generally: LF disturbance emissions and immunity, transient immunity, and RF emissions and immunity), check the generic standards to see if the tests specified in them are applicable. The environmental classification will depend on the intended power supply connection.
3. If you cannot apply any harmonised standards, or can only partly apply them, then you will need to do a detailed EMC assessment of the equipment to cover the essential requirements, which may or may not involve other test methods.
4. Having determined what standards you will use, decide on the test levels and to what ports of the equipment (enclosure, power leads, signal/control leads) they will apply. In some cases there will be no choice, but in others the test applicability will depend on factors such as length of cable, EUT configuration and class of environment.
5. From this information you will be able to draw up a test plan, which specifies in detail the version and configuration of the EUT and any associated apparatus, the tests that will be applied to it and the pass/fail criteria. Test plans are covered in greater depth in Chapter 9. You can discuss this with your selected test house or your in-house test facility staff, and it will form the basis for your agreement with them and also for the technical documentation required by the provisions of the Directive.
6. Knowing the requirements of the test plan will enable you to some degree to incorporate cost-effective EMC measures into the product design, since the test limits and the points to which they will be applied will have been specified.
7. As the design progresses through prototype and pre-production stages you can make pre-compliance confidence tests to check the performance of the product and also the validity of the test plan. It is normal for both design and test plan to undergo iterative modifications during this stage.
8. Once the design has been finalized and shortly before the product launch

you can then perform, or get a test house to perform, full compliance tests against harmonised standards or against your own assessment as appropriate, on one or more production samples, the results of which are recorded in the technical documentation. Provided that confidence tests were satisfactory this should be no more than a formality.

9. You are then at liberty to mark the product, and/or its packaging or documentation with the CE mark (if there is no other Directive to satisfy) and your empowered signatory can sign the Declaration of Conformity, to be kept for ten years. If other New Approach Directives apply, a parallel process needs to be followed for each. The product can be placed on the market.
10. Once the product is in series production you must take steps to ensure that any changes to its build state allow it to continue to comply with the protection requirements.

B. Involvement of a Notified Body

This approach is in addition to the steps shown above.

11. In theory, once you have produced the full package of documentation at step 8 above, you select a Notified Body capable of assessing your type of product based on their published scope of designation, hand over the documentation to them, and expect a statement confirming compliance within a few weeks.
12. In practice, you will have chosen an NB at a much earlier stage, quite possibly from step 1 above, and confirmed with them what standards and/or what test plan content would be acceptable to them in order for them to provide a positive statement; also you will have agreed what the scope of their involvement will be, i.e. what aspects of the essential requirements they will be asked to review.

Chapter 3

The R&TTE Directive

The R&TTED is given a separate chapter in this book because of the proliferation of all types of product which incorporate wireless or telecom connectivity. As soon as any product includes a wired or wireless modem or a short-range receiver or transmitter, no matter how low its power, it becomes a radio or telecom device and the EMCD ceases to apply, to be replaced by the provisions and conformity assessment procedures of the R&TTED. Of course, for conventional radio equipment this is easy enough to understand, but many manufacturers who have never before had anything to do with radio are taken aback when they learn just how wide the scope of this Directive is.

3.1 The implementation of the R&TTE Directive

The Radio & Telecommunications Terminal Equipment Directive (99/5/EC) [187] went into effect on 8th April 2000, with a transition period to 7th April 2001; after this date all equipment within its scope had to comply with its provisions. It is a development of the earlier telecoms equipment Directive, 98/13/EC. Included in its scope is all telecoms terminal equipment (TTE), and all radio equipment, and it supersedes the EMC Directive for this equipment – although the EMC requirements are maintained, so that on that score at least there is little change.

It represents a fairly fundamental shift in the way that radio and telecom equipment, previously subject to national and pan-European type approval regimes, is regulated. The goals which the R&TTE Directive addresses were, basically, simplified and relaxed procedures, minimum essential requirements, consistency with the EC's approaches and a responsiveness to market needs. These aspects had not been successfully addressed by the TTE Directive 98/13/EC and its predecessor 91/263/EC, partly because of differences in national implementation, and partly because of a burdensome conformity assessment regime and disproportionately severe essential requirements. There was also a lack of mutual recognition of the various national radio type approval regimes within Europe. The Commission saw these factors as stifling growth and innovation within the telecoms marketplace and the R&TTE Directive has been aimed at dealing with these issues.

Largely, the Directive has been successful in this, as witness the dramatic growth within Europe of both wired and wireless telecom systems. Furthermore, it has achieved this success apparently without an observed increase in interference problems between radio systems and without affecting the integrity of telecomms networks, suggesting that the earlier type approval regimes were indeed unnecessarily strict.

3.1.1 Scope

The main definitions defining the Directive's scope are:

- TTE (Telecommunications Terminal Equipment): a product or relevant component thereof, which is intended to be connected by any means whatsoever to interfaces of public telecommunications networks
- RE (Radio Equipment): a product or relevant component thereof, capable of communication by means of the emissions and/or reception of radio waves utilizing the spectrum allocated to terrestrial/space radio communications
- Radio waves: electromagnetic waves of frequency from 9kHz to 3000GHz, propagated in space without artificial guide
- Interface:
  i) a network termination point, that is, a physical connection point at which a user is provided with accesss to a public telecommunications network, and/or
  ii) an air interface, specifying the radio path between items of RE, and their technical specifications

Note that under the above definitions, a product can be both TTE and RE: a mobile phone or wireless LAN card, for instance. Explicit exceptions from the scope are:

- apparatus exclusively used for public security, defence, state security, and state activities in the area of criminal law;
- marine equipment, civil aviation equipment and air traffic management equipment (all covered by their own regulations);
- amateur radio equipment, broadcast radio receivers, and cabling and wiring.

3.1.1.1 Modules

The definitions of RE and TTE include the notion of a "relevant component". Therefore, any terminal equipment module or radio module, when placed on the Community market must comply with the essential requirements of the R&TTED. Such a module could for instance be a modem card for a PC, as TTE. Examples of radio modules are any component (IC, hybrid circuit, plug-in unit, etc), which together with an antenna, constitutes the transmitter RF circuit of a radio communications device, which has well defined RF parameters and which can clearly be identified. It is the responsibility of the manufacturer to declare if a product is a module. Modules which are not "placed on the market" are by default not covered by the Directive, though a finished product which uses one, is.

In the common case where a terminal equipment module or radio module is integrated into an otherwise non-R&TTE product, the person integrating the module becomes the manufacturer of the final product, and is therefore responsible for demonstrating its compliance with the essential requirements of the R&TTED. Assessed radio modules installed in equipment in conformance with the manufacturer's installation instructions require no further evaluation under Article 3.2 (effective use of spectrum, see 3.1.2.2) of the Directive and do not require further involvement of a Notified Body for the final product. In all other cases, or if the manufacturer of the final product is in doubt, then the equipment integrating the radio module must be assessed against Article 3.2 [204].

3.1.2 Requirements

There are three sets of essential requirements:

a) applying to both TTE and radio equipment:
   - the objectives with respect to the safety requirements contained in the Low Voltage Directive 73/23/EEC but with no lower voltage limit applying;
   - the protection requirements with respect to EMC contained in 89/336/EEC.

b) Radio Equipment shall effectively use the spectrum allocated to terrestrial or space radio communications and orbital resources so as to avoid harmful interference.

c) Essential requirements which may be applied at the Commission's discretion:
   - interworking via networks with other apparatus to allow connection to interfaces of the appropriate type throughout the EC;
   - prevention of harm to the network or misuse of network resources, that may cause an unacceptable degradation of service;
   - safeguards to ensure data protection and user privacy;
   - support of certain features to ensure avoidance of fraud;
   - support of certain features to ensure access to emergency services;
   - support of certain features to facilitate use by users with a disability.

3.1.2.1 Safety & EMC

The R&TTE requirements incorporate the safety requirements of the LVD and the EMC requirements of the EMCD. An important extension is the removal of the lower voltage limit (50V AC or 75V DC) for application of the LVD (73/23/EEC). This means that safety requirements apply even to handheld, battery powered apparatus.

For example, this calls for mandatory application of radiation limits to prevent hazard to human health, so that mobile handheld transmitters should be subject to a SAR (specific absorption rate, see section 1.4.1.2) assessment. And not just cellphones: there is a strong move towards applying such assessment to any radio transmitter, and even other types of electrical and electronic apparatus. Standards harmonized under both the R&TTE and Low Voltage Directives have been published for exactly this purpose. Such standards do not necessarily demand an actual test of emitted radiation, but there will be many circumstances where this is the simplest method of demonstrating conformity, even if not the cheapest. Whether such standards actually contribute to the protection of human health in a way that is proportionate to their cost of implementation is open to debate.

Because of the removal of the lower voltage limit, safety under the R&TTED extends beyond just an assessment of the mains supply to the apparatus. The threat of electric shock is only one aspect of safety: fire or other hazards due to batteries, for instance, may have to be considered.

The EMC requirements for TTE will normally be covered, as with other information technology equipment under the EMCD, by application of EN 55022, 55024 and 61000-3-2, as discussed in Chapter 4. Radio EMC is generally dealt with by the EN 301 489 series of standards, also mentioned in Chapter 4. These make allowance for semi-specific radio performance criteria and the application of exclusion bands around the transmit and receive frequencies.

3.1.2.2 Effective use of spectrum

Type approval of radio transmitters has been converted into the additional requirement for effective use of the spectrum so as to avoid harmful interference. This does not preclude national authorities from applying restrictions on the grounds of local spectrum management through the licensing process, but they must not attempt to enforce a type-approval regime in this context. The technical requirements for spectrum use include such parameters as transmitter power and frequency control, spurious emissions and responses of receivers, and occupied bandwidth. These requirements are detailed in harmonised ETSI standards for the more common types of radio equipment; the conformity assessment regime allows for the creation of "essential radio test suites" for those types of radio that are not covered, or inadequately covered, in harmonised standards, and this is intended to be one function of Notified Bodies as discussed in the next section.

The definition of an air interface refers to "technical specifications" of the interface; Member States are required to notify the Commission of the types of interface offered by public network operators, who in turn must publish interface specifications of a sufficient level to enable design of apparatus, and to allow manufacturers to carry out (if they so wish) relevant tests. At the time of writing, some six years after the inception of the Directive, this obligation has not been carried out by many Member States. A list of links to each country's published specifications and other documents is available on the Commission's website[†], but the list is not complete, and by no means all of the links are in working order.

There is a requirement to inform the relevant national authorities whenever it is intended to place on the market equipment that uses non-harmonized spectrum allocations. The authorities then have a four-week period within which to raise objections.

3.1.2.3 Further requirements

The Directive also allows the Commission to impose extra requirements for certain classes of equipment, but to date this has not been applied. A particular requirement for terminal equipment is the prevention of harm to the network or its functioning, which causes an unacceptable degradation of service to persons other than the user of the apparatus. This aspect was traditionally handled by the type approval process. There were concerns that leaving the network requirement specifications hanging, as it were, in mid-air would damage the pan-European harmonization of the wired sector of the telecoms industry, but despite the continuing lack of availability of some Member States' network specifications in the public domain this does not seem to have become a serious issue.

The Directive is related to placing products on the market, and does not change the licensing and license exemption regimes in place in each Member State, which continue to be the mechanism for authorizing individual use of radio equipment.

3.2 The process of conformity assessment

Article 10.5 of the EMC Directive used to require radio transmitters (which may also be TTE, such as cellphone transmitters) to undergo EMC-specific type examination, which needed certification from a Notified Body. This was different from a Competent Body. With the entry into force of the R&TTE Directive, radio transmitters and

† http://ec.europa.eu/enterprise/rtte/weblinks.htm

receivers and TTE still follow different routes to compliance compared to other electronic apparatus, since they are complying with a different Directive.

Conformity assessment annexes

ii: Internal Production Control with technical documentation

iii: Annex ii plus specific tests

iv: Annex iii plus Technical Construction File submitted to a notified body

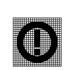
v: Full Quality Assurance assessed by a notified body

Figure 3.1 Compliance with the R&TTE Directive

3.2.1 Procedures

The conformity assessment procedures allowed under the R&TTE Directive are outlined in Figure 3.1 and Table 3.1. Their applicability varies depending on whether the equipment is telecoms terminal, or radio equipment; the receiving part of radio equipment is treated as telecoms equipment.

Table 3.1 Applicability of the conformity assessment procedures

<table>
<tr><th rowspan="2">Annex</th><th colspan="2">Applicable to</th><th rowspan="2">Role of the Notified Body</th><th rowspan="2">Marking</th></tr>
<tr><th>Without radio</th><th>With radio</th></tr>
<tr><td>ii</td><td>Terminal equipment</td><td>Receivers</td><td></td><td rowspan="2">CE only</td></tr>
<tr><td rowspan="2">iii</td><td rowspan="2"></td><td rowspan="2">Radio equipment including a transmitter complying with harmonised standards</td><td></td></tr>
<tr><td>Identification of the series of essential radio test suites</td><td rowspan="3">CE + NB number</td></tr>
<tr><td>iv</td><td>Terminal equipment</td><td>Radio equipment including a transmitter not complying or only partially complying with harmonised standards</td><td>Opinion on the conformity of the equipment based on a review of the manufacturer's TCF</td></tr>
<tr><td>v</td><td colspan="2">All equipment covered by the R&TTE directive</td><td>Certification of manufacturer's quality system</td></tr>
<tr><td colspan="4">The alert sign for Class 2 equipment must be indicated if a restriction on use applies to the equipment, and it must follow the CE marking</td><td></td></tr>
</table>

Annexes ii and iii

Annex ii can only be applied to TTE and radio receivers, but it requires only that the manufacturer puts together technical documentation, using harmonised standards at his own option, to support his own declaration of conformity. This is equivalent to the harmonised standards route in the original EMC Directive, and the EMC assessment method in the new EMCD. Annex iii applies to radio transmitters for which harmonised standards are available, in which case again the manufacturer applies these harmonised standards to support his own declaration. In some cases the harmonised standards will not include “essential radio test suites” and it is then the job of a Notified Body to identify these on a case-by-case basis, at the behest of the manufacturer. But it is not mandatory for the Notified Body to carry them out itself; the manufacturer can do this in any way he chooses, although the resulting CE Mark has to carry the identification of the Notified Body who specified the tests.

In either of the cases of annexes ii or iii, there is no absolute requirement for the involvement of a notified body in the *assessment* process. The specific tests in annex iii must be identified by a Notified Body unless they are already defined in the harmonized standard(s). But otherwise, these annexes represent pure self certification on the part of the manufacturer.

Annex iv

Annex iv involves the generation of a Technical Construction File, similarly to the mechanism of Article 10.2 in the original EMC Directive, which must then be assessed by a Notified Body. It is used for radio transmitters for which there are no harmonised standards, or for which harmonised standards have been only partly applied. It can also be applied to TTE instead of the Annex ii route. The TCF should contain similar documentation as for the earlier Annexes as well as the results of essential radio test suites (for radio transmitters) as agreed with the NB. The NB must issue an opinion within four weeks of receipt of the TCF; after these four weeks are up, or once he has received the opinion, the manufacturer can place the product on the market.

Annex v

The Full Quality Assurance method of Annex v may be an attractive route for a large manufacturer of radio equipment, since it is the only option that avoids the case-by-case involvement of a notified body for radio terminals whose tests are not defined in harmonized standards. Instead, the NB assesses the manufacturer’s own QA system for design, manufacture and final product inspection and test.

It must be appreciated that in all cases it is the manufacturer’s responsibility to ensure the product complies and to declare accordingly. Where an NB is used it has no power to *prevent* a product being marketed. Even with the Annex iv TCF mechanism, the NB only issues an opinion, but it is the manufacturer who makes the final decision on compliance. Clearly, a manufacturer will be taking a risk if he goes against an NB’s opinion; the risk will be defined by his view of the likelihood of suffering a serious non-compliance versus the effectiveness of enforcement action.

It has to be said that in fact, leaving aside the Annex iii process of specifying essential radio tests, which are in any case ill-defined, the actual experience of the Annex iv route has been almost opposite to that intended. Many smaller manufacturers, confused by the way in which they are forced into taking a responsibility for which they are ill-prepared and poorly briefed, will happily accept guidance from an authority who appears to be familiar with the system and who moreover is designated for that purpose – in the case of the UK, authorized by the accreditation agency UKAS, on behalf of the government. And who just happens to have laboratory testing as its main business activity. This leads to the situation described by one commentator as follows [45]:

> ... the unexpected effect is that lack of clarity has led to opportunity for commercial exploitation and a side-lining of established accreditation. We end up with a situation where “notification”

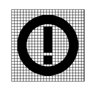
is of greater commercial value than "accreditation" and the norms for accreditation are flexed on an ad-hoc national basis for regulatory purposes. All too often, the de-facto scenario becomes third party certification by notified bodies in the guise of "opinions" based on testing they themselves have performed! This lack of rigour in the notified body world begins to undermine confidence not only in their activities but also in the expected reliance on simple manufacturers' declarations against harmonised standards as the main route to compliance.

3.2.2 Classes of radio transmitter

The Commission Decision of 6 April 2000 (2000/299/EC) [188] established two classes of equipment under the R&TTED:

- Class 1 equipment: Radio Equipment and TTE which can be placed on the market and be put into service without restrictions;
- Class 2 Radio Equipment: All Radio Equipment not falling into the definition of Class 1.

Class 1 equipment does not have a separate identifier: just the CE Mark (perhaps with a Notified Body number) is all that is needed. Class 2 equipment must be additionally marked with an "equipment class identifier" which has the same height as the initials "CE". This "alert symbol" is an exclamation mark in a circle:



In order to decide whether they have to comply with this requirement, manufacturers need to know what equipment falls into what class. Class 1 is divided by the Commission into a number of sub-classes of which sub-classes 1 to 6 and 8 are varieties of TTE:

- Sub-class 1. ISDN (ISDN Basic Rate, ISDN Primary Rate, ISDN U, Broadband ISDN ATM);
- Sub-class 2. PSTN (Analogue single line, Analogue multi-line (with/ without DDI), equipment attached to Centrex interfaces or Virtual Private Networks);
- Sub-class 3. Leased lines (2w and 4w analogue (baseband), 2w and 4w analogue (voiceband), Digital, SDH, optical);
- Sub-class 4. Wired data equipment (X.21, X.25, ethernet, token ring, token bus, TCP/IP, frame relay);
- Sub-class 5. Wired interactive broadcast equipment (unswitched vision/ sound, switched vision/sound);
- Sub-class 6. Telex (single line equipment, multiple line equipment);
- Sub-class 7. Receive-only radio equipment;
- Sub-class 8. Other terminal equipment attached to fixed networks.

Higher sub-classes refer to various radio transmitters. The ERO (the European Radiocommunications Office) in Denmark maintains a list on its website, along with a tool called EFIS which gives the basic technical description of each type of apparatus represented in a sub-class. Table 3.2 shows a summary of these sub-classifications published as at summer 2006; consult the website† for a more up-to-date listing.

† http://www.ero.dk/rtte

Table 3.2 Sub-classes for Class 1 Radio Equipment

Sub-classes	Frequency bands	Service
9	890–915/935–960 MHz 880–890/925–935 MHz 876–880/921–925 MHz 1710–1785/1805–1880 MHz	GSM, GSM-R
11, 12	1525.0–1544.0 MHz 1555.0–1559.0 MHz 1631.5–1634.5 MHz 1656.5–1660.5 MHz 10.7–11.7 GHz 12.5–12.75 GHz 14.0–14.25 GHz	Land Mobile Earth Stations
13	380–385 MHz 390–395 MHz	TETRA
14, 15	1610–1613.5 MHz 1613.8–1626.5 MHz 2483.5–2500 MHz 2483.5–2500 MHz 1980–2010 MHz 2170–2200 MHz	Satellite-Personal Communications (PCN) Earth Stations
16	1525.0–1544.0 MHz 1555.0–1559.0 MHz 1626.5–1645.5 MHz 1656.5–1660.5 MHz	Land Mobile Satellite Service
18	1880–1900 MHz	DECT
19, 20, 21, 25, 28, 29, 30, 31, 43	40.665 MHz, 40.675 MHz, 40.685 MHz, 40.695 MHz 433.05–434.79 MHz 2400–2483.5 MHz 26.995 MHz, 27.045 MHz, 27.145 MHz, 27.195 MHz 868.0–868.6 MHz 868.7–869.2 MHz 869.4–869.65 869.7–870 MHz 5725–5875 MHz	Non-specific Short Range Devices
22	2400–2483.5 MHz	Wideband Data Transmission Systems including RLANs
24, 36, 37, 38, 39, 40, 41, 42, 44, 45	13.553–13.567 MHz 20.05–59.75 kHz 59.750–60.250 kHz 60.250–67 kHz 67–70 kHz 70–119 kHz 119–127 kHz 127–135 kHz 6765–6795 kHz 7400–8800 kHz	Inductive applications

Table 3.2 Sub-classes for Class 1 Radio Equipment (Continued)

Sub-classes	Frequency bands	Service
26, 27	2446–2454 MHz 24.15–24.175 GHz	Movement Detection
32, 33, 34	868.6–868.7 MHz 869.25–869.3 MHz 869.65–869.7 MHz	Alarms
35	869.2–869.25 MHz	Social Alarms
46	863–865 MHz	Radio Microphones
47	402–405 MHz	Ultra Low Power Active Medical Implants
48	863–865 MHz	Wireless Audio
49	457 kHz	Avalanche Victims
50	76–77 GHz	Road Transport & Traffic Telematics (RTTT)
51	446.0–446.1 MHz	PMR446

3.2.3 Notification

If a radio product uses a frequency band or bands which is not harmonised throughout the EC, the manufacturer has to notify any Member State's national authority of their intention to place the equipment on the market in their country. Member States have the authority to prohibit the placing on the market, or to withdraw from the market, radio equipment which they consider has caused or is likely to cause harmful interference.

The person responsible must notify their intention at least four weeks before the equipment is first placed on the national market. The notification must include information on the radio characteristics such as operating frequency, channel spacing, modulation, RF power and so on. The notification form has to be sent to the responsible national authority. In the UK, the responsible authority is Ofcom, who accept notifications by e-mail; other administrations do the same.

One practical difficulty with the notification requirements is for manufacturers to know, for any given equipment, whether a notification is required or not. The information needed on the use of frequencies has to be provided by many different administrations spread across the EC. Manufacturers would find it difficult to compile this information, and so a mechanism is needed to allow manufacturers to discover when notifications are not required. Details of the technologies for which harmonised frequency bands are available are maintained by ERO (see Table 3.2 above), and can be accessed via the internet from the ERO website.

In the UK, Ofcom's position is that it will alert manufacturers when they intend marketing products in the UK that cannot be legally used. In the vast majority of cases an alert to the manufacturer will be all that is necessary. It will then be the manufacturer's responsibility to provide the necessary warning to its customers on the equipment packaging and instructions for use. But, manufacturers should not rely on the notification procedure to identify countries where equipment can and cannot be legally used or licensed. Administrations have no obligation under the Directive to respond to a notification: lack of response does not mean acceptability.

3.2.4 Information requirements

Where Radio Equipment is involved, the packaging and user instructions must identify the Member State or geographical area within a Member State where the apparatus is intended to be used. In addition, the marking on Radio Equipment must warn the user that there may be restrictions of use of the apparatus in certain Member States, using the alert symbol shown in section 3.2.2; on TTE it should indicate the interfaces of the networks to which the equipment is intended to be connected.

Unlike the EMC Directive, which requires the manufacturer to make a Declaration of Conformity but only mandates its availability to enforcement officers, the R&TTED requires that the D of C is provided to the user. This can be either a full (multilingual) copy, or an abbreviated version:

> Hereby, [Name of manufacturer], declares that this [type of equipment] is in compliance with the essential requirements and other relevant provisions of Directive 1999/5/EC.

as long as the abbreviated version is accompanied by details of how to obtain a copy of the D of C, or by a copy of the actual D of C in the language of the manufacturer.

3.2.4.1 Indication of the intended use of the equipment

It may seem obvious, but the R&TTED requires that the user must be provided with information on the intended use of the apparatus. The Commission envisages that this could be done in a number of ways:

- written description: e.g. cordless telephone with answering machine, Garage door remote control;
- description in visual form: illustration on the packaging, photo in the user manual, pictogram, equipment visible through the packaging;
- by the use of terms known to the public: e.g. baby monitor, modem, PMR, GSM terminal, etc.

If the description is in written form, it has to be in the official language(s) of the region in which the equipment is placed on the market. For a multilingual region this means all languages. Since the EC is multilingual, and since most suppliers want market access to the EC as a whole, this explains the preponderance of pictograms.

3.2.4.2 Countries where the equipment is intended to be used

As above, this information can be presented in written form, as a pictogram, or in abbreviated fashion using the following two-letter country identifiers (from ISO 3166):

Austria	AT	Malta	MT
Belgium	BE	Netherlands	NL
Cyprus	CY	Poland	PL
Czech Republic	CZ	Portugal	PT
Denmark	DK	Slovakia	SK
Estonia	EE	Slovenia	SI
Finland	FI	Spain	ES
France	FR	Sweden	SE
Germany	DE	United Kingdom	GB

Greece	GR	Iceland	IS
Hungary	HU	Liechtenstein	LI
Ireland	IE	Norway	NO
Italy	IT	Switzerland	CH
Latvia	LV	Bulgaria	BG
Lithuania	LT	Romania	RO
Luxembourg	LU	Turkey	TR

3.2.5 Marking of equipment and documentation

The requirements for marking under the R&TTED are rather more involved than just the simple CE mark as is used for other Directives. There are inconsistencies between it and other Directives in respect of administrative requirements.

3.2.5.1 General requirements

Annex vii of the Directive requires of the CE Mark:

> The CE marking must be affixed to the product or to its data plate. Additionally it must be affixed to the packaging, if any, and to the accompanying documents.

The word "additionally" here distinguishes this Directive from other CE Marking Directives, which allow affixing *either* to the product *or* to its documentation and packaging. This single requirement is responsible for more non-compliances with the Directive than any other, since it is so easy to overlook.

Extra requirements in this Directive are the need for an identification number for the Notified Body used in Annexes iii, iv and v; and the need for the Class 2 equipment class identifier (the exclamation mark in a circle) for equipment subject to restrictions on use. It's also a requirement that the equipment itself shall be identified, with the model, manufacturer's name and serial or batch number.

3.2.5.2 Equipment containing an R&TTE component

According to the Commission's guidance, equipment containing an R&TTE component must be marked in the following way:

- Equipment, which at the time of placing on the market contains as an integral part an R&TTE component, which should not be removed by the user, should be marked according to the R&TTE Directive. In addition, in its user manual it should declare compliance with the R&TTE Directive and if necessary indicate geographic limitations of use;
- Equipment, which provides for the capability that users insert R&TTE components, but in themselves are not covered by the R&TTE Directive (e.g. laptop computers) should not be marked according to the R&TTE Directive.

By "marking according to the R&TTE Directive" it is understood that this Directive does apply. So, for example, a laptop with an integral wireless modem falls under the R&TTED, but a laptop with no wireless connectivity does not, even though a plug-in wireless modem card could be inserted. Such a card should itself be declared under the R&TTED.

Chapter 4

Commercial standards

4.1 The standards making bodies

The structure of the bodies which are responsible for defining EMC standards for the purposes of the EMC Directive is shown in Figure 4.1.

4.1.1 The International Electrotechnical Commission

The IEC operates in close co-operation with the International Standards Organization (ISO), and in 2006 had 68 member countries including 16 associate members. It is

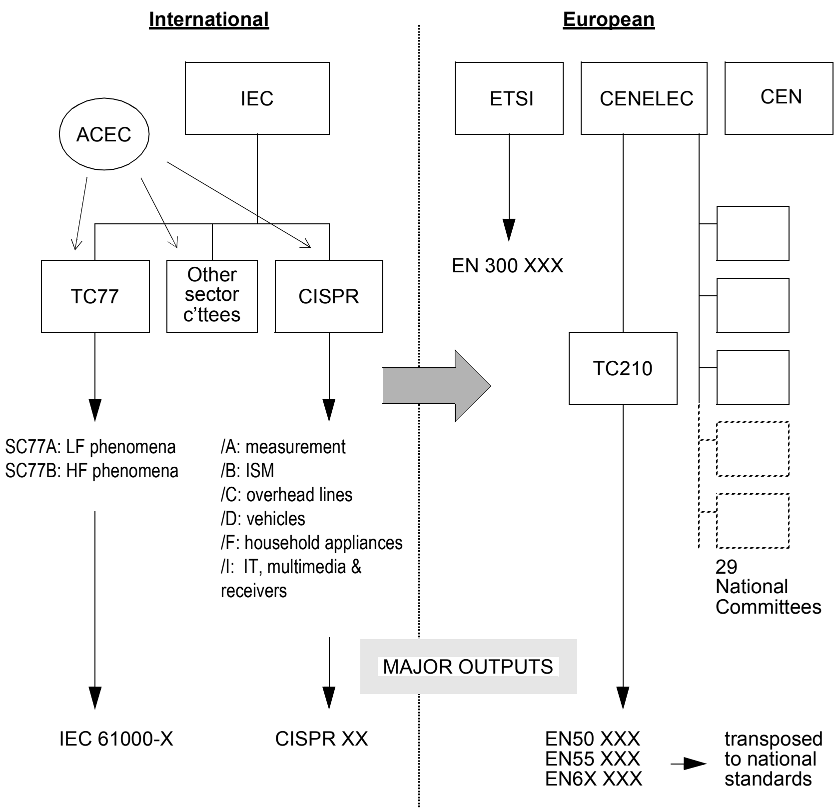


Figure 4.1 EMC standards structure

composed of National Committees which are expected to be fully representative of all electrotechnical interests in their respective countries. Work is carried out in technical committees and their sub-committees addressing particular product sectors, and the secretariat of each technical committee is the responsibility of one of the National Committees, which appoints a Secretary with the necessary resources. The IEC's objectives are "to promote international co-operation on all questions of standardization.... (this is) achieved by issuing publications including recommendations in the form of international standards which the National Committees are expected to use for their work on national standards."[88]

Two IEC technical committees are devoted full-time to EMC work, although nearly 40 others have some involvement with EMC as part of their scope. The two full-time committees are TC77, *Electromagnetic compatibility between equipment including networks*, and the *International Special Committee on Radio Interference* or CISPR, which is the acronym for its French title. There is also the Advisory Committee on EMC (ACEC), which is expected to prevent the development of conflicting standards.

IEC standards themselves have *no legal standing* with regard to the EMC Directive. If the National Committees do not agree with them, they need not adopt them; although in the UK, 85% of IEC standards are transposed to British Standards. The real importance of the IEC standards is that they may either be transposed directly into harmonised EN standards, in which case they become applicable for the self certification route, or they may be referred to by product-specific or generic harmonised standards.

Table 4.1 Plan of IEC 61000

IEC 61000-1	Part 1: General General considerations (introduction, fundamental principles, functional safety) Definitions, terminology
IEC 61000-2	Part 2: Environment Description of the environment Classification of the environment Compatibility levels
IEC 61000-3	Part 3: Limits Emission limits Immunity limits (if not the responsibility of product committees)
IEC 61000-4	Part 4: Testing and measurement techniques Measurement techniques Testing techniques
IEC 61000-5	Part 5: Installation and mitigation guidelines Installation guidelines Mitigation methods and devices
IEC 61000-6	Part 6: Generic standards
IEC 61000-9	Part 9: Miscellaneous (none published to date)

IEC 61000 is published in separate parts by IEC TC77 according to the above plan. Each part is further subdivided into sections which can be published either as international standards or as Technical Reports.

4.1.1.1 TC77

TC77 has been characterized as "The United Nations for EMC" [99]; certainly it attempts to cover most aspects of the subject on a worldwide basis. The structure of TC77 is shown in Figure 4.2. It is a large and influential group, and liaises with several

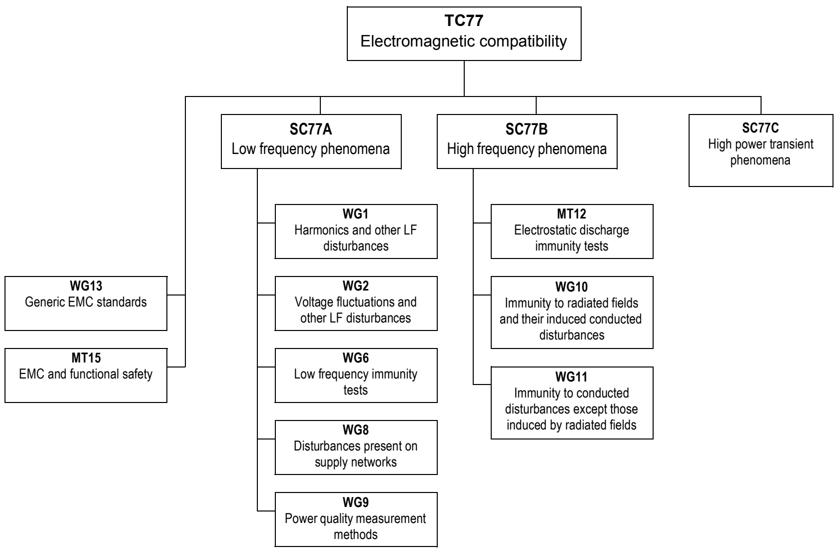


Figure 4.2 Structure of IEC TC77

other product-related committees within IEC including CISPR, as well as with outside bodies such as CENELEC, ITU and several electric power-related groups.

The major output of TC77 now is the various parts of IEC Publication 61000, *Electromagnetic Compatibility*. This document has been published in stages as defined by the plan shown in Table 4.1, and incorporates all non-CISPR and non-product-specific EMC material. Previous EMC standards such as IEC 555 and IEC 801 have been subsumed within IEC 61000. A detailed description of some sections of IEC 61000 parts 3 and 4 can be found later in this chapter under section 4.5, but meanwhile, a complete list of the parts of this mammoth standard that have been published or are in preparation up to autumn 2006 is given in Table 4.3.

4.1.1.2 CISPR

CISPR publications deal with limits and measurement of the radio interference characteristics of potentially disturbing sources, and look set to continue to co-exist with IEC 61000. There are a number of sub-committees as shown in Table 4.2. Most of these relate to particular product groups and have a historical basis; before the advent of pan-European legislation these products were the major ones subject to legislative control on their emissions. CISPR/A has an important role as the developer and guardian of common test methods and instrument specifications. CISPR/E and /G were amalgamated into CISPR/I in 2001.

Although all the output of CISPR sub-committees (except A and H) is nominally product related, several of the emissions standards – particularly CISPR 11, 14 and 22 – have assumed wider importance since their limits and test methods are referenced in many more product standards. In general, the limits versus frequency are rationalised into two classifications, A and B (see section 4.8), which are common across most of the standards. Although CISPR is not in general interested in immunity standards, two anomalous instances exist: CISPR 20 for broadcast receivers and associated apparatus, and CISPR 24 for information technology equipment.

Table 4.2 Structure of CISPR

Committee	Title/Scope	Main publications
CISPR/A	Radio interference measurements and statistical methods	CISPR 16, CISPR 17
CISPR/B	Industrial, Scientific and Medical radio-frequency apparatus	CISPR 11, CISPR 19, CISPR 23, CISPR 28
CISPR/C	Overhead power lines, high-voltage equipment and electric traction systems	CISPR 18
CISPR/D	Electrical/electronic equipment on vehicles and internal combustion engine powered devices	CISPR 12, CISPR 21, CISPR 25
CISPR/F	Household appliances, tools, lighting equipment and similar apparatus	CISPR 14, CISPR 15
CISPR/H	Limits for the protection of radio services	CISPR/TR 31, IEC 61000-6-3, -6-4
CISPR/I	Information technology, multimedia equipment and receivers	CISPR 13, CISPR 20, CISPR 22, CISPR 24; future CISPR 32, 35

4.1.1.3 The IEV

One further important document is Chapter 161 of IEC Publication 60050 [152], the International Electrotechnical Vocabulary. This contains definitions of EMC terminology in English, French and Russian, with equivalent terms in Dutch, German, Italian, Polish, Spanish and Swedish.

Table 4.3 Published and planned parts of IEC 61000
(Shaded sections were not yet published in autumn 2006)

Part	Section	Title
1	General	
	1	Application and interpretation of fundamental definitions and terms
	2	Methodology for the achievement of the functional safety of electrical and electronic equipment with regard to electromagnetic phenomena
	3	The effects of high-altitude EMP (HEMP) on civil equipment and systems
	4	Historical rationale for the limitation of power-frequency conducted harmonic current emissions from equipment, in the frequency range up to 2kHz
	5	High power electromagnetic (HPEM) effects on civil systems
2	Environment	
	1	Electromagnetic environment for low-frequency conducted disturbances and signalling in public power supply systems
	2	Compatibility levels for low-frequency conducted disturbances and signalling in public low-voltage power supply systems
	3	Radiated and non-network-frequency-related conducted phenomena
	4	Compatibility levels in industrial plants for low-frequency conducted disturbances
	5	Classification of electromagnetic environments
	6	Assessment of the emission levels in the power supply of industrial plants as regards low-frequency conducted disturbances

Table 4.3 Published and planned parts of IEC 61000 (Continued)
(Shaded sections were not yet published in autumn 2006)

Part	Section	Title
	7	Low frequency magnetic fields in various environments
	8	Voltage dips and short interruptions on public electric power supply systems with statistical measurement results
	9	Description of HEMP environment. Radiated disturbance
	10	Description of HEMP environment. Conducted disturbance
	11	Classification of HEMP environments
	12	Compatibility levels for low-frequency conducted disturbances and signalling in public medium-voltage power supply systems
	13	High-power electromagnetic (HPEM) environments – Radiated and conducted
	14	Overvoltages on public electricity distribution networks
3	Limits	
	1	Overview of emission standards and guides – Technical Report
	2	Limits for harmonic current emissions (equipment input current <= 16 A per phase)
	3	Limitation of voltage changes, voltage fluctuations and flicker in public low-voltage supply systems, for equipment with rated current <= 16 A per phase and not subject to conditional connection
	4	Limitation of emission of harmonic currents in low-voltage power supply systems for equipment with rated current greater than 16 A
	5	Limitation of voltage fluctuations and flicker in low-voltage power supply systems for equipment with rated current greater than 16 A
	6	Assessment of emission limits for distorting loads in MV and HV power systems
	7	Assessment of emission limits for fluctuating loads in MV and HV power systems
	8	Signalling on low-voltage electrical installations. Emission levels, frequency bands and electromagnetic disturbance levels
	9	Limits for interharmonic current emissions (equipment with input power <=16 A per phase and prone to produce interharmonics by design)
	10	Emission limits in the frequency range 2 ... 9 kHz
	11	Limitation of voltage changes, voltage fluctuations and flicker in public low-voltage supply systems for equipment with rated current <= 75A and subject to conditional connection.
	12	Limits for harmonic currents produced by equipment connected to public low-voltage systems with input current >16 A and <=75 A per phase
	13	Assessment of emission limits for the connection of unbalanced installations to MV, HV and EHV power systems
	15	Assessment of electromagnetic immunity and emission requirements for dispersed generation in LV networks
4	Testing and measurement techniques	
	1	Overview of IEC 61000-4 series
	2	Electrostatic discharge immunity test
	3	Radiated radio frequency electromagnetic field immunity test
	4	Electrical fast transient/burst immunity test
	5	Surge immunity test
	6	Immunity to conducted disturbances induced by radio frequency fields
	7	General guide on harmonics and interharmonics measurements and instrumentation, for power supply systems and equipment connected thereto
	8	Power frequency magnetic field immunity test
	9	Pulse magnetic field immunity test
	10	Damped oscillatory magnetic field immunity test
	11	Voltage dips, short interruptions and voltage variations immunity test
	12	Oscillatory waves immunity test (to become Ring wave immunity test, with damped oscillatory wave test moved to part 18)

Table 4.3 Published and planned parts of IEC 61000 (Continued)
(Shaded sections were not yet published in autumn 2006)

Part	Section	Title
	13	Harmonics, interharmonics including mains signalling at AC power port, low frequency immunity tests
	14	Voltage fluctuation immunity test
	15	Flickermeter – functional and design specifications
	16	Test for immunity to conducted, common mode disturbances in the frequency range 0 Hz to 150 kHz
	17	Ripple on DC input power port immunity test
	18	Oscillatory waves immunity test (new edition)
	20	Emission and immunity testing in transverse electromagnetic (TEM) waveguides
	21	Reverberation chamber test methods
	22	Radiated emissions and immunity measurements in fully anechoic rooms (FARs)
	23	Test methods for protective devices for HEMP and other radiated disturbances
	24	Test methods for protective devices for HEMP conducted disturbance
	25	HEMP immunity test methods for equipment and systems
	27	Unbalance immunity test
	28	Variation of power frequency immunity test
	29	Voltage dips, short interruptions and voltage variations on DC input power port immunity tests
	30	Power quality measurement methods
	31	Measurements in the frequency range 2kHz to 9kHz
	32	High-altitude electromagnetic pulse (HEMP) simulator compendium
	33	Measurement methods for high-power transient parameters
	34	Voltage dips, short interruptions and voltage variations immunity tests for equipment with input current more than 16 A per phase
	35	Intentional Electromagnetic Interference (IEMI) Simulator Compendium
5	Installation and mitigation guidelines	
	1	General considerations
	2	Earthing and cabling
	3	HEMP protection concepts
	4	Specification for protective devices against HEMP radiated disturbance
	5	Specification of protective devices for HEMP conducted disturbance
	6	Mitigation of external EM influences
	7	Degrees of protection by enclosures against electromagnetic disturbances (EM code)
	8	HEMP protection methods for the distributed civil infrastructure
	9	System-level susceptibility assessments for HEMP and HPEM
6	Generic standards	
	1	Immunity for residential, commercial and light-industrial environments
	2	Immunity for industrial environments
CISPR	3	Emission standard for residential, commercial and light-industrial environments
CISPR	4	Emission standard for industrial environments
	5	Immunity for power station and substation environments
	6	HEMP immunity for indoor equipment
	7	Generic emission standard for in-situ measurements

4.1.2 CENELEC and ETSI

CENELEC (the European Committee for Electrotechnical Standardization) is the European standards making body, which has (among many other things) been mandated by the Commission of the EC to produce EMC standards for use with the

European EMC Directive. For telecommunications equipment ETSI (the European Telecommunications Standards Institute) is the mandated standards body. ETSI generates standards for telecoms network equipment – that is, equipment not intended for the subscriber, in contrast to terminal equipment, which is – and for radio communications equipment and broadcast transmitters.

CENELEC and ETSI use IEC/CISPR documents wherever possible as a basis for preparation of such standards, through a mechanism known as "parallel voting". This is so that European standards do not stray far out of line with international requirements, which would create difficulties for global trade. The committee charged with preparing the EMC standards is TC210. Representatives of National Committees meet in TC210 about once a year to discuss the technical implementation of the drafts. TC210 has a sub-committee, SC210A, which is concerned specifically with immunity of Information Technology Equipment (ITE), and other working groups.

CENELEC is made up of the National Committees of each of the EEA countries; adoption of standards is based on a qualified weighted voting by the 29 National Committees [44][136]. Of these member committees France, UK, Germany and Italy have 10 votes, Spain has 8 votes and smaller countries have one or two votes. There are two requirements for a standard to be approved: the vote must yield a majority of National Committees in favour, and at least 71% of the weighted votes cast must be positive.

Unlike the position with international standards, a country must accept a new CENELEC standard even if it voted against it. Formal national conditions may be attached to the standard to ameliorate this situation, such as the occasion when CENELEC decided to harmonize on a 230V mains supply, and the UK declared to stay at 240V as a special national condition.

In the UK the BSI committee GEL210 generates the British position on TC210 papers. The BSI has an obligation to invite all organizations which have an interest in EMC to be members of GEL210 – in practice this is done mostly through representation by trade organizations.

Once CENELEC has produced and agreed a European EMC standard (prefixed with EN or HD) all the CENELEC countries are required to implement identical national standards. The EN will be transposed word for word, while the HD (harmonisation document) does not need to be reproduced verbatim as long as it reflects the technical content. In the context of European Directives, the standard is notified to the Commission and the reference number of the EN and the equivalent national standards will then be published in the *Official Journal of the European Union* (OJEU), and once this is done the standard is deemed to be a "relevant standard" for the purpose of demonstrating compliance with the appropriate Directive (section 2.3.4.4). Conflicting national standards must be withdrawn within a limited time frame.

Draft standards and amendments to existing standards are made available for public comment, through the National Committees, for some time before the standard is actually published. Apart from being the mechanism by which industry can (if it has sufficient resources and interest) influence the content of the standards, this has the further advantage of permitting manufacturers to make an informed decision on the testing and limit levels to which they may choose to submit their products in advance of the actual publication date, even though it is not possible to make an official declaration of compliance with an unpublished standard. There is of course some risk that the final published version will differ in detail, and sometimes quite substantially, from the draft.

4.1.2.1 Product standards

As mentioned in section 2.3.4, the intent of the EMC Directive is that self certification should be serviced primarily by a whole range of product standards. When published and harmonised, these take precedence over the generic standards and may either be drafted specifically to cover the EMC aspects of a particular range of product types, or they could be EMC sections added to an existing product performance standard. The general intention is that these standards should refer to basic standards (such as the IEC 61000-4-X series or their EN equivalent) for test methods wherever possible, and the product-specific aspects should consist mainly in defining what tests to carry out, with what levels or limits, and what operating conditions and performance criteria to apply. The impetus to develop such standards should come from the industry sectors themselves.

Since this approach means that non-EMC committees can (and indeed are expected to) contribute, there is a wide range of standards organizations that can participate in generating such documents. It includes CEN, CENELEC and ETSI product committees as well as IEC and ISO committees – the latter feeding into the European regime through the process of parallel voting, whereby a draft is circulated within both CENELEC and IEC for consideration at the same time. To be sure whether there is, or will be, an EMC product standard which covers your particular activities, you have to continually monitor the standards development process – trade associations, and the websites of the standards agencies, are usually the most useful route for this purpose.

The following sections (4.2 *et seq*) outline those standards which form harmonised standards or basic standards for the purposes of the EMC Directive, which have been announced in the *Official Journal of the EU* [185]. They only briefly refer to the ETSI radio standards, which form a large group in themselves.

4.1.2.2 ETSI radio standards

The listing for the R&TTE Directive in the OJEU includes 25 EMC standards in the EN 301489 series for radio equipment. These appeared in the latter half of the 1990s, and with the implementation of the R&TTED were redrafted and reorganized. Because ETSI were able to start from scratch in developing EMC standards, there is much greater consistency and co-ordination between and within these documents than is the case for the other product standards, which come from several sources and often carry a great deal of historical baggage.

If your product involves a radio device then you will need to have regard to one of these standards, and it can normally be used as a stand-alone document since it will typically cover both emissions and immunity. It will also cover the particular issues, such as exclusion bands, that arise when a general EMC requirement is applied to a radio receiver or transmitter.

4.1.2.3 CEN

A few harmonised EMC standards are published by CEN, which is the European standards body for non-electrotechnical subjects. The main products covered by these documents are machines that have some electrical aspect. They can be recognized by their numbering, which although prefixed by EN does not fit into the 50XXX, 55XXX or 6XXXX series used by CENELEC.

4.1.2.4 The timescale for adoption of standards

Because standards are introduced or amended frequently, there has to be a formal mechanism for deciding by what date changes become mandatory. Clearly it would be impossible for a change to be enforced on the date of publication in the OJEU. The method is implemented by a column in the table published in the OJEU and headed "Date of cessation of presumption of conformity of the superseded standard". Generally the date of cessation of presumption of conformity will be the date of withdrawal (DOW), set by the European standards body and published in the EN version of the standard, but in certain exceptional cases this can be otherwise. The DOW will be typically 2–3 years after the date of publication of the new standard or amendment.

In the period between the publication of the new version and the DOW, you are entitled to choose either old or new versions for your self certification. By the time of the date published in the OJEU, you should be sure to have updated your declaration of conformity, including carrying out any new testing that the new version requires (it is very rare for new standards to be more relaxed!). Figure 4.3 illustrates this graphically.

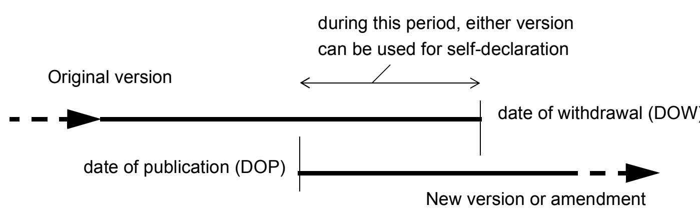


Figure 4.3 Applying changes in standards

If the new standard has a narrower scope than the superseded standard, on the date stated the (partially) superseded standard ceases to give presumption of conformity for those products that fall within the scope of the new standard. Presumption of conformity for products that still fall within the scope of the (partially) superseded standard, but that do not fall within the scope of the new standard, is unaffected. In the case of amendments, the referenced standard is EN XXXXX:YY, its previous amendments, if any, and the new, quoted amendment. The superseded standard therefore consists of EN XXXXX:YY and its previous amendments, if any, but without the new quoted amendment. On the date stated, the superseded standard ceases to give presumption of conformity with the essential requirements of the Directive.

4.1.2.5 Dated references

The structure of the standards that are harmonised for the EMC Directive is, as explained above, two-tier: the product and generic standards list the tests and levels, but refer to basic standards for the test method. This means that to get the complete picture you must build up a library of documents, easily exceeding 10 and sometimes 20 separate standards, in order to cover all your requirements. But each of these documents is subject to change, and you must then keep up with all the amendments and revisions as well. Since CENELEC/IEC standards are expensive, this process can consume a not insignificant budget. But just as importantly, it introduces a potentially serious source of confusion, because a basic standard may change but this change will not be immediately reflected in the product standards which reference it.

To attempt to deal with the confusion, CENELEC created the distinction between dated and undated references. Each CENELEC standard includes an Annex ZA, entitled “Normative references to international publications with their corresponding European publications”. This cross-references the IEC documents referred to in the text (which is usually word-for-word the same as the IEC text) with those documents which are to be used in a European context. Hence, even if the text refers to an IEC standard, say IEC 61000-4-3, you should actually use the EN version, i.e. EN 61000-4-3 as quoted in Annex ZA. Importantly, Annex ZA may quote dates against the EN standard, and it includes a standard phrasing which says:

> For dated references, subsequent amendments to or revisions of any of these publications apply to this European Standard only when incorporated in it by amendment or revision. For undated references the latest edition of the publication referred to applies (including amendments).

This therefore should resolve the question, when you are using a particular product or generic standard, of whether to pay attention to later revisions to the test methods in the basic standards – which can have potentially dramatic consequences for your compliance position, for instance when a later revision extends the frequency range or tightens up the specification of a test waveform. If the reference is undated, you should use the latest version; if it is dated you should stick with the specified version.

Warning: do not rely solely on the following information in this book to make a legal declaration of conformity. Obtain and refer to the appropriate standard directly.

4.2 Generic standards – emissions

CENELEC put great urgency on the development of generic standards [128][129], but they are now being gradually superseded by a whole raft of new product standards. There is a particular significance to ENs 55011, 55014-1 and 55022: as well as being product standards in themselves, they also specify RF emissions test methods that are applied very much more widely, and they are included here in the same section as the generics. The generic standards listed below all claim to represent essential EMC emissions or immunity requirements.

4.2.1 EN 61000-6-3: 2001 + A11: 2004

Title	Generic emission standard: Residential, commercial and light industrial environment
Equivalents	IEC 61000-6-3
Scope	All apparatus intended for use in the residential, commercial and light industrial environment for which no dedicated product or product-family emission standards exist NB equipment installed in the residential, commercial and light industry environment is considered to be directly connected to the public mains supply or to a dedicated DC source. Typical locations are residential properties, retail outlets, laboratories, business premises, outdoor locations, etc.
Tests	Enclosure: radiated emissions from 30 to 1000MHz as per EN 55022

Class B; applicable only to apparatus containing processing devices operating above 9kHz

AC mains port: conducted emissions from 150kHz to 30MHz as per EN 55022 Class B

Discontinuous interference on AC mains port measured at spot frequencies as per EN 55014-1, if relevant

Mains harmonic emission measured as per EN 61000-3-2, mains flicker measured as per EN 61000-3-3

Signal, control and DC power ports: conducted current from 150kHz to 30MHz using current probe, according to EN 55022 Class B

4.2.2 EN 61000-6-4: 2001

Title Generic emission standard: industrial environment

Equivalents IEC 61000-6-4

Scope Apparatus intended for use in the industrial environment, for which no dedicated product or product-family immunity standard exists, but excluding radio transmitters

NB equipment installed in the industrial environment is not connected to the public mains network but is considered to be connected to an industrial power distribution network with a dedicated distribution transformer

Tests Enclosure: radiated emissions from 30 to 1000MHz as per EN 55011 Class A

AC mains port: conducted emissions from 150kHz to 30MHz as per EN 55011 Class A; impulse noise appearing more often than five times per minute is also covered. Applicable only for apparatus operating at less than 1000V_{rms} AC

4.3 Main product standards: emissions

4.3.1 EN 55011: 1998 + A1: 1999 + A2: 2002

Title Industrial, scientific and medical (ISM) radio-frequency equipment – Radio disturbance characteristics – Limits and methods of measurement

Equivalents CISPR 11 third edition

Scope Equipment designed to generate RF energy for industrial, scientific and medical (ISM) purposes, including spark erosion; excluding applications in telecomms and IT, or covered by other CISPR publications

Class A equipment is for use in all establishments other than domestic; Class B equipment is suitable for use in domestic establishments

Group 1 equipment is that in which the RF energy generated is necessary for its internal functioning; Group 2 equipment is that in which RF energy is intentionally generated and/or used for material inspection, analysis or treatment, and spark erosion

Comment The scope of EN 55011/CISPR 11 is the subject of some confusion; it is sometimes applied more widely than was intended, and there is an amendment pending to clarify the intent

Tests Mains terminal disturbance voltage from 150kHz to 30MHz measured on a test site using 50Ω/50μH CISPR artificial mains network; Group 2 Class A equipment, including equipment with mains supply currents exceeding 100A per phase subject to less stringent limits

Radiated emissions from 30MHz to 1000MHz on a test site (Class A or B) or in situ (Class A only); Group 2 Class A equipment to be measured from 0.15 to 1000MHz but with relaxed limits, below 30MHz measurement performed with loop antenna

Specific limits for magnetic field strength from induction cooking appliances from 0.15 to 30MHz, and for emissions between 1 and 18GHz from Group 2 Class B equipment operating above 400MHz

4.3.2 EN 55014-1: 2000 + A1: 2001 + A2: 2002

Title Electromagnetic compatibility – Requirements for household appliances, electric tools and similar apparatus – Part 1: Emission – Product family standard

Equivalents CISPR 14-1 fourth edition

Scope Appliances whose main functions are performed by motors and switching or regulating devices

Excluding apparatus covered by other CISPR standards (except for multi-function equipment), semiconductor regulating controls of more than 25A per phase, stand-alone power supplies.

Tests Mains terminal disturbance voltage, quasi-peak and average detection from 148.5kHz to 30MHz measured using 50Ω/50μH CISPR artificial mains network; less stringent limits for electric tools and the load terminals of regulating controls. Discontinuous interference (clicks) must also be measured at spot frequencies for appliances which generate such interference through switching operations

Interference power from 30MHz to 300MHz on mains lead, quasi-peak and average detection, measured by means of the absorbing clamp; battery-operated appliances which cannot be mains connected, regulating controls incorporating semiconductor devices, rectifiers, battery chargers and convertors excluded

Radiated tests 30MHz to 1000MHz, as per EN 55022, for toys only

4.3.3 EN 55022: 1998 + A1: 2000 + A2: 2003

Title Information technology equipment – Radio disturbance characteristics – Limits and methods of measurement

Equivalents CISPR 22 third edition

Scope Equipment whose primary function is either (or a combination of) data entry, storage, display, retrieval, transmission, processing, switching or control, and which may be equipped with one or more terminal ports typically operated for information transfer, and with a rated supply voltage not exceeding 600V

Class A equipment is for use in other than Class B environments; Class B equipment is suitable for use in domestic establishments

This standard is particularly widely referenced by other product standards, outside its own scope

Tests Mains terminal interference voltage, quasi-peak and average detection from 150kHz to 30MHz measured using 50Ω/50μH CISPR artificial mains network

Radiated interference field strength using quasi-peak detection from 30MHz to 1000MHz measured at 10m (preferred) or 3m (alternative) on an open area or alternative test site. Tests above 1GHz are covered in an amendment to the fifth edition which is pending at the time of writing; see also Table 4.4 on the US FCC rules for testing above 1GHz, and section 6.4.3. Amendment 1 added the requirement for ferrite absorbers on leads leaving the test site, which provoked a storm of protest and has been omitted from later drafts

Conducted current or voltage (limits related by a common mode impedance of 150Ω) from 150kHz to 30MHz at telecommunication ports, defined as those "which are intended to be connected to telecommunications networks (e.g. public switched telecommunications networks, integrated services digital networks), local area networks (e.g. Ethernet, Token Ring) and similar networks". Various measurement methods are defined for different types of cable connections

Comment EN 55022 and CISPR 22 have suffered considerable disarray over the last few years. At the time of writing they have to be regarded as two separate standards; the fifth edition of CISPR 22 has been published in IEC, but the similar but non-equivalent fifth edition of EN 55022 has not yet been accepted in CENELEC. Hence the above review refers to the third edition (the fourth edition of CISPR 22 didn't make it to an EN) which although published in 1998 is still current in the OJ listing. Various amendments to the fifth edition are in train which are intended to allow the CISPR and CENELEC documents to re-synchronize, but these are not yet finalized. The date of withdrawal of EN 55022: 1994 (the second edition) has been put back several times because of perceived difficulties with the third edition, at present (August 2006) it is August 2007; so that manufacturers have the option of complying with 1994 plus amendments or 1998 plus amendments (including ferrites) or the fifth edition when it appears. In 2007 the option of compliance with the 1994 version should cease. But experience suggests that you should treat all predictions about the future course of development of this standard with some scepticism. An added difficulty is that experience also suggests that, whatever the intentions of the CENELEC committee responsible for the standard, they can easily be disrupted by the demonstrated inability of the European Commission to publish updated lists in the OJEU to a predictable timetable.

4.4 Generic standards – immunity

4.4.1 EN 61000-6-1: 2001

Title Generic immunity standard, Part 1: residential, commercial and light industry environment

Equivalents IEC 61000-6-1

Scope All apparatus intended for use in the residential, commercial and light industrial environment for which no dedicated product or product-family immunity standards exist

NB such apparatus is intended to be directly connected to the public mains supply or to a dedicated DC source. It also includes battery-operated apparatus. Typical locations are residential properties, retail outlets, laboratories, business premises, areas of public entertainment, outdoor locations, etc.

Tests Electrostatic discharge to enclosure as per EN 61000-4-2, at 8kV (air discharge) or 4kV (contact discharge)

Radiated RF field from 80MHz to 1000MHz as per EN 61000-4-3, at 3V/m

Electrical fast transients 5/50ns common mode as per EN 61000-4-4, applied to all functional earth and power ports and some I/O ports, amplitude 0.5 or 1kV dependent on type of port and method of coupling

Surge as per EN 61000-4-5, applied to AC power input ports at 2kV line to earth and 1kV line to line, and to some DC power input ports at 0.5kV

Radio frequency in common mode applied to all power ports and the earth port and some I/O ports, amplitude 3V rms from 150kHz to 80MHz as per EN 61000-4-6

Power frequency magnetic field, as per EN 61000-4-8, 50 or 60Hz at 3A/m, only for apparatus containing magnetically susceptible devices

Voltage dips and interrupts on the AC power input ports, as per EN 61000-4-11

NB the applicability of many of the above tests depends on the allowable length of line that may be connected to the port in question

4.4.2 EN 61000-6-2: 2005

Title Generic immunity standard, Part 2: industrial environment

Equivalents IEC 61000-6-2

Scope Apparatus intended for use in the industrial environment, for which no dedicated product or product-family immunity standard exists, but excluding radio transmitters

NB equipment installed in the industrial environment is not connected to the public mains network but is considered to be connected to an industrial power distribution network with a dedicated distribution transformer. Battery powered equipment intended for this environment is also covered

Tests Electrostatic discharge to enclosure as per EN 61000-4-2, at 8kV (air discharge) or 4kV (contact discharge)

Radiated RF field from 80MHz to 1000MHz as per EN 61000-4-3, at 10V/m except in the broadcast bands, 87–108MHz, 174–230MHz and 470–790MHz, where the level is 3V/m; also from 1.4 to 2.0GHz at 3V/m and 2.0 to 2.7GHz at 1V/m, all 80% AM 1kHz. Testing of small EUTs to IEC 61000-4-20, in a GTEM or other TEM cell, is also allowed as an option

Power frequency magnetic field, as per EN 61000-4-8, 50 or 60Hz at 30A/m, only for apparatus with magnetically susceptible devices

Electrical fast transients 5/50ns common mode as per EN 61000-4-4, applied to some signal and DC power and all AC power ports, amplitude 1 or 2kV dependent on type of port and method of coupling

Radio frequency in common mode applied to some signal and all power ports, amplitude 10V rms from 150kHz to 80MHz with 80% AM 1kHz, except in the broadcast band 47–68MHz where the level is 3V rms, as per EN 61000-4-6

Surges as per EN 61000-4-5, to signal ports with long cables and some DC power ports at 500V, and AC power ports at 1kV or 2kV

Voltage dips and interrupts on the AC power input ports, as per EN 61000-4-11

NB the applicability of many of the above tests depends on the allowable length of line that may be connected to the port in question

4.5 Basic standards – EN 61000-3-X and -4-X

This section only considers those parts of IEC/EN 61000 which are directly relevant for testing equipment (Table 4.3 gives the full picture). Part 2 (The EM environment) is useful for understanding the many environmental aspects of EMC but does not specify tests. Part 5 (Installation and mitigation guidelines) is primarily aimed at systems installers. Note that the European equivalent number of any IEC standard is obtained by writing EN 6XXXX instead of IEC 6XXXX. The standards are (mostly) technically equivalent – there may be so-called European "common modifications" – but the European versions have an additional foreword which specifies how the standard is to be applied for certification purposes.

Currently, EN 61000-3-2, -3, -11 and -12 are harmonised under the EMCD and therefore can and should be applied directly, according to their scope. Those in the EN 61000-4 series are not harmonised and only describe general test methods, but are applied widely through reference in the generic or product standards.

4.5.1 EN 61000-3-X

Title Electromagnetic compatibility – Part 3: Limits†

Equivalents IEC 61000-3-X

Section 2: 2006 Limits for harmonic current emissions

Scope Electrical and electronic equipment having an input current up to and including 16A per phase, and intended to be connected to public low-voltage distribution systems (nominal voltage 220V or higher)

Tests Measurement of 50Hz harmonic currents up to 2kHz using a wave analyser and current shunt or transformer (see section 8.1.1)

Limits Class A (balanced 3-phase equipment and everything outside Classes B, C or D): absolute limits on even and odd harmonics up to the 40th harmonic
Class B (portable tools and non-professional arc welding equipment): as Class A but 1.5 times higher
Class C (lighting equipment, excluding dimmers which are Class A): relative limits expressed as a percentage of the input current for odd harmonics and the second harmonic only, up to the 39th harmonic; discharge lighting equipment with an active input power ≤25W must either meet Class D limits or specific limits on 3rd and 5th harmonics as a percentage of the fundamental current, with a waveform restriction
Class D (personal computers and their monitors, and TV receivers, with a specified power less than 600W): limits expressed in mA per watt for odd harmonics only, up to the 39th harmonic
Transitory harmonics are allowed a relaxation of 1.5 times under certain restricted conditions. No limits apply to:

- equipment with a rated power of 75W or less, other than lighting equipment;
- professional equipment with a total rated power > 1kW;
- symmetrically controlled heating elements with a rated power less than or equal to 200W;

† Although EN 61000-3 has the all-inclusive title of "Limits", it does not refer to radio frequency emission limits, which are the province of CISPR, but only to LF emissions.

- independent dimmers for incandescent lamps with a rated power less than or equal to 1kW;
- incandescent lamp luminaires with no electronic transformer or dimming device, which are deemed to fulfil the requirements without testing.

Section 3: 1995 + A1: 2001 + A2: 2005 Limitation of voltage changes, voltage fluctuations and flicker in public LV supply systems

Scope Electrical and electronic equipment having an input current up to and including 16A per phase, and intended to be connected to public low-voltage distribution systems (nominal voltage 220V or higher)

Tests Measurement of voltage fluctuations using a flickermeter as per IEC 61000-4-15 or by analytical methods, with the EUT supplied from a defined reference impedance (see section 8.1.4)

Limits Limits apply to magnitude of maximum permissible percentage voltage changes (d) with respect to number of voltage changes per second or per minute (P_{st})

A1: 2001 revises some of the voltage change limits and makes it clearer that they apply to the voltage fluctuation at the moment of switch-on, i.e. the standard places a limit on allowable inrush current

Section 11: 2000 Limitation of voltage changes, voltage fluctuations and flicker in public low-voltage supply systems – Equipment with rated current <= 75 A and subject to conditional connection

This is the equivalent standard to IEC 61000-3-3 for higher powered equipment than 16A per phase. It applies the limits of IEC 61000-3-3 but with greater freedom to set the test source impedance, with the actual requirement for conditional connection subject to the result. It is based on IEC 61000-3-4, which is still relevant for equipment with a rated input current >75A

Section 12: 2005 Limits for harmonic currents produced by equipment connected to public low-voltage systems with input current > 16A and ≤75A per phase

This is the equivalent standard to IEC 61000-3-2 for higher-powered equipment than 16A per phase. It has three classes of harmonic current limits, corresponding to three "stages" of connection: Stage 1 allows connection with no need to contact the supply utility company; Stage 2 is based on availability of supply network and equipment data and requires the user to consult with the supply utility; Stage 3 does not apply limits but requires the user to seek supply utility acceptance for connection. It becomes mandatory under the EMCD in February 2008.

4.5.2 EN 61000-4-X

This section merely covers those parts of EN 61000-4 which are in widespread use for testing; many more parts are published (see Table 4.3 on page 68).

Scope Testing and measurement techniques for immunity of electrical and electronic equipment: basic EMC standards

Criteria Test results to be classified as follows:

- normal performance within specification limits;
- temporary degradation or loss of function or performance which is self recoverable;
- temporary degradation or loss of function or performance which requires operator intervention or system reset;
- degradation or loss of function which is not recoverable due to hardware or software damage or loss of data.

Section 1: Overview of immunity tests

Not a test standard itself, its intention is to give "a general and comprehensive reference to the technical committees of IEC or other bodies, users and manufacturers of electrical and electronic equipment on EMC immunity specifications and tests, and to give general guidance on selection and application of these tests." A third edition is expected to be published in 2006.

Section 2: 1995 + A1: 1998 + A2: 2001 Electrostatic discharge

Equivalent IEC 61000-4-2

Tests At least ten single discharges to preselected points, accessible to personnel during normal usage, in the most sensitive polarity. Contact discharge method to be used unless this is impossible, in which case air discharge used. Also ten single discharges to be applied to horizontal and vertical coupling planes

Levels Severity levels from 2kV to 15kV (8kV contact discharge) depending on installation and environmental conditions

Section 3: 2006 Radiated radio frequency field

Equivalent IEC 61000-4-3

Tests Radiated RF field generated by antennas in a shielded anechoic enclosure using the substitution method (pre-calibrated field), swept from 80MHz to 1000MHz with a step size not more than 1% of preceding frequency and dwell time sufficient to allow the EUT to respond, minimum 0.5 seconds. Eight (twelve) tests are needed, one in each polarization with the antenna facing each of the four sides of the EUT (and top and bottom if these might be affected). Field uniformity within –0/+6dB over 12 out of 16 points within a 1.5 x 1.5m square area at the front face of the EUT is required of the chamber

Testing from 800 to 960MHz and 1.4 to 6GHz (though not necessarily the whole of this range) is included for protection against digital mobile phones.

Levels Severity levels of 1, 3 or 10V/m unmodulated (or greater) depending on the expected EMR environment; the actual applied signal is modulated to 80% with a 1kHz sinewave

Section 4: 2004 Electrical fast transient burst

Equivalent IEC 61000-4-4

Tests Bursts of 5ns/50ns pulses at a repetition rate of 5kHz or 100kHz with a duration of 15ms and period of 300ms, applied in both polarities between power supply terminals (including the protective earth) and a reference ground plane, or via a capacitive coupling clamp onto I/O circuits and communication lines

Levels Severity levels of 0.5, 1, 2 and 4kV on power supply lines, and half these values on signal, data and control lines, depending on the expected environmental and installation conditions

Section 5: 1995 + A1: 2001 Surge

Equivalent IEC 61000-4-5 (NB there was a second edition of this standard in 2005 but it had not been accepted in CENELEC by mid-2006)

Tests At least 5 positive and 5 negative surges, at a repetition rate no faster than 1 per minute, of 1.2/50μs voltage or 8/20μs current waveshape surges from a surge generator of 2Ω output impedance, line-to-line on AC/DC power lines; 12Ω output impedance, line-to-earth on AC/DC power lines; 42Ω output impedance, capacitively coupled or via gas-filled arrestors line-to-line and line-to-earth on I/O lines

Levels Severity levels of 0.5, 1, 2 and 4kV, selected according to installation conditions and type of line; all lower test level voltages must also be applied

Section 6: 1996 + A1: 2001 Conducted disturbances induced by radio frequency fields

Equivalent IEC 61000-4-6 (NB there was a second edition of this standard in 2005 but it had not been accepted in CENELEC by mid-2006)

Tests RF voltage swept at slower than $1.5 \cdot 10^{-3}$ decades/s, or with a step size not more than 1% of fundamental and dwell time sufficient to allow the EUT to respond, over the frequency range 150kHz to 80MHz (possibly 230MHz), applied via coupling/decoupling networks (CDNs) to cable ports of the EUT. When CDNs are not suitable or are unavailable, the alternative methods of EM-clamp or current injection probe can be used (except on supply lines)

NB: applicability of tests over the frequency range 80MHz to 230MHz overlaps with IEC 61000-4-3, and may be used instead of the tests specified in that document, depending on the EUT dimensions

Levels Severity levels of 1, 3 or 10V emf unmodulated depending on the EMR environment on final installation; the actual applied signal is modulated to 80% with a 1kHz sinewave

Section 8: 1993 + A1: 2001 Power frequency magnetic field

Equivalent IEC 61000-4-8

Tests Continuous and short duration power frequency magnetic field, applied via an induction coil adequately sized to surround the EUT in three orthogonal positions

Levels Continuous: 1, 3, 10, 30 or 100 A/m; short duration (1 to 3s): 300 or 1000A/m, for the higher severity levels only

Section 9: 1993 + A1: 2001 Pulse magnetic field

Equivalent IEC 61000-4-9

Tests Mainly applicable to electronic equipment to be installed in electrical plants. At least 5 positive and 5 negative 6.4/16µs pulses applied via an induction coil adequately sized to surround the EUT in three orthogonal positions, repetition period no less than 10s

Levels 100, 300 and 1000 A/m

Section 10: 1993 + A1: 2001 Damped oscillatory magnetic field

Equivalent IEC 61000-4-10

Tests Mainly applicable to electronic equipment to be installed in electrical plants. Oscillatory wave of 0.1 or 1MHz damped to 50% of peak after three to six cycles at a repetition rate of 40 or 400 per second, applied via an induction coil adequately sized to surround the EUT in three orthogonal positions for 1 second

Levels 10, 30 and 100 A/m

Section 11: 2004 Voltage dips, short interruptions and voltage variations

Equivalent IEC 61000-4-11

Scope Electrical and electronic equipment fed by low-voltage power supply networks and having an input current not exceeding 16A per phase, but not equipment which is connected to DC networks or 400Hz AC networks

Tests Dips and short interruptions initiated at any phase angle of the input voltage, to a level of 0%, 40%, 70% and 80% of the nominal voltage for a duration of 0.5 to 250 50Hz cycle periods

Short-term variations (optional test) to a level of 70% nominal voltage

Section 12: 1995 + A1: 2001 Oscillatory waves

Equivalent IEC 61000-4-12

Tests Ring wave: 100kHz decaying at 60% per peak, initial voltage rise time 0.5µs, applied at a rate of 1 to 6 transients per minute from a generator with output impedance of 12, 30 or 200Ω via a coupling-decoupling network in common or differential mode to power supply, signal and control ports
Damped oscillatory wave: same characteristics as damped field of IEC 61000-4-10, applied for not less than 2 seconds from a generator with output impedance 200Ω via CDNs as for ring wave

Levels	Ring wave: 0.5, 1, 2 and 4kV common mode, half these values for differential mode Damped oscillatory wave: 0.5, 1 and 2kV common mode, half these values for differential mode

4.6 Product standards

A variety of standards (with the exceptions of ENs 55011, 55014-1 and 55022, discussed separately because they operate by default more like basic standards, see sections 4.3.1 to 4.3.3) are listed here, separated into "principal" and "other". This is of course an entirely arbitrary distinction; a taxi driver would quite naturally regard EN 50148 as the most important EMC standard in the world. The basis for listing "principal" standards in more detail than "others" is simply that they would appear to cover a fairly large range of products, by value or quantity. All non-radio standards that appear in the current (at the time of writing) OJEU listing [185] are included, either in the main part of this section or in short form at the end of the section.

Choice of product standard

When you choose a product standard for use in self certification, you can only do so initially on the basis of its title. Sometimes this is enough, but sometimes it is not at all obvious which of several is the most likely. Without consulting an expert, the only way to proceed is to obtain all of the likely ones, read at least through their scopes (more often than not you will have to persevere to the annexes at the back), and then make your selection from a more informed standpoint. ETSI standards are freely available for download on the web, but because of copyright issues IEC-based standards are not, and you have to pay for them. With sustained pressure from concerned standards-users, this situation might change in the future. So far, the only concession has been for the IEC to make the scopes of its standards available for preview on its website.

Many of the product standards had dates of cessation of presumption of conformity of the superseded standard (usually the generic standard) in 2000 or 2001. Experience suggests that while a few forward-thinking manufacturers were testing their products to the newer product standards in advance, most were not. There seems to be a commonly-held misconception that once you have settled on a particular set of standards, you can use them in perpetuity. It is all too usual to find declarations of conformity of new products boasting compliance to, say, EN 50082-1:1992, when this standard was superseded by a second edition in 1997, by EN 61000-6-1 in 2004, and maybe by a new product standard after that, and its tests are limp-wristed by comparison with the later version and the product standards. There will be three possibilities to correct this situation:

- companies may be actively keen to make the best use of the new standards;
- their more aware customers will demand compliance with the new standards;
- rigorous market surveillance, by enforcement officers familiar with the significance of the dating of standards, may happen.

4.6.1 Broadcast receivers and associated equipment

Emissions: EN 55013: 2001 + A1: 2003

Title	Limits and methods of measurement of radio disturbance characteristics of broadcast receivers and associated equipment
Equivalents	CISPR 13
Scope	Broadcast sound and television receivers, and associated equipment intended to be connected directly to these or to generate or reproduce audio or visual information, for example audio equipment, video cassette recorders, compact disc players, electronic organs. Information technology equipment as defined in EN55022 is excluded
Tests	Mains terminal interference voltage from 150kHz to 30MHz measured using 50Ω/50μH CISPR artificial mains network
	Antenna terminal disturbance voltage over the range 30–1000MHz due to local oscillator and other sources, higher limits for car radios
	Radiated disturbance field strength of local oscillator and harmonics in the range 80 to 1000MHz measured on an open area test site at a distance of 3m; A1 adds methods for digital receivers
	Disturbance power of associated equipment excluding video recorders on all leads of length 25cm or more, over the range 30 to 300MHz, measured by means of the absorbing clamp
Limits	Limits for mains terminal disturbance voltage and disturbance power are the same as those in EN55014-1. Radiated field limits for local oscillator and harmonics are 12–20dB higher than equivalent Class B emissions limits for other products

Immunity: EN 55020: 2002 + A1: 2003 + A2: 2005

Title	Sound and TV broadcast receivers and associated equipment – immunity characteristics – limits and methods of measurement
Related to	CISPR 20 (Not equivalent)
Scope	Broadcast sound and television receivers, including direct-to-home satellite receivers, and associated equipment intended to be connected directly to these or to generate or reproduce audio or visual information, for example audio equipment, video cassette recorders, compact disc players, electronic organs. Information technology equipment as defined in EN 55022 is excluded
	No immunity requirements apply (they are "under consideration") to battery powered sound and TV receivers or those without an external antenna connection
Tests	Immunity from unwanted signals present at the antenna terminal: VHF band II receivers tested with in-band and out-of-band signals up to 85dBμV; TV receivers and VTRs tested with adjacent channel

modulated signals up to 80dBμV

Immunity from conducted voltages at the mains input, audio input and output terminals of receivers (except AM sound and car radios) and multi-function equipment over the range 150kHz to 150MHz; audio input & output terminals have less stringent low frequency levels than mains, loudspeaker and headphone terminals; the tuned channel and IF channel frequencies are excluded

Immunity from conducted currents of receivers (including car radios and AM sound) and multi-function equipment over the range 26 to 30MHz applied to the antenna terminal

Immunity from radiated fields from 150kHz to 150MHz of receivers and multi-function equipment, as tested in an open stripline test set-up, at 125dBμV/m except at IF and in-band frequencies; 900MHz 3V/m modulated by 217Hz keyed carrier, to EN 61000-4-3

Electrostatic discharge to EN 61000-4-2, 4kV contact, 8kV air

Electrical fast transient bursts to EN 61000-4-4, 1kV to the AC mains power input port

Criteria Wanted to unwanted audio signal ratio of ≥40dB, or just perceptible degradation of a standard picture

Amendment A1 covers broadcast receivers for digital signals; A2 covers objective evaluation of picture quality

4.6.2 Household appliances, electric tools and similar apparatus

Emissions: EN 55014-1

See 4.3.2 on page 76

Immunity: EN 55014-2: 1997 + A1: 2001

Title Electromagnetic compatibility – Requirements for household appliances, electric tools and similar apparatus – Part 2: immunity – product family standard

Scope Electromagnetic immunity of appliances and similar apparatus for household and similar purposes as well as electric toys and tools. This standard is the immunity counterpart to EN 55014-1

Tests Apparatus is classified into four categories:

Category I: Apparatus containing no electronic control circuitry

Category II: Mains powered appliances containing electronic control circuitry with no internal frequency higher than 15MHz

Category III: Battery powered apparatus containing electronic control circuitry with no internal frequency higher than 15MHz

Category IV: All other apparatus within the scope

Levels for ESD, electrical fast transients, conducted RF, radiated RF, surges, and voltage dips and interruptions are defined with the test methods as per the basic standards. Permissible performance criteria are also defined. For each category, applicable tests and criteria are then specified. Category I apparatus is deemed to fulfil the requirements without testing

Amendment A1 expands requirements for toys, as well as making other changes

4.6.3 Lighting equipment

Emissions: EN 55015: 2000 + A1, A2

Title Limits and methods of measurement of radio disturbance characteristics of electrical lighting and similar equipment

Equivalents CISPR 15

Scope Conduction and radiation of radio frequency disturbances from all lighting equipment with a primary function of generating and/or distributing light intended for illumination purposes, including the lighting part of multi-function illumination equipment and independent auxiliaries exclusively for use with lighting equipment; but excluding aircraft and airport lighting and apparatus explicitly covered by other IEC/CISPR standards, e.g. built-in lighting devices in other equipment, photocopiers or slide projectors

Tests For luminaires intended for fluorescent lamps, insertion loss is measured between 150kHz and 1605kHz between terminals on a dummy lamp (construction specified in the standard) and the mains terminals of the luminaire

All other types of lighting equipment, including independent auxiliaries and self-ballasted fluorescent lamps, must meet quasi-peak and average limits for mains terminal disturbance voltage in the range 9kHz to 30MHz

In addition, lighting equipment with lamp operating frequencies in excess of 100Hz must meet quasi-peak limits for radiated magnetic field in the range 9kHz to 30MHz, measured with a Van Veen loop

Incandescent lamps are deemed to fulfil requirements without testing

Uniquely among CISPR-based standards, there is a "chimney" of + 17dB in the conducted mains emission limits between 2.51 and 3MHz (except in Japan), and more in the radiated limits, between 2.2MHz and 3MHz. No explanation is provided in the standard for this departure, but a 1995 draft proposing the change makes clear that it is intended to allow the marketing of RF compact fluorescent lamps, on the grounds that "no broadcasting exists in this frequency band", and "a relaxation would allow the introduction of a relatively cheap energy saving lamp of light weight and smaller dimensions".

Immunity: EN 61547: 1995 + A1: 2000

Title Equipment for general lighting purposes – EMC immunity requirements

Equivalent IEC 61547

Scope Lighting equipment within the scope of IEC TC 34, such as lamps, auxiliaries and luminaires; exclusions similar but not identical to those of CISPR 15

Tests Levels for ESD, electrical fast transients, conducted RF, radiated RF, surges, and voltage dips and interruptions are defined with the test methods mostly as per the basic standards. Lighting-related performance criteria are also defined. For self-ballasted lamps, independent auxiliaries and luminaires, applicable tests and criteria are then specified. Non-electronic lighting equipment (except emergency lighting luminaires) is deemed to fulfil the requirements without testing

4.6.4 Information technology equipment

Emissions: EN 55022: 1998

See 4.3.3 on page 77

Immunity: EN 55024: 1998 + A1: 2001 + A2: 2003

Title Information technology equipment – Immunity characteristics – Limits and methods of measurement

Equivalent CISPR 24

Scope Information technology equipment as defined in CISPR 22

Tests Electrostatic discharge, electrical fast transients, radiated RF, conducted RF, power frequency magnetic field, surge, voltage dips and interruptions. There are some differences quoted from the basic test methods, for instance the ESD requirement is for at least 200 discharges to a minimum of four points. More than half of the standard is taken up with annexes giving particular test conditions and performance criteria for different types of apparatus.

4.6.5 Professional AV and entertainment lighting equipment

Emissions: EN 55103-1: 1996

Title Electromagnetic compatibility – Product family standard for audio, video, audio-visual and entertainment lighting control apparatus for professional use – Part 1: Emission

Scope	Professional audio, video, audio-visual and entertainment lighting control apparatus, intended for use in the residential, commercial and light industrial, urban outdoors, controlled EMC and rural, and heavy industrial environments. Consumer and security system apparatus is specifically excluded. The emphasis is on the word "control", so that for instance dimmers and luminaires (even if designed for stage use) are excluded. Annex G of the standard clarifies the scope in more detail.
Tests	RF radiated emissions 30MHz–1GHz, magnetic fields 50Hz–50kHz, harmonics and flicker as given by EN 61000-3-2, -3, -4 or -5, AC power port conducted RF emissions and discontinuous interference 150kHz–30MHz, inrush currents on the AC power port, conducted emissions 30–1000MHz on the antenna terminals of broadcast receivers according to EN 55013, conducted emissions 150kHz–30MHz on signal, control and DC power ports. Applicability and limits vary depending on the environment.

Immunity: EN 55103-2: 1996

Title	Electromagnetic compatibility – Product family standard for audio, video, audio-visual and entertainment lighting control apparatus for professional use – Part 2: Immunity
Scope	As given above for EN 55103-1
Tests	RF radiated field 80–1000MHz, electrostatic discharge, magnetic fields 50Hz–10kHz, fast transients and conducted RF 150kHz–80MHz on all ports, voltage dips, interruptions and surge on AC power input ports, AF common mode 50Hz–10kHz on signal and control ports. Applicability and levels vary depending on the environment.

4.6.6 Equipment for measurement, control and laboratory use

Emissions and immunity: EN 61326-1: 2006

Title	Electrical equipment for measurement, control and laboratory use – EMC requirements
Equivalents	IEC 61326-1
Scope	Electrical equipment operating from a supply of less than 1kV AC or 1.5kV DC, intended for professional, industrial process and educational use, for measurement and test, control or laboratory use. It includes accessories intended for use with the above. Particular requirements are found in sub-parts of Part 2 of the standard for equipment intended for use in specific applications, such as sensitive test and measurement equipment for EMC unprotected applications, e.g. oscilloscopes, logic analysers, etc., transducers with signal conditioning, or in-vitro diagnostic medical equipment

Tests Emissions: mains port conducted RF 150kHz–30MHz, harmonics and flicker according to IEC 61000-3-2, -3 (Class B only), radiated RF 30MHz–1000MHz

Immunity: Electrostatic discharge, radiated RF, voltage interruptions, electrical fast transient bursts, surge, conducted RF, power frequency magnetic field. Applicability and levels depend also on the particular requirements in Part 2.

4.6.7 Fire, intruder and social alarm systems

Emissions

There is no explicit emissions standard for this product family; use of the appropriate generic emissions standard is recommended

Immunity: EN 50130-4: 1995 + A1: 1998 + A2: 2003

Title Alarm systems – Part 4: Electromagnetic compatibility – Product family standard: Immunity requirements for components of fire, intruder and social alarm systems

Scope Components of the following alarm systems, intended for use in and around buildings in residential, commercial, light industrial and industrial environments: intruder alarm systems, hold-up alarm systems, fire detection and fire alarm systems, social alarm systems, CCTV systems and access control systems for security applications, and alarm transmission systems (the last added by Amendment 1).

Tests Mains supply voltage variations, dips and short interruptions, electrostatic discharge, radiated electromagnetic field, conducted RF, fast transient bursts and surge. There are some significant differences from the usual test requirements hidden in these specifications, for instance the RF tests extend to 2GHz and require an extra set of sweeps with 1Hz pulse modulation. For each test, fairly explicit performance criteria are provided.

4.6.8 Telecommunication network equipment

Emissions and immunity: EN 300386

Title Electromagnetic compatibility and radio spectrum matters (ERM); Telecommunication network equipment; Electromagnetic compatibility (EMC) requirements

Scope Equipment intended to be used within a telecommunications network including switching equipment, non-radio transmission equipment and ancillaries, power supply equipment and supervisory equipment. Excludes cable TV equipment and optical amplifiers. The definition of a telecommunications network excludes terminal equipment beyond the network termination points.

Tests Emissions: mains port and DC power port conducted RF 150kHz–30MHz, telecom port conducted RF 150kHz–30MHz, harmonics and flicker according to EN 61000-3-2, -3, radiated RF 30MHz–1000MHz

Immunity: electrostatic discharge, radiated RF, conducted RF, voltage interruptions, electrical fast transient bursts, surge. Extra surge and power induction requirements apply to ports for outdoor signal lines. Two environments are defined with different emissions and immunity levels, one being "telecommunication centre", the other being "other than telecommunication centre". The standard defines specific performance criteria and operating conditions.

Comment This standard illustrates perfectly the volatile nature of product standards. Version 1.3.1 is listed in the OJEU with a date of cessation of presumption of conformity of the superseded standard (Version 1.2.1) of 31st August 2005. In the same listing, version 1.3.2. is listed with exactly the same date of cessation; and version 1.3.3 is listed with a date of cessation of 31st January 2006.

4.6.9 Radio equipment

Emissions and immunity: EN 301489-1

Title Electromagnetic compatibility and Radio spectrum Matters (ERM); ElectroMagnetic Compatibility (EMC) standard for radio equipment and services; Part 1: Common technical requirements

Scope Radio communications equipment and any associated ancillary equipment operating in the residential, commercial, light industrial, telecommunications centre and vehicular environments, but not the maritime environment. With the inception of the R&TTE Directive (Chapter 3), the radio standards listed under the EMC Directive migrated to parts of EN 301489 where they were rewritten to refer to that Directive. Part 1 gives common technical requirements, while other parts give product-related information.

Tests RF conducted and radiated emissions as per EN 55022, including modified application to DC power input ports

ESD, electrical fast transients, surges, supply voltage variation, dips and interruptions, vehicular supply transients, conducted and radiated RF: as per the basic standards. Three classes of equipment are established, for fixed, vehicular and portable use, and the various tests are applied as appropriate. The standard also establishes minimum performance criteria, and makes specific provision for input and output transmitter/receiver test signals, narrowband responses and exclusion bands for RF immunity testing. The antenna port is specifically excluded from EMC tests. Frequency spectrum utilization parameters such as spurious emissions and spurious responses are also excluded

4.6.10 Marine navigation equipment

Emissions and immunity: EN 60945: 2002

Title Maritime navigation and radiocommunication equipment and systems – General requirements – Methods of testing and required test results

Related to IEC 60945

Scope Shipborne radio and navigational equipment forming part of the global maritime distress and safety system required by the International Convention for Safety of Life at Sea (SOLAS); for EMC only, all other bridge-mounted equipment, equipment in close proximity to receiving antennas, and equipment capable of interfering with safe navigation of the ship and with radiocommunications (i.e., most marine electronics)

Tests This standard specifies general performance requirements, of which EMC is only a part. It is harmonised under the EMCD for its EMC provisions. These are:

• RF emissions, conducted on the AC or DC supply port from 10kHz to 30MHz using a CISPR 16-1 LISN; radiated, from 150kHz to 30MHz using a magnetic field loop; radiated, 30MHz to 2GHz at 3m on a CISPR 16-1 compliant test site; the limits do not correspond to any other CISPR limits, and from 156 to 165MHz (the VHF marine band) the limits are much tighter than other CISPR limits. The standard also defines other aspects of the measurement methods

• immunity: conducted RF, 3V 150kHz–80MHz plus certain spot frequencies at 10V, to EN 61000-4-6; radiated RF, 10V/m 80MHz–2GHz to EN 61000-4-3; fast transient bursts to EN 61000-4-4, 1kV or 2kV; surges to EN 61000-4-5, 1kV/0.5kV on AC power lines only; power supply short-term variation and failure, to EN 61000-4-11; ESD at 6kV contact, 8kV air, to EN 61000-4-2

4.6.11 Medical electrical equipment

Emissions and immunity: EN 60 601-1-2 : 2001

Title Medical electrical equipment – part 1: General requirements for safety – 2. Collateral Standard: Electromagnetic compatibility – requirements and tests

Scope Medical electrical equipment and systems, information technology equipment used in medical electrical application

This standard defines the general EMC requirements and tests for all such equipment; requirements for particular classes of equipment are contained in the particular requirements of Part 2 of this standard, which is fundamentally a safety standard. NB it is not a harmonised standard for the EMC Directive, since EMC of medical electrical equipment is covered by the Medical Devices Directive and not the

EMC Directive; this standard is only harmonised for the MDD. Being a safety standard, it is considerably more comprehensive than, and has a different format to most EMC standards

Tests RF emissions as per CISPR 11 (EN 55011) with some modifications, using the classifications of Group 1 or 2 and Class A or B

Mains harmonics and flicker to EN 61000-3-2 and -3

Immunity tests:
ESD: 6kV contact, 8kV air discharge to EN 61000-4-2
Radiated RF: 3V/m from 80MHz to 2.5GHz, except for life-support equipment and systems, tested at 10V/m; to EN 61000-4-3
Electrical fast transients: 1kV for signal and interconnecting cables, 2kV for AC and DC power lines, as per EN 61000-4-4
Surge: 1kV line to line, 2kV line to ground, at the mains port, as per EN 61000-4-5
Conducted RF: 3V rms from 150kHz to 80MHz as per EN 61000-4-6 in general; some equipment may enjoy a higher start frequency; life-support equipment should also be tested to 10V rms in the ISM frequency bands
Voltage dips and interruptions: as per EN 61000-4-11, to the AC power input
Power frequency magnetic field: 3A/m to EN 61000-4-8.

NB all the above immunity tests, whilst referring to the EN 61000-4 basic standards, are subject to extensive modifications and clarifications which must be carefully considered

4.6.12 Future multi-media

Before leaving this section, it is worth looking ahead to what is arguably the greatest change in CISPR-based standards for a decade or more. CISPR 32 and 35, for emissions and immunity respectively of multi-media equipment, are being created by CISPR/I as an amalgamation of, and replacements for, CISPRs 13, 20, 22 and 24. The development of consumer products which are a combination of information technology and entertainment technology – the “convergence” of functions previously seen as separate – has driven this move, but the committee’s working groups have seized on it as an opportunity to inject a number of needed changes into the test regime that might otherwise have foundered through inertia.

The pair of standards are at the first draft stage for public comment at the time of writing of this book, but are likely to be published during its lifetime, hence their inclusion here. It is too early to be specific about their contents, but a number of features have emerged which are worth mentioning:

- They are based on the current best practice for EMC measurements rather than on the older standards. Hence anyone familiar with CISPR 13 and 20 in particular may not find all of the anachronistic tests described in those two documents in the new drafts.
- The radiated emissions tests allow alternative measurement methods – a classical OATS, a screened room meeting the NSA requirements, a fully anechoic room, or a reverberation chamber as per IEC 61000-4-21 – with

different limits as appropriate. In other words, the principle is of equivalent protection of the radio spectrum rather than strict equivalence of test methods and results. It is explicitly stated in both drafts that if different methods give different results (as they might), any method which results in compliance is acceptable.

- The drafts are structured more like the generic standards, that is a relatively short main body with tables on a port-by-port basis giving the requirements, and normative annexes describing the test methods in detail; this makes the documents easier to use.
- For the immunity tests, particular performance criteria are to be determined, and are given in normative annexes on a function-oriented rather than equipment-oriented basis.
- Tolerances are given for relevant measurement parameters: a small point perhaps, but important for establishing measurement uncertainty.

4.6.13 Other product standards

The following list details other product EMC standards not covered above which have been harmonised in the OJEU at the time of writing.

Standard	**Product sector**	**Comment**
EN 50065-1, -2	Mains signalling equipment	
EN 50083-2	Cable sound and TV distribution network equipment	
EN 50090-2-2	Home and building electronic systems	
EN 50091-2	Uninterruptible power systems	
EN 50148	Electronic taximeters	
EN 50240	Resistance welding equipment	
EN 50263	Measuring relays and protection equipment	
EN 50270	Gas detection and measurement equipment	
EN 50293	Road traffic signal systems	
EN 50295	LV switchgear and control gear	
EN 50370-1, -2	Machine tools	
EN 55012	Vehicles, boats and internal combustion engine devices	
EN 60034-1	Rotating electrical machines	
EN 60204-31	Sewing machines, units and systems	
EN 60439-1	Low-voltage switchgear and control gear assemblies	
EN 60669-2-X	Switches for household etc. fixed electrical installations	Various parts
EN 60730-X	Automatic electrical controls for household etc. use	Various parts
EN 60870-2-1	Telecontrol equipment and systems	
EN 60947-X	Low-voltage switchgear and control gear	Various parts

Standard	Product sector	Comment
EN 60974-10	Arc welding equipment	
EN 61008-1	Residual current operated circuit breakers (RCCBs)	
EN 61009-1	Residual current operated circuit breakers (RCBOs)	
EN 61037	Electricity metering – electronic ripple control receivers	Replaced by EN 62052, EN 62054
EN 61038	Electricity metering – time switches	
EN 61131-2	Programmable controllers	
EN 61204-3	Low voltage DC power supplies	
EN 61543	Residual current-operated protective devices (RCDs)	
EN 61800-3	Adjustable speed electrical power drive systems	
EN 61812-1	Specified time relays for industrial use	
EN 62052, 53, 54	Electricity metering equipment	
EN 617 – 620	Continuous handling equipment and systems	CEN
EN 12015, 12016	Lifts, escalators and passenger conveyors	
EN 12895	Industrial trucks	
EN 13241-1	Garage doors and gates	
EN 13309	Construction machinery	
EN 14010	Equipment for power driven parking of motor vehicles	
EN ISO 14982	Agricultural and forestry machines	

4.7 Other standards not related to the EMC Directive

4.7.1 FCC Rules

In the USA, radio frequency interference requirements are controlled by the FCC (Federal Communications Commission), which is an independent government agency responsible for regulating inter-state and international communications by radio, television, satellite and cable. The requirements are detailed in CFR (Code of Federal Regulations) 47. Part 15 of these regulations applies to unintentional and intentional radiators.

Part 15 subpart B, applying to unintentional radiators, includes clauses which cover specific classes of device such as power line carrier systems, TV receivers and TV interface devices. Industrial, scientific and medical devices which intentionally generate RF energy are covered under Part 18 of the rules. But the major impact of Part 15 is on those products which incorporate digital devices.

4.7.1.1 Approval routes

A “digital device” (previously defined as a computing device) is any electronic device or system that generates and uses timing signals or pulses exceeding 9kHz and uses digital techniques. Two classes are defined, depending on the intended market: Class A for business, commercial or industrial use, and Class B for residential use. These classes are subject to different limits, Class B being the stricter. Before being able to market his equipment in the USA, a manufacturer must follow one of three routes:

- *verification*, which is totally a self certification process;
- *declaration of conformity (DoC)*, similar to verification except that testing must be carried out in a US-accredited test laboratory;
- *certification*, where the manufacturer must send a package of information including test data, installation and operating instructions, and fees to a Telecoms Certification Body (TCB), which issues the approval.

Which route is to be followed depends on the type of product. (Note that the certification route applies to more than just telecom products.) Since June 2000, the FCC itself has declined to become involved directly in any of these routes. Under the EU/US Mutual Recognition Agreement, European laboratories are capable of acting as TCBs for certification or as Certification Bodies for the DoC route.

There are some quite broad exemptions from the rules depending on application. These include digital devices used in transport vehicles, industrial plant or public utility control systems, industrial, commercial and medical test equipment, specialized medical computing equipment, and a digital device used in an appliance.

4.7.1.2 Test requirements

Limits apply to conducted interference on the mains lead between 150kHz and 30MHz, and radiated interference measured either at 10m or 3m from 30MHz to 960MHz and above. The limits were similar but not identical to those laid down in CISPR-derived standards, and the conducted limits have now been aligned with CISPR 22; but the test procedures of ANSI C63.4 [206] must be followed and the US mains voltage must be used during the tests†. The upper frequency limit is extended to a possible maximum of 40GHz, depending on the frequencies used within the device. The relationship between internal clock (or other) frequencies and the maximum measurement frequency is shown in Table 4.4. From this you can see that devices with clock frequencies exceeding 108MHz must be tested for emissions well into the microwave region.

Highest frequency generated or used in the device or on which the device operates or tunes (MHz)	Upper frequency of measurement range (MHz)
Below 1.705	30
1.705–108	1000
108–500	2000
500–1000	5000
Above 1000	5th harmonic of highest frequency or 40GHz, whichever is lower

Table 4.4 Maximum measurement frequency for digital devices, FCC Rules Part 15

† In early 2006 the FCC fined a German company $1m for non-compliance with the Rules, despite the company's claim that its products were "CE-compliant". The FCC says that "such testing neither is the equivalent of nor demonstrates compliance with the Commission's technical standards".

4.7.2 Measurement standards

Some very important EMC standards do not appear in the sections above because they do not refer to products and do not directly give measurement methods or limits. Instead they define measuring instrumentation, facilities or methods:

CISPR 16-1-X	Specification for radio disturbance and immunity measuring apparatus and methods – Part 1: Radio disturbance and immunity measuring apparatus (see section 6.1)
CISPR 16-2-X	Specification for radio disturbance and immunity measuring apparatus and methods – Part 2: Methods of measurement of disturbances and immunity
CISPR 16-3-X	Specification for radio disturbance and immunity measuring apparatus and methods – Part 3: Reports and recommendations of CISPR (contains recommendations on statistics of disturbance complaints, on the significance and determination of CISPR limits, etc.)
CISPR 16-4-X	Specification for radio disturbance and immunity measuring apparatus and methods – Part 4: Uncertainty in EMC measurements
IEC 61000-4-7	Electromagnetic compatibility (EMC) – Part 4-7: Testing and measurement techniques – General guide on harmonics and interharmonics measurements and instrumentation, for power supply systems and equipment connected thereto (see section 8.1.1)
IEC 61000-4-15	Electromagnetic compatibility (EMC) – Part 4: Testing and measurement techniques – Section 15: Flickermeter – Functional and design specifications (see section 8.1.4)
EN 50147-1	Anechoic chambers, Part 1: Shield attenuation measurement
EN 50147-2	Anechoic chambers, Part 2: Alternative test site suitability with respect to site attenuation

4.8 RF emissions limits

Most of the standards within the EN 550XX series have harmonised limit levels for conducted and radiated emissions. Since they derive from CISPR, the limit levels are set in each case for the same purpose, to safeguard the radio spectrum for other users. A minimum separation distance is assumed between source and susceptible equipment.

Figure 4.4 and Figure 4.5 show the limits in graphical form for the emissions standards discussed above. FCC levels are now the same as the CISPR levels. All radiated emission levels are normalized to a measuring distance of 10m.

In these figures, EN Class A refers to EN 55011, EN 55022 Class A and EN 61000-6-4, and EN Class B refers to EN 55011, EN 55022 Class B, EN 55013, EN 55014-1, and EN 61000-6-3. All values are measured with the CISPR 16-1-1 quasi-peak detector, but the standards also require conducted emissions to be measured with an average detector. The limits for the average measurement are 13dB (Class A) and 10dB (Class B) below the quasi-peak limits.

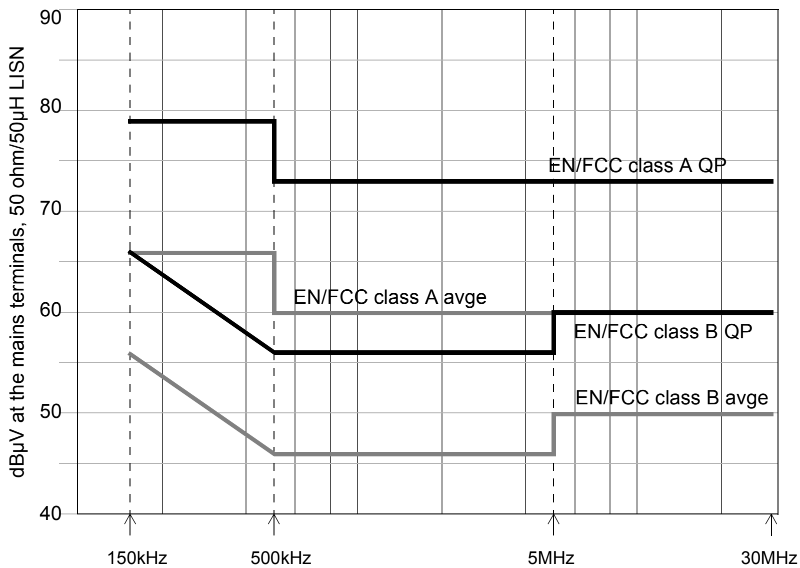


Figure 4.4 Conducted emission limits
(QP = quasi-peak, avge = average)

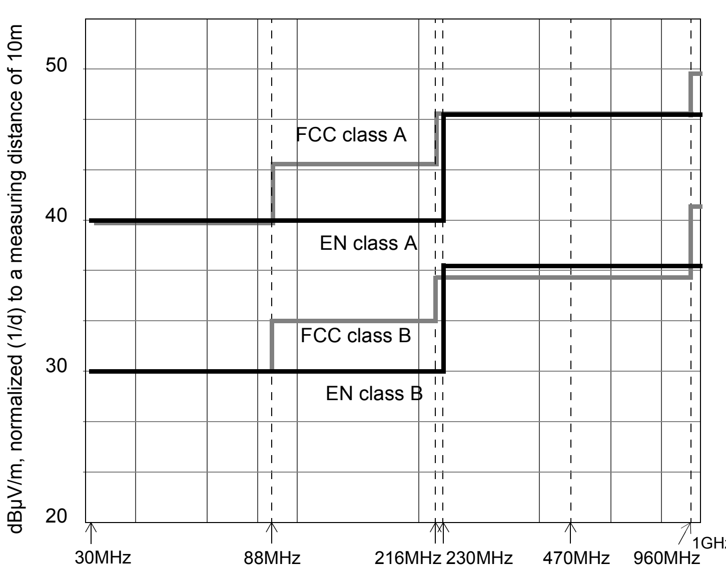


Figure 4.5 Radiated emission limits

Chapter 5

Other standards and legislation

Chapter 4 covered the commercial standards, deriving from the IEC, that are relevant for the EMC and R&TTE Directives. Although comprehensive, these are by no means the only test standards or legislation that can be found for EMC. This chapter looks at a range of other sectors that have their own EMC framework: automotive, military, civil aerospace and rail.

5.1 Automotive

The automotive sector has for a long time had its own EMC requirements.

5.1.1 The Automotive EMC Directive

In Europe, the EMC both of whole vehicles and their electronic sub-assemblies is covered by the Automotive EMC Directive. This is a type-approval Directive, not a New Approach one, and instead of the CE Mark requires that products which comply with its requirements are 'e' marked.

5.1.1.1 The first edition

The original Automotive EMC Directive 95/54/EC required type approval for EMC of all vehicles and electronic vehicle sub-assemblies. It was an amendment to the early Directive 72/245/EEC which controlled ignition interference emissions. Unlike the EMC Directive, it included within its annexes all the applicable technical requirements and test methods, many of which are quite different to the commercial standards discussed in Chapter 4 of this book. Automotive electronic products within its scope should be automatically excluded from the scope of the EMC Directive. This was clear enough for systems that are intended to be mounted in new vehicles which are themselves within the scope, but for aftermarket products (i.e. items which are sold for vehicular use but not supplied as original equipment) the situation was confused. Sub-assemblies were exempted from the Automotive Directive until 1st October 2002.

5.1.1.2 The second edition

The second edition Automotive EMC Directive 2004/104/EC [184] was published on 13th November 2004. The timescale for adoption laid out in the second edition is as follows:

- entry into force: 3rd December 2004
- transposition: Member States shall adopt and publish the provisions necessary to comply with this Directive by 31st December 2005, and apply them as from 1st January 2006.

- from 1st January 2006, Member States must recognize equipment or vehicles which comply with the new Directive.
- transitional provisions: from 1st July 2006, new vehicles and equipment must comply; existing vehicles and equipment need not comply till 1st January 2009.

What changes?

The main modifications are:

- The Directive allows for aftermarket equipment, not related to safety critical functions, to be provided with a self-declaration according to the procedures of Directive 89/336/EEC (the EMC Directive, first edition) or 1999/5/EC (the R&TTE Directive) from manufacturers, replacing conventional third party type approval but still subject to a Technical Service assessment. But, "Part of this declaration must be that the ESA fulfils the limits defined in ... this Directive." So a simple CE-marking to the EMC or R&TTE Directive's own standards is not sufficient: the applicable requirements must be matched to the vehicle environment.
- The test provisions and reference limits take into account the work done in international harmonisation. The Directive now refers to CISPR standards for emissions tests (CISPR 12 and 25) and ISO standards for immunity tests (ISO 7637, 11451 and 11452). Extra conditions for applying the tests are given (including, for instance, the requirement for immunity to GSM pulse modulation in the 800–2000 MHz frequency range). This mostly replaces the explicit but confused methods for immunity and emissions testing given in the text of the first Directive. Testing must, though, be carried out by a test house accredited to ISO 17025.
- The new Directive contains for the first time provisions relating to transient disturbances conducted along supply lines, specifically emissions of and immunity from transients according to ISO 7637-2.
- The Directive gives particular attention to safety related functions and components. It introduces the concept of “immunity-related functions”, which are:
  - functions related to the direct control of the vehicle,
  - functions related to driver, passenger and other road-user protection,
  - functions which, when disturbed, cause confusion to the driver or other road users,
  - functions related to vehicle data bus functionality, and
  - functions which, when disturbed, affect vehicle statutory data.

  The vehicle or equipment is considered as complying with immunity requirements if, during the immunity tests, there is no degradation of performance of these “immunity-related functions”. This is more stringent than 95/54/EC which did not require immunity testing at all if equipment did not affect the driver's direct control of the vehicle.

There are other detailed issues which should not be overlooked. Particularly for vehicle manufacturers, “The vehicle manufacturer must provide a statement of frequency bands, power levels, antenna positions and installation provisions for the installation of

RF transmitters, even if the vehicle is not equipped with an RF transmitter at time of type approval. This should cover all mobile radio services normally used in vehicles. This information must be made publicly available following the type approval. Vehicle manufacturers must provide evidence that vehicle performance is not adversely affected by such transmitter installations." The issue of after-market fitment of radio transmitters (not just mobile cellphones) will cause some difficulties for both vehicle manufacturers, who cannot expect to know all types of radio transmitter that will "normally" be fitted to their products throughout their life, and for radio manufacturers, who certainly don't want to have their products approved for individual vehicles.

Spot frequency testing is only required if the Technical Service has to validate test evidence supplied for inclusion in the type approval.

5.1.1.3 Tests in the second edition

The Directive refers to various international standards for the measurement methods, but with modifications and extensions. Table 5.1 shows the tests which are required for electronic sub-assemblies (whole vehicle tests are also mandated but are not discussed here).

Table 5.1 Automotive Directive tests for ESAs

Test type	Frequency range	Method
Radiated broadband emissions, Annex VII	30–1000MHz	CISPR 25: 2002, semi-anechoic chamber or OATS
Radiated narrowband emissions, Annex VIII	30–1000MHz	CISPR 25: 2002, semi-anechoic chamber or OATS
Radio frequency immunity, Annex IX	20–2000MHz; full levels apply over 90% of this range, 5/6 of the full levels apply over whole range	Any combination of the following: — Absorber chamber test: according to ISO DIS 11452-2: 2003 (30V/m) — TEM cell testing: according to ISO 11452-3: 3rd edition 2001 (75V/m) — Bulk current injection testing: according to ISO DIS 11452-4: 2003 (60mA) — Stripline testing: according to ISO 11452-5: 2nd edition 2002 (60V/m) — 800 mm stripline: according to paragraph 4.5 of Annex IX (15V/m)
Conducted transients, Annex X	N/A	Both emissions and immunity to be tested, using method and test pulses 1, 2a, 2b, 3a, 3b and 4 of ISO 7637-2:2002

5.1.2 ISO, CISPR and SAE standards

There are several international standards which give general test methods for vehicles and their components but which are not of themselves mandatory. These are referenced in the Automotive EMC Directive as shown in the table above. CISPR has two vehicle-related emissions standards, and there is a series of immunity standards published by ISO with a similar series published by SAE, the American Society of Automotive Engineers. These are more closely related to the military standards than to commercial ones, but are equivalent to neither. The more important immunity tests are shown in Table 5.2, and the two CISPR emissions tests in Table 5.3.

Table 5.2 ISO and SAE automotive immunity tests

Test	ISO	SAE	Requirements	
ESD	TR 10605	J1113-13	±4, 6, 7, 8kV – direct (contact) discharge; ±4, 8, 14, 15kV – air discharge (extra ±25kV required on vehicle test, test points accessible from outside vehicle)	
Transients	ISO 7637 -1, -2, -3	–	Voltage pulses on supply and signal lines: inductive load supply disconnection and current interruption, switching transients, supply voltage droop, load dump, ignition coil current interruption, alternator field decay	
Conducted RF	ISO 11452-4	J1113-4	Bulk current injection	1MHz – 400MHz, 25 – 40 – 60 – 80 – 100mA
	ISO 11452-7	J1113-3	Direct RF injection	250kHz – 500MHz, 0.05 – 0.1 – 0.2 – 0.3 – 0.4 – 0.5W
Radiated RF	ISO 11452-2	J1113-21	Free field absorber lined chamber	10kHz – 18GHz, 25 – 40 – 50 – 60 – 80 – 100V/m
	ISO 11452-3	J1113-24	TEM cell	10kHz – 200MHz, 50 – 100 – 150 – 200V/m
	ISO 11452-5	–	Stripline	10kHz – 200MHz, 50 – 100 – 150 – 200V/m
	–	J1113-25	Tri-plate	10kHz – 500MHz, 50 – 100 – 150 – 200V/m
	–	J1113-27	Reverberation chamber	500MHz – 2GHz, 25 – 40 – 60 – 80 – 100V/m
All RF immunity tests use unmodulated CW and 80% AM 1kHz with an equivalent peak test level				

Table 5.3 CISPR automotive emissions tests

	Frequency	Description
CISPR 12: protection of off-board receivers	Radiated, 30–1000MHz	Both broadband and narrowband (Class B) emissions limits given at 3 or 10m distance, measured on an outdoor test site (not a standard CISPR OATS) or within an anechoic shielded room with a fixed antenna height
CISPR 25: protection of on-board receivers	Conducted: bands within 0.15–108MHz	5µH/50Ω ±10% artificial network, used up to 108MHz
	Radiated: bands within 0.15–1000MHz, to be raised to 2.5GHz	0.15–30MHz, 1m vertical monopole; 30–1000MHz, biconical/log periodic or equivalent, at 1m from EUT on ground plane bench in anechoic shielded room; also includes alternative TEM cell method, and measurement of emissions received by an antenna on the vehicle

CISPR 12 applies to vehicles and boats driven by an internal combustion engine or electrically, and "devices" (machines such as compressors and chainsaws) with an internal combustion engine. It is a whole vehicle test rather than applying to sub-assemblies, and is intended to protect radio reception away from the vehicle. CISPR 25 by contrast is for protecting radio reception on the vehicle and therefore has various limits set out in bands, for different radio services. It includes both component or module emissions measurements, very similar to the methods in the military standards (see section 5.2), and whole vehicle emissions measurements using the antenna of the type to be supplied with the vehicle.

5.1.3 Vehicle manufacturers

From the point of view of the automotive electronic equipment manufacturer, the legislative EMC requirements are almost secondary; these suppliers have principally to consider their customers. Every major vehicle manufacturer has developed their own EMC test requirements, including levels and limits, through a historical process dependent on their exposure to actually-occurring field EMC problems, and filtered through their own EMC specialists. These are mostly based on the international or US documents but often with significant variations, and often more stringent. Every vehicle component supplier therefore has to negotiate a maze of detailed requirements and test methods in order to be able to supply their apparatus to a number of vehicle manufacturers (VMs). As an example, the requirements for one vehicle manufacturer are outlined in Table 5.4. Types of apparatus are broken down as follows:

- Passive Modules:
  - P: a passive electrical module consisting of only passive components, connected to the vehicle power supply.
- Inductive Devices:
  - R: relays, solenoids and horns.
- Electric Motors:
  - BM: a brush commutated dc electric motor.
  - EM: an electronically controlled electric motor.
- Active Electronic Modules:
  - A: a component that contains active electronic devices.
  - AS: an electronic component or module operated from a regulated power supplier located in another module. This is usually a sensor providing input to a controller.
  - AM: an electronic component or module that contains magnetically sensitive elements or is connected to an external magnetically sensitive element.
  - AX: an electronic module that contains an electric or electronically controlled motor within its package or controls an external inductive device including electric or electronically controlled motor(s).
  - AY: an electronic module that contains a magnetically controlled relay within its package.

Within this classification of devices, each test has varying degrees of applicability to each type. Also, different levels are applied depending on various aspects of the use of the device. It’s probably fair to say that documents such as these (and other vehicle manufacturers have similar specifications) represent the most sophisticated implementation of standardized test methods in any sector.

Table 5.4 Ford Motor Co Component and Subsystem EMC specification [211]

Type	ID	Applicable	Description
Emissions			
Radiated RF	RE 310	BM, EM, all A	0.15–2500 MHz, based on CISPR 25 Ed. 2, ALSE method
Conducted RF	CE 420	BM, EM, all A	LW, MW, FM broadcast bands, based on CISPR 25 Ed. 2
Conducted Transient	CE 410	R, BM, EM, AX, AY	Based on ISO 7637-2 emissions
Radiated immunity			
RF Immunity	RI 112, RI 114	EM, all A	BCI (ISO 11452-4), 1–400MHz; ALSE method (ISO 11452-2) or reverberation chamber method (IEC 61000-4-21), 400–3100 MHz
Magnetic Field	RI 140	AM	Based on MIL-STD-461E, RS101 50Hz–10kHz
Coupled transients			
Inductive	RI 130	EM, all A	In-house chattering relay method
Charging System	RI 150	EM, all A	In-house 0.6–10kHz (sinewave) inductively coupled to test harness
Conducted immunity			
Continuous	CI 210	EM, A, AM, AX, AY	In-house AF ripple 50Hz–10kHz on power and control circuits
Transient	CI 220	P, EM, A, AM, AX, AY	Similar to ISO 7637-2 with additional test pulses
Power Cycle	CI 230	EM, A, AM, AX, AY	In-house, voltage fluctuations on engine start
Ground Offset	CI 250	EM, A, AM, AX, AY	In-house, AC ground voltage offset, 50Hz–1kHz (sinewave)
Voltage Dropout	CI 260	EM, all A	In-house, verification of controlled recovery of hardware and software from power interruptions, various waveforms
Over-voltage	CI 270	All except BM, AS	-14, +19 and +24V on power supply or control circuits
ESD	CI 280	P, EM, all A	Based on ISO 10605

5.1.4 Specialist requirements

There are a number of specialist vehicle applications which demand greater attention to EMC. Foremost among these are the police and emergency services, and the procurement agencies for these bodies place their own technical constraints on equipment which will be used in their vehicles. Some of these requirements are particularly severe. The main concerns are:

- emissions within the frequency bands used for communications; since safety-of-life issues ride on these communications, the allowable levels are much lower than you will find in normal CISPR-based standards. With the implementation of TETRA, compliance with these requirements has become more difficult because the whole of each of the TETRA receive bands must be kept clear, with no relaxation allowable for narrowband spot frequencies.
- immunity to on-board radio transmitters, particularly for safety related and law enforcement equipment, and considering the presence of a wide variety of transmitters both personal and vehicle mounted, with the added complication of TETRA communications which can transmit outside the control of the vehicle user.

5.2 Military

Military equipment is generally not subject to EMC regulation as such. Instead, since military equipment is procured through contract, the EMC specifications can be applied at the contract negotiation stage and this is the usual procedure. This allows the specifications to be fine-tuned and negotiated for a particular application, rather than applied in blanket fashion as happens for commercial products where such pre-purchase technical liaison is unusual. Even so, it is becoming more common for military contracts to specify CE compliance in addition.

As a result the military standards have evolved to cover certain widely-encountered phenomena in a consistent fashion, but without being too prescriptive. Test methods are defined so that test laboratories can use common equipment and procedures, but parameters such as limits, levels to be applied and frequency ranges are left partly open so that they can be adjusted to suit the purpose and product being tested.

Because the US military is a major customer for most Western manufacturers, the US MIL STD series have become *de facto* procurement standards. MIL-STD-461D specified a variety of levels and limits for different purposes, and MIL-STD-462D defined the corresponding test methods. The two were amalgamated into MIL-STD-461E, with a number of changes, in August 1999.

Other countries do have their own variants of the military tests, and in the UK the DEF STAN 59-41 series published by the Ministry of Defence provides a similar variety of tests to the US documents but in a different format, and also gives project planning and documentation requirements along with installation guidelines. DEF STAN 59-41 is at the time of writing undergoing a substantial revision (to become DEF STAN 59-411), which will make detailed changes to some of the tests but is more focussed on revising the structure to incorporate other defence-related EMC subjects, particularly systems trials.

Historically there have been more severe EMC requirements for both emissions and immunity in this sector, since equipment must operate on the "platform" – ship, aircraft, satellite or land vehicle – in close proximity to other apparatus. The internal electromagnetic environment of the platform can usually be closely defined and will typically include several radio transmitters on known frequency bands as well as power supply disturbances due to switching and operation of large motors, actuators, and so on. In addition, the external radio frequency environment, while less predictable, can be quite extreme since the platform can find itself in near proximity and in direct line of sight to very powerful transmitters such as radars and electronic warfare transmitters.

This history, along with the different environmental constraints, explains many of the variations that exist between the military and the commercial standard tests. Even so, it is noticeable that the military tests are far less well controlled in many ways than are their commercial equivalents. Two examples are the use of near field radiated emissions and immunity testing – the antenna is invariably 1m from the EUT, which is mounted on a ground plane bench – and the wide use of current injection probes for conducted emissions and immunity, at frequencies for which the cable looms will be resonant. Both of these practices make for very large uncertainties, which have been much more carefully considered in the IEC/CISPR test methods.

Sections 6.4.4 and 7.3 later describe the most significant technical features of the military emissions and immunity tests; the requirements are summarized below.

5.2.1 DEF STAN 59-41

At present this group of standards is organized as follows [202]:

Control and management:

- Part 1: **Introduction & Guide** – provides definitions and advice on the selection and specification of EMC requirements, including selection of limits; classifies three types of equipment:
  - Type 1: equipment containing electronic components – requires all tests
  - Type 2: motors, generators and electromechanical units (excluding items under Type 3) – requires conducted and radiated emissions, and transient tests
  - Type 3: Relays, solenoids and transformers – requires only imported/exported transients and power frequency magnetic field
- Part 2: **Management & Planning** – discusses test plans and control plans

Testing:

- Part 3: **Technical Requirements, Test methods & Limits** – divided into three categories:
  - Section 1: Man Worn Man Portable Equipment
  - Section 2: Military Support Equipment for Use in a Civilian Environment
  - Section 3: LRU and Sub Systems – generally the most comprehensive of the three, contains a large section on general test requirements, but also refers to Part 5 for certain test equipment
- Part 4: **Large equipment testing** – gives changes from Part 3 for large equipment, likely to be extensively revised and incorporated with section on systems trials
- Part 5: **Specialized EMC test equipment** – covers transient and pulse generators and calibration jigs, monitor loops and voltage probes, Line Impedance Stabilization Networks (LISN), method of damping and verification of screened rooms, and a supply frequency filter for the voltage probe

EMC Design Guidelines:

- Part 6: **Military Vehicles Installation Guidelines**
- Part 7: **HM Ships Installation Guidelines**

In the new format of the standard, whose revision is ongoing at the time of writing [60], Parts 1 and 2 are combined into Volume 1, *Management and Planning*; a new Volume 2, *Electromagnetic Environments*, will be created; Parts 3 and 5 will be combined into Volume 3, *Test Methods and Limits for Sub Systems*; the existing Part 4 and other documents will be combined into a greatly expanded Volume 4, *Platform and System Tests and Trials*; and Parts 6 and 7 along with other DEF STANs will be combined into Volume 5, *Code of Practice for Tri-Service Design and Installation*.

The following tables give the menu of tests currently available in Part 3 Section 3.

Table 5.5 DEF STAN 59-41 emissions tests

Test	Type	Frequency range	Method
DCE 01	Conducted emissions on primary power lines	20Hz to 150MHz	Differential mode, current probe and LISN
DCE 02	Conducted emissions on control, signal and power lines	20Hz to 150MHz	Common mode, current probe on all harnesses
DCE 03	Exported transients on primary power lines	N/A	Oscilloscope, with contactor and functional switching
DRE 01	Electric field radiated emissions	14kHz to 18GHz	Various antennas, 1m from EUT in screened room
DRE 02	Magnetic field radiated emissions	20Hz to 100kHz	Search coil 70mm from each face of the EUT
DRE 03	Radiated emissions – installed antenna, land systems	1.6 to 76MHz	Clansman rod or "L" antenna at 1m from EUT

Table 5.6 DEF STAN 59-41 transient susceptibility tests

Test	Transient type	Applicability	Levels
DCS 04	Type 1: 2 to 30MHz switching	Air	500V, 20A peak, all cables
	Type 2: 100kHz switching	Air	700V, 30A peak, power; 100V, 5A peak, signal lines
DCS 05	Type 1N: 0.5 to 50MHz switching	Land and sea	10A peak, all cables
	Type 1N: NEMP	Land and sea	100A peak, all cables
DCS 06	Type 2: 100kHz switching	Land and sea	2kV, 100A peak, power
DCS 08	Lightning EMP/NEMP	Air	3kV, 30A peak, all cables
DCS 09	Direct Lightning	Air	5kV, 10kA peak, all cables
DCS 10	Electrostatic Discharge	All	Up to 8kV
DCS 12	LF switching	Sea	Up to 2.5kV peak (power)

Table 5.7 DEF STAN 59-41 continuous susceptibility tests

Test	Description	To test immunity	Method
DCS 01	Power 20Hz to 50kHz	Ripple on power supply waveform	Coupling transformer in series with power line
DCS 02	Power, signal & control 50kHz to 400MHz	Disturbances induced by local transmitters	Pre-calibrated current injected by current probe onto each cable bundle, primary power lines also tested individually
DCS 03	Control & signal 20Hz to 50kHz	Ripple on adjacent cables	Current passed through test wire, three turns wrapped around cable under test
DRS 01	H Field 20Hz to 100kHz	Magnetic field from e.g. transformers and power cables	Calibrated radiating loop, 5cm from the EUT face
DRS 02	E Field 14kHz to 18GHz	Transmitted fields	Anechoic screened room, E-field sensor monitoring field during test, transmit antenna 1m from boundary of EUT; or alternative method using mode-stirred reverberation chamber above 100MHz for aircraft equpt
DRS 03	Magnetostatic (H Field DC)	High power DC current, degaussing fields	DC current passed through Helmholtz coil assembly

5.2.2 MIL STD 461

This is the principal US military EMC standard for equipment [209], although not by any means the only one, and has been in widespread use for many years. As can be seen, some of its tests are similar to some of the DEF STAN 59-41 tests, but there are a number of different methods and even the similar tests have detailed differences.

Table 5.8 MIL-STD-461E emissions tests

Test	Type	Frequency range	Method
CE 101	Conducted emissions on power leads	30Hz to 10kHz	Differential mode, current probe and LISN
CE 102	Conducted emissions on power leads	10kHz to 10MHz	Voltage measurement on LISN port, each power lead
CE 106	Conducted emissions, antenna terminal	10kHz to 40GHz depending on EUT operation	Direct connection or via coupler, to antenna port
RE 101	Magnetic field radiated emissions	30Hz to 100kHz	Search coil 70mm from each EUT face and connector
RE 102	Electric field radiated emissions	10kHz to 18GHz	Various antennas, 1m from EUT; screened room preferred

Table 5.8 MIL-STD-461E emissions tests (Continued)

Test	Type	Frequency range	Method
RE 103	Radiated emissions – antenna spurious and harmonic outputs	10kHz to 40GHz depending on EUT operation	Alternative to CE106 for transmitters with integral antennas

Table 5.9 MIL-STD-461E susceptibility tests

Test	Description	To test immunity	Method
CS 101	Power leads, 30Hz to 150kHz	Ripple on power supply	Coupling transformer in series with power line
CS 103	Antenna port, intermodulation, 15kHz to 10GHz	Presence of intermodulation products	Determined on a case-by-case basis
CS 104	Antenna port, undesired signal rejection, 30Hz to 20GHz	Presence of spurious responses	Determined on a case-by-case basis
CS 105	Antenna port, cross modulation, 30Hz to 20GHz	Presence of cross-modulation products	Determined on a case-by-case basis
CS 109	Structure current, 60Hz to 100kHz	Currents flowing in the EUT structure	Currents injected by transformer at diagonal extremes across surfaces
CS 114	Bulk cable injection, 10kHz to 200MHz	RF signals coupled onto EUT associated cabling	Pre-calibrated current injected by current probe onto each cable bundle, including power leads with returns and grounds excluded
CS 115	Bulk cable injection, impulse excitation	Impulse signals coupled onto EUT associated cabling	As CS114, but with impulse generator giving 30ns pulses at 30Hz repetition rate
CS 116	Damped sinusoidal transients, cables and power leads, 10kHz to 100MHz	Damped sinusoidal transients due to excitation of wiring, coupled onto cables and power leads	As CS114, but with damped sinewave generator giving pulses at least once per second at a minimum of 0.01, 0.1, 1, 10, 30, and 100 MHz
RS 101	Magnetic field 30Hz to 100kHz	Magnetic field from e.g. transformers and power cables	Calibrated radiating loop, 5cm from the EUT face, or place EUT within calibrated Helmholtz coils
RS 103	Electric Field 2MHz to 40GHz	Transmitted fields	Anechoic screened room, E-field sensor monitoring field during test, transmit antenna 1m from boundary of EUT; or alternative method using mode-stirred reverberation chamber

Table 5.9 MIL-STD-461E susceptibility tests (Continued)

Test	Description	To test immunity	Method
RS 105	Transient electromagnetic field	Unidirectional pulsed radiated field, 2.3/23ns 50kV/m	Transient pulse generator feeding TEM cell, parallel plate transmission line or similar

5.3 Aerospace

EMC test requirements for civil aerospace equipment are defined in two documents:

- EUROCAE/ED-14E in Europe
- RTCA/DO-160E in the USA.

They describe a series of minimum standard environmental test conditions and applicable test procedures for airborne equipment, intended for everything from light aircraft and helicopters through to large passenger airliners. The documents are worded identically and both are subject to regular changes: at the time of writing the current issue is E, published in September 2004. There are 26 sections, of which sections 15 through to 23 and section 25 cover various EMC phenomena.

Aircraft themselves are not covered by the EMC Directive within Europe. The legislation that applies to them, and hence to their equipment, is in terms of airworthiness acceptance by the national aviation authorities, overseen by the European Joint Aviation Authorities (JAA). One exception to the scope of the R&TTE Directive is radio transmitters for aircraft and air traffic management systems, for which the original EMC Directive required an EC Type-Examination route to compliance, which in the UK was administered by the CAA as a Notified Body. Passenger-carried electronic devices (PEDs) are of course not covered by these standards, which has led to ongoing concern about their possible effects on aircraft control and navigation systems (see section 1.1.1).

5.3.1 DO-160/ED-14

The various sections of DO-160 [210] that are relevant for EMC purposes are given in Table 5.10. Some of the sections are relatively undemanding and well-established; the ones that have undergone the most significant changes in the last few years are section 20 on RF susceptibility and section 22 on lightning induced transients [35][145]. These are in response to the challenges associated with the introduction of both composite structures and fly-by-wire systems in modern aircraft; greater reliability is needed, but at the same time less protection from radiated fields and lightning currents is enjoyed by many of the electrical sub-systems. The radiated RF susceptibility test now refers to the mode-stirred reverberation chamber as an alternative method to the traditional semi-anechoic chamber, which is just as well, since the maximum test levels are impressive: the highest level for Category F from 4-6GHz is 7200V/m pulsed.

5.3.1.1 Lightning induced transients

The lightning requirements are split between sections 22 and 23, but section 23 is mostly applicable for whole aircraft tests with externally-mounted equipment, in very large high-voltage facilities. The tests in section 22 have five levels, the highest of which are far more stringent than levels appearing in the aircraft requirements of military standards. Section 22 suggests four installation protection zones:

- Zone a: well protected environment, for example within the passenger cabin. Because equipment located in this environment is furthest from the aircraft skin and should be protected through the systems it communicates with, the lowest test level (level 1) applies.
- Zone b: partially protected environments such as equipment electronic bays are distributed around the airframe with cables linking equipment in other zones or to another electronic bay. Cables linking electronic bays, regardless of whether they run through a well protected environment, should be considered as belonging to the equipment bay category. Such equipment requires level 2.
- Zone c: moderately protected environments are considered to be those areas potentially subject to direct electromagnetic interference effects. Cockpit areas fall into this category and equipment mounted here should be subjected to tests at level 3.
- Zone d: severe electromagnetic effects are most likely in airframes with significant amounts of composite material without wire meshing. Equipment in this category could be landing gear or engine or flight controls. They are recommended to be tested to level 4 or 5. The values for level 4 are more than double, and for level 5 more than five times, the values for level 3. Testing at these levels (up to 3200V and 3200A, though not at the same time) requires a significant test equipment investment.

All cable bundle tests take into account the potential influence of EUT cabling on the impulse, with respect to its amplitude, by defining parameters of I_{Test} and V_{Limit} or V_{Test} and I_{Limit} values. A "Test" value is the ideal that should be reached if possible. The "Limit" value is the maximum allowable value measured in a cable bundle to prevent over-stressing the EUT. When this occurs, the test is deemed to have been completed. The "Test" and "Limit" values do not define the generator impedance; this is given only by the pin injection requirements. Because the cable bundle impedance is so significant, the type and routing of the cable influences whether the "Test" or "Limit" value is reached first.

Table 5.10 DO-160E/ED-14E EMC tests

Part	Test	Description of requirement
Section 15	Magnetic effect	Compass deflection measurement: separation distance for one degree deflection
Section 16	Power input	Normal and emergency power conditions: variable voltage, frequency and phase unbalance, ripple, surge, dips and interruptions, harmonic emissions
Section 17	Voltage spike	Two categories of transient waveform applied to primary power inputs; similar to MIL-STD-462D method CS06
Section 18	AF conducted susceptibility	Differential mode injected AF on power inputs, up to 10Hz–150kHz; similar to MIL-STD-461E method CS101

Table 5.10 DO-160E/ED-14E EMC tests (Continued)

Part	Test	Description of requirement
Section 19	Induced signal susceptibility	400Hz–15kHz electric and magnetic field, and unsuppressed relay coil spikes, induced via near-field coupling to cable bundle; 400Hz 20A magnetic field for EUT; similar to MIL-STD-462D method RS02
Section 20	RF susceptibility: conducted	10kHz–400MHz pre-calibrated bulk current injection on all interconnecting cable bundles; similar to MIL-STD-461E method CS114
	RF susceptibility: radiated	100MHz–18GHz with antenna in semi-anechoic chamber, similar to MIL-STD-461E method RS103, or mode-stirred reverberation chamber
Section 21	RF emissions: conducted	150kHz–30MHz using LISN or current probe, on primary power lines and interconnecting cable bundles
	RF emissions: radiated	2MHz–6GHz, various antennas at 1m from EUT
Section 22	Lightning induced transient susceptibility	Five levels depending on location (levels 4 and 5 most relevant for exposed locations in composite aircraft), several waveforms including multiple stroke and multiple burst; pin injection, cable bundle induction and ground injection techniques
Section 23	Lightning direct effects	Applies to externally mounted equipment, categories defined depending on lightning protection zones, equipment normally unpowered, generator must create 100's of kV
Section 25	Electrostatic discharge	As IEC 61000-4-2 except EUT bonded to ground plane and air discharge only at 15kV applied

5.4 Rail

The railway industry in Europe is in a somewhat unusual position regarding EMC; on the one hand it has to ensure operational safety, but on the other its installations are covered under the EMC Directive, which has nothing to do with safety. This leads to parallel compliance requirements both for Network Rail and the train operators themselves, and for the suppliers of equipment into the rail industry [138].

5.4.1 Railway Group Standards

On the UK mainline system, Railway Group Standards (RGS) are mandatory on all members of the Railway Group, which comprises the infrastructure controller (Network Rail) and the train operating companies as well as associated organizations. Compliance with the route acceptance process is normally demonstrated by way of the Railway Safety Case which is based on the successful application of all relevant RGSs. For EMC, the relevant top-level RGS is GE/RT8015, *Electromagnetic Compatibility between Railway Infrastructure and Trains* [197]. This places generalized requirements for emissions and susceptibility limits on both the trains and the infrastructure. The default levels are those set out in EN 50121 parts 3, 4 and 5, but with the caveat that the rolling stock may have to meet extra requirements based on known emissions or susceptibilities of the infrastructure. This in turn means that the infrastructure controller must establish and maintain information on its systems, and GE/RT8015 devotes considerable space to codifying its responsibilities for this, including:

- identifying all safety-related infrastructure systems which could have their safety performance reduced through EMI;
- analysing the susceptibility of such systems, particularly to determine the nature and level of train emissions which would affect the safety perfomance;
- documenting the analysis and making it available to train operators and other stakeholders, particularly so that the train operator can demonstrate compatibility with the infrastructure system;
- implementing a maintenance and testing regime to ensure that the susceptibility of the system is not worsened.

Since implementing these processes, the railway industry has demonstrated a voracious appetite for consuming all available types of EMC expertise.

5.4.1.1 Typical safety-critical infrastructure and train systems

GE/RT8015 gives an indication (not exhaustive) of the types of systems and equipment whose safety performance could be affected by EMI, and which should be included in the analysis.

For infrastructure:

- train detection systems (including track circuits and axle counters);
- interlocking systems;
- signals and point operating equipment and their controlling circuits;
- train warning and protection systems;
- telecommunications systems (including voice and data transmission, and supervisory control and data acquisition (SCADA) systems);
- radio systems (including voice and data transmission, fixed and mobile systems).

For trains:

- braking systems;
- traction control systems;
- tilt control systems;
- door control systems;
- coupling systems;
- communications systems;
- lighting systems (internal and external);
- train-borne elements of command and control systems.

5.4.2 London Underground standards

The basic framework which gives the EMC requirements for equipment installed within the London Underground (LU) network are set out in LU CED Engineering Standard 2-01018-001 A2 (formerly E 1027). The objective of the standard is to ensure that LU meets the requirements of the EMC Directive and corresponding UK Regulations. 2-01018-001 A2 is very much a top-level document setting out responsibilities and procedures. A companion guideline document, M1027 (5-01018-001) *Manual of EMC best practice*, provides greater technical detail regarding EMC requirements. The top level document includes the requirement for an EMC Control Plan which is detailed in M1027 together with the requirement for an EMC Test Plan.

LU engineering documents are issued by the Chief Engineer's Directorate and a number of these relate to EMC. These documents emphasise the need to provide EMC assurance for the management and improvement of assets on LU, including rolling stock. Unlike the Railway Group Standards for the mainline railway, the top-level standard 2-01018-001 A2 makes it clear from the outset that the main objective is to ensure that LU meets the requirements of the EMC Directive and the corresponding UK Regulations. It also states that compliance with appropriate parts of EN 50121 is a requirement together with any special LU needs.

5.4.3 EN 50121

This document was first published as a pre-standard in 1996, and then as a full standard in 2000, with the overall title of "Railway applications – Electromagnetic compatibility". It has a total of six parts.

5.4.3.1 Parts of EN 50121

Part 1: General

- Describes the EM behaviour of the railway system, gives the (generic) immunity performance criteria, and also discusses the management of EMC at the interface between the infrastructure and the trains.

Part 2: Emission of the whole railway system to the outside world

- Sets the RF emissions limits at 10m from the railway track from 9kHz to 1GHz, and gives the methods of measurement. This is not a trivial matter, because the measurement must take place on a real railway, or at least on a section of test track which is representative of a real railway. Ambient and weather conditions can seriously affect the measurement. Most importantly, it may be expected that maximum emissions will occur with the traction unit either at maximum power or at maximum speed (the two conditions may not be the same). But a train at maximum speed will be past the measurement point in a few seconds. So some quite sophisticated test methods have to be developed to allow the capture of the worst case emissions in a realistic measurement time.

Part 3-1: Rolling stock: Train and complete vehicle

- Sets the emission and immunity requirements for all types of rolling stock; its scope ends at the interface of the stock with its respective energy inputs and outputs, i.e. for locomotives it is the sliding contact, for coaches and wagons, it is the AC or DC power connector. The standard does not in fact give any immunity requirements, but it says:

> the immunity tests and limits in Part 3-2 of this standard were selected in the knowledge that the vehicle should be immune to a level of 20V/m over the frequency range 0.15MHz to 1GHz. It is expected that the assembly of the apparatus into a complete vehicle will give adequate immunity, provided that an EMC plan has been prepared and implemented, using the limits in Part 3-2 of this standard.

The emissions limits are similar to but slightly lower than those specified in part 2.

Part 3-2: Rolling stock: Apparatus

- Applies emission and immunity requirements for apparatus intended for use on rolling stock.

Part 4: Emission and immunity of the signalling and telecommunications apparatus

- Applies emission and immunity requirements for signalling and telecom.

Part 5: Emission and immunity of fixed power supply installations and apparatus

- Applies emission and immunity requirements for apparatus and systems intended for use in the railway power supply: this includes the power feed, the supply equipment itself with protection and control circuits, and trackside items such as switching stations, transformers and switchgear.

In all of parts 3-2, 4 and 5 the usual basic standards in the IEC 61000-4 series are referenced, as with the generic standards, with in the main industrial or higher levels of stress: for instance, the RF immunity level in Part 4 is 20V/m, in Part 5 the power line EFT/Burst and surge immunity level is 4kV.

5.4.3.2 EN 50121 and the EMC Directive

This standard was written principally to allow railway operators to demonstrate compliance with the EMC Directive, via the standards route. It is also intended to achieve EMC between various parts of the railway. So far, the first of these purposes has not been achieved, for a mixture of reasons: it deals in part with installations, and installations so far do not require CE Marking; and there is considerable disquiet within CISPR and at the European level that the RF emissions levels it sets are far too lax, and out of step with other CISPR-based limits. As a result it has not been harmonised, and therefore equipment and installations using it under the first edition EMCD could only use the Technical Construction File route. Now that the second edition EMCD has swept away the TCF route and requires simply an EMC Assessment, which may or may not use harmonised standards, the way seems to be clearer for the railway industry to use EN 50121 together with an EMC Assessment for compliance purposes.

Unfortunately, it is noticeable on looking through the various parts of the standard how many omissions there are, both explicit and implicit. Very often these seem to fall under the heading of “too difficult”: or at least, too complex to be put into a European standard, and best left to individual circumstances. Industry sources suggest that the limits, particularly those in Part 2 for emissions to the outside world, were arrived at simply by a process of measuring what was actually occurring in real life, and drawing an envelope which encompassed all the measurements, on the basis that this would continue to be achievable in the future and there hadn’t been enough complaints of radio interference so far to worry about. A comparison of some of these against comparable CISPR limits is shown in Figure 5.1. It is perhaps unsurprising that the CENELEC Technical Board have not yet seen fit to recommend this series of standards as suitable for declaring compliance with the EMC Directive.

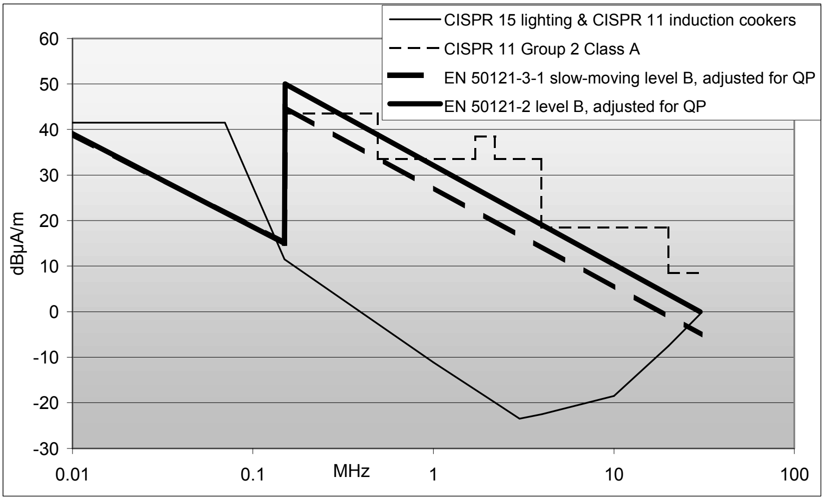


Figure 5.1 Comparison between CISPR and EN 50121 magnetic field emission limits

NB all limits are expressed at a distance of 10m; the values for EN 50121 were reduced by 20dB to allow for the difference between peak and quasi-peak detection, as assumed by the standard; Level B is for 15kV AC, 3kV DC and 1.5kV DC systems, level A for 25kV AC systems is 5dB higher

Chapter 6

RF emissions measurements

One of the aspects of electromagnetic compatibility that is most difficult to grasp is the raft of techniques that are involved in making measurements. EMC phenomena extend in frequency to well beyond 1GHz and this makes conventional and well-known techniques, established for low frequency and digital work, quite irrelevant. Development and test engineers must appreciate the basics of high frequency measurements in order to perform, or at least understand, the EMC testing that will be demanded of them. This chapter and the next will serve as an introduction to the equipment, the test methods and some of the causes of error and uncertainty that attend high frequency EMC testing.

For ease of measurement and analysis, in the commercial tests radiated emissions are assumed to predominate above 30MHz and conducted emissions are assumed to predominate below 30MHz. There is of course no magic changeover at 30MHz. But typical cable lengths tend to resonate above 30MHz, leading to anomalous conducted measurements, while measurements of radiated fields below 30MHz will of necessity be made in the near field if closer to the source than $\lambda/2\pi$ (see section 10.1.4.2), which gives results that do not necessarily correlate with real situations. In practice, investigations of interference problems have found that controlling the noise voltages developed at the mains terminals has been successful in alleviating radio interference in the long, medium and short wave bands [85]. At higher frequencies, mains wiring becomes less efficient as a propagation medium, and the dominant propagation mode becomes radiation from the equipment or wiring in its immediate vicinity. If you are considering military, aerospace or automotive tests, the supply wiring is relatively less important as a principal route, and both conducted and radiated emissions tests are performed over a wider and overlapping frequency range.

Emissions testing requires that the equipment under test (EUT) is set up within a controlled electromagnetic environment under its normal operating conditions. If the object is to test the EUT alone, rather than as part of a system, its ancillary support equipment (if any) must be separately screened from the measurement (see section 9.2.4.4). Any ambient signals should be well below the levels to which the equipment will be tested.

6.1 Emissions measuring instruments

6.1.1 Measuring receiver

Conformance test measurements are normally taken with a measuring receiver, which is optimized for the purpose of taking EMC measurements. Typical costs for a complete receiver system to cover the range 10kHz to 1GHz can be anywhere between £15,000–£60,000.

Early measuring receivers were manually tuned and the operator had to take readings from the meter display at each frequency that was near to the limit line. This was a lengthy procedure and prone to error. The current generation of receivers are fully automated and can be software controlled via an IEEE-488 standard bus; this allows a PC-resident program to take measurements with the correct parameters over the full frequency range of the test, in the minimum time consistent with gap-free coverage. Results are stored in the PC's memory and can be processed or plotted at will.

The distinguishing features of a measuring receiver compared to a spectrum analyser are:

- the receiver output is provided at a spot frequency, although high-end units can also provide a spectrum display;
- very much better sensitivity, allowing signals to be discriminated from the noise at levels much lower than the emission limits;
- robustness of the input circuits, and resistance to overloading;
- intended specifically for measuring to CISPR standards, with bandwidths, detectors and signal circuit dynamic range tailored for this purpose;
- frequency and amplitude accuracy is better than the cheaper spectrum analysers;
- two units may be required, one covering up to 30MHz and the other covering 30–1000MHz or higher.

6.1.2 Spectrum analyser

A fairly basic spectrum analyser can be cheaper than a measuring receiver (typically £7,000–£15,000) and is widely used for "quick-look" testing and diagnostics. The instantaneous spectrum display is extremely valuable for confirming the frequencies and nature of offending emissions, as is the ability to narrow-in on a small part of the spectrum. When combined with a tracking generator, a spectrum analyser is useful for checking the HF response of circuit networks.

Basic spectrum analysers are not an alternative to a measuring receiver in a full compliance set-up because of their limited sensitivity and dynamic range, and susceptibility to overload. Figure 6.1(a) shows the block diagram of a typical spectrum analyser. The input signal is fed straight into a mixer which covers the entire frequency range of the analyser with no advance selectivity or preamplification. The consequences of this are threefold: firstly, the noise figure is not very good, so that when the attenuation due to the transducer and cable is taken into account, the sensitivity is hardly enough to discriminate signals from noise at the lower emission limits (see section 6.2.1.1 later). Secondly, the mixer diode is a very fragile component and is easily damaged by momentary transient signals or continuous overloads at the input. If you take no precautions to protect the input, you will find your repair bills escalating quickly. Thirdly, the energy contained in broadband signals can overload the mixer and drive it into non-linearity even though the energy within the detector bandwidth is within the instrument's apparent dynamic range; this means you could be making an artificially low measurement, due to overloading, without realizing it.

6.1.2.1 Preselector

You can find instruments which offer a performance equivalent to that of a measuring receiver, but the price then becomes roughly equivalent as well. A more satisfactory

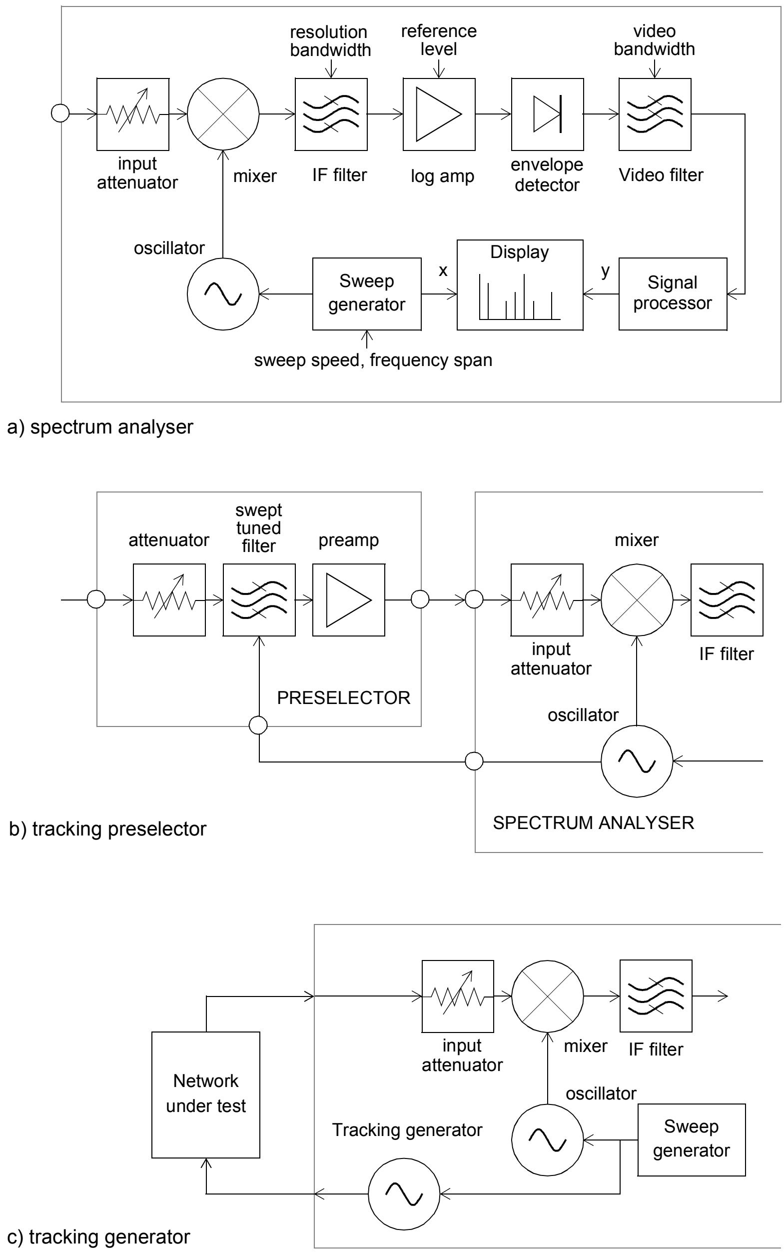


Figure 6.1 Block diagram of spectrum analyser

compromise is to enhance the spectrum analyser's front-end performance with a tracking preselector. The preselector (Figure 6.1(b)) is a separate unit that contains input protection, preamplification and a swept tuned filter which is locked to the spectrum analyser's local oscillator. The preamplifier improves the system noise performance to that of a test receiver. Equally importantly, the input protection allows the instrument to be used safely in the presence of gross overloads, and the filter reduces the energy content of broadband signals that the mixer sees, which improves the effective dynamic range.

The negative side of a preselector is that it can cost virtually as much as the spectrum analyser itself, doubling the cost of the system. There are manual preselectors available, but these are clumsy to use. But you can treat it as an upgrade. The analyser can be used on its own for diagnostics testing, and you can add a preselector when the time comes to make compliance measurements. Like the measuring receiver, modern spectrum analysers and preselectors can be software controlled via an IEEE-488 bus, and provided the system hardware is adequate you can perform the compliance testing in exactly the same way (using a PC for the data processing).

6.1.2.2 Tracking generator

Including a tracking generator with the spectrum analyser greatly expands its measuring capability without greatly expanding its price. With it, you can make many frequency-sensitive measurements which are a necessary feature of a full EMC test facility.

The tracking generator (Figure 6.1(c)) is a signal generator whose output frequency is locked to the analyser's measurement frequency and is swept at the same rate. The output amplitude of the generator is maintained constant within very close limits, typically less than ±1dB over 100kHz to 1GHz. If it provides the input to a network whose output is connected to the analyser's input, the frequency–amplitude response of the network is instantly seen on the analyser. Whilst it is not as capable as a proper vector network analyser, it gives some of the same functions at a fraction of the cost. The dynamic range could be theoretically equal to that of the analyser (up to 120dB), but in practice it is limited by stray coupling which causes feedthrough in the test jig.

You can use the tracking generator/spectrum analyser combination for several tests related to EMC measurements:

- characterize the loss of RF cables. Cable attenuation versus frequency must be accounted for in an overall emissions measurement;
- perform open site attenuation calibration (section 6.3.1.2). The site loss between two calibrated antennas versus frequency is an essential parameter for open area test sites;
- characterize components, filters, attenuators and amplifiers. This is a vital tool for effective EMC remedies;
- make tests of shielding effectiveness of cabinets or enclosures;
- determine structural and circuit resonances.

6.1.3 Receiver specifications

Whether you use a measuring receiver or spectrum analyser, there are certain requirements laid down in the relevant standards for its performance. The "relevant standards" are CISPR 16-1-1 for CISPR-related tests, and MIL STD 461E or DEF STAN 59-41 for military tests.

6.1.3.1 Bandwidth

The actual value of an interference signal that is measured at a given frequency depends on the bandwidth of the receiver and its detector response. These parameters are rigorously defined in a separate standard that is referenced by all the commercial emissions standards that are based on the work of CISPR, notably EN 55011, 55013, 55014-1 and 55022. This standard is CISPR publication 16-1-1 [162].

CISPR 16-1-1 splits the measurement range of 9kHz to 1000MHz into four bands, and defines a measurement bandwidth for quasi-peak detection which is constant over each of these bands (Table 6.1). Sources of emissions can be classified into *narrowband*, usually due to oscillator and signal harmonics, and *broadband*, due to discontinuous switching operations, commutator motors and digital data transfer. The actual distinction between narrowband and broadband is based on the bandwidth occupied by the signal compared with the bandwidth of the measuring instrument. A broadband signal is one whose occupied bandwidth exceeds that of the measuring instrument. Thus a signal with a bandwidth of 30kHz at 20MHz (CISPR band B) would be classed as broadband, while the same signal at 40MHz (band C) would be classed as narrowband.

Quasi-peak detector	Frequency band A 9–150kHz	 B 0.15–30MHz	 C 30–300MHz	 D 300–1000MHz
6dB bandwidth	200 Hz	9 kHz	120 kHz	
Charge time constant, ms	45	1	1	
Discharge time constant, ms	500	160	550	
Pre-detector overload factor, dB	24	30	43.5	

Table 6.1 The CISPR 16-1 quasi-peak detector and bandwidths

Noise level versus bandwidth

The indicated level of a broadband signal changes with the measuring bandwidth. As the measuring bandwidth increases, more of the signal is included within it and hence the indicated level rises. The indicated level of a narrowband signal is not affected by measuring bandwidth. Noise, of course, is inherently broadband, and therefore there is a direct correlation between the "noise floor" of a receiver or spectrum analyser and its measuring bandwidth: minimum noise (maximum sensitivity) is obtained with the narrowest bandwidth. The relationship between noise and bandwidth is given by equation (6.1):

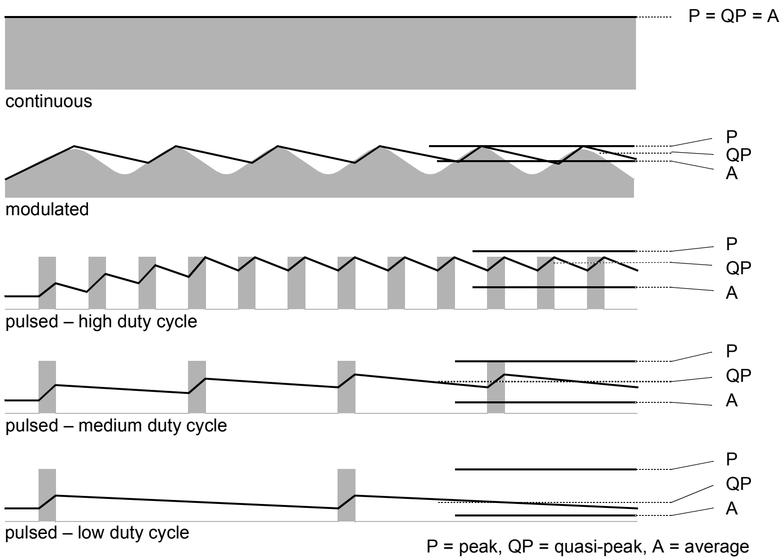
$$\text{Noise level change (dB)} = 10\log_{10}(BW_1/BW_2) \tag{6.1}$$

For instance, a change in bandwidth from 10kHz to 120kHz would increase the noise floor by 10.8dB.

6.1.3.2 Detector function

There are three kinds of detector in common use in RF emissions measurements: peak, quasi-peak and average. The characteristics are defined in CISPR 16-1-1 and are different for the different frequency bands.

Interference emissions are rarely continuous at a fixed level. A carrier signal may be amplitude modulated, and either a carrier or a broadband emission may be pulsed. The measured level which is indicated for different types of modulation will depend on the type of detector in use. Figure 6.2 shows the indicated levels for the three detectors with various signal modulations.



NB the detector function refers to the signal's modulation characteristics. All detectors respond to the RMS value of the unmodulated RF voltage.

Figure 6.2 Indicated level versus modulation waveform for different detectors

Peak

The peak detector responds near-instantaneously to the peak value of the signal and discharges fairly rapidly. If the receiver dwells on a single frequency the peak detector output will follow the "envelope" of the signal, hence it is sometimes called an envelope detector. Military specifications make considerable use of the peak detector, but CISPR emissions standards do not require it at all for frequencies below 1GHz. However its fast response makes it very suitable for diagnostic or "quick-look" tests, and it can be used to speed up a proper compliance measurement as is outlined in section 6.4.2.

Average

The average detector, as its name implies, measures the average value of the signal. For a continuous signal this will be the same as its peak value, but a pulsed or modulated signal will have an average level lower than the peak. The main CISPR standards call for an average detector measurement on conducted emissions, with limits which are 10–13dB lower than the quasi-peak limits. The effect of this is to penalize continuous emissions with respect to pulsed interference, which registers a lower level on an average detector [86]. A simple way to make an average measurement on a spectrum analyser is to reduce the post-detector "video" bandwidth to well below the lowest expected modulation or pulse frequency [90].

Quasi-peak

The quasi-peak detector is a peak detector with weighted charge and discharge times (Table 6.1) which correct for the subjective human response to pulse-type interference. Interference at low pulse repetition frequencies (PRFs) is said to be subjectively less annoying on radio reception than that at high PRFs. Therefore, the quasi-peak response de-emphasizes the peak response at low PRFs, or to put it another way, pulse-type emissions will be treated more leniently by a quasi-peak measurement than by a peak measurement. But to get an accurate result, the measurement must dwell on each frequency for substantially longer than the QP charge and discharge time constants.

Since CISPR-based tests have historically been intended to protect the voice and broadcast users of the radio spectrum, they lay considerable emphasis on the use of the QP detector. There is a point of view which suggests that with the advent of digital telecommunications and broadcasting this will change, since digital signals are affected by impulsive interference in a quite different way.

Possible future detectors

A study group within CISPR is looking at other means of detector weighting, and various documents have been circulated in the last few years. At the time of writing the favourite [169] appears to be a weighting detector which is a combination of an RMS detector (for pulse repetition frequencies above a corner frequency f_c) and the average detector (for pulse repetition frequencies below the corner frequency f_c), which achieves a pulse response curve with the following characteristics: 10dB/decade above the corner frequency and 20dB/decade below the corner frequency. The draft goes on to say:

> Nowadays the majority of disturbance sources may not contain repeated pulses, but still a great deal of equipment contains broadband emissions (with repeated pulses) and pulse modulated narrowband emissions. In addition, the transition from analog radiocommunication services to digital radiocommunication services has happened to a great deal and is partially still going on. The introduction of a new detector type may follow the transition from analog to digital radiocommunication systems. This transition may be regarded as a matter of frequency ranges: above 1 GHz, the use of digital radiocommunication systems is more frequent than below.

6.1.3.3 Overload factor

A pulsed signal with a low duty cycle, measured with a quasi-peak or average detector, should show a level that is less than its peak level by a factor which depends on its duty cycle and the relative time constants of the quasi-peak detector and PRF. To obtain an accurate measurement the signal that is presented to the detector must be undistorted at very much higher levels than the output of the detector. The lower the PRF, the higher will be the peak value of the signal for a given output level (Figure 6.3). Conventionally, the input attenuator is set to optimize the signal levels through the receiver, but the required pulse response means that the RF and IF stages of the receiver must be prepared to be overloaded by up to 43.5dB (for CISPR bands C and D) and remain linear. This is an extremely challenging design requirement and partially accounts for the high cost of proper measuring receivers, and the unsuitability of spectrum analysers for pulse measurements.

The same problem means that the acceptable range of PRFs that can be measured by an average detector is limited. The overload factor of receivers up to 30MHz is only

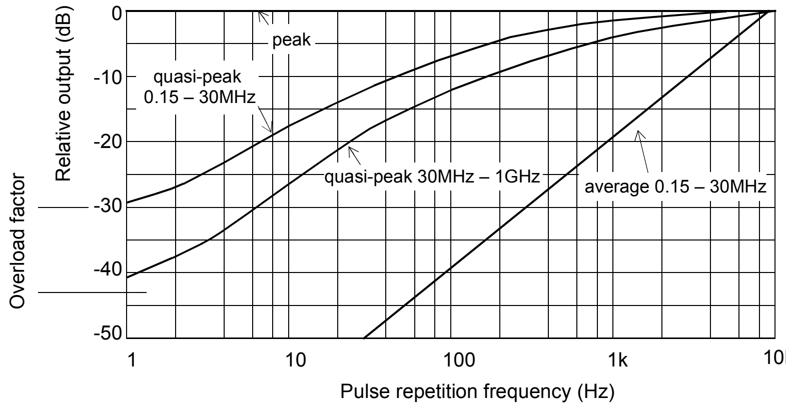


Figure 6.3 Relative output versus PRF for CISPR 16-1 detectors

required to be 30dB, and this degree of overload would be reached on an average detector with a pulsed signal having a PRF of less than 300Hz. For this reason average detectors are only intended for measurement of continuous signals to allow for modulation or the presence of broadband noise, and are not generally used to measure impulsive interference.

6.1.3.4 Measurement time

Both the quasi-peak and the average detector require a relatively long time for their output to settle on each measurement frequency. This time depends on the time constants of each detector and is measured in hundreds of milliseconds. When a range of frequencies is being measured, the conventional method is to step the receiver at a step size of around half its measurement bandwidth, in order to cover the range fully without gaps (for the specified CISPR filter shape, the optimum step size is around 0.6 times the bandwidth, to give the largest step size consistent with maintaining the accuracy of measurement of any signal). For a complete measurement scan of the whole frequency range, as is required for a compliance test, the time taken is given by:

$$T = (\text{frequency span/step size}) \cdot \text{dwell time per spot frequency} \qquad (6.2)$$

If the dwell time is restricted to three time constants, the time taken to do a complete quasi-peak sweep from 150kHz to 30MHz turns out to be 53 minutes. For an average measurement the scan time would be even longer, were it not for the way in which the average limit is applied. The difference between the quasi-peak and average limits is at most 13dB (in CISPR 22, Class A) and this difference occurs at a PRF of 1.8kHz. The decisive value for lower PRFs is always the QP indication. Therefore the average indication only has to be accurate for modulated or pulsed signals above this PRF, and this can be ensured with only a short dwell time, such as 1ms.

But if you were to sweep with the quasi-peak detector, the dwell time would have to be increased to ensure that the peaks are captured and indicated correctly. It should be at least 1 second, if the signals to be measured are unknown. This has repercussions on the test method, as is discussed later in section 6.4. It places correspondingly severe restrictions on the sweep rate when you are using a spectrum analyser [40].

6.1.3.5 Input VSWR

For any RF measuring instrument, the input impedance is crucial since amplitude accuracy depends on the maximum power transfer from the antenna, through the connecting cable to the receiver input. Invariably the system impedance is specified as 50Ω. As long as the receiver input is exactly 50Ω resistive, all the power is transferred without loss and hence without measurement error. Any departure from 50Ω causes some power to be reflected and there is said to be a "mismatch error".

In practice the receiver input impedance cannot maintain a perfect 50Ω across the whole frequency range, and the degree to which it departs from this is called its VSWR (Voltage Standing Wave Ratio, see appendix D section D.2.4 and section 6.5.2.1). VSWR can also be expressed differently as return loss or input reflection coefficient. A VSWR of 1:1 means no mismatch; CISPR 16-1-1 requires the receiver to have better than 2:1 VSWR with no input attenuation, and better than 1.2:1 with 10dB or greater input attenuation. There is a trade-off between input attenuation and sensitivity. As far as possible, receivers should be operated with at least 10dB input attenuation since this gives a better match and greater accuracy, but this will bring the noise floor closer to the limit line, which may degrade accuracy. The EMC test engineer has to apply receiver settings which balance these two aspects, and may require different settings at different frequencies. Both sources of error should be accounted for in the measurement uncertainty budget, as discussed in section 6.5.

6.1.3.6 Other measuring instruments

Instruments have appeared on the market which fulfil some of the functions of a spectrum analyser or receiver at a much lower price. These may be units which convert an oscilloscope into a spectrum display, or which act as add-ons to a PC that performs the majority of the signal processing and display functions. Such devices are useful for diagnostic purposes provided that you recognize their limitations – typically frequency range, stability, bandwidth and/or sensitivity. The major part of the cost of a spectrum analyser or receiver is in its bandwidth-determining filters and its local oscillator. Cheap versions of these simply cannot give the performance that is needed of an accurate measuring instrument.

Even for diagnostic purposes, frequency stability and accuracy are necessary to make sense of spectrum measurements, and the frequency range must be adequate (150kHz–30MHz for conducted, 30MHz–1GHz for radiated diagnostics). Sensitivity matching that of a spectrum analyser will be needed if you are working near to the emission limits. The inflexibility of the cheaper units soon becomes apparent when you want to make detailed tests of particular emission frequencies.

6.2 Transducers

For any RF emissions measurement you need a device to couple the measured variable into the input of the measuring instrumentation. Measured variables take one of four forms:

- radiated electric field
- radiated magnetic field
- conducted cable voltage
- conducted cable current

and the transducers for each of these forms are discussed below.

6.2.1 Antennas for radiated field

6.2.1.1 VHF-UHF antennas

The basics of electromagnetic fields are outlined in section 10.1.4.1. Radiated field measurements can be made of either electric (E) or magnetic (H) field components. In the far field the two are equivalent, and related by the impedance of free space:

$$E/H = Z_o = 120\pi = 377\Omega \tag{6.3}$$

but in the near field their relationship is complex and generally unknown. In either case, an antenna is needed to couple the field to the measuring receiver. Electric field strength limits are specified in terms of volts (or microvolts) per metre at a given distance from the EUT, whilst measuring receivers are calibrated in volts (or microvolts) at the 50Ω input. The antenna must therefore be calibrated in terms of volts output into 50Ω for a given field strength at each frequency; this calibration is known as the *antenna factor*.

CISPR 16-1-4 defines transducers for radiated measurements. Historically its reference antenna has been a tuned dipole, but it also allows the use of broadband antennas, which remove the need for retuning at each frequency. Up until the mid-1990s the two most common broadband devices were the biconical, for the range 30–300MHz, and the log periodic, for the range 300–1000MHz. Some antennas have different frequency ranges.

However, it is possible to combine a biconical and a log periodic to cover the range 30–1000MHz or even wider. The two structures have been amalgamated with a means of ensuring that the feed is properly defined over the whole frequency range, and this type, originally designed at York University in the UK, is now available commercially. It is known, unsurprisingly, as the BiLog. Its major advantage, particularly appreciated by test houses, is that an entire radiated emissions (or immunity) test can be done without changing antennas, with a consequent improvement in speed and reliability. This has meant that the type is now almost universally used and versions are available from all the main EMC antenna manufacturers.

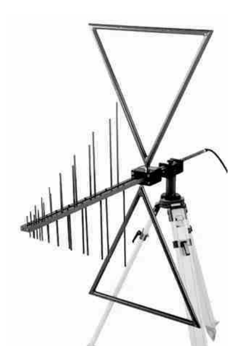


Figure 6.4 The BiLog (courtesy Schaffner)

The advantage of the tuned dipole is that its performance can be accurately predicted, but because it can only be applied at spot frequencies it is not used for everyday measurement but is reserved for calibration of broadband antennas, site surveys, site attenuation measurements and other more specialized purposes.

Antenna factor

Those who use antennas for radio communication purposes are familiar with the specifications of gain and directional response, but these are of only marginal importance for EMC emission measurements. The antenna is always oriented for maximum response. Antenna factor is the most important parameter, and each calibrated broadband antenna is supplied with a table of its antenna factor (in dB/m, for E-field antennas) versus frequency. Antenna calibration is treated in more detail in section 6.5.2.3. Typical antenna factors for a biconical, a log periodic and two varieties of BiLog are shown in Figure 6.5. From this you can see that there is actually very little difference for the log periodic section (above 300MHz) in particular: any LP design with the same dimensions will give substantially the same performance. The biconical (30–300MHz) section can be "tweaked" but again, antennas with the same basic dimension give largely similar performance. This has the particular consequence that if all labs use pretty much the same design of antenna (which they do), the inter-lab variations in radiated field measurement due to the antenna itself are minimized.

To convert the measured voltage at the instrument terminals into the actual field strength at the antenna you have to add the antenna factor and cable attenuation (Figure 6.6). Cable attenuation is also a function of frequency; it can normally be regarded as constant with time, although long cables exposed to wide temperature variations, such as on open sites, may suffer slight variations of loss with temperature.

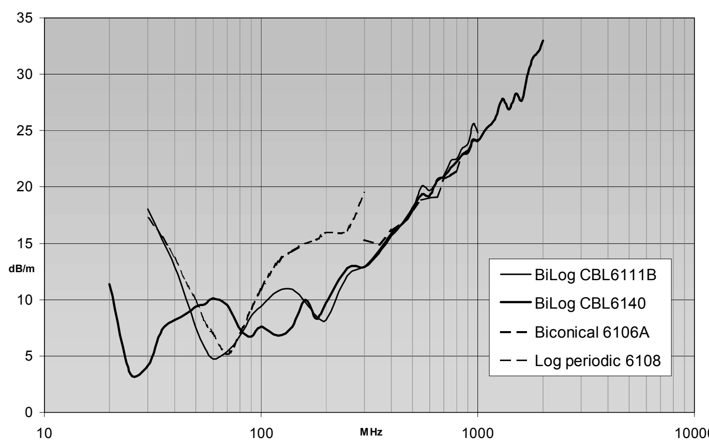


Figure 6.5 Typical antenna factors (courtesy Schaffner)

System sensitivity

A serious problem can arise when using an antenna with a spectrum analyser for radiated tests. Radiated emission compliance tests can be made at 10m distance and the most severe limit in the usual commercial standards is the EN Class B level, which is 30dBμV/m below 230MHz and 37dBμV/m above it. The minimum measurable level will be determined by the noise floor of the receiver or analyser (see section 6.1.3.1), which for an analyser with 120kHz bandwidth is typically +13dBμV. To this must be added the antenna factor and cable attenuation in order to derive the overall measurement system sensitivity. Taking the antenna factors already presented, together with a typical 3dB at 1GHz due to cable attenuation, the overall system noise floor rises to 41dBμV/m at 1GHz as shown in Figure 6.7, which is 4dB *above* the limit line.

The CISPR 16-1-1 requirement on sensitivity is that the noise contribution should affect the accuracy of a compliant measurement by less than 1dB. This implies a noise floor that is below the measured value by at least 6dB.

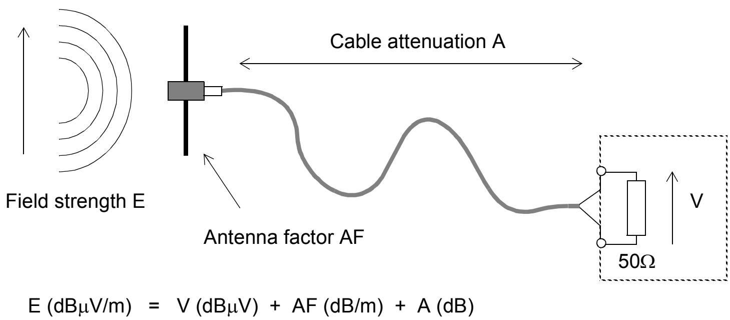


Figure 6.6 Converting field strength to measured voltage

Thus full radiated Class B compliance measurements *cannot be made with a spectrum analyser alone*. Three options are possible: reduce the measuring distance to 3m, which may raise the limit level by 10dB, but this increases the measurement uncertainty and still gives hardly enough margin at the top end; or, use a preamplifier or preselector to lower the effective system noise floor, by a factor equal to the preamp gain less its noise figure, typically 20–25dB; or use a test receiver, which has a much better inherent sensitivity.

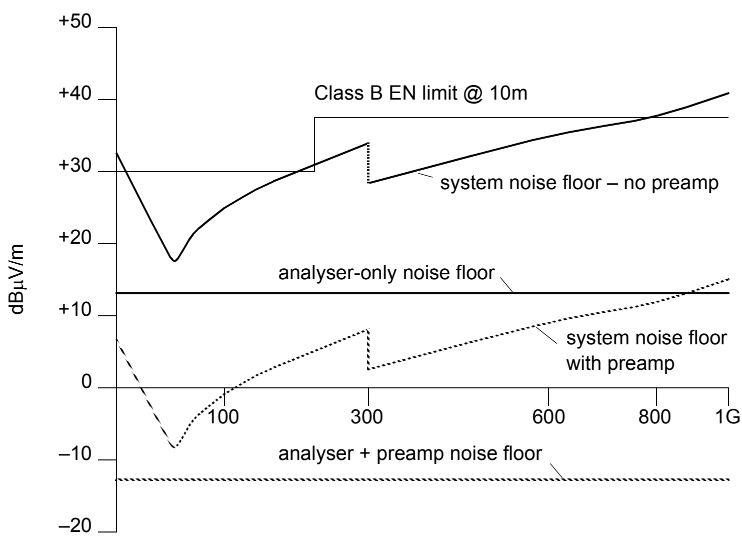


Figure 6.7 System sensitivity

Polarization

In the far field the electric and magnetic fields are orthogonal (Appendix D, section D.3.8). With respect to the physical environment each field may be vertically or horizontally polarized, or in any direction in between. The actual polarization depends on the nature of the emitter and on the effect of reflections from other objects. An antenna will show a maximum response when its plane of polarization aligns with that of the incident field, and will show a minimum when the planes are at right angles. The plane of polarization of biconical, log periodic and BiLog is in the plane of the elements. CISPR emission measurements must be made with "substantially plane polarized" antennas; circularly-polarized antennas, such as the log spiral, a broadband type once favoured for military RF immunity testing, are outlawed.

6.2.1.2 The loop antenna

The majority of radiated emissions are measured in the range 30 to 1000MHz. A few CISPR standards call for radiated measurements below 30MHz. In these cases the magnetic field strength is measured, using a loop antenna. Measurements of the magnetic field give better repeatability in the near field region than do measurements of the electric field, which is easily perturbed by nearby objects. The loop (Figure 6.8(a)) is merely a coil of wire which produces a voltage at its terminals proportional to frequency, according to Faraday's law:

$$E = 4\pi \cdot 10^{-7} \cdot N \cdot A \cdot 2\pi F \cdot H \tag{6.4}$$

where N is the number of turns in the loop
A is the area of the loop, m^2
F is the measurement frequency, Hz
H is the magnetic field, Amps/metre

The low impedance of the loop does not match the 50Ω impedance of typical test instrumentation. Also, the frequency dependence of the loop output makes it difficult to measure across more than three decades of frequency, typically 9kHz to 30MHz. Passive loops deal with this latter problem by switching in different numbers of turns to cover smaller sub-ranges in frequency, but naturally this does not lend itself to test automation.

These disadvantages are overcome by including as part of the antenna a preamplifier which corrects for the frequency response and matches the loop output to 50Ω. The preamp can be battery powered or powered from the test instrument. Such an "active" loop has a flat antenna factor across its frequency range. Its disadvantage by comparison with a passive loop is that it can be saturated by large signals, and some form of overload indication is needed to warn of this.

The Van Veen loop

A disadvantage of the loop antenna as it stands is its lack of sensitivity at low frequencies. An alternative method [32] is to actually surround the EUT with the loop; in its practical realization, three orthogonal loops of 2–4m diameter are used with the current induced in each being sensed by a current transformer, and the three signals are measured in turn by the test receiver. This is the large loop antenna (LLA) or Van Veen loop, named after its inventor, and it is specified in EN 55015, the standard for lighting equipment.

6.2.1.3 The electric monopole

The complementary antenna to the loop is the monopole (Figure 6.8(b)). Covering the frequency range again typically up to 30MHz, the monople is simply a single vertical rod of length 1m referenced against a ground plane or "counterpoise", and it measures the electric field in vertical polarization. It's not used in many CISPR-based tests, only

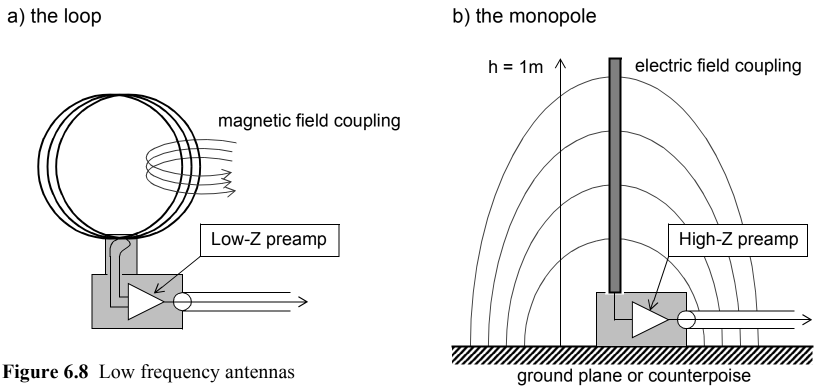


Figure 6.8 Low frequency antennas

the automotive emissions standards CISPR 12 and CISPR 25 call it up, but it is used fairly widely in military testing to DEF STAN 59-41 and its American equivalent, MIL STD 461E. These tests require low frequency measurement of both the E-field and the H-field strengths.

The monopole is electrically short (its length is much less than a wavelength) and its source impedance looks like a capacitance of a few pF. So just as with the loop it is not suitable for connecting directly to a 50 ohm measuring system, and it should be fitted with a high impedance pre-amplifier to give impedance matching and to give a flat antenna factor. This makes it sensitive to damage and to electrical overload by large signals during the measurement – including the all-pervasive 50Hz mains E-field – so again it needs an overload indicator, and a high-pass filter to remove the mains field. Because the near electric field can be affected by the presence of virtually any conducting object, the accuracy and repeatability of measurements made with this antenna is poor even by the standards of EMC testing.

6.2.2 LISNs and probes for cable measurements

6.2.2.1 Artificial mains network

To make conducted voltage emissions tests on the mains port, you need an Artificial Mains Network (AMN) or Line Impedance Stabilizing Network (LISN) to provide a defined impedance at RF across the measuring point, to couple the measuring point to the test instrumentation and to isolate the test circuit from unwanted interference signals on the supply mains. The most widespread type of LISN is defined in CISPR 16-1-2 and presents an impedance equivalent to 50Ω in parallel with 50μH + 5Ω across each line to earth (Figure 6.9). This is termed a “V-network” since for a single-phase supply the impedance appears across each arm of the V, where the base of the V is the reference earth.

Note that its impedance is not defined above 30MHz, partly because commercial mains conducted measurements are not required above this frequency (although military and automotive standards are) but also because component parasitic reactances make a predictable design difficult to achieve. An alternative 50Ω/5μH network is available in the CISPR specification, and a similar version is widely used in military and automotive tests according to DEF STAN 59-41 and MIL STD 461E. Since it uses a smaller inductor, it can carry higher currents and its impedance can be controlled up

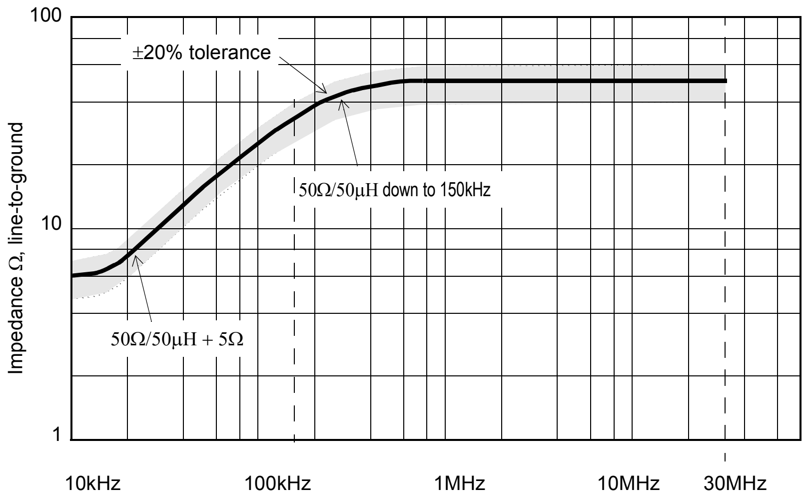


Figure 6.9 LISN impedance versus frequency

to 100MHz and beyond. The military LISN's impedance is specified down to 1kHz and up to 400MHz. It is principally intended to simulate DC supplies but can also be used for mains tests when higher current ratings, typically above 50A, are needed.

CISPR 16-1-2 includes a suggested circuit (Figure 6.10) for each line of the LISN, but it only actually *defines* the impedance characteristic. The main impedance determining components are the measuring instrumentation input impedance, the 50μH inductor and the 5Ω resistor. The remaining components serve to decouple the incoming supply. The 5Ω resistor is only effective at the bottom end of the frequency range, and in fact a "cut-down" version of LISN is defined which omits it and the 250μH inductor but is restricted to frequencies above 150kHz. Most commercial LISNs, though, include the whole circuit and cover the range down to 9kHz. A common addition is a high-pass filter between the LISN output and the receiver, cutting off below 9kHz, to prevent the receiver from being affected by high-level harmonics of the mains supply itself. Of course, this filter has to maintain the 50Ω impedance and have a defined (preferably 0dB) insertion loss at the measured frequencies.

An extension to the LISN specification is underway at the time of writing in the form of an amendment to CISPR 16-1-2 [167]. This tightens various aspects of the impedance requirement, in particular that it should have a phase angle tolerance of better than ±11.5° across the whole frequency range, even up to 30MHz. This new requirement will be difficult to meet for some existing LISNs, particularly larger ones at the high frequency end, because of stray capacitance. The justification for it is that if you are going to calculate a complete measurement uncertainty budget for the conducted emissions test, a boundary needs to be placed on the possible range of impedance in both magnitude and phase presented by the LISN, otherwise the uncertainty due to impedance variations is unknown. It seems, though, debatable whether the greater complexity in calibration will be worthwhile when other, larger sources of uncertainty in what is already a reasonably well-specified test remain unaddressed (section 6.5.2.2).

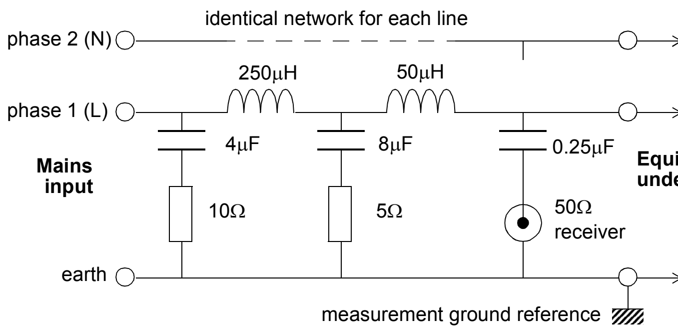


Figure 6.10 LISN circuit per line

Current rating

For the user, the most important parameter of the LISN is its current rating. Since the inductors are in series with the supply current, their construction determines how much current can be passed; if they are air cored, then magnetic saturation is not a problem but the coils may still overheat with too much current. Most LISNs use air cored coils, but if magnetic cores are used (typically iron powder) for a smaller construction then you also need to be concerned about saturation, which will affect the impedance characteristic of the device.

Another effect of the impedance of the inductors is the voltage drop that may appear between the supply feed and the EUT terminals. At 50Hz the total in-line inductance of 600μH gives an impedance of about 0.2Ω, which may itself give a significant voltage drop, added to the resistance through the unit, at high currents; but if, as may happen with an electronic power supply, your unit draws substantial harmonics of 50Hz (see section 10.4), then the inductive impedance of a few ohms at a few hundred Hz can give a much greater voltage drop with attendant waveform distortion. This in turn could affect the functional performance of the EUT.

Limiter

The spectrum analyser's input mixer is a very fragile component. As well as being affected by high-level continuous input signals it is also susceptible to transients. Unfortunately the supply mains is a fruitful source of such transients, which can easily exceed 1kV on occasion. These transients are attenuated to some extent by the LISN circuitry but it cannot guarantee to keep them all within safe limits. More importantly, switching operations within the EUT itself are likely to generate large transients due to interruption of current through the LISN chokes, and these are fed directly to the analyser without attenuation.

For this reason it is essential to include a transient limiter in the signal cable between the LISN and the spectrum analyser. This adds an extra 10dB loss (typically) in the signal path which must be added to the LISN's own transducer factor, since the limiting devices need to be fed from a well-defined impedance, but this can normally be tolerated and is a much cheaper option than expensive repair bills for the analyser front end. The limiter circuit generally uses a simple back-to-back diode clipping scheme; some limiters also incorporate a filter to restrict the frequency range transmitted to the analyser.

A limiter is less necessary, though still advisable, when a measuring receiver is used since the receiver's front end should be already protected. The limiter does have one particular danger: low frequency signals that are outside the measurement range and therefore not subject to limits may legitimately have amplitudes of volts, which will drive the limiter into continuous clipping and create harmonics which are then incorrectly measured as in-band signals. If you suspect such an eventuality, be prepared to place extra attenuation or high-pass filtering before the limiter to check.

Earth current

A large capacitance (in total around 12μF) is specified between line and earth, which when exposed to the 230V line voltage results in around 0.9A in the safety earth. This level of current is lethal, and the unit must therefore be solidly connected to earth for safety reasons. If it is not, the LISN case, the measurement signal lead and the equipment under test (EUT) can all become live. As a precaution, you are advised to bolt your LISN to a permanent ground plane and not allow it to be carried around the lab! A secondary consequence of this high earth current is that LISNs cannot be used directly on mains circuits that are protected by earth leakage or residual current circuit breakers. Both of these problems can be overcome by feeding the mains to the LISN through an isolating transformer, provided that this is sized adequately, bearing in mind the extra voltage drop through the LISN itself.

Diagnostics with the LISN

As it stands, the LISN does not distinguish between differential mode (line-to-line) and common mode (line-to-earth) emissions (see section 10.2.2); it merely connects the measuring instrument between phase and earth. A modification to the LISN circuit (Figure 6.11) allows you to detect either the sum or the difference of the live and neutral voltages, which correspond to the common mode and differential mode voltages respectively [111]. This is not required for compliance measurements but is very useful when making diagnostic tests on the mains port of a product. Transformers TX1 and 2, which can be nothing more than 50 ohm 1:1 broadband balun parts, are switched to add or subtract the two signals.

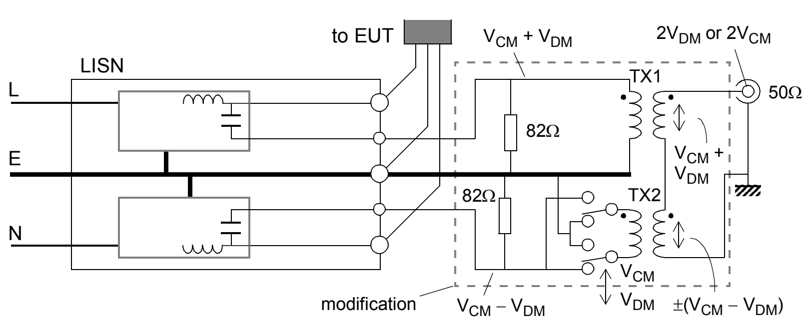


Figure 6.11 Modifying the LISN to measure differential or common mode

6.2.2.2 *Absorbing clamp and CMAD*

As well as measuring the emissions above 30MHz directly as a radiated field you can also measure the disturbances that are developed in common mode on connected cables. Standards which apply primarily to small apparatus connected only by a mains cable – notably EN 55014-1 – specify the measurement of interference power present on the mains lead. This has the advantage of not needing a large open area for the tests, but it should be done inside a fairly large screened room and the method is somewhat clumsy. The transducer is an absorbing device known as a ferrite clamp.

The ferrite absorbing clamp (often referred to as the MDS-21 clamp, and different from the EM-clamp used in immunity tests despite its similar appearance) consists of a current transformer using two or three ferrite rings, split to allow cable insertion, with a coupling loop (Figure 6.12). This is backed by further ferrite rings forming a power absorber and impedance stabilizer, which clamps around the mains cable to be measured. The device is calibrated in terms of output power versus input power, i.e. insertion loss. The purpose of the ferrite absorbers is to attenuate reflections on the lead under test downstream of the current transformer; this is not 100% effective, and a full compliance test requires the clamp to be traversed for a half wavelength along the cable, i.e. 5m, to find a maximum. The lead from the current transformer to the measuring instrument is also sheathed with ferrite rings to attenuate screen currents on this cable. Because the output is proportional to current flowing in common mode on the measured cable, it can be used as a direct measure of noise power, and the clamp can be calibrated as a two-port network in terms of output power versus input power.

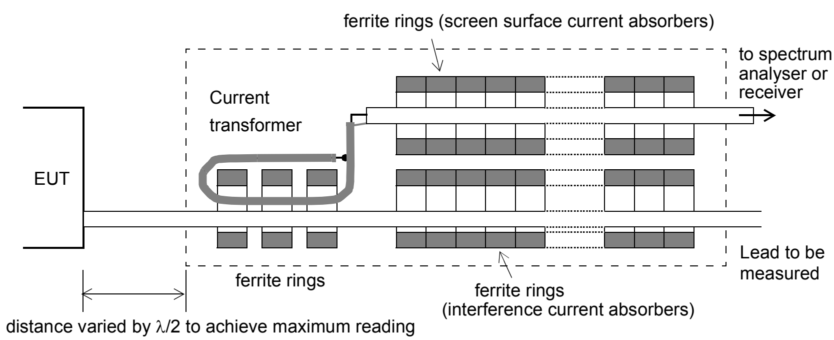


Figure 6.12 The ferrite absorbing clamp

CISPR 16-1-3 specifies the construction, calibration and use of the ferrite clamp. As well as its use in certain compliance tests, it also lends itself to diagnostics as it can be used for repeatable comparative measurements on a single cable to check the effect of circuit changes. It has been suggested [127] that the clamp should be used to "pre-test" small EUTs, with the application of a suitable empirically-derived correction factor, so that they need only be subjected to a radiated test at the most critical emitting frequencies, but this idea hasn't been pursued within the standards.

The CMAD

A further common use for the clamp is to be applied to the further end of connected cables, both in radiated emissions and immunity tests, to damp the cable resonance and reduce variations due to cable termination. The clamp output is not connected when it is used in this way. Although this is a convenient method if a clamp is already to hand, using a string of 6–10 large snap-on ferrite sleeves is almost as effective. Alternatively, you can obtain commercially a unit known as a "Common Mode Absorbing Device"

(CMAD) which is the same thing. Amendment A1 to the third edition of CISPR 22 specified such a device to be applied to cables leaving the radiated emissions test site, whose purpose was simply to stabilize the far-end cable impedance and therefore make for a more repeatable test.

Despite this worthy aim, the amendment sparked such a howl of protest that it was abandoned and later editions of the standard make no reference to cable common mode impedance stabilization. Many of the reasons against using CMADs in this way demonstrated a lack of understanding of the purpose of the amendment: for instance it was claimed that such clamps would never be used in real installations and therefore the test would not be representative; but there is really no way that such test set-ups can ever be both truly representative *and* repeatable. The merit of the CMAD is that it would ensure a high impedance at a fixed distance from the EUT and therefore improve repeatability, and at the same time represent one installation condition where such an impedance did occur.

A more reasonable objection to the amendment was that at the time, no method was published for verifying the impedance of the CMAD. This is now being addressed and a draft amendment to CISPR 16-1-4 is in progress. This author is wholeheartedly in favour of using some kind of CMAD to control the cable impedances in the radiated tests; even if the compliance test stubbornly refuses to address the issue, your pre-compliance measurements will be easier to repeat if you do.

6.2.2.3 Current probe

Also useful for diagnostics is the current probe, which does the same thing as the absorbing clamp except that it doesn't have the absorbers. It is simply a clamp-on, calibrated wideband current transformer. Military specifications call for its use on individual cable looms, and the third edition of EN 55022 giving test methods for telecoms ports also requires a current probe for some versions of the tests. CISPR 16-1 includes a specification for the current probe. Because the current probe does not have an associated absorber, the RF common mode termination impedance of the line under test should be defined by an impedance stabilizing network, which must be transparent to the signals being carried on the line.

Both the ferrite clamp and the current probe have the great advantage that no direct connection is needed to the cable under test, and disturbance to the circuit is minimal below 30MHz since the probe effect is no more than a slight increase in common mode impedance. But at higher frequencies the effect of the common mode coupling capacitance between probe and cable becomes significant, as does the exact position of the probe along the cable, because of standing waves on the line. Your test plan and report should note this position exactly, along with the method of bonding the current probe case to the ground plane, which controls the stray capacitance.

6.2.2.4 ISNs and other methods for telecom port conducted emissions

The third edition of EN 55022 has provisions for tests on telecommunications ports. The preferred method uses a particular variant of impedance stabilizing network (ISN) which is designed to mimic the characteristics of balanced, unscreened ISO/IEC 11801 data cables. This network has a common mode impedance of 150Ω and a carefully controlled longitudinal conversion loss (LCL, see section 13.1.9.1), that is, the parameter which determines the conversion from differential mode signal to common mode interference currents. Both current and voltage limits are published, related by the 150Ω impedance.

The telecom port test has been controversial since the third edition was published. The basic problem is that, since it applies to ports such as Local Area Network (LAN) interfaces, the signal that is intended to be passed through the port – data up to, say, 100Mb/s – is in the same frequency range as the interference to be measured. The wanted signal is in differential mode while the interference is in common mode. Therefore, for a given data amplitude as determined by the network in use, such as Ethernet, the interference you measure is at least partly determined by the LCL of the network that is used for the measurement to represent a particular type of cable. There will be two components to the measurement:

- the wanted data converted from differential to common mode by the LCL;
- any extraneous common mode noise added by imperfections in the design of the port.

Of course, the second of these must be controlled, and so the test is necessary, but the first should not be allowed to spoil the measurement unnecessarily. Hence a need for a very careful specification of the LCL of the impedance stabilizing network, since all other parameters are invariant. The initial version of the standard paid insufficient attention both to this question and to the proper calibration of the ISN, and it took some time for the issue to be sorted out. The fifth edition of EN 55022 (see section 4.3.3) has apparently corrected the problems and its values of LCL are shown graphically in Figure 6.13, but in the meantime earlier versions were not acceptable for full compliance purposes in respect of the telecom port test, and for that reason their dates of mandatory implementation kept being postponed.

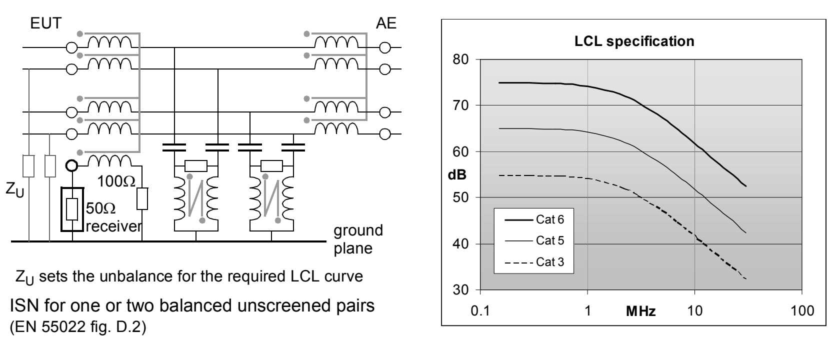


Figure 6.13 An example ISN circuit and the LCL specification

As well as using ISNs with specified parameters for signal lines with balanced, unscreened pairs, the telecom port test is also applied to signal lines that use other kinds of cable. It is not always possible to specify an ISN for every one of these, and so alternative and preferably non-invasive methods of measurement have been and still are needed. These are described in Annex C of the standard and include:

- using alternative ISNs, such as the CDNs used for conducted immunity tests to IEC 61000-4-6 (section 7.1.4), as long as the EUT can operate normally with this inserted, and as long as the CDN has a calibrated minimum LCL (C.1.1);

- for shielded cables, using a 150Ω load to the outside surface of the shield in conjunction with a ferrite decoupler (C.1.2);
- using a combination of current probe and capacitive voltage probe, and comparing the result to both current and voltage limits (C.1.3);
- using a current probe only, but with the common mode impedance on the ancillary equipment side of the probe explicitly set to 150Ω at each test frequency with a ferrite decoupler (C.1.4).

The various methods are illustrated in Figure 6.14. In the fifth edition of EN 55022 improvements have been made over the original, although even so the method of C.1.4 is so cumbersome that the standard itself says “If the method in C.1.4 is combined with the method of C.1.3, it is possible to use the advantages of both methods, without suffering too much from the disadvantages” and in fact it is wise to avoid it if at all possible. The EMC Test Laboratories Association published a Technical Guidance Note (TGN42, [205]) dealing with the various difficulties and questions raised by the test as originally published, in an attempt to help labs to arrive at a common interpretation of the methods.

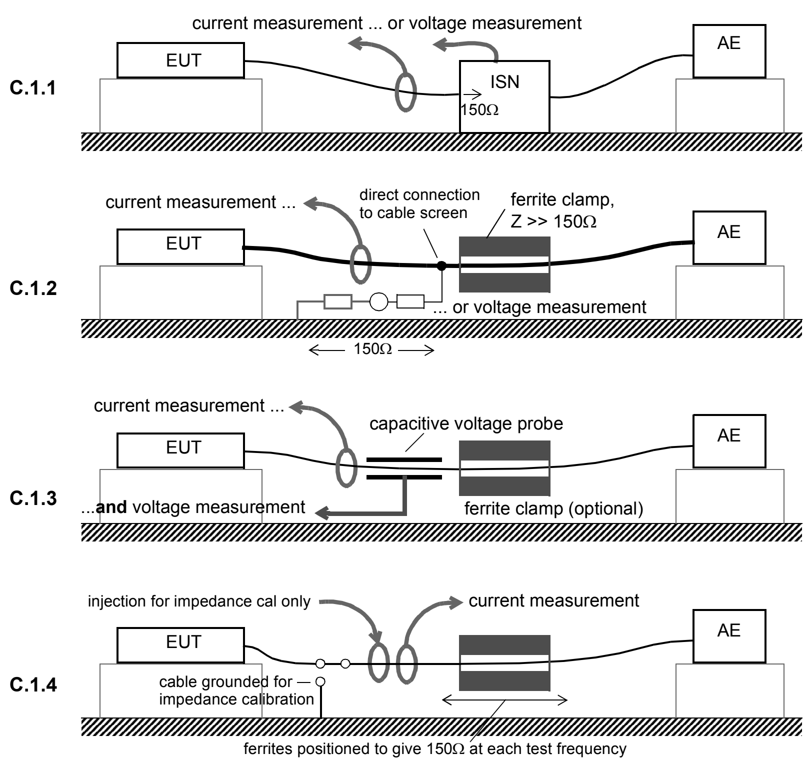


Figure 6.14 Alternative telecom port measurement methods

6.2.3 Near field probes

Very often you will need to physically locate the source of emissions from a product. A set of near field (or "sniffer") probes is used for this purpose. These are so-called because they detect field strength in the near field, and therefore two types of probe are needed, one for the electric field (rod construction) and the other for the magnetic field (loop construction). It is simple enough to construct adequate probes yourself using coax cable (Figure 6.15), or you can buy a calibrated set. A probe can be connected to a spectrum analyser for a frequency domain display, or to an oscilloscope for a time domain display.

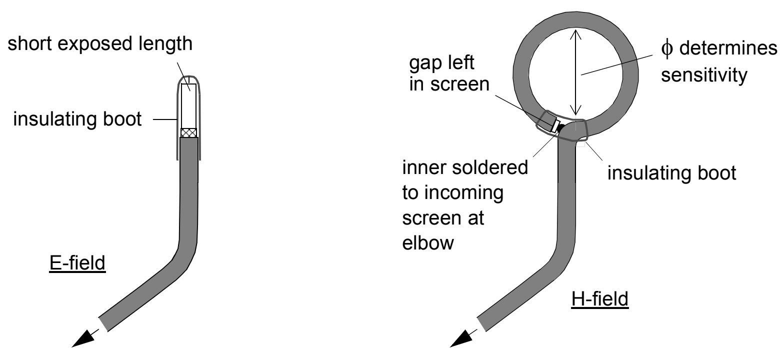


Figure 6.15 Do-it-yourself near field probes

Probe design is a trade-off between sensitivity and spatial accuracy. The smaller the probe, the more accurately it can locate signals but the less sensitive it will be. You can increase sensitivity with a preamplifier if you are working with low-power circuits. A good magnetic field probe is insensitive to electric fields and vice versa; this means that an electric field probe will detect nodes of high dv/dt but will not detect current paths, while a magnetic field probe will detect paths of high di/dt but not voltage points.

Near field probes can be calibrated in terms of output voltage versus field strength, but these figures should be used with care. Measurements cannot be directly extrapolated to the far field strength (as on an open site) because in the near field one or other of the E or H fields will dominate depending on source type. The sum of radiating sources will differ between near and far fields, and the probe will itself distort the field it is measuring. Perhaps more importantly, you may mistake a particular "hot spot" that you have found on the circuit board for the actual radiating point, whereas the radiation is in fact coming from cables or other structures that are coupled to this point via an often complex path. Probes are best used for tracing and for comparative rather than absolute measurements.

6.2.3.1 Near field scanning devices

A particular implementation of a near field probe is the planar scanning device, such as EMSCAN. This was developed and patented at Bell Northern Research in Canada and is now marketed by EmScan Corporation of Canada, as well as having competitors from other manufacturers. In principle it is essentially a planar array of tiny near field current probes arranged in a grid form on a multilayer PCB [65]. The output of each current probe can be switched under software control to a frequency selective measuring instrument, whose output in turn provides a graphical display on the controlling workstation. An alternative method uses a single probe positioned by X-Y stepper motors, in the same way as an X-Y plotter.

The device is used to provide a near-instantaneous two-dimensional picture of the RF circulating currents within a printed circuit card placed over the scanning unit. It can provide either a frequency versus amplitude plot of the near field at a given location on the board, or an x-y co-ordinate map of the current distribution at a given frequency. For the designer it can quickly show the effect of remedial measures on the PCB being investigated, while for production quality assurance it can be used to evaluate batch samples which can be compared against a known good standard.

6.2.4 The GTEM for emissions tests

Use of the GTEM for radiated RF immunity testing is covered in section 7.1.1.5. It can also be used for emissions tests with some caveats. The GTEM is a special form of enclosed TEM (transverse electromagnetic mode) transmission line which is continuously tapered and terminated in a broadband RF load. This construction prevents resonances and gives it a flat frequency response from DC to well beyond 1GHz. An EUT placed within the transmission line will couple closely with it, and its radiated emissions can be measured directly at the output of the cell. Since far field conditions also describe a TEM wave, any test environment that provides TEM wave propagation should be acceptable as an alternative. The great advantages of this technique are that no antenna or test site is needed, the frequency range can be covered in a single sweep, and ambients are eliminated.

However, compliance tests demand a measure of the radiated emissions as they would be found on an OATS (see next section). This requires that the GTEM measurements are correlated to OATS results. This is done in software and the model was originally described in [143]. In fact, three scans are done with the EUT in orthogonal orientations within the cell. The software then derives at each frequency an equivalent set of elemental electric and magnetic dipole moments, and then re-calculates the far field radiation at the appropriate test distance from these dipoles.

The limitation of this model is that the EUT must be "electrically small", i.e. its dimensions are small when compared to a wavelength. Connected cables pose a particular problem since these often form the major radiating structure, and are of course rarely electrically small, even if the EUT itself is. Good correlation has been found experimentally for small EUTs without cables [41][109], but the correlation worsens as larger EUTs, or EUTs with connected cables, are investigated.

There is now a standard for measurements of both emissions and immunity in TEM cells, IEC 61000-4-20 [155]. This expands on the OATS correlation and gives a detailed description of the issues involved in using any kind of TEM cell, including the GTEM, for radiated tests, but its limitation is apparent in Clause 6:

> **6.1 Small EUT**
>
> An EUT is defined as a small EUT if the largest dimension of the case is smaller than one wavelength at the highest test frequency (for example, at 1 GHz λ = 300 mm), and if no cables are connected to the EUT. All other EUTs are defined as large EUTs.
>
> **6.2 Large EUT**
>
> An EUT is defined as a large EUT if it is
>
> • a small EUT with one or more exit cables,

• a small EUT with one or more connected non-exit cables,

• an EUT with or without cable(s) which has a dimension larger than one wavelength at the highest test frequency,

• a group of small EUTs arranged in a test set-up with interconnecting non-exit cables, and with or without exit cables.

For emissions tests, it then goes on to state that "Large EUTs and specific considerations regarding EUT arrangements and cabling are deferred for elaboration in the next edition of this standard", in other words virtually all real EUTs cannot be treated under the current version.

In practice the GTEM is a useful device for testing small EUTs without cables, and can be applied in a number of specialized applications such as for measuring direct emissions from integrated circuits. But, although IEC 61000-4-20 has been published, it does not specify the tests to be applied to any particular apparatus or system. Instead it is meant to provide a general basic reference for all interested product committees, and its popularity will depend on it being referenced widely in conventional product emissions standards. The OATS correlation is limited in its full applicability to only a few types of EUT, and the likelihood that the GTEM will inhabit more than a restricted niche in EMC testing is small. Meanwhile, the National Physical Laboratory Electromagnetic Metrology Group has published a Best Practice Guide [104] for the GTEM which expands on the practicalities of IEC 61000-4-20 measurements. Anybody who uses a GTEM should have both of these documents as a reference.

6.3 Sites and facilities

6.3.1 Radiated emissions

6.3.1.1 The CISPR OATS

For CISPR-based standards, the reference for radiated emissions compliance testing is an Open Area Test Site (OATS). The characteristics of a minimum standard OATS are defined in EN 55022 and CISPR 16-1-4 and some guidance for construction is given in ANSI C63.7 [207]. Such a site offers a controlled RF attenuation characteristic between the emitter and the measuring antenna (known as "site attenuation"). To avoid influencing the measurement there should be no objects that could reflect RF within the vicinity of the site. The original CISPR test site dimensions are shown in Figure 6.16.

The ellipse defines the area which must be flat and free of reflecting objects, including overhead wires. In practice, for good repeatability between different test sites a substantially larger surrounding area free from reflecting objects is advisable. This means that the room containing the control and test instrumentation needs to be some distance away from the site. An alternative is to put this room directly below the ground plane, either by excavating an underground chamber (as long as your site's water table will allow it) or by using the flat roof of an existing building as the test site. Large coherent surfaces will be a problem, but not smaller pieces of metal for door hinges, door knobs, light fixtures at ground level, bolts, and metal fittings in primarily non-conductive furniture. Don't put your site right next to a car park!

Ground plane

Because it is impossible to avoid ground reflections, these are regularized by the use of a ground plane. The minimum ground plane dimensions are also shown in Figure 6.16.

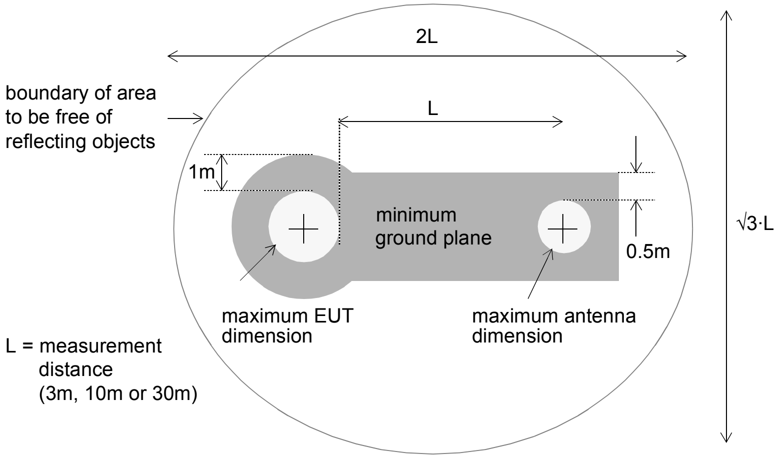


Figure 6.16 The CISPR OATS

Again, an extension beyond these dimensions will bring site attenuation closer to the theoretical; scattering from the edges contributes significantly to the inaccuracies, although these can be minimized by terminating the edges into the surrounding soil [91]. Close attention to the construction of the ground plane is necessary. It should preferably be solid metal sheets welded together, but this may be impractical over the whole area. Bonded wire mesh is suitable, since it drains easily and resists warping in high temperatures if suitably tensioned. For RF purposes it must not have voids or gaps that are greater than 0.1λ at the highest frequency (i.e. 3cm). Ordinary wire mesh is unsuitable unless each individual overlap of the wires is bonded. CISPR 16-1 suggests a ground plane surface roughness of better than 4.5cm. The surface should not be covered with any kind of lossy dielectric – floor paint is acceptable, but nothing much more than that.

Measuring distance

The measurement distance d between EUT and receiving antenna determines the overall dimensions of the site and hence its expense. There are three commonly specified distances: 3m, 10m and 30m, although 30m is rarely used in practice. In EN 55022 and related standards the measuring distance is defined between the boundary of the EUT and the reference point of the antenna. Although the limits are usually specified at 10m distance, tests can be carried out on a 3m range, on the assumption that levels measured at 10m will be 10.5dB lower (field strength should be proportional to 1/d). This assumption is not entirely valid at the lower end of the frequency range, where 3m separation is approaching the near field, and indeed experience from several quarters shows that a linear 1/d relationship is more optimistic than is found in practice. Nevertheless because of the expense of the greater distance, especially in an enclosed chamber, 3m measurements are widely used.

Weather proofing

The main environmental factor that affects open area emissions testing, particularly in Northern European climates, is the weather. Some weatherproof but RF-transparent structure is needed to cover the EUT to allow testing to continue in bad weather. The structure can cover the EUT alone, for minimal cost, or can cover the entire test range; a half-and-half solution is sometimes adopted, where a 3m range is wholly covered but the ground plane extends outside and the antenna can be moved from inside to out for a full 10m test. Fibreglass is the favourite material. Wood is not preferred, because the reflection coefficient of some grades of wood is surprisingly high [93] and varies with moisture content, leading to differences in site performance between dry and wet weather. You may need to make allowance for the increased reflectivity of wet surfaces during and after precipitation, and a steep roof design which sheds rain and snow quickly is preferred.

6.3.1.2 Validating the site: NSA

Site attenuation is the insertion loss measured between the terminals of two antennas on a test site, when one antenna is swept over a specified height range, and both antennas have the same polarization. This gives an attenuation value in dB at each frequency for which the measurement is performed. Transmit and receive antenna factors are subtracted from this value to give the Normalized Site Attenuation (NSA), which should be an indication only of the performance of the site, without any relation to the antennas or instrumentation.

NSA measurements are performed for both horizontal and vertical polarizations, with the transmit antenna positioned at a height of 1m (for broadband antennas) and the receive antenna swept over the appropriate height scan. CISPR standards specify a height scan of 1 to 4m for 3m and 10m sites. The purpose of the height scan, as in the test proper (section 6.4.2), is to ensure that nulls caused by destructive addition of the direct and ground reflected waves are removed from the measurement. Note that the height scan is *not* intended to measure or allow for elevation-related variations in signal emitted directly from the source, either in the NSA measurement or in a radiated emissions test.

Figure 6.17 shows the geometry and the basic method for an NSA calibration. Referring to that diagram, the procedure is to record the signal with points [1] and [2] connected, to give V_{DIRECT}, and then via the antennas over the height scan, to give V_{SITE}. Then NSA in dB is given by

$$NSA = V_{DIRECT} - V_{SITE} - A_{FT} - A_{FR} \qquad (6.5)$$

where A_{FT} and A_{FR} are the antenna factors

CISPR 16-1-4, and related standards, includes the requirement that

> A measurement site shall be considered acceptable when the measured vertical and horizontal NSAs are within ±4dB of the theoretical normalized site attenuation

and goes on to give a table of theoretical values versus frequency for each geometry and polarization (the values differ between horizontal and vertical because of the different ground reflection coefficients). This is the yardstick by which any actual site is judged; the descriptions in 6.3.1.1 above indicate how this criterion might be achieved, but as long as it *is* achieved, any site can be used for compliance purposes. Vice versa, a site which does not achieve the ±4dB criterion cannot be used for compliance purposes no matter how well constructed it is. Note that the deviation from the theoretical values

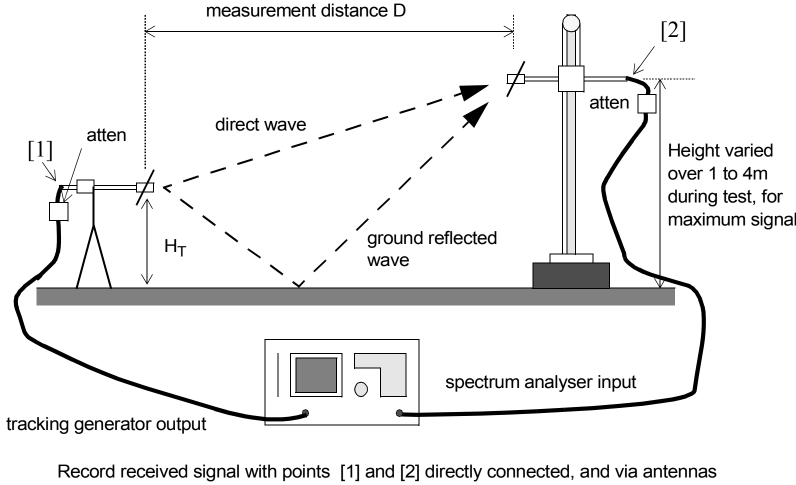


Figure 6.17 Geometry and set-up for an NSA measurement

cannot be used as a "correction factor" to "improve" the performance of a particular site. This is because the NSA relates to a specific emitting source, and the site attenuation characteristics for a real equipment under test may be quite different.

In choosing the ±4dB criterion, it is assumed by CISPR that the instrumentation uncertainties (due to antenna factors, signal generator and receiver, cables, etc.) account for three-quarters of the total and that the site itself can be expected to be within ±1dB of the ideal. This has a crucial implication for the method of carrying out an NSA measurement. If you reduce the instrumentation uncertainties as far as possible you can substantially increase the chances of a site being found acceptable. Conversely, if your method has greater uncertainties than allowed for in the above table, *even a perfect site will not meet the criterion*. Clearly, great attention must be paid to the method of performing an NSA calibration. The important aspects are:

- antenna factors – suitable for the geometry of the method, not free-space;
- antenna balance and cable layout – to minimize the impact of the antenna cables;
- impedance mismatches – use attenuator pads on each antenna to minimize mismatch error.

6.3.1.3 Radiated measurements in a screened chamber

Open area sites have two significant disadvantages, particularly in a European context – ambient radiated signals, and bad weather. Ambients are discussed again in section 6.5.2.5. These disadvantages create a preference for using sheltered facilities, and in particular screened chambers.

Alternative sites to the standard CISPR open area test site are permitted provided that errors due to their use do not invalidate the results. As you might expect, their adequacy is judged by performing an NSA measurement. However, an extra requirement is added, which is to insist that the NSA is checked over the *volume* to be occupied by the largest EUT. This can require up to 20 separate NSA sweeps – five positions in the horizontal plane (centre, left, right, front and back) with two heights and for two polarizations each. As before, the acceptability criterion is that none of the measurements shall exceed ±4dB from the theoretical.

The problem with untreated screened chambers for radiated measurements is that reflections occur from all six surfaces and will substantially degrade the site attenuation from EUT to measuring antenna [80]. For any given path, significant nulls and peaks with amplitude variations easily exceeding 30dB will exist at closely-spaced frequency intervals. Equally importantly, different paths will show different patterns of nulls and peaks, and small changes within the chamber can also change the pattern, so there is no real possibility of correcting for the variations. If you have to look for radiated emissions within a screened chamber, do it on the basis that you will be able to find frequencies at which emissions exist, but will not be able to draw any firm conclusions as to the amplitude of those emissions.

To be able to make anything approaching *measurements* in a screened chamber, the wall and ceiling reflections must be damped. The floor remains reflective, since the test method relies on the ground plane reflection, and so this kind of chamber is called "semi-anechoic". This is achieved by covering these surfaces with radio absorbing material (RAM). RAM is available as ferrite tiles, carbon loaded foam pyramids, or a combination of both, and it is quite possible to construct a chamber using these materials which meets the volumetric NSA requirement of ±4dB. Enough such chambers have been built and installed that there is plenty of experience on call to ensure that this is achieved. The snag is that either material is expensive, and will at least double the cost of the installed chamber. A comparison between the advantages and disadvantages of the three options is given in Table 6.2.

Table 6.2 Comparison of absorber materials

	Ferrite tiles	**Pyramidal foam**	**Hybrid**
Size	No significant loss of chamber volume	Substantial loss of chamber volume	Some loss of chamber volume
Weight	Heavy; requires ceiling reinforcement	Reinforcement not needed	Heavy; requires ceiling reinforcement
Fixing	Critical – no gaps, must be secure	Not particularly critical	Critical – no gaps, must be secure
Durability	Rugged, no fire hazard, possibility of chipping	Tips can be damaged, possible fire hazard; can be protected	Some potential for damage and fire
Performance	Good mid-frequency, poor at band edges	Good at high frequency, poor at low frequency	Can be optimized across whole frequency range

Partial lining of a room is possible but produces partial results [28][59]. It may, though, be an option for pre-compliance tests. Figure 6.18 shows an example of a chamber NSA which falls substantially outside the required criterion but is still quite a lot better than a totally unlined chamber.

The FAR proposal

So far, we have discussed chambers which mimic the characteristics of an open area

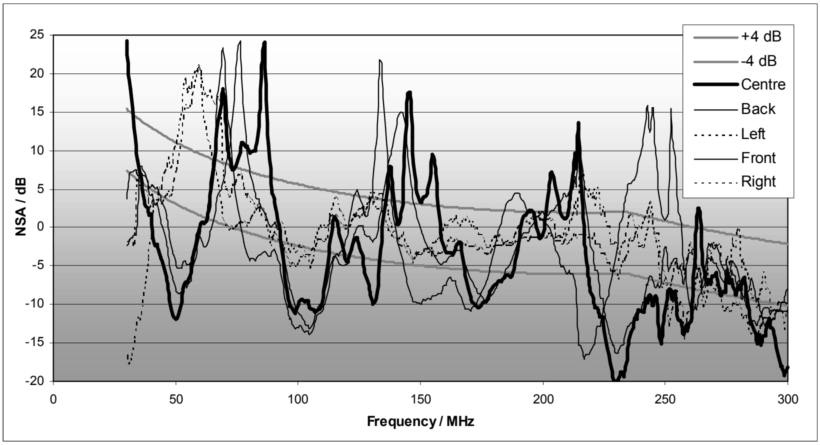


Figure 6.18 Example of a poor chamber NSA (vertical polarization, 1m height)

site, that is they employ a reflective ground plane and a height scan. This makes them a direct substitute for an OATS, allows them to be used in exactly the same way for the same standards, and generally avoids the question of whether the OATS is the optimum method for measuring radiated emissions.

In fact, it isn't. It was originally proposed as a means of dealing with the unavoidable proximity of the ground in practical test set-ups, in the USA, where difficulties with ambient signals and the weather are less severe than in Europe. However, developments over the past ten years in absorber materials have made it quite practical and cost-effective to construct a small fully-anechoic room (FAR), that is, with absorber on the floor as well, which can meet the volumetric ±4dB NSA criterion. This environment is as near to free space as can be achieved. Its most crucial advantage is that, because there is no ground plane, there is no need for an antenna height scan. This eliminates a major source of uncertainty in the test (see section 6.5) and generally allows for a faster and more accurate measurement.

Considerable work has been put in during the last few years to develop a standard for defining the test method in a FAR, and the detailed criteria that must be met by the room itself. This resulted in a European document, prEN 50147-3, which was never published but handed over to CISPR. It has eventually appeared as a combination of an amendment to CISPR 16-1-4 [163] (which describes the validation of the FAR) and another to CISPR 16-2-3 [164] (which describes the test method). Since it will have to co-exist with the standard CISPR test method for some time to come, much of the concern surrounding its development has been to ensure as far as possible that it produces results that are comparable to the OATS method, which means that the necessary adjustment to the limit levels (because of the elimination of the reflected signal) is carefully validated, and that the termination of the off-site cables is dealt with in an appropriate way. Neither of the above documents sets emissions limits.

6.3.1.4 Conducted emissions

By contrast with radiated emissions, conducted measurements need the minimum of extra facilities. The only vital requirement is for a ground plane of at least 2m by 2m, extending at least 0.5m beyond the boundary of the EUT. It is convenient but not essential to make the measurements in a screened enclosure, since this will minimize the amplitude of extraneous ambient signals, and either one wall or the floor of the room can then be used as the ground plane. Non-floor-standing equipment should be placed on an insulating table 40cm away from the ground plane.

Testing cable interference power with the absorbing clamp, as per EN55014-1, requires that the clamp should be moved along the cable by at least a half wavelength, which is 5m at 30MHz. This therefore needs a 5m “racetrack” along which the cable is stretched; the clamp is rolled the length of the cable at each measurement frequency while the highest reading is recorded. There is no guidance in the standards as to whether the measurement should or should not be done inside a screened room. There are likely to be substantial differences one to the other, since the cable under test will couple strongly to the room and will suffer from room-induced resonances in the same manner as a radiated test, though to a lesser extent. For repeatability, a quasi-free space environment would be better, but will then suffer from ambient signals.

6.3.1.5 Pre-compliance and diagnostic tests

Full compliance with the EMC Directive can be achieved by testing and certifying to harmonized European standards. However the equipment and test facilities needed to do this are quite sophisticated and often outside the reach of many companies. The alternative is to take the product to be tested to a test house which is set up to do the proper tests, but this itself is expensive, and to make the best use of the time some preliminary if limited testing beforehand is advisable.

This consideration has given rise to the concept of “pre-compliance” testing. “Pre-compliance” refers to tests done on the production unit (or something very close to it) with a test set-up and/or test equipment that may not fully reflect the standard requirements. The purpose is to:

- avoid or anticipate unpleasant surprises at the final compliance test;
- adequately define the worst-case EUT configuration for the final compliance test, hence saving time;
- substitute for the final compliance test if the results show sufficient margin.

Diagnostics

Although you will not be able to make accurate radiated measurements in a laboratory environment, it is possible to establish a minimum set-up in one corner of the lab at which you can perform emissions diagnostics and carry out comparative tests. For example, if you have done a compliance test at a test house and have discovered one particular frequency at 10dB above the required limit, back in the lab you can apply remedial measures and check each one to see if it gives you a 15dB improvement (5dB margin) without being concerned for the absolute accuracy. While this method is not absolutely foolproof, it is often the best that companies with limited resources and facilities can do.

The following checklist suggests a minimum set-up for doing this kind of in-house diagnostic work:

- unrestricted floor area of at least 5m x 3m to allow a 3m test range with 1m beyond the antenna and EUT;
- no other electronic equipment which could generate extraneous emissions (especially computers) in the vicinity – the EUT’s support equipment should be well removed from the test area;

- no mobile reflecting objects in the vicinity, or those which are mobile should have their positions carefully marked for repeatability;
- an insulating table or workbench at one end of the test range on which to put the EUT, with a LISN bonded to the ground plane beneath it;
- equipment consisting of a spectrum analyser, limiter, antenna set and insulating tripod;
- antenna polarization generally horizontal, and the EUT cables stretched out horizontally and taped to the table facing it, since this reduces errors due to reflections and ground proximity.

Once this set-up is established it should not be altered between measurements on a given EUT. Since the antenna is at a fixed height, there should be no ground plane and the floor should not be metallic, since floor reflections should be attenuated as far as possible. This will give you a reasonable chance of repeatable measurements even if their absolute accuracy cannot be determined.

6.4 Test methods

The major part of all the basic standards referred to in Chapter 4 consists of recipes for carrying out the tests. Because the values obtained from measurements at RF are so dependent on layout and method, these have to be specified in some detail to generate a standard result. This section summarizes the issues involved in full compliance testing, but to actually perform the tests you are recommended to consult the relevant standard carefully.

6.4.1 Test set-up

6.4.1.1 Layout

Conducted emissions

For conducted emissions, the principal requirement is placement of the EUT with respect to the ground plane and the LISN, and the disposition of the mains cable and earth connection(s). Placement affects the stray coupling capacitance between EUT and the ground reference, which is part of the common mode coupling circuit, and so must be strictly controlled; in most cases the standards demand a distance of 0.4m. Cable connections should have a controlled common mode inductance, which means a specified length and minimum possible coupling to the ground plane. Figure 6.19 shows the most usual layout for conducted emissions testing.

Radiated emissions

Radiated emissions to EN 55022 require the EUT to be positioned so that its boundary is the specified distance from the measuring antenna. “Boundary” is defined as “an imaginary straight line periphery describing a simple geometric configuration” which encompasses the EUT. A tabletop EUT should be 0.8m, and a floor-standing EUT should be insulated from and up to 15cm above the ground plane. The EUT will need to be rotated through 360° to find the direction of maximum emission, and this is usually achieved by standing it on a turntable. If it is too big for a turntable, then the antenna must be moved around the periphery while the EUT is fixed. Figure 6.20 shows the general layout for radiated tests.

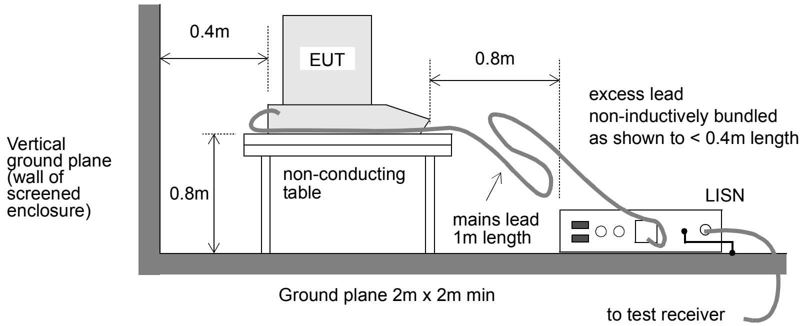


Figure 6.19 Layout for conducted emission tests

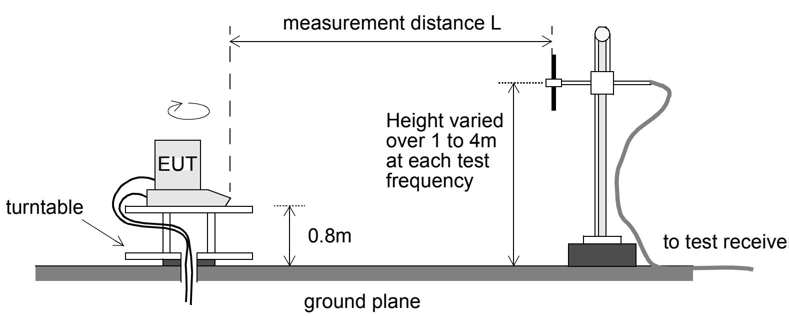


Figure 6.20 Layout for radiated emission tests

6.4.1.2 Configuration

Once the date for an EMC test approaches, the question most frequently asked of test houses is "what configuration of system should I test?" The configuration of the EUT itself is thoroughly covered in the current version of CISPR 22:/EN 55022: it specifies both the layout and composition of the EUT in great detail, especially if the EUT is a personal computer or peripheral. Factors which will affect the emissions profile from the EUT, and which if not specified in the chosen standard should at least be noted in the test plan (see Chapter 9) and report, are:

- number and selection of ports connected to ancillary equipment: you must decide on a "typical configuration". Where several different ports are provided each one should be connected to ancillary equipment. Where there are multiple ports for connection of identical equipment, only one need be connected provided that you can show that any additional connections would not take the system out of compliance;
- disposition of the separate components of the EUT, if it is a system; you should experiment to find the layout that gives maximum emissions within

the confines of the supporting table top, or within typical usage if it is floor standing

- layout, length, disposition and termination practice of all connecting cables; excess cable lengths should be bundled (not looped) near the centre of the cable with the bundle 30 to 40cm long. Lengths and types of connectors should be representative of normal installation practice;
- population of plug-in modules, where appropriate; as with ancillary equipment, one module of each type should be included to make up a minimum representative system. Where you are marketing a system (such as a data acquisition unit housed in a card frame) that can take many different modules but not all at once, you may have to define several minimum representative systems and test all of them;
- software and hardware operating mode; all parts of the system should be exercised, e.g. equipment powered on and awaiting data transfer, and sending/receiving data in typical fashion. You should also define displayed video on VDUs and patterns being printed on a printer;
- use of simulators for ancillary equipment is permissible provided that its effects on emissions can be isolated or identified. Any simulator must properly represent the actual RF electrical characteristics of the real interface;
- EUT grounding method should be as specified in your installation instructions. If the EUT is intended to be operated ungrounded, it must be tested as such. If it is grounded via the safety earth (green and yellow) wire in the mains lead, this should be connected to the measurement ground plane at the mains plug (for conducted measurements, this will be automatic through the LISN).

The catch-all requirement in all standards is that the layout, configuration and operating mode *shall be varied so as to maximize the emissions*. This means some exploratory testing once the significant emission frequencies have been found, varying all of the above parameters – and any others which might be relevant – to find the maximum point. For a complex EUT or one made up of several interconnected sub-systems this operation is time-consuming. Even so, you must be prepared to justify the use of whatever final configuration you choose in the test report.

Information technology equipment

The requirements for testing information technology equipment and peripherals are specified in some depth. The minimum test configuration for any PC or peripheral must include the PC, a keyboard, an external monitor, an external peripheral for a serial port and an external peripheral for a parallel port. If it is equipped with more than the minimum interface requirements, peripherals must be added to all the interface ports unless these are of the same type; multiple identical ports should not all need to be connected unless preliminary tests show that this would make a significant difference. The support equipment for the EUT should be typical of actual usage.

6.4.2 Test procedure

The procedure which is followed for an actual compliance test, once you have found the configuration which maximizes emissions, is straightforward if somewhat lengthy.

Conducted emissions require a continuous sweep from 150kHz to 30MHz at a fixed bandwidth of 9kHz, once with a quasi-peak detector and once with an average detector – the more expensive test receivers can do both together. If the average limits are met with the quasi-peak detector there is no need to perform the average sweep. Radiated emissions require only a quasi-peak sweep from 30MHz to 1GHz with 120kHz bandwidth, with the receiving antenna in both horizontal and vertical polarization. EN 55022 requires that the frequencies, levels and antenna polarizations of at least the six highest emissions closer to the limit than 20dB are reported.

Maximizing emissions

But most importantly, for each significant radiated emission frequency, the EUT must be rotated to find the maximum emission direction *and* the receiving antenna must be scanned in height from 1 to 4m to find the maximum level, removing nulls due to ground reflections. If there are many emission frequencies near the limit this can take a very long time. With a test receiver, automatic turntable and antenna mast under computer control, software can be written to perform the whole operation. This removes one source of operator error and reduces the test time, but not substantially.

A further difficulty arises if the operating cycle of the EUT is intermittent: say its maximum emissions only occur for a few seconds and it then waits for a period before it can operate again. Since the quasi-peak or average measurement is inherently slow, with a dwell time at each frequency of hundreds of milliseconds, interrupting the sweep or the azimuth or height scan to synchronize with the EUT's operating cycle is necessary and this stretches the test time further. If it is possible to speed up the operating cycle to make it continuous, as for instance by running special test software, this is well worthwhile in terms of the potential reduction in test time.

Fast pre-scan

A partial way around the difficulties of excessive test time is to make use of the characteristics of the peak detector (see section 6.1.3.2 and 6.1.3.4). Because it responds instantaneously to signals within its bandwidth the dwell time on each frequency can be short, just a few milliseconds at most, and so using it will enormously speed up the sweep rate for a whole frequency scan. Its disadvantage is that it will overestimate the levels of pulsed or modulated signals (see Figure 6.2). This is a positive asset if it is used on a qualifying pre-scan in conjunction with computer data logging. The pre-scan with a peak detector will only take a few seconds, and all frequencies at which the level exceeds some pre-set value lower than the limit can be recorded in a data file. These frequencies can then be measured individually, with a quasi-peak and/or average detector, and subjecting each one to a height and azimuth scan. Provided there are not too many of these spot frequencies the overall test time will be significantly reduced, as there is no need to use the slow detectors across the whole frequency range.

You must be careful, though, if the EUT emissions include pulsed narrowband signals with a relatively low repetition rate – some digital data emissions have this characteristic – that the dwell time is not set so fast that the peak detector will miss some emissions as it scans over them. The dwell time should be set no less than the period of the EUT's longest known repetition frequency. It is also necessary to do more than one pre-scan, with the EUT in different orientations, to ensure that no potentially offending signal is lost, for instance through being aligned with a null in the radiation pattern.

A further advantage of the pre-scan method is that the pre-scan can be (and usually is) done inside a screened room, thereby eliminating ambients and the difficulties they introduce. The trade-off is that to allow for the amplitude inaccuracies, a greater margin below the limit is needed.

6.4.3 Tests above 1GHz

Up until recently, CISPR emissions tests have not included measurements above 1GHz except for specialized purposes such as microwave ovens and satellite receivers. The US FCC has required measurements above 960MHz for some time, for products with clock frequencies in excess of 108MHz. This has now fed through into the CISPR regime and the test is becoming more common. The method in CISPR 22 (fifth edition, amendment 1, yet to be incorporated in the EN) is based on the approach of conditional testing depending on the highest frequency generated or used within the EUT or on which the EUT operates or tunes [166]:

> If the highest frequency of the internal sources of the EUT is less than 108 MHz, the measurement shall only be made up to 1 GHz.
>
> If the highest frequency of the internal sources of the EUT is between 108 MHz and 500 MHz, the measurement shall only be made up to 2 GHz.
>
> If the highest frequency of the internal sources of the EUT is between 500 MHz and 1 GHz, the measurement shall only be made up to 5 GHz.
>
> If the highest frequency of the internal sources of the EUT is above 1 GHz, the measurement shall be made up to 5 times the highest frequency or 6 GHz, whichever is less.

This aligns with the FCC requirements for unintentional radiators in Part 15 of the FCC rules.

6.4.3.1 Instrumentation and antennas

At microwave frequencies, the sensitivity of the receiving instrument deteriorates. Measurements are often taken with spectrum analyser/preamplifier combinations, sometimes with the addition of preselection or filtering. The field strength is calculated by adding the antenna factor, cable loss, and any other amplification or attenuation to the measured voltage at the receiver/spectrum analyser. The receiver noise floor, determined by the thermal noise generated in the receiver's termination, therefore sets a lower bound on the field strength that can be measured (see Figure 6.7).

The measurement bandwidth above 1GHz is normally set to 1MHz as a compromise between measurement speed and noise floor. A typical microwave spectrum analyser with a mixer front-end may have a noise floor at a resolution bandwidth of 1MHz ranging from approximately 25 dBμV at 1GHz to 43 dBμV at 22GHz. You can use a low noise preamplifier to improve this poor noise performance. The noise figure of a two-stage system is given by the following equation:

$$\text{Total NF} = NF_1 + (NF_2 - 1)/G1, \tag{6.6}$$

where NF_1 is the preamplifier noise figure, G1 is the preamplifier gain, and NF_2 is the noise figure of the spectrum analyser (all in linear units, not dB).

This shows that the overall system noise figure is dominated by the first stage's noise figure and gain. A preamplifier is virtually a necessity for these measurements; it should be located very close to the measuring antenna with a short low-loss cable connecting the two, since any loss here will degrade the system noise performance irretrievably, while loss incurred after the preamp has much less effect.

Antennas

Dipole-type antennas become very small and insensitive as the frequency increases above 1GHz. They have a smaller “aperture,” which describes the area from which energy is collected by an antenna. It is possible to get log periodics up to 2 or even 3GHz, and the BiLog types can have a range extended to 2GHz, but above this it is normal to use a horn antenna. This type converts the 50 ohm coax cable impedance directly to a plane wave at the mouth of the horn, and depending on construction can have either a wide bandwidth or a high gain and directivity, though it is rare to get both together.

The high gain gives the best system noise performance, but the directivity can be both a blessing and a curse. Its advantage is that it gives less sensitivity to off-axis reflections, so that the anechoic performance of a screened room or the effect of reflection from dielectric materials – which worsens at these higher frequencies – becomes less critical; but it will also cover less area at a given distance, so large EUTs cannot be measured in one sweep but must have the antenna trained on different parts in consecutive sweeps.

6.4.3.2 Methods

These considerations affect the test methods by comparison with those used below 1GHz. The measurement system is less sensitive, but this is compensated by a closer preferred measurement distance (3m) and a higher limit level. The quasi-peak detector is not used; only peak and average detectors are used, with a measurement bandwidth of 1MHz. Spectrum analyser measurements can use a reduced video bandwidth to implement the average detection, although this will substantially increase the sweep time, so a peak scan and spot frequency average measurements would be usual. Because the antenna is more directional, reflections from the ground plane or from the walls are less troublesome and a height scan to deal with the ground reflection is not required, although absorbers on all chamber surfaces will nevertheless be needed. But the same principle applies to the EUT: at these frequencies, the EUT’s own emissions are more directional and so more care is needed in finding the maximum in EUT azimuth, and 15° increments for the turntable rotation steps are recommended.

Generally, the set-up conditions of the EUT are the same as those for tests below 1GHz; the tests can be performed on the same arrangement. CISPR 16-1-4 includes a method to validate a test volume and the EUT should be wholly located within this volume. CISPR 16-2-3, which describes the method, defines a dimension *w* formed by the minimum 3dB beamwidth of the receiving antenna at the measurement distance *d* actually used (Figure 6.21):

$$w = 2 \cdot d \cdot \tan(0.5 \cdot \theta_{3db}) \qquad (6.7)$$

It requires a minimum value of *w* over the frequency range 1 to 18GHz, which in turn implies careful selection of antenna type and measurement distance; and if the EUT’s height is larger than *w* then it has to be scanned in height in order to cover the whole height of the EUT. A width scan is not needed as the EUT will be rotated on its turntable to find the maximum emission in azimuth.

6.4.4 Military emissions tests

The foregoing has concentrated on emissions tests to CISPR standards. Military and aerospace methods as set out in DEF STAN 59-41 and MIL STD 461E (section 5.2) are

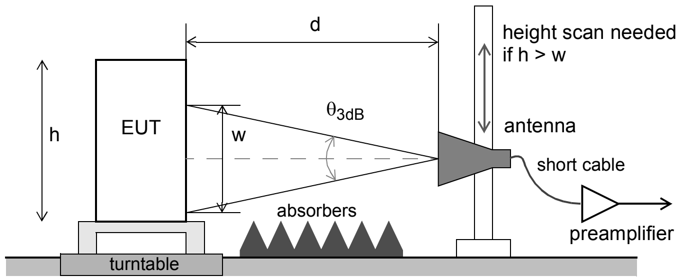


Figure 6.21 Arrangement of test above 1GHz

significantly different, to the extent that making a one-to-one comparison of the results, as is necessary if military applications are to use commercial-off-the-shelf (COTS) products, is very difficult. This section briefly summarizes the most important differences.

6.4.4.1 Instrumentation

The measuring receiver has a different detector function and bandwidth specification. The CISPR quasi-peak and average detectors are avoided; only the peak detector is used. The bandwidths are compared in Table 6.3.

Table 6.3 Comparison of DEF STAN 59-41 and CISPR bandwidths

Military frequency range	6dB bandwidth	CISPR frequency range	6dB bandwidth
20Hz–1kHz	10Hz		
1kHz–50kHz	100Hz		
		9kHz–150kHz	200Hz
50kHz–1MHz	1kHz		
1MHz–30MHz	10kHz *		
		150kHz–30MHz	9kHz
30MHz–18GHz	100kHz *		
		30MHz–1GHz	120kHz
* some deviations for certain applications			

6.4.4.2 Transducers

For conducted emissions, DEF STAN tests use a LISN on the power lines, but it is the 50Ω/5µH version with a frequency range from 1kHz to 400MHz over which the impedance is defined. It is permanently connected to the power supplies and used in all tests, not only those which measure power supply emissions. For dc supplies an additional 30,000 µF capacitor is connected between positive and negative on the power supply side of the two LISNs to improve the low frequency performance.

For the radiated tests, all antennas are situated at a separation distance of 1m from the closest surface of the EUT to the antenna calibration reference point. This is probably the single most important difference from the commercial tests under CISPR. To cover the extended frequency range required by the radiated emissions measurement, the antennas used for E-field tests are:

14kHz–1.6MHz (land systems), 30MHz (air and sea systems): active or passive 41” vertical monopole (rod) antenna with counterpoise or ground plane
1.6MHz–76MHz (land systems): antenna as used in installed systems
25MHz–300MHz: biconical
200MHz–1GHz: log periodic
1GHz–18GHz: waveguide or double ridged waveguide horns

6.4.4.3 Test site

Radiated emission tests are all conducted in a screened room: there is no “open area” test site as such. The EUT is laid out on a ground plane bench which is bonded to the rear wall of the screened room, and the exact distances from the antenna to the bench and to the EUT are specified. The screened room itself has to comply with Part 5 of the DEF STAN, which includes the requirement for partial lining with anechoic material, and that the maximum dimensions of the room give a lowest chamber resonance (see equation (14.3) on page 386) not below 30MHz. It should be demonstrated that the room’s normalized site insertion loss (NSIL) is representative of free-space theoretical values: the maximum permitted tolerances are ±10dB over the frequency range 80 to 250MHz and ±6dB from 250MHz to 1GHz. Measurements are to be made with both vertical and horizontal polarizations. The concept is similar to the CISPR ±4dB requirement but for a single position of the antenna (no height scan), and of course the tolerances are much wider.

6.5 Measurement uncertainty

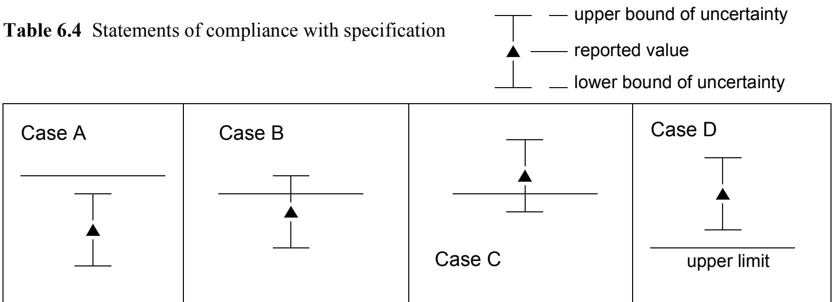
EMC measurements are inherently less accurate than most other types of measurement. Whereas, say, temperature or voltage measurement can be refined to an accuracy expressed in parts per million, field strength measurements in particular can be in error by 10dB or more, partly due to uncertainties in the measuring instrumentation and method and partly due to uncertainties introduced by the EUT set-up. It is always wise to allow a margin of about this magnitude between your measurements and the specification limits, not only to cover measurement uncertainty but also tolerances arising in production.

6.5.1 Applying measurement uncertainty

UKAS, the body which accredits UK EMC test houses, issues guidelines on determining measurement uncertainty in LAB 34 [195] and it requires test houses to calculate and if necessary to report their own uncertainties, but for EMC tests it does not define acceptable levels of uncertainty. Amongst other things this document suggests that, if there is no other specification criterion, guidance or code of practice, test houses express their results in one of four ways, as shown in Table 6.4.

Cases B and C in the table, whilst being metrologically sound, are clearly not helpful to manufacturers who want a simple statement of pass or fail. However, CISPR 16-4-2 (“Uncertainty in EMC measurements”) [165] prescribes that for emissions tests the measurement uncertainty should be taken into account in determining compliance. But it goes on to give a total uncertainty figure U_{cispr} for each of the principal emissions tests (Table 6.5), based only on the instrumentation and test method errors and not

Table 6.4 Statements of compliance with specification



Case A	Case B	Case C	Case D
The product complies	The measured result is below the specification limit by a margin less than the measurement uncertainty; it is not therefore possible to determine compliance at a level of confidence of 95%. However, the measured result indicates a higher probability that the product tested complies with the specification limit.	The measured result is above the specification limit by a margin less than the measurement uncertainty; it is not therefore possible to determine compliance at a level of confidence of 95%. However, the measured result indicates a higher probability that the product tested does not comply with the specification limit.	**The product does not comply**

taking into account any contribution from the EUT. If the test house's declared uncertainty is less than or equal to this value, then direct comparison with the limit is acceptable (cases A and D with an effective measurement uncertainty of zero). If the uncertainty is greater, then the test result must be increased by the excess before comparison with the limit – effectively penalizing manufacturers who use test houses with large uncertainties.

6.5.2 Sources of uncertainty

This section discusses how measurement uncertainties arise (Figure 6.22).

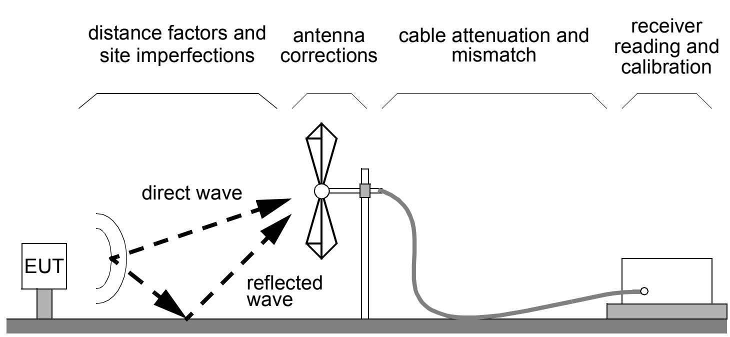


Figure 6.22 Sources of error in radiated emissions tests

Table 6.5 CISPR uncertainties according to CISPR 16-4-2

Measurement	U_{cispr}
Conducted disturbance, mains port, 9–150kHz	4.0dB
Conducted disturbance, mains port, 150kHz–30MHz	3.6dB
Disturbance power, 30–300MHz	4.5dB
Radiated disturbance, 30–300MHz	5.1dB

6.5.2.1 Instrument and cable errors

Modern self-calibrating test equipment can hold the uncertainty of measurement at the instrument input to within ±1dB. To fully account for the receiver errors, its pulse amplitude response, variation with pulse repetition rate, sinewave voltage accuracy, noise floor and reading resolution should all be considered. Input attenuator, frequency response, filter bandwidth and reference level parameters all drift with temperature and time, and can account for a cumulative error of up to 5dB at the input even of high quality instrumentation. To overcome this a calibrating function is provided. When this is invoked, absolute errors, switching errors and linearity are measured using an in-built calibration generator and a calibration factor is computed which then corrects the measured and displayed levels. It is left up to the operator when to select calibration, and this should normally be done before each measurement sweep. Do not invoke it until the instrument has warmed up – typically 30 minutes to an hour – or calibration will be performed on a "moving target". A good habit is to switch the instruments on first thing in the morning and calibrate them just before use.

The attenuation introduced by the cable to the input of the measuring instrument can be characterized over frequency and for good quality cable is constant and low, although long cables subject to large temperature swings can cause some variations. Uncertainty from this source should be accounted for but is normally not a major contributor. The connector can introduce unexpected frequency-dependent losses; the conventional BNC connector is particularly poor in this respect, and you should perform all measurements whose accuracy is critical with cables terminated in N-type connectors, properly tightened (and not cross-threaded) against the mating socket.

Mismatch uncertainty

When the cable impedance, nominally 50Ω, is coupled to an impedance that is other than a resistive 50Ω at either end it is said to be mismatched. A mismatched termination will result in reflected signals and the creation of standing waves on the cable. Both the measuring instrument input and the antenna will suffer from a degree of mismatch which varies with frequency and is specified as a Voltage Standing Wave Ratio (VSWR). Appendix D (section D.2.4) discusses VSWR further. If either the source or the load end of the cable is perfectly matched then no errors are introduced, but otherwise a *mismatch error* is created. Part of this is accounted for when the measuring instrument or antenna is calibrated. But calibration cannot eliminate the error introduced by the phase difference along the cable between source and load, and this leaves an uncertainty component whose limits are given by:

$$\text{uncertainty} = 20\log_{10}(1 \pm \Gamma_L \cdot \Gamma_S) \tag{6.8}$$

where Γ_L and Γ_S are the source and load reflection coefficients

As an example, an input VSWR of 1.5:1 and an antenna VSWR of 4:1 gives a mismatch uncertainty of ±1dB. The biconical in particular can have a VSWR exceeding 15:1 at the extreme low frequency end of its range. When the best accuracy is needed, minimize the mismatch error by including an attenuator pad of 6 or 10dB in series with one or both ends of the cable, at the expense of measurement sensitivity.

6.5.2.2 Conducted test factors

Mains conducted emission tests use a LISN/AMN as described in section 6.2.2.1. Uncertainties attributed to this method include the quality of grounding of the LISN to the ground plane, the variations in distances around the EUT, and inaccuracies in the LISN parameters. Although a LISN theoretically has an attenuation of nearly 0dB across most of the frequency range, in practice this can't be assumed particularly at the frequency extremes and you should include a voltage division factor derived from the network's calibration certificate. In some designs, the attenuation at extremes of the frequency range can reach several dB. Mismatch errors, and errors in the impedance specification, should also be considered.

Other conducted tests use a telecom line ISN instead of a LISN, or use a current probe to measure common mode current. An ISN will have the same contributions as the LISN with the addition of possible errors in the LCL (see section 6.2.2.4). A current probe measurement will have extra errors due to stray coupling of the probe with the cable under test, and termination of the cable under test, as well as calibration of the probe factor.

6.5.2.3 Antenna calibration

One method of calibrating an antenna is against a reference standard antenna, normally a tuned dipole on an open area test site [23]. This introduces its own uncertainty, due to the imperfections both of the test site and of the standard antenna – ±0.5dB is now achievable – into the values of the antenna factors that are offered as calibration data. An alternative method of calibration known as the Standard Site Method [123] uses three antennas and eliminates errors due to the standard antenna, but still depends on a high quality site.

Further, the physical conditions of each measurement, particularly the proximity of conductors such as the antenna cable, can affect the antenna calibration. These factors are worst at the low frequency end of the biconical's range, and are exaggerated by antennas that exhibit poor balance [24]. When the antenna is in vertical polarization and close to the ground plane, any antenna imbalance interacts with the cable and distorts its response. Also, proximity to the ground plane in horizontal polarization can affect the antenna's source impedance and hence its antenna factor. Varying the antenna height above the ground plane can introduce a height-related uncertainty in antenna calibration of up to 2dB [93].

These problems are less for the log periodic at UHF because nearby objects are normally out of the antenna's near field and do not affect its performance, and the directivity of the log periodic reduces the amplitude of off-axis signals. On the other hand the smaller wavelengths mean that minor physical damage, such as a bent element, has a proportionally greater effect. Also the phase centre (the location of the active part of the antenna) changes with frequency, introducing a distance error, and since at the extreme of the height scan the EUT is not on the boresight of the antenna its directivity introduces another error. Both of these effects are greatest at 3m distance. An overall uncertainty of ±4dB to allow for antenna-related variations is not unreasonable, although this can be improved with care.

The difficulties involved in defining an acceptable and universal calibration method for antennas that will be used for emissions testing led to the formation of a CISPR/A working group to draft such a method. It has standardized on a free-space antenna factor determined by a fixed-height 3-antenna method on a validated calibration test site [76]. The method is fully described in CISPR 16-1-5.

6.5.2.4 Reflections and site imperfections

The antenna measures not only the direct signal from the EUT but also any signals that are reflected from conducting objects such as the ground plane and the antenna cable. The field vectors from each of these contributions add at the antenna. This can result in an enhancement approaching +6dB or a null which could exceed −20dB. It is for this reason that the height scan referred to in section 6.4.2 is carried out; reflections from the ground plane cannot be avoided but nulls can be eliminated by varying the relative distances of the direct and reflected paths. Other objects further away than the defined CISPR ellipse will also add their reflection contribution, which will normally be small (typically less than 1dB) because of their distance and presumed low reflectivity.

This contribution may become significant if the objects are mobile, for instance people and cars, or if the reflectivity varies, for example trees or building surfaces after a fall of rain. They are also more significant with vertical polarization, since the majority of reflecting objects are predominantly vertically polarized. With respect to the site attenuation criterion of ±4dB, CISPR 16-4-2 states:

> ... measurement uncertainty associated with the CISPR 16-1 site attenuation measurement method is usually large, and dominated by the two antenna factor uncertainties. Therefore a site which meets the 4 dB tolerance is unlikely to have imperfections sufficient to cause errors of 4 dB in disturbance measurements. In recognition of this, a triangular probability distribution is assumed for the correction δSA.
>
> [165], clause A.5

Antenna cable

With a poorly balanced antenna, the antenna cable is a primary source of error [92][93]. By its nature it is a reflector of variable and relatively uncontrolled geometry close to the antenna. There is also a problem caused by secondary reception of common mode currents flowing on the sheath of the cable. Both of these factors are worse with vertical polarization, since the cable invariably hangs down behind the antenna in the vertical plane. They can both be minimized by choking the outside of the cable with ferrite sleeve suppressors spaced along it, or by using ferrite loaded RF cable (section 13.1.8.4). If this is not done, measurement errors of up to 5dB can be experienced due to cable movement with vertical polarization. However, modern antennas with good balance, which is related to balun design, will minimize this problem.

6.5.2.5 The measurement uncertainty budget

Some or all of the above factors are combined together into a budget for the total measurement uncertainty which can be attributed to a particular method. The detail of how to develop an uncertainty budget is beyond the scope of this book, but you can refer to LAB 34 [195] or CISPR 16-4-2 [165] for this. Essentially, each contribution is assigned a value and a probability distribution. These are derived either from existing evidence (such as a calibration certificate) or from estimation based on experience, experiment or published information. Contributions can be classified into two types: Type A contributions are random effects that give errors that vary in an unpredictable way while the measurement is being made or repeated under the same conditions. Type

B contributions arise from systematic effects that remain constant while the measurement is made but can change if the measurement conditions, method or equipment is altered.

The "standard uncertainty" for each contribution is obtained by dividing the contribution's value by a factor appropriate to its probability distribution. Then the "combined standard uncertainty" is given by adding the standard uncertainties on a root-sum-of-squares basis; and the "expanded uncertainty" of the method, defining an interval about the measured value that will encompass the true value with a specified degree of confidence, and which is reported by the laboratory along with its results, is calculated by multiplying the combined standard uncertainty by a "coverage factor" k. In most cases, k = 2 gives a 95% level of confidence.

A simplified example budget for a straightforward conducted emissions test is given in Table 6.6. This is derived from LAB 34; the contributions are typical values, but each test lab should derive and justify its own values to arrive at its own overall uncertainty for the test. Each measurement method (conducted emissions, radiated emissions, disturbance power) needs to have its own budget created, and it is reasonable to sub-divide budgets into, for example, frequency sub-ranges when the contributions vary significantly over the whole range, such as with different antennas.

Table 6.6 Example uncertainty budget for conducted measurement 150kHz to 30MHz

Contribution	Value	Prob. dist.	Divisor	ui(y)	ui(y)^2
Receiver sinewave accuracy	**1.00**	Rectangular	1.732	0.577	0.333
Receiver pulse amplitude response	**1.50**	Rectangular	1.732	0.866	0.750
Receiver pulse repetition response	**1.50**	Rectangular	1.732	0.866	0.750
Receiver indication	**0.05**	Rectangular	1.732	0.029	0.001
Frequency step error	**0.00**	Rectangular	1.732	0.000	0.000
Noise floor proximity	**0.00**	Rectangular	1.732	0.000	0.000
LISN attenuation factor calibration	**0.20**	Normal k=2	2.000	0.100	0.010
Cable loss calibration	**0.40**	Normal k=2	2.000	0.200	0.040
LISN impedance	**2.70**	Triangular	2.449	1.102	1.215
Mismatch	-0.891	U-shaped	1.414	-0.630	0.397
Receiver VRC	**0.15**				
LISN + cable VRC	**0.65**				
Measurement system repeatability	**0.50**	Normal k=1	1.000	0.500	0.250
Combined standard uncertainty		Normal		1.936	3.746
Expanded uncertainty		Normal, k = 2.0		**3.87**	

Entered	**0.15**
Calculated	0.5
Result	**3.87**

It is important to realize that this is, strictly speaking, a measurement *instrumentation and method* uncertainty budget. It doesn't take into account any uncertainty contributions attributable to the EUT itself or to its set-up, because the lab cannot know what these contributions are, yet they are likely to be at least as important in determining the outcome of the test as the visible and calculable contributions.

6.5.2.6 Human and environmental factors

The test engineer

It should be clear from section 6.4 that there are many ways to arrange even the simplest EUT to make a set of emissions measurements. Equally, there are many ways in which the measurement equipment can be operated and its results interpreted, even to perform measurements to a well-defined standard – and not all standards are well defined. In addition, the quantity being measured is either an RF voltage or an electromagnetic field strength, both of which are unstable and consist of complex waveforms varying erratically in amplitude and time. Although software can be written to automate some aspects of the measurement process, still there is a major burden on the experience and capabilities of the person actually doing the tests.

Some work has been reported which assesses the uncertainty associated with the actual engineer performing radiated emission measurements [119]. Each of four engineers was asked to evaluate the emissions from a desktop computer consisting of a processor, VDU and keyboard. This remained constant although its disposition was left up to the engineer. The resultant spread of measurements at various frequencies and for both horizontal and vertical polarization was between 2 and 15dB – which does not generate confidence in their validity! Two areas were recognized as causing this spread, namely differences in EUT and cable configurations, and different exercising methods.

The tests were repeated using the same EUT, test site and test equipment but with the EUT arrangement now specified and with a fixed antenna height. The spread was reduced to between 2 and 9dB, still an unacceptably large range. Further sources of variance were that maximum emissions were found at different EUT orientations, and the exercising routines still had minor differences. The selected measurement time (section 6.1.3.4) can also have an effect on the reading, as can ancillary settings on the test receiver and the orientation of the measurement antenna.

Ambients

The major uncertainty introduced into EMC emissions measurements by the external environment, apart from those discussed above, is due to ambient signals. These are signals from other transmitters or unintentional emitters such as industrial machinery, which mask the signals emitted by the EUT. On an OATS they cannot be avoided, except by initially choosing a site which is far from such sources. In a densely populated country such as the UK, and indeed much of Europe, this is wishful thinking. A "green-field" site away from industrial areas, apart from access problems, almost invariably falls foul of planning constraints, which do not permit the development of such sites – even if they can be found – for industrial purposes.

Another Catch-22 situation arises with regard to broadcast signals. It is important to be able to measure EUT emissions within the Band II FM and Bands IV and V TV broadcast bands since these are the very services that the emission standards are meant to protect. But the *raison d'être* of the broadcasting authorities is to ensure adequate field strengths for radio reception throughout the country. The BBC publish their requirements for the minimum field strength in each band that is deemed to provide coverage [1] and these are summarized in Table 6.7. In each case, these are (naturally) significantly higher than the limit levels which an EUT is required to meet. In other words, assuming country-wide broadcast coverage is a fact, *nowhere* will it be possible to measure EUT emissions on an OATS at all frequencies throughout the broadcast bands because these emissions will be masked by the broadcast signals themselves.

The only sure way around the problem of ambients is to perform the tests inside a screened chamber, which is straightforward for conducted measurements but for radiated measurements is subject to severe inaccuracies introduced by reflections from the wall of the chamber as discussed earlier. An anechoic chamber will reduce these inaccuracies and requirements for anechoic chambers are now in the standards, as mentioned in section 6.3.1.3, but a fully compliant anechoic chamber will be

Service	Frequency range	Minimum acceptable field strength
Long wave	148.5–283.5kHz	5mV/m
Medium wave	526.5–1606.5kHz	2mV/m
VHF/FM band II	87.5–108MHz	54dBμV/m
TV band IV	471.25–581.25MHz	64dBμV/m
TV band V	615.25–853.25MHz	70dBμV/m

Source: [1]

Table 6.7 Minimum broadcast field strengths in the UK

prohibitively expensive for many companies. (Major blue-chip electronics companies have indeed invested millions in setting up such facilities in house.) The method of pre-scan in a non-anechoic chamber discussed in section 6.4.2 goes some way towards dealing with the problem, but doesn't solve the basic difficulty that a signal that is underneath an ambient on an OATS cannot be accurately measured.

Emissions standards such as EN 55022 recognize the problem of ambient signals and in general require that the test site ambients should not exceed the limits. When they do, the standard allows testing at a closer distance such that the limit level is increased by the ratio of the specified distance to the actual distance. This is usually only practical in areas of low signal strength where the ambients are only a few dB above the limits. Some relief can be gained by orienting the site so that the local transmitters are at right angles to the test range, taking advantage of the antennas' directional response at least with horizontal polarization.

When you are doing diagnostic tests the problem of continuous ambients is less severe because even if they mask some of the emissions, you will know where they are and can tag them on the spectrum display. Some analysis software performs this task automatically. Even so, the presence of a "forest" of signals on a spectrum plot confuses the issue and can be unnerving to the uninitiated. Transient ambients, such as from portable transmitters or occasional broadband sources, are more troublesome because it is harder to separate them unambiguously from the EUT emissions. Sometimes you will need to perform more than one measurement sweep in order to eliminate all the ambients from the analysis.

Ambient discrimination by bandwidth and detector

Annex A to CISPR 16-2-3 [164] attempts to address the problem of ambients from another angle. This distinguishes between broadband and narrowband EUT emissions in the presence of broadband or narrowband ambient noise (Figure 6.23). If both the ambient noise and the EUT emissions are narrowband, a suitably narrow measurement bandwidth is recommended, with use of the peak detector. The measurement bandwidth should not be so low as to suppress the modulation spectra of the EUT emission. If the EUT noise is broadband, the measurement cannot be made directly underneath a narrowband ambient but can be taken either side, and the expected actual level interpolated.

When the ambient disturbance is broadband, bandwidth discrimination is not possible, but a narrowband EUT emission may be extracted by using the average detector with a narrower measuring bandwidth that maximizes the EUT disturbance-to-ambient ratio. The average detector should reduce the broadband level without affecting the desired EUT narrowband signal, as long as the EUT signal is not severely amplitude or pulse modulated; if it is, some error will result.

Broadband EUT disturbances in the presence of broadband ambients cannot be directly measured, although if their levels are similar (say, within 10dB) it is possible to estimate the EUT emission through superposition, using the peak detector.

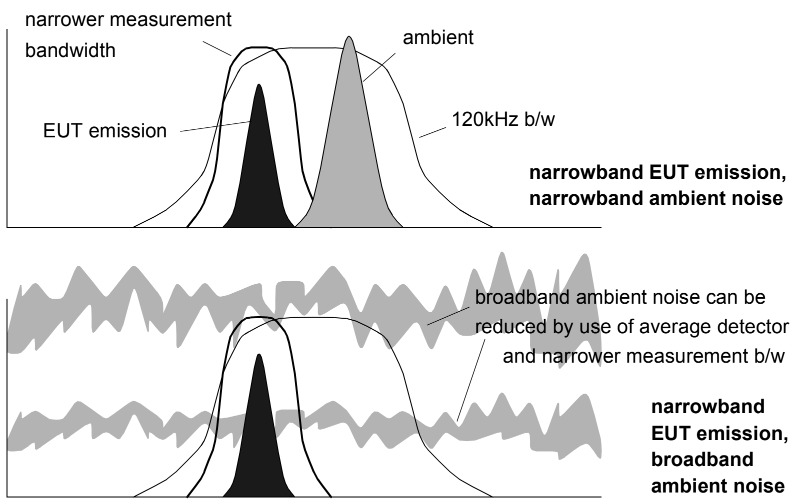


Figure 6.23 Ambient discrimination on the basis of bandwidth

Chapter 7
Immunity tests

7.1 RF immunity

Until the EMC Directive, most commercial immunity testing was not mandatory, but driven by customer requirements for reliability in the presence of interference. Military and aerospace immunity test standards have been in existence for some time, and were occasionally called up in commercial contracts in default of any other available or applicable standards. These allow for both conducted and radiated RF immunity test methods. The major established commercial standard tests were originally listed until the mid-90s in IEC 801. These have long since been superseded; we now have IEC 61000-4-3 third edition and IEC 61000-4-6 second edition, for radiated and conducted tests respectively. CISPR 20 requires both conducted and radiated immunity tests but applies only to broadcast receivers and related equipment.

7.1.1 Equipment

Figure 7.1 shows the components of a typical radiated immunity test system using a screened room.

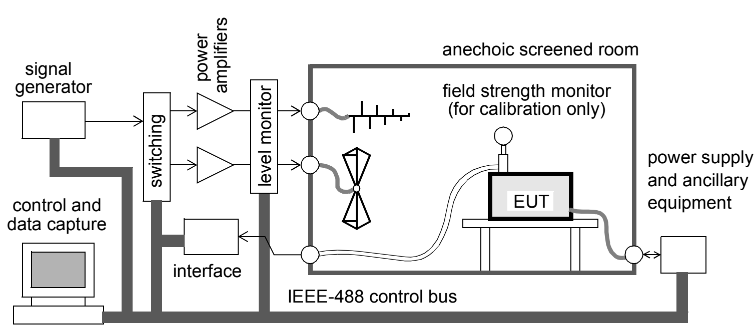


Figure 7.1 RF immunity test system

The basic requirements are an RF signal source, a broadband power amplifier and a transducer. The latter may be a set of antennas, a transmission line cell or a stripline. These will enable you to generate a field at the EUT's position, but for accurate control of the field strength there must be some means to control and calibrate the level that is fed to the transducer. A test house will normally integrate these components with computer control to automate the frequency sweep and levelling functions and to meet the field calibration requirements of the standard.

7.1.1.1 Signal source

Any RF signal generator that covers the required frequency range (80–1000MHz and above for IEC 61000-4-3, 150kHz–80MHz for IEC 61000-4-6) will be useable. Its output level must match the input requirement of the power amplifier with a margin of a few dB. This is typically 0dBm and is not a problem.

IEC 61000-4-3 calls for the RF carrier to be modulated at 1kHz to a depth of 80%. This will normally be done within the signal generator; other more specialized modulation, such as the 1Hz pulse requirement for testing alarm products, may need a separate modulator. Typically, a synthesized signal generator will be used for stepped application. Control software will set the frequency in steps across the band to be covered. The required frequency accuracy depends on whether the EUT exhibits any narrowband responses to interference. A manual frequency setting ability is necessary for when you want to investigate the response around particular frequencies. Be careful that no transient level changes are caused within the signal generator by range changing or frequency stepping, since these will be amplified and applied as transient fields to the EUT, possibly causing an erroneous susceptibility.

7.1.1.2 Power amplifier

Most signal sources will not have sufficient output level on their own, and you will require a set of power amplifiers to increase the level. The power output needed will depend on the field strength that you have to generate at the EUT, and on the characteristics of the transducers you use to do this. In contrast to the antenna factor for emissions measurements, an antenna for RF immunity will be characterized for the power needed to provide a given field strength at a set distance. This can be specified either directly or as the gain of the antenna. The relationship between antenna gain, power supplied to the antenna and field strength in the far field is:

$$P_t = (r \cdot E)^2/(30 \cdot G) \tag{7.1}$$

where P_t is the antenna power input
r is the distance from the antenna in metres
E is the field strength at r in volts/metre
G is the numerical antenna gain [= antilog(G_{dB}/10)] over isotropic

The gain of a broadband antenna varies with frequency and hence the required power for a given field strength will also vary with frequency. Figure 7.2 shows a typical power requirement versus frequency for an unmodulated field strength of 10V/m at a distance of 1m. Less power is needed at high frequencies because of the higher gain of the log periodic antenna. You can also see the large increase in power required by the conventional biconical below 80MHz; it is partly because of this that the lowest frequency for radiated immunity testing was chosen to be 80MHz, although subsequent developments in broadband antennas have improved the situation (see section 7.1.1.4).

The [power output · bandwidth] product is the most important parameter of the power amplifier you will choose, and it largely determines the cost of the unit. Very broad band amplifiers (1–1000MHz) are available with powers of a few watts, but this may not be enough to generate required field strengths from a biconical antenna in the low VHF region. A higher power amplifier with a bandwidth restricted to 30–300MHz will also be needed. If you can use two amplifiers, each matched to the bandwidth and power requirements of the two antennas you are using, this will minimize switching requirements to cover the whole frequency sweep. Note that the power delivered to the antenna (net power) is not the same as power supplied by the amplifier unless the

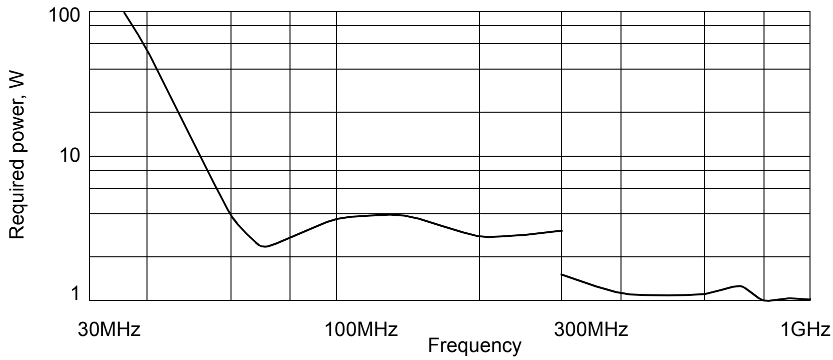


Figure 7.2 Required power versus frequency for 10V/m at 1m, biconical and log periodic antennas

antenna is perfectly matched, a situation which does not occur in practice. With high VSWR (such as a biconical or standard bilog below 70MHz) most of the power supplied to the amplifier is reflected back to it, which is inefficient and can be damaging to the amplifier.

Modulation

Some over-rating of the power output is necessary to allow for modulation, system losses and for the ability to test at a greater distance. Modulation at 80%, as required by IEC 61000-4-3, increases the instantaneous power requirement by a factor of 5.2dB (3.3 times) over the unmodulated requirement, as shown in Figure 7.2. If you will be

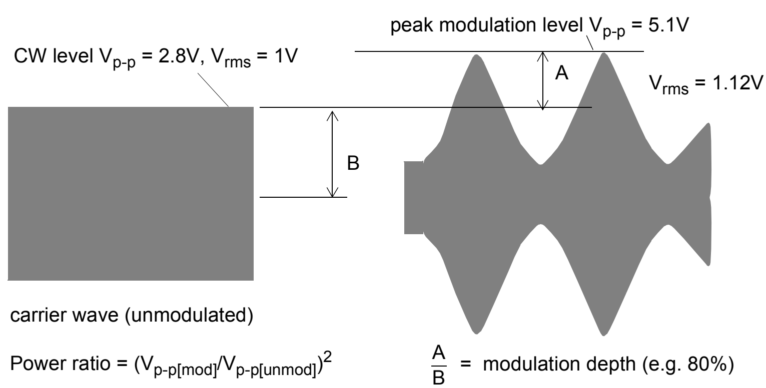


Figure 7.3 Modulated versus unmodulated waveforms

using the system in a non-anechoic screened room the system should be further over-rated by at least 6dB (four times power) to allow for field nulls at certain frequencies due to room reflections. If the system uses other transducers such as a TEM cell or stripline (discussed in section 7.1.1.4) rather than a set of antennas, then the power output requirement for a given field strength will be significantly less. Thus there is a direct cost trade-off between the type of transducer used and the necessary power of the amplifier.

The 1kHz modulation requirement is common to most tests that reference IEC 61000-4-3. Other standards may take a different approach; for instance in the automotive immunity tests of ISO 11452 the modulation is also 1kHz at 80% depth, but the specification field strength is quoted at the peak of the modulation, rather than on an unmodulated waveform. Historically, some product standards specified a particular test for immunity to GSM phones at a spot frequency of 900MHz, which used pulsed modulation at 200Hz to simulate more accurately the effects of the GSM signal; this has been dropped in most later versions, on the grounds that experimental results on several different types of test object showed that the 1kHz sinewave always gave the most severe results and it was of universal applicability. There are a few standards that specify different frequencies for their own reasons, for instance the alarm immunity standard EN 50130-4 requires a 1Hz pulsed modulation, since many alarm detectors will be sensitive to a slow rate of change of the RF stress.

Note that when you are setting the applied power level during calibration for any RF immunity test, the modulation should be disabled; RF power meters and field strength meters give inaccurate results on modulated signals.

Secondary parameters

Other factors that you should take into account (apart from cost) when specifying a power amplifier are:

- linearity: RF immunity testing can tolerate some distortion but this should not be excessive, since it will appear as harmonics of the test frequency and may give rise to spurious responses in the EUT; according to the standard, distortion products should be at least −15dB relative to the carrier;
- ruggedness: the amplifier should be able to operate at full power continuously, without shutting itself down, into an infinite VSWR, i.e. an open or short circuit load. Test antennas are not perfect, and neither are the working practices of test engineers!
- power gain: full power output must be obtainable from the expected level of input signal, with some safety margin, across the whole frequency band;
- reliability and maintainability: in a typical test facility you are unlikely to have access to several amplifiers, so when it goes faulty you need to have assurance that it can be quickly repaired.

7.1.1.3 Field strength monitor and levelling

It is essential to be able to ensure the correct field strength at the EUT. Reflections and field distortion by the EUT will cause different field strength values from those which would be expected in free space, and these values will vary as the frequency band is swept. You are recommended to re-read section 6.5 on sources of uncertainty in emissions measurements, as the issues discussed there apply equally to measurements of field strength used for immunity tests.

RF fields can be determined by a broadband field sensor, normally in the form of a small dipole and detector replicated in three orthogonal planes so that the assembly is sensitive to fields of any polarization. In the simplest extreme, the unit can be battery powered with a local meter so that the operator must continuously observe the field strength and correct the output level manually. A more sophisticated set-up, and one that is essential for calibrating to IEC 61000-4-3, uses a fibre optic data link from the sensor, so that the field is not disturbed by an extraneous cable.

There are two major methods of controlling the applied field strength: by closed loop levelling, or by substitution. In non-anechoic screened rooms, closed loop levelling as specified in the military tests is necessary. In this method, the field sensor is placed next to the EUT and the power applied to the transducer is adjusted to provide the correct field strength value, while the sweep is in progress. While this method seems intuitively correct, in practice it has several disadvantages:

- the sensor measures the field only at one point; at other points around the EUT, the field can change significantly, especially when the EUT is large compared to a wavelength;
- if the sensor by chance is positioned in a null at a particular frequency, the result will be an increase in applied power to attempt to correct the field strength, with a consequent increase, often well over the intended value, at other locations;
- with a stepped frequency application, attempting to find the correct field strength at each step may result in over-correction of the applied power and hence a transient excess of field strength.

Clearly it is possible to inadvertently over-test the EUT by this method. In an anechoic chamber and with transducers such as TEM cells, the substitution method is preferred, and is the only method allowed in the current standard IEC 61000-4-3. This involves pre-calibrating the empty chamber or cell by measuring, at each frequency, the power required to generate a given field strength. The EUT is then introduced and the same power is applied. The rationale for this method is that any disturbances in field caused by the EUT are taken at face value, and no attempt is made to correct for them by monitoring the actual field at the EUT; instead the field which would be present in the *absence* of the EUT is used as the controlled parameter. The method is only viable as long as the field uniformity is closely defined (see section 7.1.2.2), but in these circumstances it is much preferable. The parameter which is best controlled in the pre-calibration is the amplifier output power (forward power) rather than the net power supplied to the antenna; this is acceptable provided that the antenna characteristics are not significantly changed with the introduction of the EUT, which in turn dictates as great a separation distance as possible.

7.1.1.4 Transducers

The radiated field can be generated by an antenna as already discussed. You may well want to use the same antennas as you have for radiated emissions tests, and in principle this is perfectly acceptable. The power handling ability of these antennas is limited by the balun transformer which is placed at the antenna's feed point. This is a wideband ferrite cored 1:1 transformer which converts the *bal*anced feed of the dipole to the *un*balanced connection of the coax cable (hence bal-un). It is supplied as part of the antenna and the antenna calibration includes a factor to allow for balun losses, which are usually very slight. Nevertheless, some of the power delivered to the antenna ends up as heat in the balun core and windings, and this sets a limit to the maximum power the antenna can take.

The high VSWR of broadband antennas (see section 6.5.2.1 and Appendix D section D.2.4), particularly of the biconical at low frequencies, means that much of the feed power is reflected rather than radiated, which accounts for the poor efficiency at these frequencies. Figure 7.4 shows a typical VSWR versus frequency plot for three types of BiLog. Much effort has been put into antenna development for immunity testing and the curves for the extended (X-Wing) models show the advances that have been made. As with radiated emissions testing, the plane polarization of the antennas calls for two test runs, once with horizontal and once with vertical polarization.

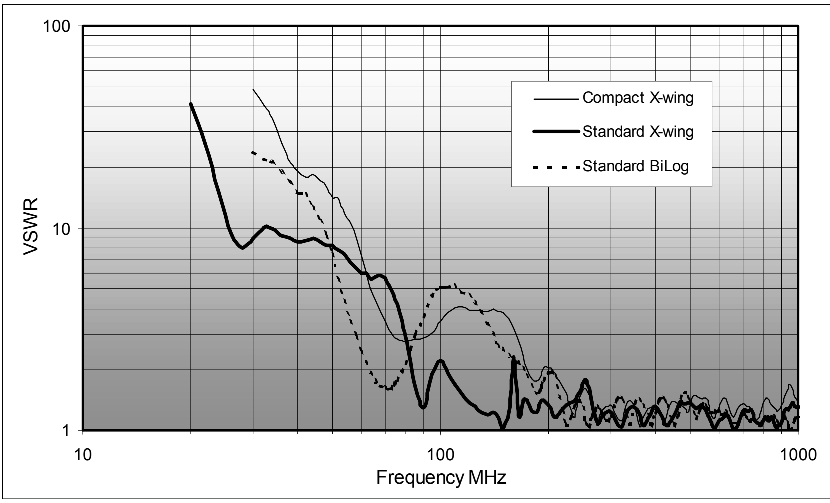


Figure 7.4 VSWR of BiLog antennas (Source: Schaffner EMC)

Two other types of transducer are available for radiated RF immunity testing of small EUTs. These are the stripline and the TEM cell.

Stripline

The difficulties of testing with antennas led to developments in the 1970s of alternative forms of irradiation of the EUT. Groenveld and de Jong [77] designed a simple transmission line construction which provides a uniform electromagnetic field between its plates over a comparatively small volume, and this was written into both IEC 801-3 (1984) and EN 55020 as a recommended method of performing part of the radiated immunity testing. It doesn't appear in the latest (third) edition of IEC 61000-4-3.

The stripline is essentially two parallel plates between which the field is developed, fed at one end through a tapered matching section and terminated at the other through an identical section. The dimensions of the parallel section of line are defined in the standards as 80 x 80 x 80cm, and the EUT is placed within this volume on an insulating support over one of the plates (Figure 7.5). The field between the plates is propagated in TEM (transverse electro-magnetic) mode. The calibration of the stripline is theoretically very simple: assuming proper matching, the field is directly proportional to the voltage at the feed point divided by the distance between the plates:

$$E = V/h \quad \text{volts per metre} \tag{7.2}$$

In practice some variations from the ideal are likely, and calibration using a short probe extending into the test volume is advisable. If the stripline test is conducted in a screened room reflections from the walls will disturb the propagation characteristics quite severely, as they do with antennas, and you will have to surround the stripline with absorbing plates to dampen these reflections. This will be cheaper than lining the walls with anechoic absorber.

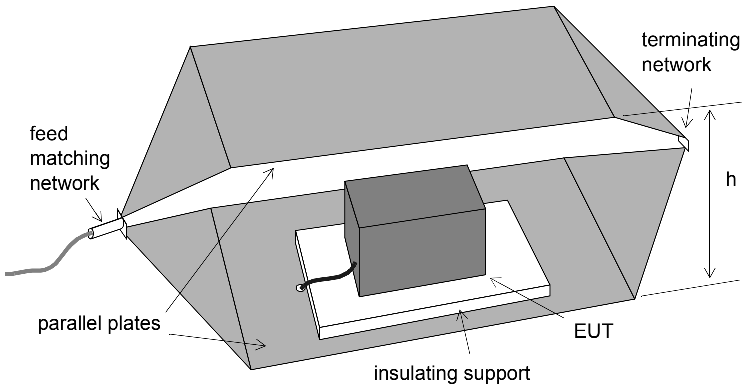


Figure 7.5 The stripline

The accuracy of the stripline's applied field depends to a large extent on the dimensions of the EUT. IEC 801-3 recommends that the dimensions should not exceed 25cm, while EN 55020 allows a height up to 0.7m with a calibration correction factor. Either way, you can only use the stripline on fairly small test objects. There is also an upper frequency restriction of 150–200MHz, above which the plate spacing is greater than a half-wavelength and the transmission mode becomes complex so that the field is subject to variability. It would be quite possible though to use the stripline for immunity testing below 200MHz (theoretically down to DC if required) along with a log periodic antenna above 200MHz, to get around the unsuitability of the biconical for low frequency immunity tests. The power requirement of the stripline for a field strength of 10V/m is no more than a few watts.

A particular characteristic of testing with the stripline is that the connecting cables for the EUT are led directly through one of the plates and are not exposed to the field for more than a few centimetres. Thus it only tests for direct exposure of the enclosure to the field, and for full immunity testing it should be used in conjunction with common mode conducted current or voltage injection. Also, you will need to be able to reorient the EUT through all three axes to determine the direction of maximum susceptibility.

The TEM cell

An alternative to the stripline for small EUTs and low frequencies is the TEM or Crawford cell. In this device the field is totally enclosed within a transmission line structure, and the EUT is inserted within the transmission line. It is essentially a parallel plate stripline in which one of the plates has been extended to completely enclose the other. Or, you can think of it as a screened enclosure forming one half of the transmission line while an internal plate stretching between the sides forms the other half. The transverse electromagnetic mode is defined as a waveguide mode in which the components of the E and H fields in the direction of propagation are much less than the primary field components across any transverse cross-section; it shares this property with the free-space plane wave.

The advantage of the TEM cell, like the stripline, is its small size, low cost and lack of need for high power drive; it can easily be used within the development lab. A further advantage, not shared with the stripline, is that it needs no further screening to attenuate external radiated fields. The disadvantage is that a window is needed in the enclosure if you need to view the operation of the EUT while it is being tested, if for example it is a television set or a measuring instrument. It is not so suitable for do-it-yourself construction as the stripline. As with the stripline, it can only be used for small EUTs (dimensions up to a third of the volume within the cell, see Table 7.1), and it suffers from a low upper frequency limit. If the overall dimensions are increased to allow larger EUTs, then the upper frequency limit is reduced in direct proportion.

Table 7.1 TEM cell dimensions versus frequency range

Cell size cm^2	Maximum EUT size W x D x H cm	Frequency range
30.5	15 x 15 x 5	DC – 500MHz
61	20 x 20 x 7.5	DC – 300MHz
91.5	30.5 x 30.5 x 10	DC – 200MHz
122	40.5 x 40.5 x 15	DC – 150MHz
183	61 x 61 x 20	DC – 100MHz

7.1.1.5 The GTEM

The GTEM cell [70][78] overcomes some of these disadvantages and holds out the promise of lower cost, well-defined testing. The restriction on upper frequency limit is removed by tapering the transmission line continuously outward from feed point to termination, and combining a tapered resistive load for the lower frequencies with an anechoic absorber load for the higher frequencies. This allows even large cells, with test volume heights up to 1.75m and potentially larger, to be made with a useable upper frequency exceeding 1GHz (hence the "G" in GTEM). The actual unit looks from the outside something like a pyramid on its side. Its use for emissions testing has already been discussed in section 6.2.4.

The GTEM has clear advantages for immunity testing since it allows the full frequency range to be applied in one sweep, without the need for a screened enclosure – or for high power amplifiers, since its efficiency is much higher than an antenna. As with the other TEM methods, the EUT must be subjected to tests in a number of orthogonal orientations, and cable dressing needs to be considered carefully. A feature of TEM cells is the intentionally transverse nature of the field, but at some frequencies it has been shown [69] that the field distribution in a GTEM includes a large longitudinal component (Figure 7.6). The amplitude and frequency of this component depends on the size of the cell and the position along the length at which it is measured. In the graph, 0.0m refers to a position opposite the centre of the door, –1.0m is close to the absorber, and +1.0m is towards the apex. The existence of a field in this orientation means that if the field strength is controlled only on the primary (vertical) component, there is the likelihood of over-testing or at least variability in the actual test field at such frequencies.

Its advantages are attractive particularly in terms of allowing one relatively inexpensive (c. £50,000) facility to perform all RF EMC testing. Considerable

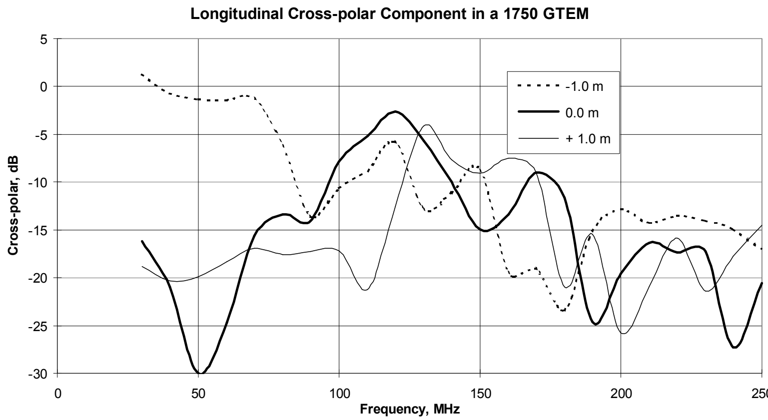


Figure 7.6 Longitudinal field components in the GTEM (dB with respect to vertical component) (Source: NPL [69] © Crown Copyright 2000. Reproduced by permission of the Controller of HMSO)

resources have been put into characterizing the GTEM's operation and in persuading the standards authorities to accept it as an alternative test method. This has resulted in a new document, IEC 61000-4-20, which covers both emissions and immunity tests in such cells. A large part of the work in producing this standard has involved finding acceptable solutions to the field uniformity and cable layout problems outlined above, that are consistent with existing test methods in screened chambers. The first edition of IEC 61000-4-20 gives a method for calibrating the uniform area and requires that the secondary electric field components are at least 6dB less than the primary component, over at least 75% of the uniform area. But it states that the test methods for "large EUTs" (i.e. those including cables) are under consideration; as with the emissions testing requirements, effectively that issue has yet to be dealt with.

7.1.2 Facilities

RF immunity testing, like radiated emissions testing, cannot readily be carried out on the development bench. You will need to have a dedicated area set aside for these tests – which may be in the same area as for the emissions tests – that includes the RF field generating equipment and, most importantly, has a screened room.

7.1.2.1 The screened room

RF immunity tests covering the whole frequency bands specified in the standards should be carried out in a screened room to comply with various national regulations prohibiting interference to radio services. Recommended shielding performance is at least 100dB attenuation over the range 10MHz to 1GHz [149]; this will reduce internal field strengths of 10V/m to less than 40dBμV/m outside. The shielding attenuation depends on the constructional methods of the room in exactly the same way as described for shielded equipment enclosures in section 14.2. It is quite often possible to trade off performance against reduced construction cost, but a typical high-performance room will be built up from modular steel-and-wood sandwich panels, welded or clamped together. Ventilation apertures will use honeycomb panels; the room will be windowless. All electrical services entering the chamber will be filtered. Lighting will be by incandescent lamps as fluorescent types emit broadband interference. The access door construction is critical, and it is normal to have a double wiping action "knife-edge" door making contact all round the frame via beryllium copper finger strip.

In addition, the screened room isolates the test and support instrumentation from the RF field. The interconnecting cables leaving the room should be suitably screened and filtered themselves. A removable bulkhead panel is often provided which can carry interchangeable RF connectors and filtered power and signal connectors. This is particularly important for a test house whose customers may have many and varied signal and power cable types, each of which must be provided with a suitable filter. As well as for RF immunity tests, a screened room is useful for other EMC tests as it establishes a good ground reference plane and an electromagnetically quiet zone. Figure 7.7 shows the features of a typical screened chamber installation.

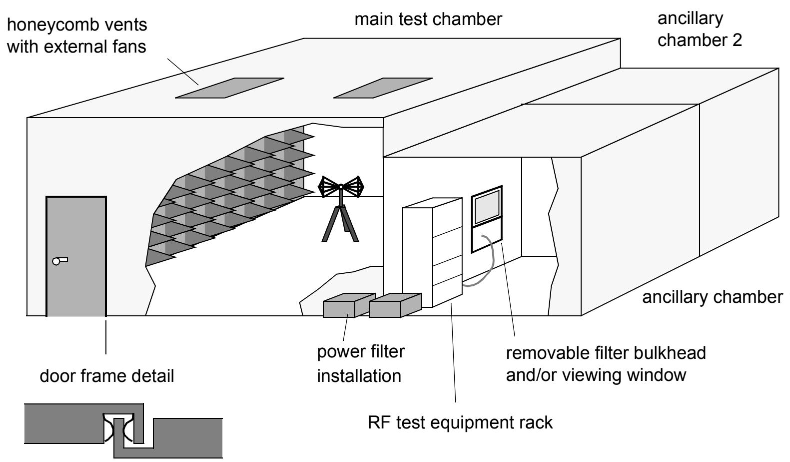


Figure 7.7 Typical screened room installation

7.1.2.2 Room resonances and field uniformity

An unlined room will exhibit field peaks and nulls at various frequencies determined by its dimensions. The larger the room, the lower the resonant frequencies; equation (14.3) on page 386 gives the lowest resonant frequency. For a room of 2.5 x 2.5 x 5m this works out to around 70MHz.

To damp these resonances the room can be lined with absorber material to reduce wall reflections, either carbon loaded foam shaped into pyramidal sections, or ferrite tiles. The room is then said to be "fully anechoic" if all walls and floor are lined, or "semi-anechoic" if the floor is left reflective. The use of screened chambers for emissions measurements is covered in section 6.3.1.3, and the same considerations regarding anechoic absorber material apply here. For radiated immunity tests, a fully anechoic chamber is needed; if you are using the same chamber for emissions and immunity, you will need to move the floor absorber in and out on a regular basis.

Field uniformity

A serious effect of these resonances is that they cause standing waves in the field distribution throughout the chamber. At the higher frequencies these standing waves can result in significant variation in the field strength over quite a small volume, certainly smaller than is occupied by the EUT. As a practical measure of the effectiveness of anechoic lining, and to calibrate the field strength that will be used in the actual test, IEC 61000-4-3 specifies a test of the field uniformity to be made at 16 points over a grid covering a plane area.

The measurements are made in the *absence* of the EUT but with the chamber, especially the antenna and absorber positions, set up as it would be in the test itself; the grid corresponds to the intended position of the front face of the EUT. The field strength over at least 75% (i.e. 12) of the grid points must be within the tolerance –0dB/+6dB to be acceptable, though a tolerance up to +10dB is allowed for no more than 3% of test frequencies provided it is stated in the test report. The 0dB reference is the field strength specified for exposure of the EUT; it may, and usually does, occur at a different position in the 16-point grid for each different frequency. The tolerance is quoted in this asymmetrical way to ensure that the applied field strength is never less than the stated level, but it does imply that over-testing by up to a factor of two (+6dB) at any other point in the grid is possible. This is not a mistake; it is the best that can be achieved with a reasonable-sized chamber. Figure 7.8 shows the geometry of the recommended field uniformity criterion.

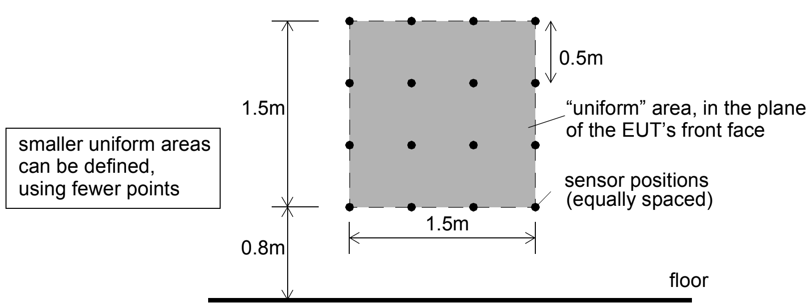


Figure 7.8 Field uniformity measurement in IEC 61000-4-3

The actual test then uses the same forward power values that have been used to create the uniform field calibration, with the EUT in place. This is why it is called a "substitution" method. No closed loop control of the *field strength* level takes place during testing, only forward power is controlled.

7.1.2.3 Ancillary equipment

You will need a range of support equipment in addition to the RF test equipment described in detail in section 7.1.1. Obviously, control and data capture computing equipment will be required for a comprehensive set-up. Various test jigs and coupling networks, depending on the type of EUT and the detail of the standards in use, must be included. Beyond that, some form of communication will be needed between the inside of the screened room and the outside world. This could take the form of RFI-proof CCTV equipment, intercoms or fibre optic data communication links.

The ancillary equipment housed outside the screened room will also include all the support equipment for the EUT. Test houses will often have two subsidiary screened chambers abutting the main one, one of which contains the RF test instrumentation, the other housing the support equipment. This ensures that there is no interaction between the external environment, the RF instrumentation and the support equipment. Provided the environment is not too noisy and the RF instrumentation is individually well screened, you do not really need these two extra screened chambers for your own EMC testing.

7.1.3 Test methods

As with radiated emissions, the major concern of standardized immunity test methods is to ensure repeatability of measurements. The immunity test is complicated by not having a defined threshold which indicates pass or failure. Instead, a (hopefully) well-defined level of interference is applied to the EUT and its response is noted. The test procedure concentrates on ensuring that the applied level is as consistent as possible and that the means of application is also consistent.

Radiated field immunity testing, in common with radiated emissions testing, suffers from considerable variability of results due to the physical conditions of the test set-up. Layout of the EUT and its interconnecting cables affects the RF currents and voltages induced within the EUT to a great extent. At frequencies where the EUT is electrically small, cable coupling predominates and hence cable layout and termination must be specified in the test procedure.

7.1.3.1 Preliminary checking

You will need to carry out some preliminary tests to find the most susceptible configuration and operating mode of the EUT. If it is expected to pass the compliance test with a comfortable margin, you may need to apply considerably greater field strengths in order to deliberately induce a malfunction. Hopefully (from the point of view of the test), with the initially defined set-up and operation there will be some frequency and level at which the operation is corrupted. This is easier to find if the EUT has some analogue functions, which are perhaps affected to a small degree, than if it is entirely digital and continues operating perfectly up to a well-defined threshold beyond which it crashes completely.

Once a sensitive point has been found, you can vary the orientation, cable layout, grounding regime and antenna polarization to find the lowest level which induces a malfunction at that frequency. Similarly, the operating mode can be changed to find the most sensitive mode. It is often worthwhile incorporating special test software to continuously exercise the most sensitive mode, if this is not part of the normal continuous operation of the instrument. Note that some changes may do no more than shift the sensitive point to a different frequency, so you should always repeat a complete frequency sweep after any fine-tuning at a particular frequency.

7.1.3.2 Compliance tests

Once the sensitive configuration has been established it should be carefully defined and rigorously maintained throughout the compliance test. Changes in configuration halfway through will invalidate the testing. If there are several sensitive configurations these should be fully tested one after the other.

The test set-up

Notwithstanding this, equipment should always be tested in conditions that are as close as possible to a typical installation – that is with wiring and cabling as per normal practice, and with hatches and covers in place. If the wiring practice is unspecified, leave a nominal length of 1m of cable "exposed to the incident field" as the standard says: this can be interpreted as leaving this length between the EUT and the floor, and/or applying ferrite clamp absorbers at this distance from the EUT.

If the EUT is floor-standing (such as a rack or cabinet) it will be placed on but insulated from the floor, otherwise it should be on a non-conductive, preferably plastic 0.8m high table. The antenna will normally be placed at least 1m from it, at a greater distance if possible consistent with generating an adequate field strength; the preferred distance is 3m. Too close a distance affects the uniformity of the generated field and also, because of mutual coupling between antenna and EUT, invalidates the basis on which the substitution method is used.

Running the test

During the compliance test the specified test level is maintained throughout the frequency sweep. This will be achieved by controlling the forward power to replicate the field uniformity calibration level. The parameters which have been chosen to represent the operation of the EUT must be continuously monitored throughout the sweep, preferably by linking them to an automatic data capture and analysis system – although the test engineer's eyeball still remains one of the most common monitoring instruments. For most EUTs, eight sweeps are needed: two for each of the four faces of the EUT, once in horizontal polarization and once in vertical. A different calibration file is needed for the two polarizations. If the EUT can be used in any orientation (e.g. hand-held equipment) then all six faces must be exposed, and twelve sweeps are necessary.

Assuming that the EUT remains correctly operational throughout the sweep, it can be useful to know how much margin there is in hand at the sensitive point(s). You can do this by repeating the sweep at successively higher levels and mapping the EUT's response. This will indicate both the margin you can allow for testing uncertainty and production variability, and the possibilities for cost reduction by removing suppression components.

7.1.3.3 Testing above 1GHz

The third edition of IEC 61000-4-3 states that tests from 80 to 1000MHz are "related to general purposes"; for the frequency ranges 800 to 960MHz and 1.4 to 6GHz they are "related to the protection against RF emissions from digital radio telephones and other RF emitting devices". Meanwhile, a number of product and generic standards have started demanding immunity testing up to at least 2GHz and so the test method has to be extended above 1GHz.

Equipment

The same basic system configuration is used irrespective of frequency, but signal sources, amplifiers and antennas have to cover the highest frequency to be tested. This will normally need investment in at least one extra amplifier and antenna, although some BiLogs can cover the range up to 2GHz. You need to make sure that other equipment such as power meters, directional couplers and field strength sensors are suitable for the higher frequencies. You may also find that cable losses become unacceptably high, forcing investment in new cables.

Method

Even with a test distance of 3 m, using an antenna with a narrow beam width or a ferrite-lined chamber at frequencies above 1GHz, you may not be able to satisfy the field uniformity requirement over the 1.5 m × 1.5 m calibration area. The standard gives an alternative method (the "independent windows method") for frequencies above 1GHz, which divides the calibration area into an array of 0.5 m × 0.5 m windows such that the whole area to be occupied by the face of the EUT is covered. The field uniformity and field strength level is independently calibrated over each window, using a variation of the procedure given for tests below 1GHz. The field generating antenna is placed 1m from the calibration area and repositioned during the test to illuminate each of the required windows in turn.

Cable length and geometry are less critical at these high frequencies; therefore, the face area of the EUT is the determining factor for the size of the calibration area. On the other hand, maintaining the position of the antenna and the uniform window areas in the chamber is more critical, since even small displacements will significantly affect the field distribution.

7.1.3.4 Sweep rate, step size and modulation

The sweep rate of the applied field may be critical to the performance of the EUT. According to earlier versions of IEC 61000-4-3, the signal generator should either be manually or automatically swept across the output range at $1.5 \cdot 10^{-3}$ decades per second or slower, depending on the speed of response of the EUT, or automatically stepped at this rate in steps of 1% – that is, each test frequency is 1.01 times the previous one, so that the steps are logarithmic. The dwell time for stepped application should be at least enough to allow time for the EUT to respond; slow responses translate directly to a longer test time. As an example, to cover the range 80–1000MHz with a step size of 1% and a dwell time of 3s takes 12.7 minutes.

The third edition recognizes that sweeping has been almost entirely replaced by stepping and has explicitly removed the sweep rate limit of $1.5 \cdot 10^{-3}$ decades per second. Instead, it mandates a minimum dwell time of 0.5 seconds with AM on at each step, but retains the need for the EUT to be exercised and to respond. It would be rare to be able to achieve this at 0.5 seconds with most EUTs and in practice a default dwell time of 2 to 3 seconds is advisable. In fact, this doesn't slow down the total test time appreciably since many test systems take several seconds to step from one frequency to the next and re-establish the correct level; this has to be done with AM off since power meters respond incorrectly to a modulated signal. The typical software-controlled sequence is:

- step to the next frequency with modulation off;
- set and check the correct power level with modulation off;
- apply modulation for the desired dwell time;
- set modulation off and step to the next frequency, etc... .

For many systems there may be little sensitivity to sweep rate or step size since demodulation of applied RF tends to have a fairly broad bandwidth; usually, responses are caused by structural or coupling resonances which are low-Q and therefore several MHz wide. On the other hand, some frequency-sensitive functions in the EUT may have a very narrow detection bandwidth so that responses are only noted at specific frequencies. This may easily be the case, for instance, with analogue-to-digital converters operating at a fixed clock frequency, near which interfering frequencies are aliased down to the baseband. If the step spacing is too great then a response may be missed. Such narrowband susceptibility may be many times worse than the broadband response. Therefore some knowledge of the EUT's internal functions is essential, or considerably more complex test procedures are needed. IEC 61000-4-3 says "The sensitive frequencies (e.g., clock frequencies) shall be analysed separately according to the requirements in product standards" but the product standards rarely if ever say anything about this aspect, so it is up to the test plan to determine such details.

7.1.3.5 Safety precautions

At field strengths not much in excess of those defined in many immunity standards, there is the possibility of a biological hazard from the RF field arising to the operators if they remain in the irradiated area for an appreciable time. For this reason a prudent test facility will not allow its test personnel inside a screened chamber while a test is in progress, making it necessary for a remote monitoring device (such as a CCTV system) to be installed for some types of EUT.

Health and safety legislation differs between countries. In the UK at the time of writing there are no *mandatory* requirements placed on maximum permissible RF field exposure, but in 1998 the International Commission on Non-Ionizing Radiation Protection (ICNIRP) published a set of guidelines [196] covering exposure to RF radiation. The ICNIRP guidelines for the general public have been incorporated in a European Council Recommendation (1999) [190], which has been agreed in principle by all countries in the European Union (EU), including the UK. The occupational exposure guidelines are incorporated into the Physical Agents (Electromagnetic Fields) Directive 2004/40/EC [189], which comes into force in 2008.

The ICNIRP guidelines take into account the known thermal and electric shock effects of RF fields. They do not consider possible athermal effects, which is a highly controversial field of study and for which no firm guidance has yet been produced. The guidelines contained in [196] for occupational exposure to continuous fields over the frequency range of interest for RF immunity testing are reproduced in Table 7.2.

Table 7.2 ICNIRP guidelines for maximum field strength exposure, occupational

Frequency range	Electric field strength
0.065 to 1MHz	610 V/m
1 to 10MHz	610 / F (MHz) V/m
10 to 400MHz	61 V/m
400MHz to 2GHz	$3 \cdot \sqrt{F}$ (MHz) V/m
2 to 300GHz	137 V/m

7.1.3.6 Short cuts in immunity testing

There will be many firms which decide that they cannot afford the expense of a full RF immunity set-up, including a screened room, as described in section 7.1.2. One possibility for reduced testing is to restrict the test frequencies to the "free radiation" frequencies as permitted by international convention, on which unrestricted emissions are allowed. These are primarily intended for the operation of industrial, scientific and medical equipment and are listed in Table 1.1 on page 12. Another course known to be taken by some firms is to use the services of a licensed radio amateur transmitting on the various amateur bands available to them – 30MHz, 50MHz, 70MHz, 144MHz and 432MHz; although on a strict interpretation this is outside the terms of the amateur radio licence. Yet another possibility is to use an actual cellular telephone transmitting on 900MHz to check for immunity to this type of signal.

In each case the use of particular frequencies removes the need for a screened room to avoid interference with other services. All of these ad hoc tests should at least use a field strength meter to confirm the field actually being applied to the EUT; bear in mind that the RF field near to the transmitting antenna will vary considerably with small changes in separation distance. If the EUT's response to RF interference was broadband across the whole frequency range then spot frequency testing would be adequate, but this is rarely so; resonances in the coupling paths emphasize some frequencies at the expense of others, even if the circuit response is itself broadband. It is therefore quite possible to believe an optimistic performance of the EUT if you have only tested it at discrete frequencies, since resonant peaks may fall between these. A compliance test must always cover the entire range.

Transient testing

In practice, it has been found that for many digital products transient and ESD performance is linked to good RF immunity, since susceptible digital circuits tend to be sensitive to both phenomena. Therefore much development work can proceed on the basis of transient tests, which are easier and less time-consuming to apply than RF tests, and are inherently broadband. Where analogue circuits are concerned then a proper RF field test is always necessary, since the demodulated offset voltage which RF injection causes cannot be simulated by a transient. But a minimal set of transient plus spot frequency RF tests may give you an adequate assessment of the product's immunity during the development stages.

7.1.4 Conducted RF immunity

The basic standard IEC 61000-4-6 defines the test method for conducted immunity testing, and it is referred to in tandem with the radiated field test in the generic standards and in product standards. In its most usual implementation it covers the frequency range from 150kHz to 80MHz. The immunity standard for broadcast receivers, EN 55020, also defines test methods for immunity from conducted RF currents and voltages. The method of bulk current injection (BCI) developed within the aerospace and military industries for testing components of aircraft systems, and adapted for application to automotive components, is a related technique. See also section 10.3.1.2.

7.1.4.1 Coupling methods

Three methods of coupling are defined in IEC 61000-4-6. The transducers for each are illustrated in Figure 7.9.

CDN

The preferred method is voltage injection via a coupling/decoupling network (CDN). This has zero insertion loss and therefore needs little power.

Any method of cable RF injection testing requires that the common mode impedance at the end of the cable remote from the EUT is defined. Thus each type of cable must have a common mode decoupling network or impedance stabilizing network

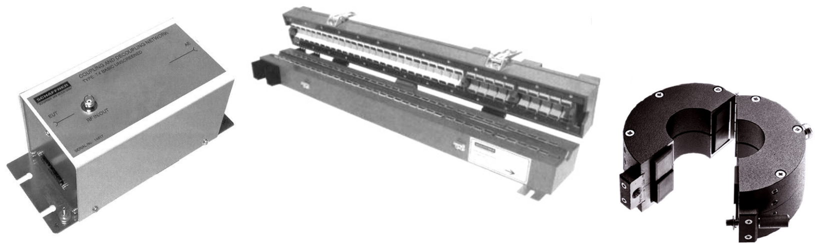


Figure 7.9 The CDN, the EM-clamp and the current probe

at its far end, to ensure this impedance and to isolate any ancillary equipment from the effects of the RF current on the cable. (This is analogous to the mains LISN used for emission testing and discussed in section 6.2.2.1. Unfortunately, the emissions LISN specification doesn't agree with that for the conducted immunity CDN, so different units are needed.) Direct voltage injection in addition requires that this network is used to couple the RF voltage onto the cable, and at the same time present a common mode source impedance of a known 150Ω. This means that the CDN is inherently invasive.

A test house which handles these methods must have a wide range of CDNs available, to cater for the variety of different cable and signal types that will come its way. If you have an in-house test lab and your company makes equipment which predominantly uses only one or two types of cable – say single-channel RS-232 data links and mains – then this is not an onerous requirement. For generalist test houses, the method is most often used for injection only onto the mains power supply port.

EM-clamp

Like the absorbing clamp used for measuring disturbance power (section 6.2.2.2), the EM-clamp consists of a tube of split ferrite rings which can be clamped over the cable to be tested, and it is therefore non-invasive and can be used on any cable type. Unlike the absorbing clamp, it provides both inductive and capacitive coupling and can be used down to 150kHz.

The signal is fed in via a single-turn loop which extends the whole length of the clamp. This loop is terminated at each end in an impedance which creates a voltage along the loop next to the cable, as well as allowing a current to flow in the loop. The voltage gives capacitive coupling and the current gives inductive coupling to the cable. The ferrite sleeve is composed of a low-μ (low permeability) grade at the EUT end and a high-μ grade at the AE end. The combination of graded ferrite and capacitive/inductive coupling gives the clamp significant directivity, particularly above 10MHz, so that substantially less signal is applied to the AE end of the cable, and the common mode impedance seen by the EUT is quite close to 150Ω across a large part of the frequency range of the test signal.

As with the CDN, the EM-clamp should be properly bonded to the ground plane to give a repeatable impedance. But also as with the CDN, variations due to cable layout on the AE-side of the test set-up, and due to the AE itself, are minimized. The coupling loss of the clamp is low enough that it does not require very much more power than a CDN for comparable stress levels.

Current injection probe

The current injection probe is an alternative to both the EM-clamp and the CDN. It is less effective than either, but is more convenient to use. The current probe is essentially a clip-on current transformer which can be applied to any cable. It is shielded, and so applies only inductive coupling, without capacitive coupling of the test signal. It has been in common use in military and automotive testing (the bulk current injection, BCI test) for many years and has been included in IEC/EN 61000-4-6 since many test laboratories are familiar with it, but this has resulted in some anomalies with respect to setting the injected level.

A principal disadvantage of the current probe is that it gives no isolation from the associated equipment (AE) end of the cable, and no control of the cable common mode impedance. The current will flow in the cable according to the ratio of the common mode impedances provided by the EUT and the AE, and at the higher frequencies, according to the cable resonances. At 80MHz, 94cm is all that is necessary to give a quarter wavelength of cable; despite the strictures in the standard, shorter lengths are impractical, and the test is often performed on much longer cables. The actual stress current applied to the EUT is therefore very variable and also very hard to repeat, because of its dependence on AE and cable impedances. At the same time, the stress current also flows through the AE, which must therefore be at least as immune as the EUT. The standard requires the AE impedance to be set to 150Ω, but recognizing that this is often impractical, it provides for a modified method whereby the level is monitored by a secondary probe and limited if it increases above the intended value.

The current probe should only be used if all other methods are either impractical or unavailable. It is best suited to system-level injection where the AE and cable layout are fixed and known, and the physical limitations make it difficult to apply CDNs or the EM-clamp. A further disadvantage is that because of the higher coupling loss, the power required for a given stress is greater for the current probe than for any other method.

Test set-up

Figure 7.10 shows the general arrangement for making conducted immunity tests. The EUT is supported by an insulator 0.1m above the ground plane. Cables leaving the EUT in close proximity or in conduit are treated as one cable. The AE impedance on the tested cable should be stabilized at 150Ω; for the EM-clamp or current probe methods, this requires extra precautions at the AE itself.

7.1.4.2 Procedure

The open circuit test level V_{stress} (NB: not the level actually applied to the EUT) is usually 3V or 10V. An attenuator of at least 6dB is placed between the power amplifier and the transducer to prevent the power amplifier's output impedance, which is poorly defined, from affecting the results; the 50Ω offered by the output impedance forms part of the 150Ω source impedance for the test stress. As with the radiated test, the applied level is actually set by substitution.

There are two basic calibration jigs, a 150Ω one for the CDN or EM-clamp and another at 50Ω for the BCI probe. The object in either case is to terminate the transducer in a known impedance and then to set the stress level applied into that impedance. The power required to give this level is then repeated in the actual test. For the 150Ω system the required power level must give a reading of $V_{stress}/6$, or V_{stress} – 15.6dB. For the 50Ω system it should be $V_{stress}/2$, or V_{stress} – 6dB. The factor of 2 is needed because

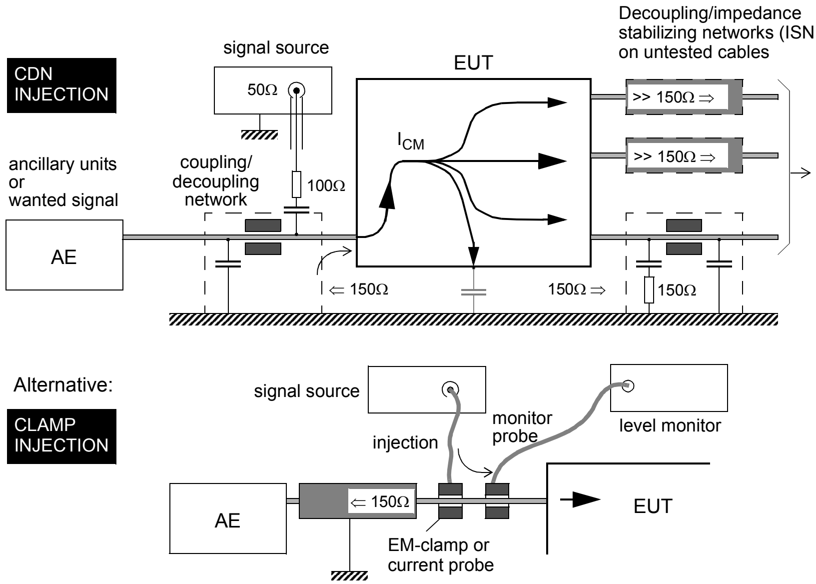


Figure 7.10 Conducted immunity test set-up

the stress voltage is given as an open circuit value, and the factor of 3 in the 150Ω system is needed because of the effect of the 50-to-150Ω resistive divider. In both cases, the level can be measured by any RF measuring device, for instance a power meter or spectrum analyser.

Having calibrated the transducer for a particular level, this is then applied to the cable to be tested, and the applied stress is then stepped across the frequency range in 1% steps with an appropriate dwell time and 80% 1kHz modulation, exactly as with the radiated RF immunity test. You then need to repeat the test for all the appropriate cable ports as mandated by the test plan.

7.1.4.3 Disadvantages and restrictions

Conducted immunity testing has the major advantages of not requiring expensive anechoic screened room facilities, and being more efficient for applying the RF stress at the lower frequencies. but it does have some disadvantages. It is particularly questionable whether it accurately represents real situations when there are several cables connected to the EUT. When the whole system is irradiated then all cables would be carrying RF currents, but in most conducted immunity test methods only one is tested at a time. Each of the other cables represents a common mode load on the test system and this must be artificially created by including extra impedance stabilizing networks on them. Networks for direct connection to cables with many signal lines are expensive to construct, bulky and may adversely affect the signal line characteristics, although clamp-on decoupling ferrites are simple and relatively cheap.

Frequency range

The major restriction on conducted immunity testing is one of frequency. For EUT sizes much less than the wavelength of the test frequency, the dominant part of the RF energy passing through equipment that is exposed to a radiated field is captured by its cables, and therefore conducted testing is representative of reality. As the frequency rises so that the EUT dimensions approach a half-wavelength, the dominance of the cable route reduces and at higher frequencies the field coupling path interacts with the EUT structure and internal circuits, as well as with its cables. For this reason the upper frequency limit is restricted in IEC 61000-4-6 to either 80 or 230MHz (corresponding to equipment dimensions of between about 0.6m and 2m). For higher frequencies, radiated testing is still necessary.

7.1.5 Measurement uncertainty for RF immunity

The application of measurement uncertainty for RF immunity tests differs from that for RF emissions, in that the uncertainty budget that a lab can create relates only to the uncertainty of the applied stress, not to that of the test result. It should be clear that if the transfer function between the applied stress and the EUT's response is unknown – and not only is it unknown in detail, but very often the only thing known about it is that it is highly non-linear [142] – then there is no way that an uncertainty can be stated for the outcome of the test. All that the lab can do is to calculate a budget for the uncertainty of the applied level.

What is then done with this value is a matter of some debate. If you simply ignore it and set the stress level to exactly the specification figure, then you have only a 50% level of confidence that the required stress level has been applied, since the actual stress could be either above or below the desired stress. Alternatively, you could increase the programmed stress above the required level by a factor related to the uncertainty value, which will then give a higher degree of confidence; a factor of 1.64 times the standard uncertainty gives a 95% confidence. As outlined in section 9.2.7.3, the CENELEC standards for RF immunity testing have been amended to avoid the increased level. But as a test lab customer, it is your choice as to what level of confidence you require in the testing you have commissioned.

7.1.5.1 Radiated immunity contributions

For radiated RF immunity to IEC 61000-4-3 it is taken as read that the –0dB/+6dB field uniformity criterion, with relaxations as allowed in the standard, has been achieved. This is *not* then a contributor to the uncertainty budget, since it is an inherent aspect of the standard method of test. The remaining uncertainty is that which applies to the actual setting of the field strength level. LAB 34 [195] gives some guidance on this and includes the following contributions:

- calibration of the field strength meter used for the uniform area measurements;
- the extent of the "window" within which the test software will accept a re-established forward power value for each frequency during the test;
- drift in the power meter;
- distortion in the power amplifier, creating harmonic content in the test signal;
- the effects of field disturbance caused by various supports and other extraneous structures in the chamber;
- measurement system repeatability from test to test, for which a value is obtained by analysis of a series of repeated readings.

7.1.5.2 Conducted immunity contributions

The method of IEC 61000-4-6 actually includes three different transducers, and also requires the level to be limited using a monitor probe if either of the clamp methods has to use an uncontrolled AE source impedance. This means that a budget should be developed for each of these situations. The factors that should be considered, for example for the CDN method, are:

- calibration of the measuring device, i.e. RF voltmeter or power meter;
- the extent of the re-established voltage level acceptability window, set in the software;
- drift of the signal generator or power meter, depending on the method used to re-establish the set level;
- distortion in the power amplifier, creating harmonic content in the test signal;
- mismatch error between the CDN and the measuring device (RF voltmeter or power meter), including the effect of the 50-to-150Ω adaptor;
- mismatch error between amplifier with 6dB attenuator and the CDN;
- measurement system repeatability.

7.2 ESD and transient immunity

By contrast with RF testing, ESD (electrostatic discharge) and transient test methods are rather less complicated and need less in the way of sophisticated test equipment and facilities. Nevertheless the bandwidth of fast transients and of the electrostatic discharge is very wide and extends into the VHF region, so many precautions that are necessary for RF work must also be taken when performing transient tests.

7.2.1 ESD

7.2.1.1 Equipment

The electrostatic discharge generator described in IEC 61000-4-2 is fairly simple. The circuit is shown in Figure 7.11. The main storage capacitor C_s is charged from the high-voltage power supply via R_{ch} and discharged to the EUT via R_d and the discharge switch. The switch is typically a vacuum relay under the control of the operator. Compliance testing uses single discharges, but for exploratory testing the capability of a fast discharge rate of 20 per second is suggested. The standard describes two modes of application, contact and air discharge. The output voltage should reach 8kV for contact discharge, or 15kV if air discharge is included, although for the tests required in the present immunity standards lower voltages are specified. Product and generic standards for most environments have settled on a level of 4kV for the contact method and 8kV for air.

The critical aspect of the ESD generator is that it must provide a well-defined discharge waveform with a rise time of between 0.7 and 1 nanosecond. This implies that the construction of the circuit around the discharge electrode is important; C_s, R_d and the discharge switch must be placed as close as possible to the discharge electrode, which itself has specified dimensions. A round tip is used for air discharge, and a sharp tip for contact. The distributed capacitance and inductance of the electrode and

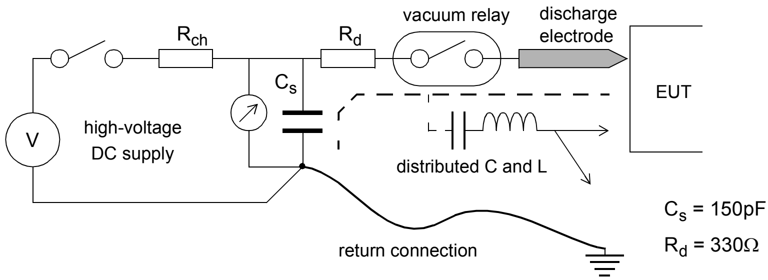


Figure 7.11 ESD generator (according to IEC 61000-4-2)

associated components forms part of the discharge circuit and essentially determines the initial rise time, since the return connection to the EUT is relatively long (2m) and its inductance blocks the initial discharge current. As these distributed parameters cannot be satisfactorily specified, the standard requires that the generator's waveform is calibrated in a special test jig using an oscilloscope with a bandwidth of at least 1GHz.

If you use a ready-built ESD generator this calibration will have already been done by the manufacturer, though it should be re-checked at regular intervals. If you build it yourself you will also have to build and use the calibration jig.

7.2.1.2 Test set-up

Because of the very fast edges associated with the ESD event, high frequency techniques are essential in ESD testing. The use of a ground reference plane is mandatory; this can of course be the floor of a screened room, or the same ground plane that you have installed for the tests outlined in section 6.3. You may want to apply ESD tests to equipment after it has been installed in its operating environment, in which case a temporary ground plane connected to the protective earth should be laid near to the equipment. Other co-located equipment may be adversely affected by the test, so it is wise not to carry out such tests on a "live" operating system.

For laboratory tests, the EUT should be set up in its operating configuration with all cables connected and laid out as in a typical installation. The connection to the ground is particularly important, and this should again be representative of installation or user practice. Tabletop equipment should be placed on a wooden table 80cm over the ground plane, with a horizontal coupling plane directly underneath it but insulated from it. Floor standing equipment should be isolated from the ground plane by an insulating support of about 10cm. Figure 7.12 illustrates a typical set-up. Any ancillary equipment should itself be immune to coupled ESD transients, which may be induced from the field generated by the ESD source/EUT system or be conducted along the connected cables.

7.2.1.3 Discharge application

The actual points of application of the discharges should be selected on the basis of exploratory testing, to attempt to discover sensitive positions. These points should only be those which are accessible to personnel in normal use, and exclude points that are only accessible under maintenance or are inaccessible after installation. Contact pins of

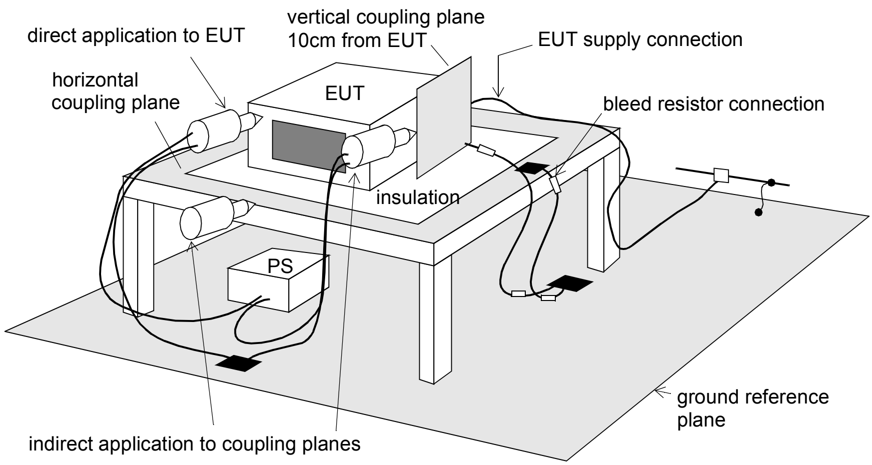


Figure 7.12 ESD test set-up

connectors that have a metallic connector shell should not need to be tested, since the shell will be expected to take the discharge in real situations – but contacts within a plastic connector *should* be tested, by air discharge only, unless there is a rationale which states that the connector will never be left open in normal operation. For the exploratory testing, use a fast repetition rate and increase the applied voltage from the minimum up to the specified test severity level, to find any threshold of malfunction. Also, select both polarities of test voltage. Compliance testing requires the specified number of single discharges with at least 1 second between them, on each chosen point at the specified test level and all lower test levels, with the most sensitive polarity, or (since you may not be able to tell which is the most sensitive) with both polarities.

The ESD generator should be held perpendicular to the EUT surface, or edge-on to the centre edge of the coupling plane. For air discharge, the discharge tip should be brought up to the EUT "as fast as possible without causing mechanical damage": this is not a test which rewards a cautious approach, you need to be vigorous and positive. Both approach speed and angle make a considerable difference to the effectiveness and repeatability of the test.

Contact versus air discharge

Contact discharge is preferred, but this can only be done against conducting surfaces or painted metal surfaces. Here, the voltage is held off the tip by the relay, the tip is applied directly to the contact point and the relay is then triggered to give a single discharge; there is therefore no air gap between the tip and the EUT to contribute to variations in the stress. For a product where this is not possible (e.g. with an overall plastic enclosure) use air discharge to investigate user accessible points where breakdown to the internal circuit might occur, such as the edges of keys or connector or ventilation openings. As the tip approaches the EUT's surface, the increasing electric field gradient causes breakdown either of the air gap, or along the surface of a plastic moulding, just before contact occurs, the distance of the breakdown depending on the applied voltage level.

Indirect discharge

To simulate discharges to objects near to the equipment in its operating environment, and to apply the test in cases where neither direct contact nor air discharge application to the EUT is possible, the discharge is also applied by contact to coupling planes located a fixed distance away from the EUT. This uses both the horizontal coupling plane and the vertical coupling plane shown in Figure 7.12. These planes are emphatically not *ground* planes: they are connected to the true ground reference plane by single wires which include 470kΩ bleed resistors at each end. The bleed resistors disconnect the coupling planes from the ground plane as far as the ESD pulse is concerned, but allow the charge applied to the coupling plane to bleed off within microseconds after the pulse is over, ready for the next one. The resistors are placed at each end so that the connecting wire is decoupled from each plane, a precaution which reduces the effect its position might otherwise have on the stress field waveform.

Ungrounded EUTs

A particular problem arises with EUTs which have no direct connection to the ground of the test set-up: handheld, battery operated devices, or mains powered safety class II devices are typical examples. Amendment A2:2001 to the standard provides for a method of discharging the EUT in between applications of the ESD pulse, in situations where the equipment cannot discharge itself because of this lack. The principal method is through use of a cable with 470kΩ bleed resistors at each end, attached at one end to the HCP or ground plane and at the other end to the point where the ESD pulse is applied. The amendment allows this bleed cable to be in place during the application of the pulse. Alternatively, a conductive earthed brush to discharge the EUT between applications is suggested, or the use of an air ioniser.

If these methods are not used, the EUT will gradually gain charge with each pulse applied during the testing, resulting in a progressive increase in static potential and a consequent reduction in stress over a series of pulses of the same level and polarity; or, if the polarity is reversed, up to twice the intended stress could be applied.

7.2.1.4 Future amendments

A full-scale revision of the standard, to create a second edition, is underway in the IEC and has been for some time. Some of the changes that are proposed in this new version are dramatic [52], and particularly include requirements on the performance of the generator that may result in most existing generators becoming obsolete. At the same time the evidence that the present standard is inherently faulty or inadequate is debatable at best. As a result there is considerable hostility to edition 2 in many quarters, and apparently little chance of it being published in the near future.

7.2.2 Electrical fast transient (EFT) bursts

The EFT-B test checks the immunity of the product against high frequency, low energy transient bursts due to switching events in the near environment.

7.2.2.1 Equipment

When you are testing equipment for immunity to conducted transients the transients themselves, and the coupling network by which the transients are fed into the ports, must be well defined. The network must decouple the side of the line furthest from the EUT and at the same time provide a fixed impedance for the coupling route. In this respect it is similar (but not identical) to the LISN used in emissions testing, and the

CDNs used for conducted RF immunity tests. IEC 61000-4-4 specifies the test generator and the coupling methods for bursts of fast transients such as are caused by local inductive load switching.

The fast transient burst is specified to have a single-pulse rise time/duration of 5ns/50ns from a source impedance of 50Ω. Bursts of 15ms duration of these pulses at a repetition rate of 5kHz are applied every 300ms (see Figure 7.13). The voltage levels are selected depending on specified severity levels from 250V to 4kV. In order to obtain these high voltages with such fast rise times, the generator was originally constructed with a spark gap driven from an energy storage capacitor, which limited the achievable repetition rate, although more modern solid-state generators have for some time superseded this approach. Now that better generators are being built, the second edition of IEC 61000-4-4 requires a burst repetition frequency of 100kHz as well as 5kHz ("Use of 5 kHz repetition rates is traditional; however, 100 kHz is closer to reality"). It also ensures that the waveform is calibrated into both 50 and 1000Ω, which means that the waveform that is actually delivered to the EUT is specified more rigorously [118]. This revision was stalled for some time because of its implications for obsolescence of the older generation of test equipment.

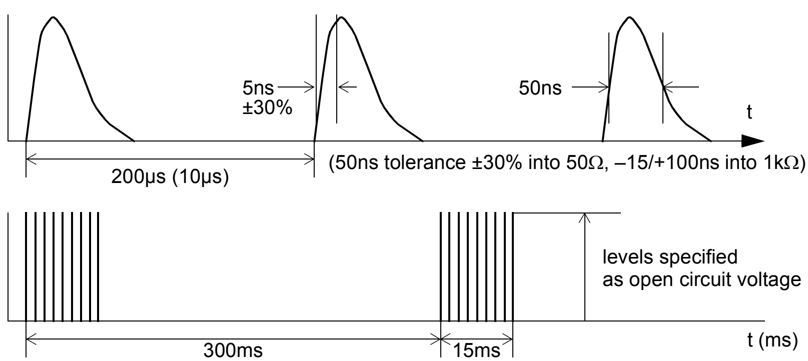


Figure 7.13 Fast transient burst specification (according to IEC 61000-4-4)

The coupling network for power supply lines applies the pulse in common mode with respect to the ground plane to each line via an array of coupling capacitors, while the source of each line is also decoupled by an LC network. Coupling onto signal lines uses a capacitive clamp, essentially two metal plates which sandwich the line under test to provide a distributed coupling capacitance and which are connected to the transient generator. Any associated equipment which may face the coupled transients must obviously be immune to them itself.

7.2.2.2 Test methods

As with ESD tests, a reference ground plane must be used. This is connected to the protective earth, and the generator ground is directly bonded to it with a short strap. Both tabletop and floor standing equipment (in the second edition) is stood off from this ground plane by a 10cm insulating block. A 0.5m length of mains cable connects the EUT to the coupling network, which itself is bonded to the ground plane. If the EUT enclosure has a separate protective earth terminal, this is connected to the ground plane via the coupling network and transients are applied directly to it also. I/O cables are fed through the capacitive clamp which is located 10cm above the ground plane. A typical set-up is shown in Figure 7.14.

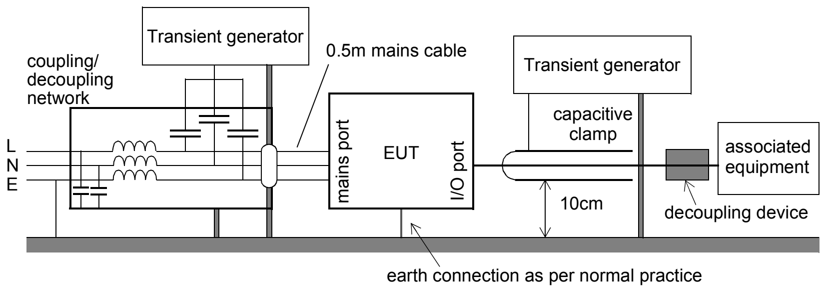


Figure 7.14 Fast transient test set-up

Actual application of the transients is relatively simple, compared to other immunity tests. No exploratory testing is necessary except to determine the most sensitive operating mode of the equipment. Typically, bursts are applied for a duration of 1 minute in each polarity on each line to be tested. Note that the second edition of IEC 61000-4-4 makes explicit that the bursts are applied only in common mode with respect to the ground plane – not to individual mains lines (L, N, or E). The required voltage levels are defined in the relevant product standard, and vary depending on the anticipated operating environment and on the type of line being tested.

7.2.3 Surge

The surge test of IEC 61000-4-5 (Figure 7.15) simulates high energy but relatively slow transient overvoltages on power lines and long signal lines, most commonly caused by lightning strikes in the vicinity of the line.

7.2.3.1 Surge waveform

The transients are coupled into the power, I/O and telecommunication lines. The surge generator called up in the test has a combination of current and voltage waveforms specified, since protective devices in the EUT (or if they are absent, flashover or component breakdown) will inherently switch from high to low impedance as they operate. The values of the generator's circuit elements are defined so that the generator delivers a 1.2/50μs voltage surge across a high-resistance load (more than 100Ω) and an 8/20μs current surge into a short circuit. These waveforms must be maintained into a coupling/decoupling network, but are not specified with the EUT itself connected.

Three different source impedances are also recommended, depending on the application of the test voltage and the expected operating conditions of the EUT. The effective output impedance of the generator itself, defined as the ratio of peak open circuit output voltage to peak short circuit output current, is 2Ω. Additional resistors of 10 or 40Ω are added in series to increase the effective source impedance as necessary.

7.2.3.2 Applying the surge

High energy surges are applied to the power port between phases, and from phase to ground. For input/output lines, again both line-to-line and line-to-ground surges are

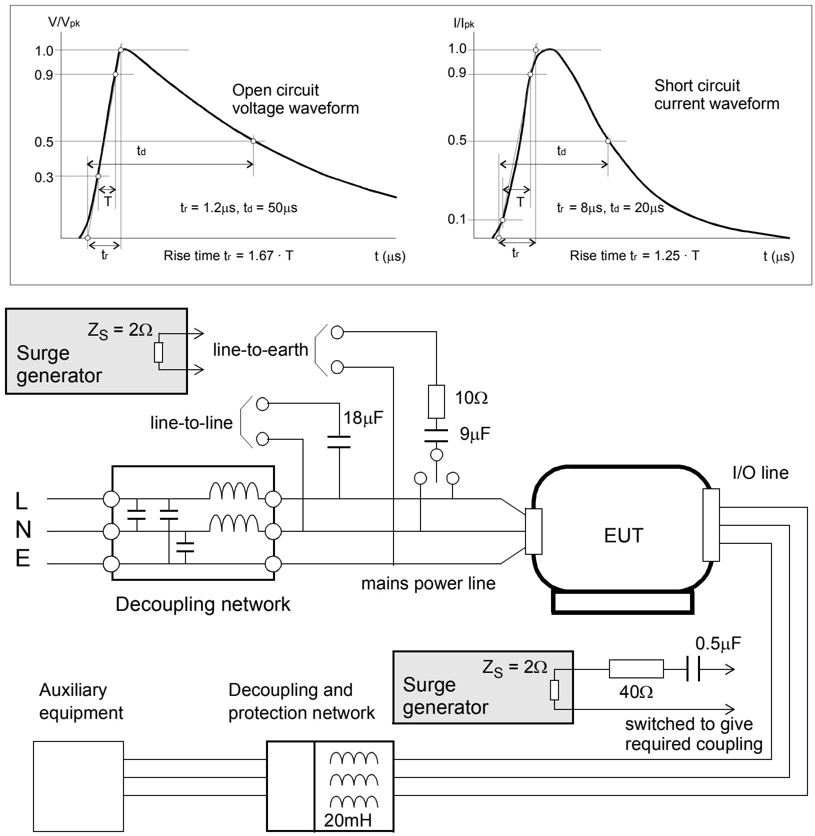


Figure 7.15 Surge waveform specification and coupling

applied, but from a higher impedance. 2Ω represents the differential source impedance of the power supply network, 12Ω represents the line-to-ground power network impedance while 42Ω represents the source impedance both line-to-line and line-to-ground of all other lines.

Power line surges are applied via a coupling/decoupling network incorporating a back filter, which avoids adverse effects on other equipment powered from the same supply, and provides sufficient impedance to allow the surge voltage to be fully developed. For line-to-line coupling the generator output must float, though for line-to-ground coupling it can be grounded. A 10Ω resistor is included in series with the output for line-to-ground coupling.

I/O line surges are applied in series with a 40Ω resistor, either via capacitive coupling with a decoupling filter facing any necessary auxiliary equipment, or by spark-gap coupling if the signals on the I/O line are of a high enough frequency for capacitive coupling to affect their operation.

The purpose of the surge immunity test at equipment level is to ensure that the equipment can withstand a specified level of transient interference without failure or upset. It is often the case that the equipment is fitted with surge protection devices (varistors, zeners, etc.). Typically such devices have low average power ratings, even though they can dissipate or handle high instantaneous currents or energies. So the maximum repetition rate of applied surges will normally be limited by the capabilities of the devices in use, and a maximum of 10 surges (5 positive and 5 negative) is recommended for any one test procedure. Overenthusiastic testing may lead to premature and unnecessary damage to the equipment, with possible consequential damage also occurring. Because of this latter risk, it is wise to physically isolate the EUT during the test. In any case, the EUT should be disconnected from other equipment where possible and the whole set-up should be well insulated to prevent flashover.

Each surge should be synchronized to the peak of the AC supply waveform to give a repeatable and maximum stress, and to the zero crossing to induce maximum follow-on energy[†] if this is likely to occur. Thus, tests are required at each of the phase angles (with respect to the 50 or 60Hz waveform) 0°, 90°, 180° and 270°. Also, the stress voltage should be increased in steps up to the maximum, to check that the protective devices do not allow upset or damage at lower levels of applied voltage whilst satisfactorily clamping high levels.

7.2.4 Other transient immunity tests

7.2.4.1 Ring and damped oscillatory waves

A potential alternative to the surge test is given in IEC 61000-4-12, whose first edition defines the ring wave and the damped oscillatory wave. The damped oscillatory wave has been deleted from the second edition of IEC 61000-4-12 and given its own standard, IEC 61000-4-18, yet to be published. It is a highly specialized waveform, intended only to represent the kind of surges found in high-voltage electrical substations. The ring wave, on the other hand, represents a very typical oscillatory transient occurring frequently in power supply networks and control and signal lines, due to load switching, power faults and lightning. The propagation of the wave in the power and signal lines is always subject to reflections, due to the mismatched line impedance. These reflections create oscillations, whose frequency is related to the propagation speed, length of line and parasitic parameters such as stray capacitance. The rise time is slowed due to the low-pass characteristic of the relevant line. The resultant phenomenon at the equipment ports is an oscillatory transient, or ring wave, which is bipolar compared to the unidirectional surge discussed above in section 7.2.3.

The parameters of the ring wave defined in the standard are shown in Figure 7.16 and the test set-up and coupling modes are shown in Figure 7.17. A minimum of 5 positive and 5 negative transients are to be applied, both line-to-ground (common mode, simultaneously between all terminals and ground) and line-to-line (differential mode), and/or between cabinets (for communication ports). The generator output impedance and minimum repetition period are varied for different applications:

- 12Ω with a minimum repetition period of 10s: EUT supply ports connected to major feeders, and for application between communication ports on cabinets interconnected with 10m long screened data comms cables;

† Follow-on occurs when the surge causes a protective device to break down, and this then puts a low impedance across the supply, which maintains the current through the protector; so that (for an AC supply) energy is dissipated in the protector not just by the surge, but also by the following half-cycle of the supply.

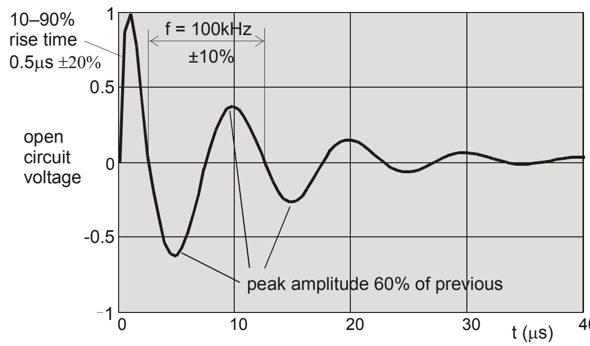


Figure 7.16 Waveform of the ring wave

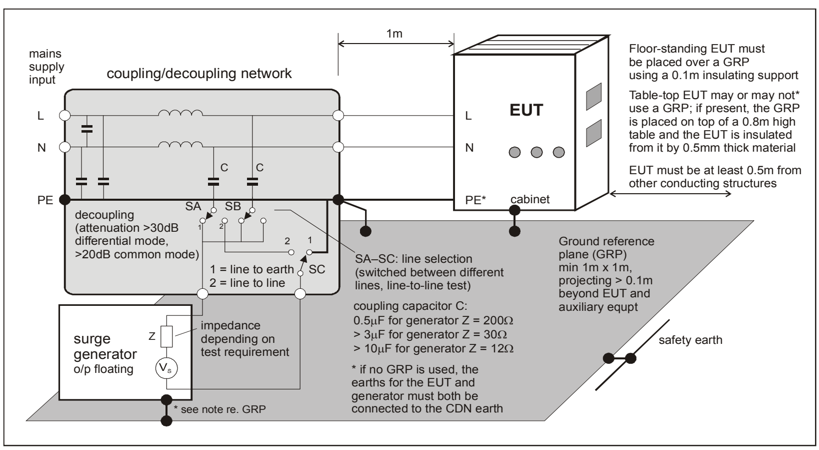


Figure 7.17 Application of the ring wave for AC and DC supplies

- 30Ω with a minimum repetition period of 6s: EUT supply ports connected to outlets;
- 200Ω with a minimum repetition period of 1s: I/O ports, unless the test involves protection devices or filters, in which case 12Ω or 30Ω is applicable.

In fact, the ring wave immunity test is rarely called up in any product standards and so although it is quite a good test for medium energy transient immunity, few products have to undergo it. This may be because of the relatively complicated evaluation of installation conditions that is needed in order to be able to select the appropriate test generator impedance, which probably encourages product committees to favour the IEC 61000-4-5 surge test.

7.2.4.2 Automotive transients

Historically, vehicle manufacturers have each had their own specifications for transient immunity of electrical and electronic parts that they sourced from outside manufacturers. An international standard provides a partial solution to the multiplicity of different requirements. ISO 7637 parts 1, 2 and 3, reorganized in 2004, refer to conducted transient immunity of equipment fitted to cars and light vehicles. Part 1 gives general requirements, and Part 2 gives technical requirements for 12V and 24V systems. Part 3 is for coupling via lines other than supply lines, and defines only the test pulses to be used for simulating switching transients.

Part 2 defines both a means of testing transient emissions (using an oscilloscope and 5μH/50Ω artificial network), without defining any emissions limits, and also a series of standard test pulses to be generated by a test pulse generator and applied to the device under test via its DC power supply leads. No test layout is specified other than that the power lead from the generator to EUT is 0.5m long and stretched straight.

Severity levels (peak voltages) of these pulses are to be agreed between manufacturer and user. Pulses 3 and 4 are called up in the ETSI radio standards with an applied value of level II. If the manufacturer states that a direct connection to the vehicle battery is not required, then pulses 1 and 2 are mandated as well, also at level II.

7.2.5 Sources of variability

Assuming that the EUT's response can be accurately characterized, the major variabilities in transient testing stem from repeatability of layout and the statistical nature of the transient application. The climatic conditions may also have some bearing on the results of air discharge ESD tests.

7.2.5.1 Layout

The wide bandwidth of the ESD and fast burst transients means that cables and the EUT structure can act as incidental radiators and receptors just as they do in RF testing. Therefore the test layout, and routing and termination of cables, must be rigorously defined in the test plan and adhered to throughout the test. Variability will affect the coupling of the interference signals into and within the EUT and may to a lesser extent affect the stray impedances and hence voltage levels. Equally, variability in the EUT's build state, such as whether metal panels are in place and tightened down, will have a major effect on ESD response.

7.2.5.2 Transient timing

In a digital product the operation is a sequence of discrete states. When the applied transient is of the same order of duration as the states (or clock period), as is the case with the ESD and fast burst transients, then the timing of application of the transient with respect to the internal state will affect the unit's immunity. If the pulse coincides with a clock transition then the susceptibility is likely to be higher than during a stable clock period. There may also be some states when the internal software is more immune than at other times, for example when an edge triggered interrupt is disabled. Under most circumstances the time relationship between the internal state and the applied transient is asynchronous and random.

Therefore, for fast transients the probability P of coincidence of the transient with a susceptible state is less than unity, and for this reason both ESD and transient test procedures specify that a relatively large number of separate transients are applied before the EUT can be judged compliant. If P is of the same order or less than the reciprocal of this number, it is still possible that during a given test run the coincidence will not occur and the equipment will be judged to have passed, when on a different run coincidence might occur and the equipment would fail. There is no way around this problem except by applying more test transients in such marginal cases.

7.2.5.3 Environment

In general, the non-electromagnetic environmental conditions do not influence the coupling of interference into or out of electronic equipment, although they may affect the operational parameters of the equipment itself and hence its immunity. The major exception to this is with air discharge ESD. In this case, the discharge waveform is heavily influenced by the physical orientation of the discharge electrode and the rate of approach to the EUT, and also by the relative humidity of the test environment. This means that the test repeatability will vary from day to day and even from hour to hour, all other factors being constant, and is one of the main reasons why the air discharge method has fallen out of favour.

7.2.6 Measurement uncertainty for transient tests

Measurement uncertainty for transient tests has to take a completely different approach to that for the other tests discussed so far. This is because the variables in transient testing include voltage or current parameters, time domain parameters and set-up parameters, and there is no meaningful way to combine these into a budget expressing a single value which could then represent the uncertainty of the applied stress. As an escape from this impasse, the accreditation standard ISO 17025 says:

> In those cases where a well-recognised test method specifies limits to the major sources of uncertainty of measurement and specifies the form of presentation of calculated results, the laboratory is considered to have satisfied this clause (on estimation of uncertainty of measurement] by following the test method and reporting instructions.
>
> [171] note to clause 5.4.6.2

Interpreting this statement, the requirements for ESD, transient and surge tests are deemed to have been satisfied (given that the lab does actually follow the test method) if the generator has been shown to meet the various individual requirements of the appropriate specification: clause 6 of IEC 61000-4-2, -4 and -5, for example. This demands that you should compare the traceable calibration details of the generator you use against the tolerances provided against all of the parameters in this clause, adjusted by the declared calibration uncertainty of the cal lab that has provided the figures. So for instance, IEC 61000-4-2 calls for a peak current at 4kV indicated of 15A ±10%; say the stated calibration uncertainty for this parameter is ±3%, then the current actually measured in your generator calibration could fall within the range 15A ±7%. As long as it does, then the requirement for estimation of measurement uncertainty has been satisfied, using the above rationale. If it doesn't you would need to adjust the generator to bring it within tolerance. This validation of calibration data should be carried out for each polarity of all test levels and all parameters. Setting up a spreadsheet for each instrument simplifies and speeds the process.

7.3 Military susceptibility tests

As with emissions measurements, military and aerospace requirements to DEF STAN 59-41 cover similar phenomena to the commercial tests but do so in different ways.

7.3.1 Continuous LF and RF susceptibility

There are two groups of tests in this category, conducted (CS) and radiated (RS). They were listed in Table 5.7 on page 109. DCS01, DCS03 and DRS01 cover similar frequency ranges and use similar test equipment. DCS01 and DCS03 are the same test applied to power line and signal line respectively.

By contrast to the commercial tests, the high frequency DEF STAN RF immunity tests use significantly different methods. The RF conducted test DCS02 applies to both power and signal lines and uses current injection only, with the bulk current injection probe. The applied stress is determined by either of two criteria, both of which must be monitored simultaneously, forward power (set up in a calibration jig) or induced current measured on the cable under test.

The radiated RF test DRS02 requires various antennas to cover the frequency range:

- Parallel plate: 14kHz – 30MHz
- Power biconical: 30MHz – 200MHz
- Log periodic/Bilog: 200MHz – 1GHz
- Horn: 1GHz – 18GHz

Power is monitored into the antenna and field strength is monitored around the EUT during the test using isotropic probes. The antenna position depends on EUT size and antenna beamwidth. There is no specification for uniform field area; the test is performed with the same set-up as for the radiated emissions test in a screened room which meets the relaxed NSIL criteria discussed in section 6.4.4.3. The modulation most likely to have the greatest effect on the EUT is to be specified in the EMC Test Plan, but the standard gives a variety of default modulations and sweep rates. If a malfunction occurs when sweeping through the frequency range the signal strength is reduced to establish the threshold level. At frequencies above 1GHz discontinuities in the screening of the EUT (connectors, displays, etc.) are presented to the transmitting antenna directly. Field strength levels depend on application and frequency range and vary up to 200V/m continuous, and 2kV/m pulsed in the microwave region.

7.3.2 Transient susceptibility

The various military transient tests were shown in Table 5.6 on page 108. These require different types of transient waveform using various generators. In most cases, levels and waveform are set up into a calibration jig before the test. A noteworthy difference from commercial testing is that all bar DCS09 and DCS10 require DCE01 (conducted power emissions) to be applied both before and after the transient test to ensure that the power filter has not been damaged.

DCS05 has two parts, switching simulation and NEMP (Nuclear ElectroMagnetic Pulse), the latter being much more severe. DCS10 ESD is similar to the commercial ESD test, IEC 61000-4-2, but otherwise there is no comparability in waveforms.

Chapter 8

Low frequency tests

The concept of EMC applies not only to high frequency phenomena. A product must be compatible with its electromagnetic environment, and there is no limitation on the frequency range encompassed by that environment, nor on the modes of coupling with it. A universal requirement is that the product should be compatible with its power supply; that is, it must be adequately immune from variations in the power supply, and it must not itself cause such variations. This is necessary whether the supply is DC, AC 50Hz mains or some other system-specific description, but for the public AC mains there are a number of common requirements which apply under the umbrella of the EMC Directive and which are discussed here.

8.1 Mains harmonic and flicker emission

Harmonic components of the AC supply input current to an item of equipment arise from non-linearities of the load over a single cycle of the input voltage. The EMC Directive includes requirements for measuring harmonic emissions as embodied in IEC 61000-3-2 (EN 61000-3-2), which covers all electrical and electronic equipment with an input current up to 16A per phase. This has a sister standard (actually a technical report), IEC 61000-3-4, for higher-powered equipment up to 75A, which has been supplemented by a full international standard, IEC 61000-3-12. The generation and control of mains harmonics are discussed further in section 10.4 and the standards themselves are surveyed in section 4.5.1.

Although the harmonic frequency range under consideration extends only up to 2kHz (the 40th harmonic of 50Hz), and therefore does not by any stretch of the imagination need to employ RF measurement techniques, there are many aspects of the measurement which are not entirely obvious and should be considered further. In the late 1990s the harmonics standard came under withering attack from several directions. There are three main interested parties: the supply authorities, who are keenly interested in preserving their networks from distortion; the manufacturers, who are equally keen to avoid expensive penalties resulting from harmonic limitation on their power supplies; and the test houses, who are keen to have a standard which will enable them to test accurately, completely and repeatably. The anomalies and gaps in the original 1995 edition of the standard allowed each of these parties ample opportunity for, to put it kindly, combative discussion. Several working groups later, a resolution was reached and the standard was drastically modified. This was first achieved in Europe by a CENELEC common modification and then the IEC document caught up, with the publication of its second edition in 2000. The third edition was published at the end of 2005 and makes relatively few changes, apart from relaxing the limits slightly for Class A equipment under certain restricted conditions and adding test conditions for some further types of apparatus.

It should be appreciated that the requirements in IEC 61000-3-2 only apply to equipment powered from a 220–240V AC mains supply. There are no requirements, at least in this document, for harmonic limitation on equipment connected to lower voltage mains supplies, for instance for the US or Japanese markets.

8.1.1 Equipment

The original IEC 61000-3-2 defined the method of measurement, and each item of test equipment is specified. Figure 8.1 shows the basic measurement circuit, and its components are:

- an AC source;
- a current transducer;
- a wave analyser.

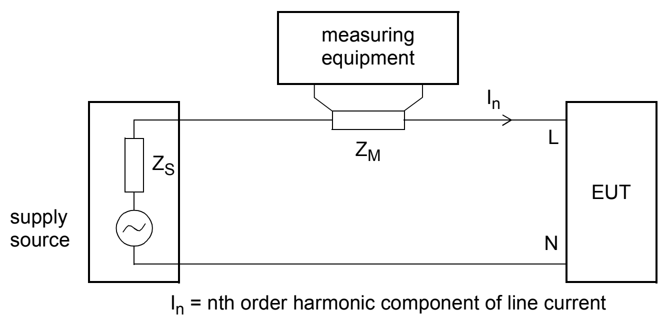


Figure 8.1 Mains harmonic emission measurement circuit

8.1.1.1 AC supply source

To make a harmonic measurement with the required accuracy you need a source with very low distortion, high voltage stability and settability and low impedance. In general the public mains supply will not be able to meet these requirements. IEC 61000-3-2 requires that the voltage must be stable to within ±2% of the selected level during the measurement, and the frequency within 0.5% of nominal. The harmonic distortion must be less than 0.9% at third harmonic, 0.4% at 5th, 0.3% at 7th, 0.2% at 9th and 0.1% at all others.

To meet these requirements typical test equipment uses a power amplifier driven by a 50Hz sinewave oscillator, with negative feedback to maintain the low output impedance. The output may be fed through a power transformer for voltage step-up purposes, but the transformer reactance must not be allowed to affect the output impedance at the higher harmonic frequencies. Variacs are not recommended for the same reason. The amplifier will need to be large to cope with the full range of loads – the standard covers equipment rated up to 16A, which is a power level of 3680W at 230V, although for in-house use your product range may not approach this level and a smaller amplifier would suffice. For high power and highly distorting loads the "model" AC source becomes quite difficult to realize. Including the maximum allowable transitory harmonics for Class B equipment, legitimate peak currents can be around 40A, although some equipment can substantially exceed this, and the source should be able to deliver this power level without distortion. If the measured harmonics are well over or under the limit then voltage distortion is a minor consideration, but it becomes important for borderline cases.

8.1.1.2 Current transducer

The current transducer couples the harmonic current I_n to the measuring instrument, and it can be either a current shunt or a current transformer. In both cases, the transducer impedance Z_M is added to the source output impedance and the two together must cause negligible variation in the load current harmonic structure. A shunt of less than 0.1Ω impedance and a time constant less than 10µs is acceptable, but does not provide any isolation from the measuring circuit. A current transformer does offer isolation, but will need to be calibrated at each harmonic frequency and may suffer from saturation if the measured current includes a DC component or has a high crest factor.

The measuring impedance Z_M should create a voltage drop of less than 0.15V rms. The total set-up is not allowed an error at any harmonic frequency of more than 5% of the permissible limits.

8.1.1.3 Wave analyser

The wave analyser measures the amplitude of each harmonic component I_n for n = 2 to 40. According to the original standard it could be either a frequency domain type, using selective filters or a spectrum analyser, or a time domain type using digital computation to derive the discrete Fourier transform (DFT). The second and third editions of IEC 61000-3-2 have deleted any requirements in this standard and have redirected them instead to IEC 61000-4-7, which is a companion standard defining the reference instrument for harmonics measurement. The intention of this standard is to outlaw frequency domain instruments and only to allow DFT types. In practice, all commercial harmonic analysers are of this sort.

For steady-state harmonics different implementations of measuring instrument will give comparable results, but this isn't necessarily the case if the harmonic components fluctuate while the measurement is being made. The response at the indicating output should be that of a first order low-pass filter with a time constant of 1.5 seconds. IEC 61000-4-7 includes more specific details of the smoothing algorithm which performs this function on the discrete data values.

8.1.1.4 Pre-compliance measurements

A general test lab must be prepared for any kind of EUT and therefore has to invest in a test system which meets the specification over the whole range of likely EUTs. If you are doing pre-compliance tests in-house you may be able to use a system with a relaxed specification, which is therefore cheaper, especially if you don't expect to sail very close to the limits. We have already mentioned the use of a lower power amplifier as a supply source if you're not developing high-power products. You could use the mains supply directly if you are prepared to accept its inherent voltage instability and likely extra harmonic distortion, which could contribute a small amount extra to the measured harmonics. It is easy enough, with a resistive (lamp) load, to measure the actual voltage distortion at the time of test and make allowance for it in your margins to the limit.

As far as the analyser and current transducer are concerned, again a lesser accuracy may be acceptable if you increase the margin from the limits, and strict compliance with the IEC 61000-4-7 specification may be less important if your products don't produce time-varying harmonics.

8.1.2 Test conditions

Special test conditions for some types of equipment are given in IEC 61000-3-2, including TV receivers, audio amplifiers, VCRs, lighting equipment and various household appliances. Independent lamp dimmers and other phase-control devices should be set for a firing angle of 90°. Information technology equipment is tested with the equipment configured to its rated current. For all equipment, on manual or automatic starting or stopping, harmonic currents and power are not taken into account for the first 10 seconds after the switching event, so that start-up conditions are generally ignored.

The original 1995 standard required that all other equipment not covered by the specific conditions – that is, most equipment within the scope of the standard – should be operated by setting its user controls or program mode to give the maximum harmonic amplitude for each successive harmonic component in turn. If followed to the letter, this procedure would require an excessive amount of time and effort for a complete test. Later editions replace this with the altogether more reasonable requirement to conduct the test in the mode *expected* to produce the maximum total harmonic current under normal operating conditions.

A general requirement, though, is for repeatability: the repeatability of the measurements must be shown to be better than ±5%, for the same EUT under identical test and environmental conditions and with the same test system. This effectively means that you have to keep taking measurements for long enough to get a series of results that fall within this criterion. For EUTs with steady-state harmonics this means that the test can be quite fast, but it also means that with fluctuating harmonics you may be testing for a long time.

8.1.3 Equipment classification and limits

The original standard established four classes of equipment:

- Class B for portable tools;
- Class C for lighting equipment, including dimmers;
- Class D for equipment having the “special wave shape” of input current, and an active input power less than or equal to 600W;
- Class A for everything else, and particularly balanced three-phase equipment.

The “special wave shape” was defined by an envelope, effectively a means of distinguishing electronic power supply circuits, which normally draw their current for less than a third of the supply half-cycle. The harmonic limits are quoted as absolute values for Class A, whatever the input power, and as a set of sliding values proportional to input power for Class D. Figure 8.2 shows these limits graphically. For equipment with an input rating greater than 600W the Class A limits, being fixed, become proportionately more severe with increasing input power.

In recognition of the fact that low-powered equipment contributes little to the overall harmonic problem, there is a blanket exemption from the limits for equipment with a rated power of 75W or less, other than lighting equipment; as there is also for symmetrically controlled heating elements with a rated power less than or equal to 200W and independent dimmers for incandescent lamps with a rated power less than or equal to 1kW.

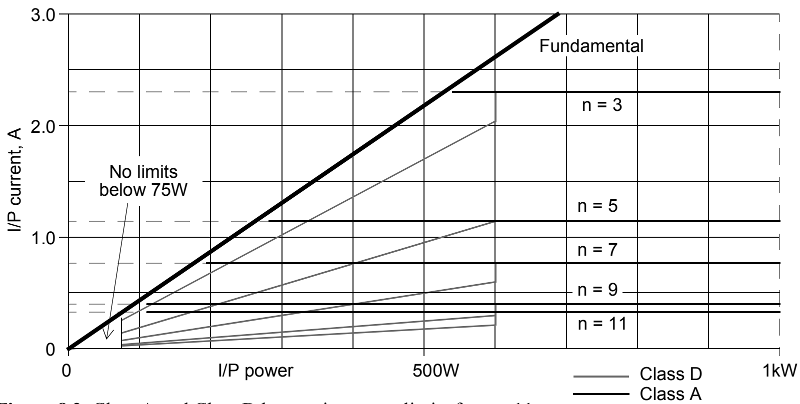


Figure 8.2 Class A and Class D harmonic current limits for n ≤ 11

8.1.3.1 Class D membership

The definition of Class D caused more problems for the standard than virtually any other aspect. If the EUT was suspected of being Class D, the test equipment had first to check the input waveform to confirm whether or not it fell inside the Class D definition, and then decide on its active power, before the limits could be applied. This led to a fundamental difficulty in deciding what current value to use, especially if the current and/or its harmonic content was fluctuating. This difficulty was decisively addressed in the CENELEC common modification, and in subsequent editions.

The working groups could not agree on an acceptable general method for applying the Class D envelope. Since Class D is intended to constrain particular types of equipment which are considered to have the greatest impact on the power network, later editions have turned the definition on its head by specifying particular types of product to which the Class D limits must apply. These are:

- personal computers and monitors
- TV receivers

with a *specified* power (see next paragraph) less than or equal to 600W. All other equipment that is not Classes B or C is to be regarded as Class A, with some modifications from the original classification. The Class D envelope is effectively removed from the discussion. The equipment classification now stands as follows:

- Class B: portable tools and non-professional arc welding equipment;
- Class C: lighting equipment, excluding dimmers for incandescent lamps;
- Class D: personal computers and their monitors, and television receivers;
- Class A: everything else.

Power basis for Class D limits

The Class D limits are given in mA per watt, and the basis for the power used for defining the limit value has been hard to pin down. In the later editions, *average* emissions are to be compared to limits based upon the *maximum* of the measured values of power in each observation time window over the entire duration of the test. The harmonic currents and active input power are measured under the same test conditions but need not be measured simultaneously.

In order not to arrive at a power at which limits change abruptly (for example, 600W or 75W), the manufacturer is allowed to specify a power level for establishing the limits, but this specified value must be within ±10% of the actual measured value. In other words, if the maximum measured power is close to the Class D cut-off point, the manufacturer has the option of specifying a power level within 10% of this value and therefore (potentially) of taking the apparatus outside the level at which severe limits apply. The purpose of this rather tortuous approach is to prevent the situation in which equipment operating near the boundary and tested under slightly different conditions might be subject to widely differing limits. The specified power for this purpose is not necessarily the same as the manufacturer's "rated" power for safety or functional purposes.

8.1.3.2 Professional equipment

A significant relaxation, present in the original standard, is that no limits apply (more correctly, limits are "not specified") for professional equipment with a total rated power of more than 1kW. Professional equipment is defined as "equipment for use in trades, professions or industries and which is not intended for sale to the general public. The designation shall be specified by the manufacturer". Later editions relax this slightly more, by allowing the connection to "certain types of low voltage supplies" of non-compliant professional equipment, if the instruction manual contains a requirement to ask the supply authority for permission to connect.

8.1.4 Flicker

A companion requirement to IEC 61000-3-2 on harmonics is that provided by IEC 61000-3-3 on flicker. Flicker is defined as the "impression of unsteadiness of visual sensation induced by a light stimulus whose luminance or spectral distribution fluctuates with time". The problem with respect to EMC is that varying loads on a power supply network can result in voltage changes at common points of connection which are of sufficient amplitude to induce flicker in connected luminaires. The affected luminaires may have nothing to do with the load equipment that is causing the variations. Therefore, IEC 61000-3-3 – which applies to the same wide range of apparatus as does IEC 61000-3-2 – regulates the degree to which a given item of equipment can cause perceptible flicker. It does so by limiting the voltage variations that are generated across a reference impedance, and it places limits on three factors:

- the relative voltage change;
- the short-term flicker value P_{st};
- the long-term flicker value P_{lt}.

These limits do not apply to emergency switching or interruptions, and the P_{st} and P_{lt} limits do not apply to manual switching. The voltage change limits do apply to manual switching, however, and this effectively places a limit on allowable switch-on inrush current for any apparatus. It has not been clear that this particular effect of the flicker standard was intended by its authoring committee. The standard also states that "tests need not be made on equipment which is unlikely to produce significant voltage fluctuations or flicker", and since the restriction on inrush current was not made explicit, this statement has been widely interpreted as meaning that most typical electronic apparatus whose steady-state load current changes only slightly can be excused testing. As a result, few manufacturers of equipment with electronic power supplies – many of which may well exceed the voltage change limit on switch-on – have been aware that they are in breach of the standard. Amendment 1: 2001 to the standard does make it slightly clearer that inrush current limitation is intended, and does in fact change the limits in this context.

Equipment that typically *will* produce flicker includes any device which switches varying loads during its operating cycle; many household appliances fall into this category, and particular offenders are products which have heaters whose temperatures are controlled by burst firing, i.e. power is provided to the heater for a few cycles of the mains supply at a time, and the on/off ratio of the bursts controls the temperature. If the heating load is at all substantial this kind of equipment easily falls foul of the flicker limits, and is effectively outlawed.

Non-compliant equipment

IEC 61000-3-3 applies only to equipment with an input current up to 16A per phase used on 220–250V 50Hz power supplies, so is not relevant for products sold in countries with lower voltages. It refers non-compliant equipment to IEC 61000-3-11; this applies to equipment with a rated input current ≤75A per phase and subject to conditional connection. The EN versions of both standards are harmonized under the EMC Directive. "Conditional connection" means "connection of equipment which requires the user's supply at the interface point to have an impedance lower than the reference impedance Z_{ref} in order that the equipment emissions comply with the limits in this standard". The consequence of this for the equipment manufacturer and its user is either:

- the maximum allowable system impedance which will allow the equipment to meet the same flicker limits as in IEC 61000-3-3 is determined; the instruction manual then quotes this figure and instructs the user that the equipment should only be connected to a supply with this impedance or less, in consultation with the supply authority if necessary; or
- the manufacturer tests to show compliance against the same flicker limits with a lower reference impedance ($0.25 + j\,0.25\Omega$ for a single phase supply), declares in the manual that the equipment is intended for use only in premises with a service current capacity ≥100A per phase, and instructs the user to determine that the service current capacity at the interface point is sufficient, in consultation with the supply authority if necessary.

EN 61000-3-11 became mandatory on 1st November 2003.

8.1.4.1 Measuring instrumentation

The basic instrumentation used to measure flicker has essentially the same block diagram and characteristics as the harmonics analyser shown in Figure 8.1, and for this reason harmonics and flicker analysers are often packaged together. The difference can be seen in Figure 8.3, which gives the circuit for a three-phase supply, and which shows that the measured variable is now the voltage across the point of supply rather than the current drawn from it. The source impedance of the supply generator is more carefully defined so that load current changes in the EUT produce a defined voltage change, which is then analysed to compare it with the various limits. In fact, the standard itself allows either a direct voltage measurement or a current measurement, with the voltage calculated from the theoretical value of the source impedance. It is arguable that the latter is potentially more accurate, since there is less error introduced by departures

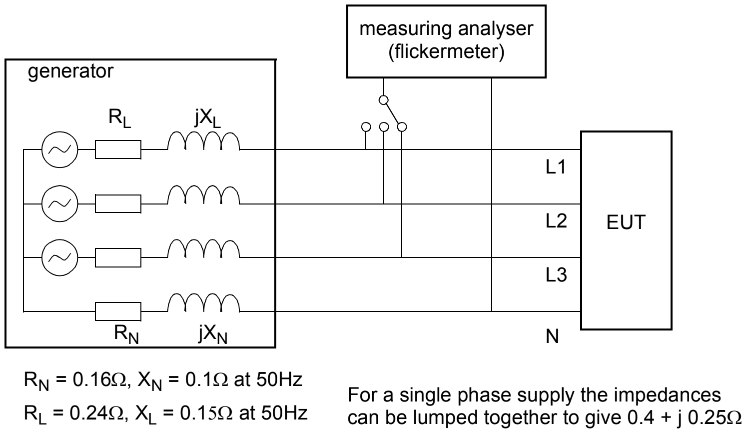


Figure 8.3 Flicker measurement circuit

from the ideal source impedance value, but virtually all commercial flicker measurement instruments use a direct voltage measurement.

The accuracy of this set-up is required to be such that the relative voltage change can be measured with a total accuracy of better than ±8% of the maximum allowed value. The measurement errors can be distributed between the reference impedance and the analyser as long as the total remains within this limit.

8.1.4.2 Relative voltage change

The RMS voltage is evaluated (typically by direct measurement, but it is also possible to calculate it given the active and reactive parts of the current waveform) over successive half-periods (each 10ms) to build up a time-dependent view of the voltage changes. The voltages are normalized to the nominal value to give d(t) and two characteristics are derived:

- the relative steady-state voltage change d_c, which is the difference between two adjacent steady-state voltages separated by at least one change (steady state is defined as persisting for at least 1 second);
- the maximum relative voltage change d_{max}, which is the difference between maximum and minimum values of the voltage change characteristics.

The standard (with A1: 2001) requires that d_c does not exceed 3.3% and d_{max} does not exceed 4%, and that the value of d(t) during a voltage change does not exceed 3.3% for more than 500ms. The value for d_{max} can be relaxed to 6% for manual switching and to 7% if the equipment is attended while in use, with some more complex relaxations for automatically switched equipment. To try and get some repeatability into measurements of switched inrush current, the standard requires 24 measurements, each taking into account the need to let inrush current limiting devices function properly, with the highest and lowest values subsequently deleted and the result being the average of the remaining 22.

Rough-and-ready calculation

If we ignore the reactive part of the load impedance and consider only the real value of 0.4Ω for single-phase supplies, and take the 4% d_{max} figure applied to a supply voltage of 230V, then the inrush current that will just touch that limit is

$$I_{inrush\text{-}max} = (0.04 \cdot 230)/0.4 = 23A \qquad (8.1)$$

Clearly, if your known inrush current is nowhere near this figure, you need not be too concerned about compliance with this standard. If it's approaching or greater, then you should do a proper measurement.

8.1.4.3 Short-term flicker

Voltage changes by themselves do not adequately characterize the flicker perceptibility. The human eye–brain combination varies in sensitivity to flicker as the flicker frequency changes. To account for this, the voltage changes must themselves be processed over a period of a few minutes to take account of the frequency of changes, the shape of the voltage change characteristic, and the cumulative irritating effect of repeated changes. Whilst in some special cases this can be done analytically, and in one case by direct comparison to a graph (see below), in general the voltage changes are passed to a "flickermeter", whose specifications are given in a separate standard, IEC 61000-4-15 (see section 4.7.2). The flickermeter applies a weighting to the voltage change characteristic depending on its waveform, and is the reference method.

The output of the flickermeter gives the short-term flicker indicator P_{st}. P_{st} is observed over a period of 10 minutes, to include that part of the operating cycle in which the EUT produces the least favourable sequence of voltage changes. P_{st} is not allowed to exceed a value of 1.

For the special case of rectangular voltage changes of the same amplitude separated by equal time intervals, the P_{st} value can be derived from a graph published in the standard and reproduced in Figure 8.4. This shows the value of d(t) versus frequency which gives a P_{st} of 1, and illustrates the maximum physiological sensitivity at around 8Hz or 1000 changes per minute.

8.1.4.4 Long-term flicker

In some cases flicker must be evaluated over a longer period, using successive values of P_{st} to give P_{lt}. The P_{st} values are averaged on a root-sum-of cubes basis:

$$P_{lt} = \sqrt[3]{\left(\sum_{1}^{12} P_{st}^3\right) / 12}$$

The standard suggests that this is necessary for equipment which is normally operated for more than 30 minutes at a time. The observation period is 2 hours, that is, 12 successive P_{st} values are recorded. P_{lt} is not allowed to exceed a value of 0.65. The justification for this, in effect, is that whereas the average human can cope with a P_{st} value of up to 1 for ten minutes, if the flicker continues for a longer time, the threshold of irritability lowers.

Annex A of the standard gives operating conditions and application of the limits for certain types of equipment, particularly white goods and consumer products. In several cases, P_{lt} does not need to be evaluated.

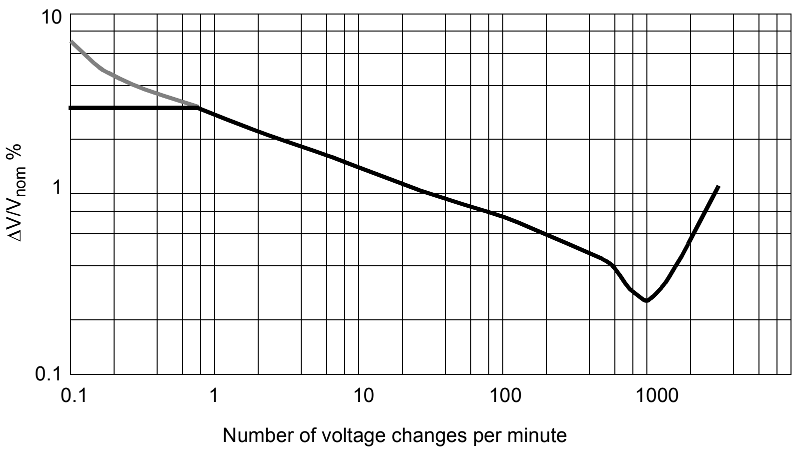


Figure 8.4 Curve for $P_{st} = 1$ for rectangular equidistant voltage changes

8.2 Magnetic field and power quality immunity

The two low frequency immunity tests that are most significant in the context of the EMC Directive are power frequency magnetic field, and voltage dips and interruptions on the power supply. Various other tests have or will become available as parts of IEC 61000-4, but there is a strong resistance to the widespread adoption of many more such tests by product committees for compliance purposes. On the other hand, for procurement purposes in specific industries such as defence, rail or automotive, there are a number of established low frequency tests which will not be covered here.

8.2.1 Magnetic field

Testing with a steady magnetic field may apply to all types of equipment intended for public or industrial mains distribution networks or for electrical installations. Testing with a short duration magnetic field related to fault conditions requires higher test levels than those for steady-state conditions; the highest values apply mainly to equipment to be installed in exposed areas of electrical plants. (IEC 61000-2-7 gives values for various environments.)

Magnetic fields at power frequencies are common in the environment but are only a threat to certain types of equipment. Although the basic test method gives no especial advice as to which products should and should not be tested, all the generic and product standards, when they refer to the magnetic field test, state that it should be applied only to equipment "containing devices susceptible to magnetic fields". Experience suggests that this certainly applies to anything with a cathode ray tube – although the humble domestic TV set isn't included, since its own product standard makes no mention of magnetic field testing – and various other specialized components such as magnetic sensors. Audio apparatus that might be sensitive to hum pick-up should also be considered, but otherwise, general electronic circuitry can be assumed not to be relevant for this test.

The test field waveform is sinusoidal at power frequency. In many cases (household areas, substations and power plant under normal conditions), the magnetic field produced by harmonics is negligible, but in special cases such as industrial areas with a concentration of large power convertors they can occur. Testing at present does not take them into account.

The magnetic field immunity test method is specified in IEC 61000-4-8. It requires the EUT to be immersed in a magnetic field of 50Hz or 60Hz sinusoidal (< 8% distortion) generated by an induction loop surrounding it, in three orthogonal orientations (Figure 8.5). Severity levels are defined as 1, 3, 10, 30 and 100A/m for continuous application, and 300 and 1000A/m for short duration (1–3 seconds) application.

The magnetic field is generated within the loop and the field uniformity is required to be 3dB within the volume occupied by the EUT. For various loop sizes, the maximum volume available is as follows:

- single square loop, 1m side: 0.6m x 0.6m x 0.5m high;
- double square loops, 1m side, 0.6m spaced: 0.6m x 0.6m x 1m high (0.8m spacing gives 1.2m height);
- single rectangular loop, 1m x 2.6m: 0.6m x 0.6m x 2m high.

The loop factor (H/I, magnetic field/current injected) is calibrated at the centre of the loop and allows a direct correlation between the current measured in series with the loop and the amplitude of the magnetic field that is produced. Although the standard describes particular designs of coil used with a ground reference plane, it does not forbid other designs provided that they meet the field homogeneity condition. For instance, multi-turn coils, which would allow a lower test current for a given field, would be acceptable. The requirements of the AC source and its output transformer turns ratio for use with a particular coil are given in Table 8.1 [82].

Table 8.1 Coil and source current and voltage for 1.5 ohms, 19 mH, 65.73 A/m/A loop factor coil

H (A/m)	Coil		Transformer turns ratio N:1	AC source		
	I (A)	V (V)		I (A)	V (V)	VA
1	0.015	0.094	240	6.34×10^{-5}	22.5	0.001
3	0.046	0.281	240	1.90×10^{-4}	67.4	0.013
10	0.152	0.936	240	6.34×10^{-4}	224.7	0.142
30	0.456	2.809	8	0.057	22.5	1.282
100	1.521	9.363	8	0.190	74.9	14.245
300	4.564	28.090	8	0.571	224.7	128.21
1000	15.214	93.635	2.5	6.086	234.1	1424.53

8.2.2 Voltage dips and interrupts

There should be a defined and controlled response of the EUT to discontinuities in the mains supply. IEC 61000-4-11 defines immunity test methods for these phenomena.

Electrical and electronic equipment may be affected by voltage dips, short interruptions or voltage variations of the power supply. Voltage dips and short interruptions are caused by faults in the network, in installations or by a sudden large

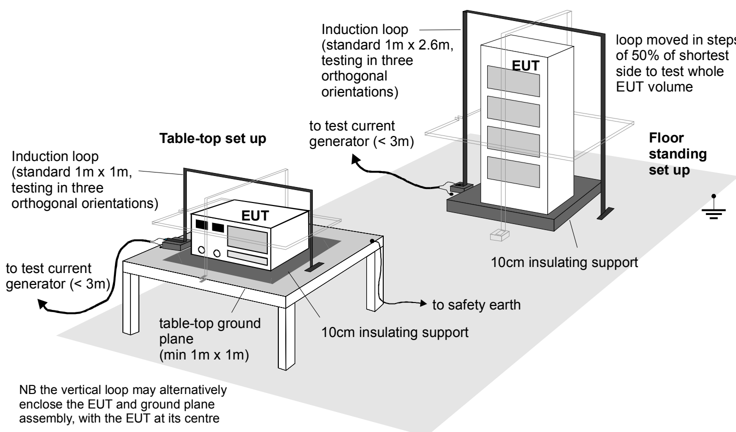


Figure 8.5 The magnetic field immunity test

change of load. In certain cases, multiple dips or interruptions may occur. Voltage variations are caused by the continuously varying loads connected to the network. These phenomena are random in nature and can be characterized in terms of their deviation from the rated voltage and their duration.

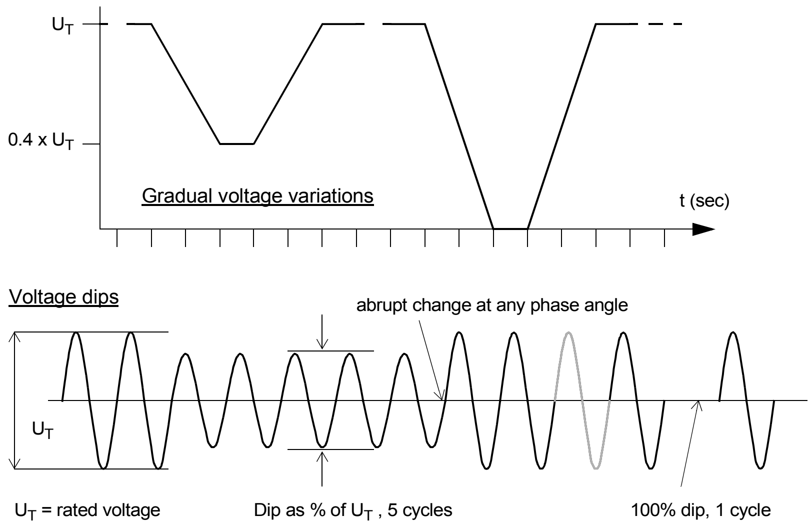
Voltage dips and short interruptions are not always abrupt. If large mains networks are disconnected the voltage will only decrease gradually due to the many rotating machines which are connected. For a short period, these will operate as generators sending power into the network. Some equipment is more sensitive to gradual variations in voltage than to abrupt change.

8.2.2.1 Applying voltage dips and interruptions

Different types of tests are specified in the standard to simulate the effects of abrupt change of voltage, and, optionally, a type test is specified also for gradual voltage change. Testing can be done either using electronically-controlled switching between the outputs of two variacs or by controlling the output of a waveform generator fed through a power amplifier. The latter is more usual for low-to-medium power applications.

Tests are given for voltage dips and short interruptions (an interruption is a dip to 0% of the supply) and for short-period voltage variations. The preferred values for period and level of dips are listed in the table in Figure 8.6. Tests of dips and interruptions are significant as these are referenced in the generic immunity standards and many product standards. The generic standard requirements are for a half-cycle dip to 70% of rated voltage, a 5-cycle dip to 40% of rated voltage and a 5 second interruption. The performance criterion which applies to the latter two tests is that temporary loss of function is allowed, provided that it is self recoverable or can be restored by operation of the controls; i.e. a latch-up or a blown fuse is unacceptable.

A significant part of the specification of the test generator is its ability to cope with the peak inrush current of the EUT without affecting the test result. If, for instance, the inrush current was high enough to pull down the source voltage significantly, the test



Preferred test level % U_T	**Preferred duration # of cycles**
0 (interruption) 40 70	0.5, 1, 5, 10, 25, 50

Figure 8.6 Supply voltage dips and variations

might not show up an effect (such as a blown fuse) that could occur on a "stiffer" power supply. The maximum capability of the generator need not exceed 500A for 230V mains, or 250A for 110V mains. If the EUT draws significantly less than this, a generator with lesser capability is allowed provided that the EUT inrush current is less than 70% of the peak drive capability. But in this case you have to verify the generator's actual capability, and the standard gives a specific way of doing this, by driving the rectified output into a 1700µF capacitor. If you are using your laboratory's mains supply, via a variac-type dip generator, you may find that it can't meet the full specification because of the combined impedances of the mains supply and the variacs.

A further quirk of the test is the effect that a half-cycle dip can have on a product with an ordinary mains transformer (as opposed to a direct-off-line switchmode supply). When only one polarity of the mains voltage is removed, the reapplication of the next half-cycle will drive the core of the transformer into saturation and create a very high inrush current, potentially 10 to 40 times the rated current, until the transformer recovers. Clearly the generator should be capable of supplying this current in the test, and it may have implications also for the design of the product's input circuitry.

Chapter 9

Test planning

9.1 The need for a test plan

An EMC test plan is a vital part of the specification of a new product. It provides the basis around which the development and pre-production stages can be structured in order for the product to achieve EMC compliance with the minimum of effort. It can act as the schedule for in-house testing, or can be used as a contractual document in dealing with an external test house. Although you should prepare it as soon as the project gets underway, it will of necessity need revision as the product itself develops and especially in the light of actual test experience on prototype versions.

9.1.1 The requirements of accreditation

You may decide to do the testing at an accredited laboratory. The accreditation standard ISO/IEC 17025 [171] actively demands that an accredited laboratory has a procedure for the review of requests, tenders and contracts, which ensures that:

- the requirements, including the methods to be used, are adequately defined, documented and understood;
- the laboratory has the capabilities and resources to meet the requirements;
- the appropriate test and/or calibration method is selected and is capable of meeting the customer's requirements.

Most such laboratories have a mechanism for dealing with this "contract review" stage by asking customers to complete a standard form with details of the equipment to be tested, its support equipment, the standards to be tested against, performance criteria and operating modes, and so forth. It's helpful if customers can anticipate this and provide full details immediately, but often it requires some interaction with and possibly advice from the lab.

As a manufacturer of products using a third party test lab, you have to remember that although the lab is a specialist in EMC testing it doesn't necessarily have any knowledge of your product. The interface between the product and the test has to be deliberately created. This is a hidden but major advantage of a company's own in-house EMC test lab – it is in a far better position to create and/or review the product test plan than is a generalist lab.

9.1.2 The requirements of standards

9.1.2.1 DEF STAN 59-41

Part 2 (Management and planning procedures) of the military EMC test standard [202] has an explicit requirement for a project test plan. To quote that document directly:

> In order to achieve consistency throughout the phases of EMC testing, it is essential to formalize the details of the test procedure for the project. The test methods described in this Standard are necessarily generalized and cannot cover the exigencies of each equipment under test. The test plan shall describe the preferred methods of interpretation of the requirements of this Standard and state in detail the equipment configuration during the test, the methods of supplying power to the equipment, the application of stimulus signals and also the application of electrical or mechanical loading.
>
> Without a formalized test plan the results of the EMC test may vary considerably due to possible variations in the test arrangement, thus obscuring the effects of any modifications during development of the equipment to the production stage.

So it is to be expected that any supply of military equipment to DEF STAN 59-41 will demand the EMC test plan as part of the contract.

9.1.2.2 IEC 61000-4-X

Clause 8 of most of the main parts of IEC 61000-4, the basic standards for immunity testing, requires that "the tests shall be carried out on the basis of a test plan that shall include the verification of the performance of the EUT". Each of the standards naturally has its own requirements for what the test plan should contain, for instance IEC 61000-4-3 for RF immunity includes:

> – the size of the EUT;
>
> – representative operating conditions of the EUT;
>
> – whether the EUT shall be tested as table-top or floor-standing, or a combination of the two;
>
> – for floor-standing equipment, the height of the support;
>
> – the type of test facility to be used and the position of the radiating antennas;
>
> – the type of antennas to be used;
>
> – the frequency range, dwell time and frequency steps;
>
> – the size and shape of the uniform field area;
>
> – whether any partial illumination is used;
>
> – the test level to be applied;
>
> – the type(s) and number of interconnecting wires used and the interface port (of the EUT) to which these are to be connected;
>
> – the performance criteria which are acceptable;
>
> – a description of the method used to exercise the EUT.

IEC 61000-4-5 for surge immunity has:

> the test plan shall specify the test set-up with:
>
> – generator and other equipment utilized;
>
> – test level (voltage);
>
> – generator source impedance;
>
> – polarity of the surge;
>
> – number of tests:
>
> – time between successive pulses;
>
> – representative operating conditions of the EUT;
>
> – locations to which the surges are applied;
>
> – actual installation conditions.

So a complete product test plan has to take account of the needs of each of these basic standards.

9.1.3 The requirements of the customer

Whilst the above headings point out that both accredited testing, and testing to certain standards, explicitly require a test plan to be created, in practice this is not always or even often the driving force. Much EMC testing is actually performed in the face of a requirement from a technically sophisticated customer. As we saw in Chapter 5, the automotive, aerospace, rail and military sectors are good examples of such situations. It is quite usual for the customer to have to agree and sign off the proposed test plan before testing can commence, if it is to regard the results as valid.

9.2 Contents of the test plan

Given that you will be writing a test plan for your product, what areas should it cover?

9.2.1 Description of the equipment under test (EUT)

A test plan could cover either a single product or a range of similar products, or a complete system. The basic description of the EUT must specify the model number and which (if any) variants are to be tested under this generic model type, to make clear what are the boundaries within which the EMC test results can apply. Important aspects of a general description for many of the tests are:

- physical size and weight;
- power supply requirements;
- number and type of connection ports.

9.2.1.1 Stand-alone or part of a larger system?

If the EUT is part of a system, or can only be tested as part of a system, for instance it is a plug-in module or a computer peripheral, then the components of the system of which it is a part must also be specified. This is because these other system components could affect the outcome of the test if they are part of the test environment, and you need to take care that the test results will not be compromised by a failure on their part.

These issues commonly arise when, for instance, third party monitors are connected to a computer-based instrument, or when different design groups in the same company are responsible for individual parts in the whole system. Not only must it be clear where the responsibilities for passes, failures and consequent remedial action in the tests are to lie, but also you may have to set up the tests so that individual contributions to emissions or immunity signatures can be identified.

9.2.1.2 System configuration and criteria for choosing it

Following on from the above, if the EUT can form part of a system or installation which may contain a variety of other different components, you will need to specify a representative system configuration which will allow you to perform the tests. The criteria on which the choice of configuration is based, i.e. how you decide what is "representative", must be made clear.

The wording of the second edition of the EMC Directive, in particular, places a burden on the manufacturer to declare compliance of the product "in all the possible configurations identified by the manufacturer as representative of its intended use". It would be typical to address this by building an EUT which included at least one of every hardware function that might be available, so that for instance the top-of-the-range model which incorporated every option would represent those lesser models that didn't have some options. If certain options are mutually exclusive then this is likely to mean testing more than one build. An example of a hypothetical "compliance test matrix analysis" for a display-based product, not necessarily definitive, might be as shown in Table 9.1.

Table 9.1 Example test matrix

Product	Processor	Interfaces	Display modes	Display size	Cover and attachments
XV200	100MHz	RS485	800 x 600	12"	Plastic, side bracket
XV250	100MHz	RS485, CAN	800 x 600, 1024 x 768	12"	Plastic, side bracket
XV300	100MHz	RS485, CAN, Ethernet	800 x 600, 1024 x 768	14"	Plastic, side bracket
XV301					Plastic, top bracket
XV400	133MHz	RS485, CAN, Ethernet	800 x 600, 1024 x 768, 1280 x 1024	17"	Metal, side bracket
XV401					Metal, top bracket
Testing choice: XV300 tested in 800 x 600 with all interfaces connected will represent XV200, 250 and 301 (same processor, largest size) XV400 tested in 1024 x 768 and 1280 x 1024 will represent XV401 (identical except for fittings)					

9.2.1.3 Revision state and acceptable revision changes

During design and development you will want to perform some confidence tests. At this stage the build state must be carefully defined, by reference to drawing status, even if revision levels have not yet been issued. Once the equipment reaches the stage of compliance testing or customer qualification, the tests must be done on a specimen that is certified as being built to a specific revision level, which should be the same as that which is placed on the market.

The issues that this raises are also addressed in section 16.3.1.2 from the point of view of project management. Essentially, you must choose the test timing such that you can be reasonably sure that any subsequent changes will not invalidate the test results. But the test plan should define the exact build state of the equipment that will be presented for testing and if possible what changes to either hardware or software can be permitted, and the following test report should document any changes that have actually been applied during the tests, usually to make the product pass. Defining the exact build can itself be non-trivial, especially for large systems with many components.

9.2.2 Statement of test objectives

Self-evidently, you will need to state why the tests are to be performed and the type and detail of report to be issued: ranging from a simple certificate, through a report of tests done without results or details, to a full report including set-ups, result values and other relevant information. For manufacturer's self certification to EU Directives, the form and level of detail of the report are not specified and are set by the manufacturer's own requirements.

The time and cost involved in preparing a report obviously will reflect its complexity. Accredited reports are required to contain at least a minimum set of information and the results are to be reported "accurately, clearly, unambiguously and objectively, and in accordance with any specific instructions in the test or calibration methods" [171] – the last clearly relating to the requirements of the basic standards of IEC 61000-4, whose section 10 lays down the content of the report "as necessary to reproduce the test" – but non-accredited reports may turn up in any format and with varying levels of detail. Possible objectives, for recording in your test plan, could be:

- to meet legislation (EMC Directive, FCC certification), as will be necessary to be legally permitted to place the equipment on the market;
- to meet voluntary standards, to improve your competitive advantage;
- to meet a contractual obligation, because EMC performance has been written in to the procurement contract for the equipment or its host system.

9.2.3 The tests to be performed

9.2.3.1 Frequency ranges and voltage levels to be covered

These are normally specified in the standard(s) you have chosen. If you are not using standards, or are extending or reducing the range of the tests, this must be made clear. Even if you are using standards, there may be options for applied levels which must be explicitly chosen; or, for instance, the choice of Class A or Class B emissions limits depends on the EUT's application environment, and you must state how the chosen limits relate to this environment. The test plan should be specific about these parameters and, if one of a range of options has been chosen, should give the rationale for this.

9.2.3.2 Test equipment and facility to be used

In theory, this will also be determined by the standard(s) in use. Most standards have specific requirements for test equipment, e.g. CISPR 16-1-1 instrumentation for emissions measurements and various generators for the IEC 61000-4-X transient tests. If you will be using an external test house they will determine the instrumentation that they will need to use to cover the required tests. If you are doing it in-house, it is your responsibility.

For some tests, the situation isn't so simple. Radiated emissions may be done on an Open Area Test Site or in a semi-anechoic screened room, or possibly in a GTEM or FAR (see sections 6.2.4 and 6.3.1 for more discussion of these options), and the test distance may also be optional. The test facility actually used will probably depend on what is available, but also on other factors such as the size of the EUT. For radiated immunity, again there may be different facilities available and again the EUT dimensions play an important part in what is chosen. For conducted RF immunity and emissions on signal lines, you may have a choice of transducers (see section 7.1.4.1). The test plan needs to pin these variables down, since the standards by themselves don't.

9.2.3.3 Choice of tested ports

The number of ports – a definition which includes the "enclosure port" as well as cable connectors – to be tested directly influences the test time. The standard(s) you choose to apply may define which lines should be tested, for instance the mains lead for conducted emissions. In some cases you can test just one representative line and claim that it covers all others of the same type.

It is important that both you and any subsequent assessment authority know why you have chosen either to apply or to omit tests to particular ports on the EUT. A decision not to test emissions or immunity of certain connected signal or I/O leads may rest on an agreed restriction of the allowable cable length that may be connected to the ports in question – for instance the generic standards apply fast transient and conducted RF immunity tests to signal ports on this basis. The use of a V-network or voltage probe rather than a LISN for supply line emissions measurement may have been due to insufficient current rating of the available coupling network.

The physical position of the test point can be critical, especially for electrostatic discharge application, and must be specified. The choice of ESD application points should be supported by an assessment of likely use of the equipment – which parts are accessible to the user – and/or some preliminary testing to determine weak points.

9.2.3.4 Number and sequence of tests and operating modes

The order in which tests are applied and the sequence of operating modes may or may not be critical, but should be specified. The results of one test may unintentionally set up conditions for the next, or the EUT could be damaged by inadvertent over-testing, especially with a large number of high energy surges in a short time – you may want to leave these till last as a precaution. An alternative and opposite argument is that surge testing could damage filter components, leaving the product with high conducted emissions but still working apparently correctly, so that the potentially damaging tests should be applied *first*, or at least the conducted tests should be repeated after the surge. Make your own decision!

9.2.4 EUT exercising software and ancillary equipment or simulators

9.2.4.1 EUT operating modes

Simple EUTs may have just one mode of operation, but any complex device will have several different operating modes. If you're lucky, you may be able to identify a worst-case mode which includes the majority of operating scenarios and emission/ susceptibility profiles. This will probably need some exploratory testing, and may also need software patches to repeatedly exercise a particular mode. The choice and number of modes has a direct influence on the testing time.

The rate at which a disturbance is applied or an emission measurement is made should depend on the cycle time of the specified operating mode. The test has to dwell on each test frequency for long enough to pick out the most susceptible response or the highest emission. If the EUT might exhibit low duty cycle frequency selective peaks in response, then the test dwell time should be extended to encompass the whole cycle – which, if it is more than a few seconds, can make the total test duration impractical or at least expensive. Sometimes, variations in test results on the same EUT between laboratories can be traced to differences in this parameter.

The EUT may benefit from special software to fully exercise its operating modes; if it is not stand-alone it will need some ancillary support equipment. Both of these should be calibrated or declared fit for purpose.

9.2.4.2 Emissions exercising

It may for example be necessary to set up a particular display pattern to give the worst case emissions: in fact, different display modes (VGA, SVGA, etc.) have different pixel clocks and all of these should be exercised, either with a "representative" picture or with a worst-case pattern. A black-white vertical bar pattern alternating on each horizontal pixel will give a video signal with the highest amplitude of emission at half the pixel clock frequency. You will normally find that any real picture will have its video RF energy distributed over a wider frequency range and therefore have a lower amplitude at any particular frequency, but this doesn't take into account possible resonances in coupling which can enhance certain frequencies, so it is preferable to test both worst-case and representative pictures.

Communications interfaces are normally configured to continually send full packets of data; if they can be set to different data rates then these should all be checked to find the worst case. Memory interfaces such as hard drives, CD-ROMs or flash cards should be exercised by continuous read-write cycles. Variable speed motor drives or similar devices such as light dimmers will need to be evaluated for worst case emissions at different settings. Typically, maximum emissions occur at half power but this isn't universal.

Once you start to explore the possible combinations in a typical complex product, it quickly becomes apparent that you could spend a very long time trying to find the worst case, and usually you have to find an acceptable compromise which adequately fulfils the tests but minimizes their expense and duration. It is often for this reason that test results vary between labs, and the need for a definite and detailed test plan becomes obvious.

9.2.4.3 Immunity exercising and monitoring

Similar issues for exercising apply for immunity tests, with the added complication that the ancillary equipment is often used to monitor for failures against the performance criteria (see section 9.3). This may be for instance checking the bit error rate of a data communications link, or measuring the breakthrough of 1kHz modulation on an audio output. Power supplies or instrumentation could be monitored by logging their output voltage on a DVM. In all these cases the measuring device that checks performance must be unaffected by the disturbance stress (RF or transient) that is being applied to the EUT; and this means either isolating the support and monitoring equipment from the test, or being sure that its EMC performance is better than the equipment being tested.

9.2.4.4 Isolating the support equipment

If the support equipment will be housed in a separate chamber during the test its cables can be interfaced via a filtered bulkhead which will reduce fortuitous emissions and isolate it from disturbances applied to the EUT (Figure 9.1). This filtering arrangement will need to be specified to ensure that, while isolating the support equipment from the test, it does not affect the passage of the wanted signals, and for unscreened cables this requirement can frequently restrict the application of this method. Screened cables are easy to deal with as long as the screen is bonded to the wall of the chamber as it passes through the bulkhead.

If the support equipment is not isolated from the test, filtering is not needed, but the support equipment's own EMC performance must then be sufficiently well specified so that it does not have a bearing on the test outcome. A compromise solution is to use ferrite clamps on the cables between the EUT and the support equipment.

Some military/aerospace tests require that the unit is tested in conjunction with the other items to which it will be connected in the final installation. Besides not affecting the outcome of the test, these items should also offer the appropriate RF terminating impedance to the connected cables. Thus the description of the ancillary equipment set-up must be detailed enough to allow the RF aspects to be reproduced.

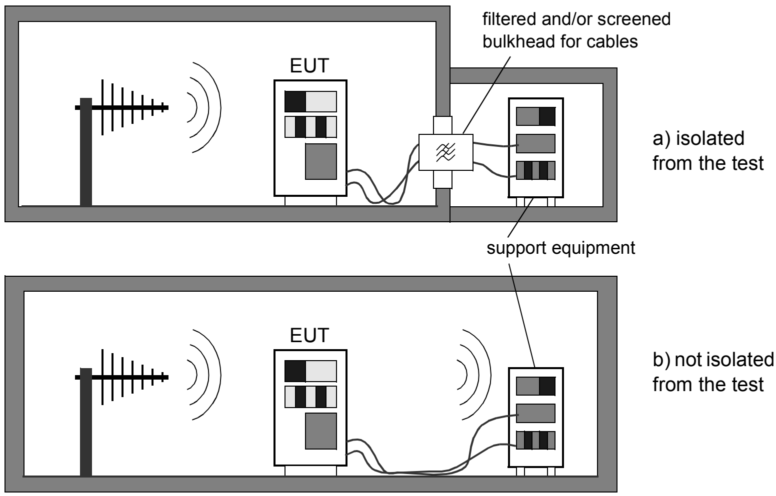


Figure 9.1 Exercising and ancillary support equipment

9.2.5 Requirements of the test facility

Some types of electronic or electrical product make special demands on the services provided by the test house, which must be clearly understood before you start testing as some test houses will be unable to provide them. Requirements may include for instance:

- environmental conditions: special requirements for temperature, humidity, or vibration;
- safety precautions needed: e.g. if the EUT uses ionizing radiations, hazardous gases or extra high voltages, is dangerously heavy or hot, or if the tests require high values of radiated field;
- special handling and functional support equipment: fork lift trucks, large turntables, hydraulic or air/water cooling services – extreme examples include a rolling road dynamometer for whole vehicle testing, or exhaust gas extraction for jet aircraft;
- power sources: AC or DC voltage, current, frequency, single- or three-phase, VA rating and surge current requirement, total number of power lines (remember that FCC certification testing will need US-standard power supplies);
- system command software: will the tests require special software to be written to integrate the test suite with the EUT operation?
- security classification – relevant for some government or military projects.

9.2.6 Details of the test set-up

9.2.6.1 Physical location and layout of the EUT and test transducers

This is defined in general terms in the various standards – see for instance the set-up diagrams in CISPR 22 – but you will usually have to interpret the instructions given in these to apply to your particular EUT. (CISPR 22 assumes that your EUT will normally be a computer, a monitor, a keyboard, a mouse and a peripheral!) The purpose of the standard set-up is to attempt to maximize repeatability between test labs. Critical points for HF tests are distances, orientation and proximity to other objects, especially the ground plane, as this determines stray capacitance coupling which can influence the results; and cable lengths to impedance stabilizing networks or coupling networks for conducted tests. An early decision has to be made as to whether the EUT is to be tested as table-top equipment or floor standing – most standards give different layouts for each. This is fine if your product is clearly intended for one or the other, but if it may be for example handheld or wall mounted you have to apply a degree of interpretation. There are many situations for which there is no “right” answer, and you have to be able to justify your decision and record the actual set-up used in detail, so that it can be replicated.

The final test report should include photographs which record the set-up as well as sketches showing relevant distances. Before the test, the laboratory will need to know the general arrangement that they are expected to implement, which should relate to representative conditions of actual use as far as possible while still being consistent with the standard.

9.2.6.2 Electrical interconnections

Cable layout and routing has a critical effect at high frequencies and must be closely defined. A cable which is run close to the ground plane and in the opposite orientation to the measuring antenna, will radiate far less than one which is suspended in free space and aligned with the antenna. Types of connector and cable to the EUT should be specified, if they would otherwise go by default, as the termination affects the coupling of interference currents between the EUT and the cable. The termination at the far end affects the radiating impedance presented to the EUT and must also be specified. Refer to Figure 10.19 on page 244 for an illustration of how currents, voltages and hence impedances are related to cable length.

This then raises the issue of what is the correct length of cable to use, and how should it be terminated. It is vital to use a type of cable and connector that will actually be used in the real situation, especially if compliance depends on cable screening, as otherwise the test won’t be representative, however closely the EUT itself represents the production model. Section 13.1.7 discusses the effect of the screen termination, from which you can deduce that if you just pick up and use any old cable that happens to be lying around, there is no guaranteeing what results you might get.

The cable length can be equally contentious. If you will be supplying a cable of a particular length with the equipment when it is marketed, by all means use that, with excess length bundled as per any instructions in the standard. But there are many circumstances where you have no control over the actual length of cable that will be used in reality, and it may vary enormously from one application to another. In this case, and if the standard gives no guidance, all that you can do is choose a length that is practical for the test set-up and that *could* be used in reality, and document it in the test plan and the test report.

Finally, the electrical loading of the EUT can be important. If it is a power supply, then the actual power that is drawn can affect its EMC profile, and it would be typical to test it at least at its rated load; but also the performance could get worse when the supply is lightly loaded (this effect has been documented with some television sets, which may have worse emissions in standby than when the TV is operating). How the load is referred to the ground reference of the test must also be defined. A simple resistor on a power supply output, floating with respect to the ground plane, can create a completely different RF impedance compared to a realistic piece of equipment which, even if it's not directly earthed, has a much greater stray capacitance to ground.

9.2.7 How to evaluate test results

9.2.7.1 Acceptance margins

The main results of the tests will be the levels of EUT emissions, or the levels of disturbance at which susceptibility effects occur. There needs to be an accepted way of deriving safety margins between these levels and specification limits, determined by known measurement uncertainties and the likely variations between production units and the EUT. These margins are essential to an interpretation of the test results. The CISPR 80/80 rule, discussed in section 2.2.4.1, can take care of the second of these if you are prepared to test several samples, but if you are only testing one instance of the product then you need to derive a margin essentially by "experienced guesswork". A typical margin that might serve as a starting point would be between 4 and 6dB below the limit, with the option to reduce this if experience shows that products are more uniform – or vice versa, to increase it if they're not.

With respect to measurement uncertainty, section 6.5 discusses emissions uncertainty and Table 6.4 on page 156 indicates the metrologically sound way of reporting pass or fail, but if you are after a compliance declaration then a qualified statement that cannot be quoted with 95% confidence (cases B or C in Table 6.4) is not of much use.

9.2.7.2 Application of measurement uncertainty to emissions

The publication of CISPR 16-4-2 has provided a formal means of dealing with this problem for the three most common emissions tests, i.e. mains conducted, disturbance power, and radiated electric field. The publication gives explicit values for an acceptable uncertainty for each of these (U_{cispr}, see Table 6.5 on page 157), and if the test house uncertainty is less than this, then [165]:

– compliance is deemed to occur if no measured disturbance exceeds the disturbance limit;

– non-compliance is deemed to occur if any measured disturbance exceeds the disturbance limit.

In other words, no account is taken of the actual uncertainty. If the test house uncertainty is greater than U_{cispr}, then the measured value must be increased by the difference between the two before being compared to the limit, so there is an incentive not to use test houses with large declared uncertainties.

The test plan, therefore, can reference CISPR 16-4-2 as a means of eliminating the application of uncertainty for these tests, although eventually we can expect that relevant CISPR documents will do so anyway. For tests for which there is no value given for U_{cispr}, then another way must be stated of applying the uncertainty. It would be ideal if standards themselves specified how to do this, and indeed standards development is moving in this direction; but meanwhile, the alternative is best illustrated by quoting UKAS LAB 34:

> ... In such cases it may be appropriate for the user to make a judgement of compliance, based on whether the result is within the specified limits with no account taken of the uncertainty. This is often referred to as 'shared risk', since the end-user takes some of the risk that the product may not meet the specification. In this case there is an implicit assumption that the magnitude of the uncertainty is acceptable and it is important that the laboratory should be in a position to determine and report the uncertainty. The shared risk scenario is normally only applicable when both the laboratory's client and the end user of the equipment are party to the decision. It would not normally apply to regulatory compliance testing unless expressly referenced by the appropriate regulatory or standards making bodies. Even in these cases the acceptable measurement uncertainty should be stated and the laboratory should demonstrate that its uncertainty meets the specified allowance...
>
> – [195] para 4.3

9.2.7.3 Application of measurement uncertainty to immunity

There is no equivalent of CISPR 16-4-2 for immunity tests. Uncertainty can be quoted for the applied stress in RF immunity tests, but it can't realistically be quoted for the applied stress in transient tests since the transient generator specification combines amplitude and time quantities, and it can't generally be calculated for the measurement of the EUT's response, which is likely to be highly non-linear [142]. LAB 34 says:

> In the case of immunity testing against a specified interference level, e.g. a radiated field strength, it is recommended that, in the absence of other guidance, the test is performed at the specified immunity level increased by the standard uncertainty multiplied by a factor k of 1.64

pointing out that this would give a confidence level in the value of the actual stress of 90%, which in turn means that there is a confidence of 95% that at least the required specification level has been applied. Thus if the lab's standard uncertainty was 1.5V/m on a test level of 10V/m, it would set the actual applied stress to be 12.46V/m, and this would ensure to a 95% confidence that 10V/m had been achieved. However, the catch is that for RF immunity testing to EN 61000-4-3 and -6 there *is* "other guidance"; CENELEC have produced "interpretation sheets" for these two standards which state that:

> The test field strengths are to be applied as stated in Tables 1 and 2, or as defined in the product standard, without any increase to take into account uncertainties in the calibration of the field

for EN 61000-4-3, and

> The test levels are to be applied as stated in Table 1, or as defined in the product standard, without any increase to take into account uncertainties in the setting of the output level at the EUT port of the coupling device. The test generator shall be adjusted to produce the nominal value of U_{mr} as defined in 6.4.1 of the standard

for EN 61000-4-6. One can only assume that the relevant standard committee understood that this was equivalent to requiring a confidence level of just 50% that the proper stress level had been applied, and this was their intention, but since they've published the documents and the ENs have now been amended, you are at liberty to take this approach for RF immunity testing if you are referencing EN 61000-4-3 or -6. It would be advisable for the test plan to state whether the test level is to be increased by (1.64 · standard uncertainty), or not. Note that there is no equivalent interpretation for the international standards IEC 61000-4-3 or -6, although amendments regarding measurement uncertainty are in train for them.

9.3 Immunity performance criteria

When you perform immunity testing, it is essential to be able to judge whether the EUT has in fact passed or failed the test. This in turn demands a statement of the minimum acceptable performance which the EUT must maintain during and after testing. Such a statement can only be related to the EUT's own functional operating specification.

9.3.1 The generic criteria

The variety and diversity of equipment and systems makes it difficult to lay down general criteria for evaluating the effects of interference on electronic products. Nevertheless, the test results can be classified on the basis of operating conditions and the functional specifications of the EUT according to the criteria discussed below. To provide a basis for the statement of acceptable performance, the generic immunity standards (see section 4.4) contain a set of guidelines for the criteria against which the operation of the EUT can be judged, and which are used to formulate the acceptance criteria for a given EUT against specific tests:

> ***Performance criterion A****: The apparatus shall continue to operate as intended. No degradation of performance or loss of function is allowed below a performance level specified by the manufacturer, when the apparatus is used as intended. In some cases the performance level may be replaced by a permissible loss of performance. If the minimum performance level or the permissible performance loss is not specified by the manufacturer then either of these may be derived from the product description and documentation (including leaflets and advertising) and what the user may reasonably expect from the apparatus if used as intended.*
>
> - This criterion applies to phenomena which are normally continuously present, such as RF interference.
>
> ***Performance criterion B****: The apparatus shall continue to operate as intended after the test. No degradation of performance or loss of function is allowed below a performance level specified by the manufacturer, when the apparatus is used as intended. During the test, degradation of performance is however allowed. No change of actual operating state or stored data is allowed. If the minimum performance level or the permissible performance loss is not specified by the manufacturer then either of these may be derived from the product description and documentation (including leaflets and advertising) and what the user may reasonably expect from the apparatus if used as intended.*
>
> - This applies to transient phenomena.
>
> ***Performance criterion C****: Temporary loss of function is allowed, provided the loss of function is self recoverable or can be restored by the operation of the controls.*
>
> - This applies to long duration mains interruption.

9.3.2 Interpreting the generic criteria

It is up to the manufacturer to specify the limits which define "degradation or loss of function" and what is a "permissible performance loss", and to decide which of these criteria should be applied to each test. Such specifications may be prompted by preliminary testing or by known customer requirements. In any case it is important that they are laid out in the final EMC test plan for the equipment, since at the very least you need to know how to monitor the EUT during the tests. If the equipment is being supplied to a customer on a one-to-one contractual basis then clearly there is room for mutual agreement and negotiation on acceptance criteria, but this is not possible for products placed on the mass market, which have only to meet the essential requirements of the EMC or R&TTE Directives. In these cases, you have to look to the immunity standards for general guidance.

An example could be, if a measuring instrument has a quoted accuracy of 1% under normal conditions, it would be reasonable to expect this accuracy to be maintained when subject to RF interference at the level specified in the standard, unless your operating manual and sales literature specifies a lower accuracy under such conditions. It may lose accuracy when transients are applied, but must recover it afterwards. A processor-based product may exhibit distortion or “snow” on the image displayed on its video monitor under transient interference, but it must not crash nor suffer corruption of data.

The second edition of the EMC Directive has made little change in the essential requirement for immunity except to require that the apparatus should “operate without unacceptable degradation of its intended use” in the presence of electromagnetic disturbance (the word “unacceptable” is new). In the same document emphasis is also placed on the information provided with the product, particularly that which is necessary to enable it to be used “in accordance with its intended purpose”. In other words, the performance criteria may be tailored to a compromise which restricts the definition of “intended use” and “unacceptable degradation”.

Of course, what your customers think of these issues will also be important to you. The generic criteria make clear that in the absence of any other information, what the user may “reasonably expect” is the defining point. The consequence of this is that the EMC test plan is of importance to the marketing department as well as to engineering; it is not acceptable for marketing to shrug off all responsibility for EMC compliance. Though they may not have a technical input, they must be prepared to sign off the performance criteria as well as other aspects, such as choice of cables, operating modes, functional configuration and so on.

Product-specific criteria

Some product-specific immunity standards can be more precise in their definition of acceptable performance levels. For example EN 55020, applying to broadcast receivers, specifies a wanted to unwanted audio signal ratio for sound interference, and a just perceptible picture degradation for vision interference. Even this relatively tight definition may be open to interpretation. Another example is EN 55024 for IT equipment, which gives particular criteria for the immunity of telephone audio links, fax machines and VDUs. Telecommunications equipment is covered by various ETSI standards; typically it might be required to comply with a defined criterion for bit error rate and loss of frame alignment.

The subjectivity of immunity performance criteria can still present a major headache in the legal context of the Directives. It will undoubtedly be open to many manufacturers to argue if challenged, not only that differing test procedures and layouts have been used in judging compliance of the same product, but that differing criteria for failure have also been applied. It will be in their own interest to be clear and precise in their supporting documentation as to what they believe the acceptance criteria are, and to ensure that these are in line as far as possible with either the generic guidelines or those given in applicable product standards.

Chapter 10

Interference coupling mechanisms

10.1 Source and victim

Situations in which the question of electromagnetic compatibility arises invariably have two complementary aspects. Any such situation must have a source of interference emissions and a victim which is susceptible to this interference. If either of these is not present, or if there is no coupling path between them, there is no EMC problem. If both source and victim are within the same piece of equipment we have an “intra-system” EMC situation; if they are two different items, such as a computer monitor and a radio receiver, it is said to be an “inter-system” situation. The standards which were discussed in Chapter 4 were all related to controlling inter-system EMC. The same equipment may be a source in one situation and a victim in another.

Knowledge of how the source emissions are coupled to the victim is essential, since a reduction in the coupling factor is often the only way to reduce interference effects, if a product is to continue to meet its performance specification. The two aspects are frequently reciprocal, that is measures taken to improve emissions will also improve the susceptibility, though this is not invariably so. For analysis, they are more easily considered separately [103].

Systems EMC

Putting source and victim together shows the potential interference routes that exist from one to the other (Figure 10.1). When systems are being built, you need to know the emissions signature and susceptibility of the component equipment, to determine whether problems are likely to be experienced with close coupling. Adherence to published emission and susceptibility standards does not guarantee freedom from systems EMC problems. Standards are written from the point of view of protecting a particular service – in the case of emissions standards, this is radio broadcast and telecommunications – and they have to assume a minimum separation between source and victim.

Most electronic hardware contains elements which are capable of antenna-like behaviour, such as cables, PCB tracks, internal wiring and mechanical structures. These elements can unintentionally transfer energy via electric, magnetic or electromagnetic fields which couple with the circuits. In practical situations, intra-system and external coupling between equipment is modified by the presence of screening and dielectric materials, and by the layout and proximity of interfering and victim equipment and especially their respective cables. Ground or screening planes will enhance an interfering signal by reflection or attenuate it by absorption. Cable-to-cable coupling can be either capacitive or inductive and depends on orientation, length and proximity. Dielectric materials may also reduce the field by absorption, though this is negligible compared with the effects of conductors in most practical situations.

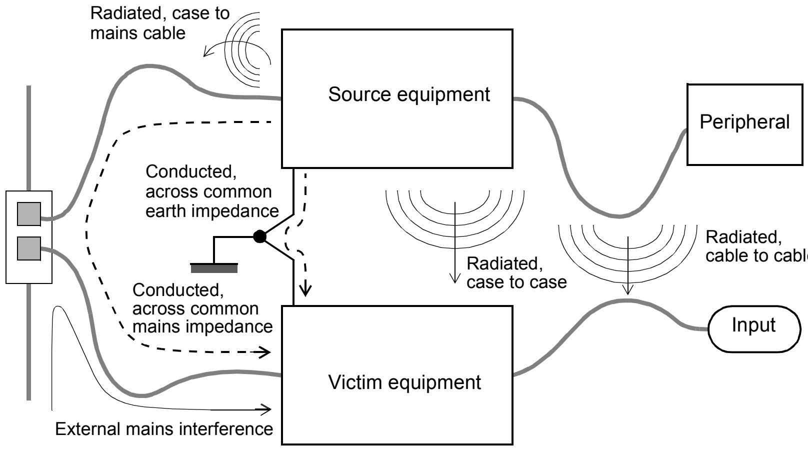


Figure 10.1 Coupling paths

10.1.1 Common impedance coupling

Common impedance coupling routes are those which are due to a circuit impedance that the source shares with the victim. The most obvious common impedances are those in which the impedance is physically present, as with a shared conductor; but the common impedance may also be due to mutual inductive coupling between two current loops, or to mutual capacitive coupling between two voltage nodes. Philosophically speaking, every node and every loop is coupled to all others throughout the universe. Practically, the strength of coupling falls off very rapidly with distance. Figure 10.4 (later) shows the variation of mutual capacitance and inductance of a pair of parallel wires versus their separation, and the field equations in Appendix D (section D.3.8) give the precise expressions for the field at any point due to a radiating element.

10.1.1.1 Conductive connection

When an interference source (output of system A in Figure 10.2) shares a ground connection with a victim (input of system B) then any current due to A's output flowing through the common impedance section X-X develops a voltage in series with B's input. The common impedance need be no more than a length of wire or PCB track. High frequency or high di/dt components in the output will couple more efficiently because of the inductive nature of the impedance (see appendix D section D.5.2 for the inductance of various conductor configurations). The voltage developed across an inductor as a result of current flow through it is given by equation (10.1). The output and input may be part of the same system, in which case there is a spurious feedback path through the common impedance which can cause oscillation.

$$V_N = -L \cdot dI_L/dt \qquad (10.1)$$

where L is the self inductance in henries

This equation, incidentally, is one of the most basic and universal in the whole EMC canon: it is the reason why digital PCBs create radio frequency emissions, why long

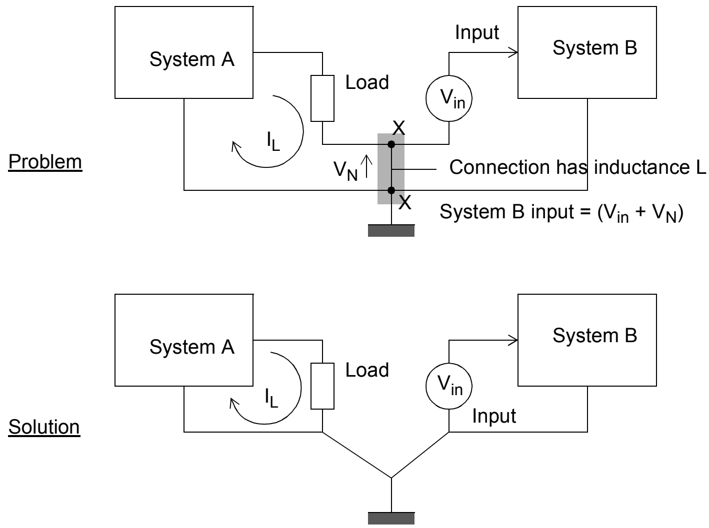


Figure 10.2 Conducted common impedance coupling

ground wires are anathema, and why switched circuits generate fast transient bursts, among many other ills for which it is responsible.

The solution as shown in Figure 10.2 is to separate the connections so that there is no common current path, and hence no common impedance, between the two circuits. The only "penalty" for doing this is the need for extra wiring or track to define the separate circuits. This applies to any circuit which may include a common impedance, such as power rail connections. Grounds are the most usual source of common impedance because the ground connection, often not shown on circuit diagrams, is taken for granted.

10.1.1.2 Magnetic induction

Alternating current flowing in a conductor creates a magnetic field which will couple with a nearby conductor and induce a voltage in it (Figure 10.3(a)). The voltage induced in the victim conductor is now given by equation (10.2):

$$V_N = -M \cdot dI_L/dt \qquad (10.2)$$

where M is the mutual inductance in henries

Notice the similarity between this and equation (10.1). M depends on the areas of the source and victim current loops, their orientation and separation distance, and the presence of any magnetic screening. Appendix D (section D.5.2) gives mutual inductance formulae, but typical values for short lengths of cable loomed together lie in the range 0.1 to 3μH. The equivalent circuit for magnetic coupling is a voltage generator in series with the victim circuit. The coupling is unaffected by whether or not there is a direct connection between the two circuits; the induced voltage would be the same if the circuits were isolated or connected to ground.

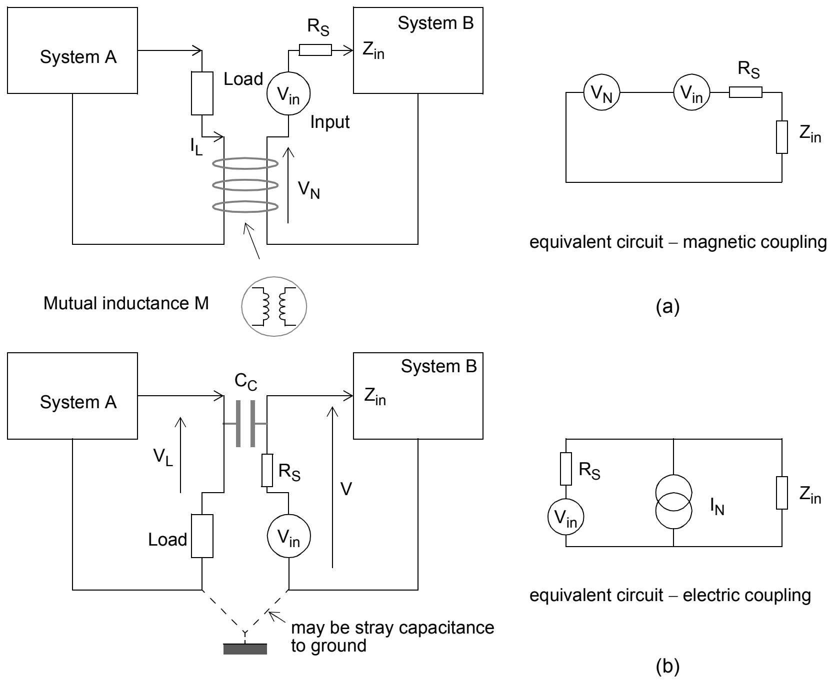


Figure 10.3 Magnetic and electric induction

10.1.1.3 Electric induction

Changing voltage on one conductor creates an electric field which may couple with a nearby conductor and induce a voltage on it (Figure 10.3(b)). The voltage induced on the victim conductor in this manner is:

$$V_N = C_C \,.\, dV_L/dt \,.\, Z_{in}//R_S, \qquad (10.3)$$

where C_C is the coupling capacitance and
$Z_{in}//R_S$ is the impedance to ground of the victim circuit

This assumes that the impedance of the coupling capacitance is much higher than that of the circuit impedances. The noise is injected as if from a current source with a value of $C_C \cdot dV_L/dt$. The value of C_C is a function of the distance between the conductors, their effective areas and the presence of any electric screening material. Typically, two parallel insulated wires 0.1" apart show a coupling capacitance of about 50pF per metre; the primary-to-secondary capacitance of an unscreened medium power mains transformer is 100–1000pF.

Floating circuits

It seems that both circuits need to be referenced to ground for the coupling path to be complete. But if either is floating, this does *not* mean that there is no coupling path: the floating circuit will exhibit a stray capacitance to ground and this is in series with the direct coupling capacitance. Alternatively, there will be stray capacitance direct from the circuit nodes of system A to B even in the absence of any ground node. The noise current will still be injected across R_L but its value will be determined by the series combination of C_C and the other stray capacitance. Morrison [9] gives a good overview of how nested capacitances interact to create capacitive coupling paths.

10.1.1.4 Effect of input impedance

The difference in equivalent circuits for magnetic and electric coupling means that their behaviour with a varying circuit input impedance is different. Electric field coupling *increases* with an increasing Z_{IN} while magnetic field coupling *remains constant* with an increasing Z_{IN}. This property can be useful for diagnostic purposes; if you are able to vary Z_{IN} while observing the coupled voltage, you can deduce which mode of coupling predominates. For the same reason, magnetic coupling is more of a problem for low-impedance circuits while electric coupling applies to high impedance circuits.

10.1.1.5 Spacing

Both mutual capacitance and mutual inductance are affected by the physical separation of source and victim conductors. Figure 10.4 shows the effect of spacing on mutual capacitance of two parallel wires in free space, and on mutual inductance of two conductors over a ground plane (the ground plane provides a return path for the current). Appendix D includes the equations from which this graph derives.

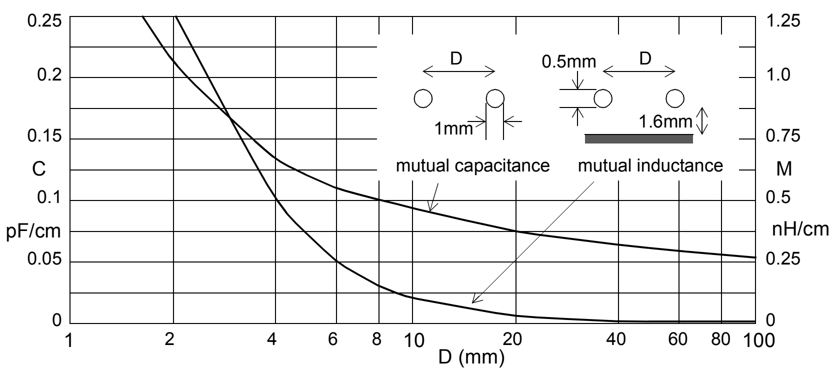


Figure 10.4 Mutual capacitance and inductance versus spacing

10.1.2 Distributed near field coupling

10.1.2.1 Low frequency model

The discussion in 10.1.1.2 and 10.1.1.3 assumes that the coupling mechanism – inductive or capacitive – occurs at a single point in the circuit, or at least can be approximated to such a point. It also assumes that inductive and capacitive coupling can be treated quite separately. In reality, and particularly in cables and on PCB tracks, neither assumption is justifiable. When near field coupling between circuits is distributed over an appreciable length then the two mechanisms interact and the circuit must be analysed in more detail.

Consider the equivalent circuit of two conductors in a cable as shown in Figure 10.5. For the purposes of the analysis we can assume that the current return paths are

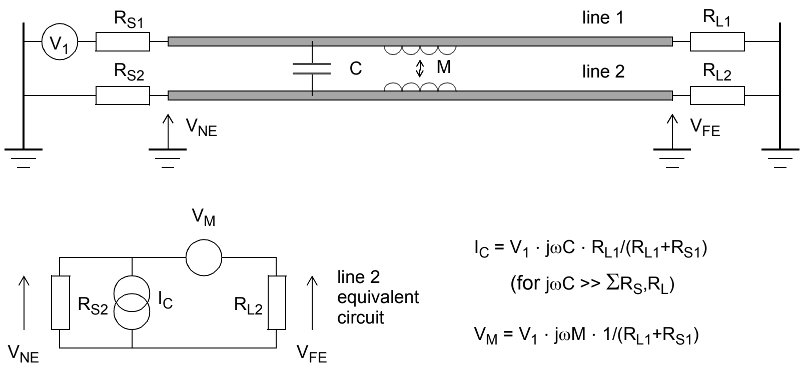


Figure 10.5 Superposition of inductive and capacitive coupling

both via a remote ground plane, and that the loads are purely resistive. The interference is induced in the second circuit magnetically, in series with the conductor (V_M), and capacitively, in parallel with it (I_C). Then the sum total of the interference at each end is given by the superposition of the two sources. But since the magnetically-induced voltage is in series with the conductor it appears with one sign at one end but the opposite sign at the far end:

$$V_{NE}(C) = I_C \cdot R_{S2} \cdot R_{L2}/(R_{S2}+R_{L2}) = V_{FE}(C) \quad \text{(Capacitive coupling)} \qquad (10.4)$$

$$V_{NE}(L) = V_M \cdot R_{S2}/(R_{S2}+R_{L2}) \quad \text{(Inductive coupling – near end)} \qquad (10.5)$$

$$V_{FE}(L) = -V_M \cdot R_{L2}/(R_{S2}+R_{L2}) \quad \text{(Inductive coupling – far end)} \qquad (10.6)$$

So

$$V_{NE}(tot) = V_{NE}(C) + V_{NE}(L) = (I_C \cdot R_{L2} + V_M) \cdot (R_{S2}/(R_{S2}+R_{L2})) \qquad (10.7)$$

$$V_{FE}(tot) = V_{FE}(C) + V_{FE}(L) = (I_C \cdot R_{S2} - V_M) \cdot (R_{L2}/(R_{S2}+R_{L2})) \qquad (10.8)$$

Since this is a crosstalk phenomenon it gives rise to the terms "near-end crosstalk" (NEXT) and "far-end crosstalk" (FEXT).

10.1.2.2 High frequency model

The above model is only valid at low frequencies, that is if the length of the coupled lines is much less than a wavelength. A more general approach treats the two conductors as transmission lines with lumped L and C parameters, and integrates the coupling contributions along the length of the line. These parameters together with the source and load terminating impedances then determine the coupling time constants. Appendix D section D.5.4 gives the full coupling equations, and Figure 10.6 shows their effect: at frequencies below the breakpoint determined by the terminated line's time constant, the LF model applies and the coupling increases monotonically with frequency. At higher frequencies the lines become resonant and a maximum coupling is reached, followed by a series of nulls and peaks at integer multiples of a half wavelength.

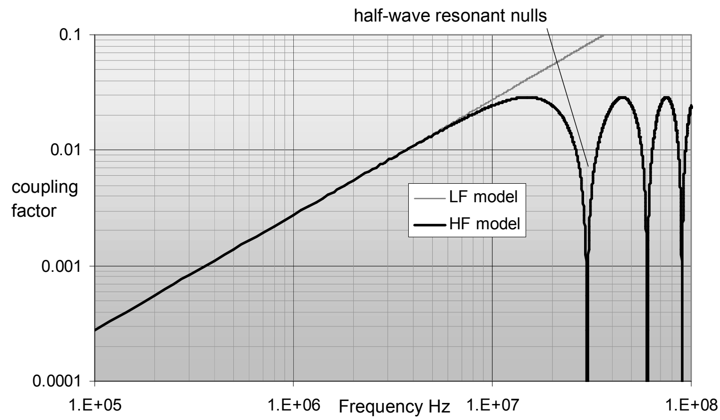


Figure 10.6 High frequency line coupling, line length 5m

10.1.3 Mains coupling

Interference can propagate from a source to a victim via the mains distribution network to which both are connected. This is not well characterized at high frequencies, especially since connected electrical loads can present virtually any RF impedance at their point of connection. We have already seen that the RF impedance presented by the mains as specified in conducted emissions tests can on the average be approximated by a network of 50Ω in parallel with 50μH between each phase and earth (section 6.2.2.1). For short distances such as between adjacent outlets on the same ring, coupling via the mains connection of two items of equipment can be represented by the equivalent circuit of Figure 10.7, assuming that the signal is injected and received between phase and earth.

Over longer distances, power cables are fairly low loss transmission lines of around 150–200Ω characteristic impedance up to about 10MHz. However, in any local power

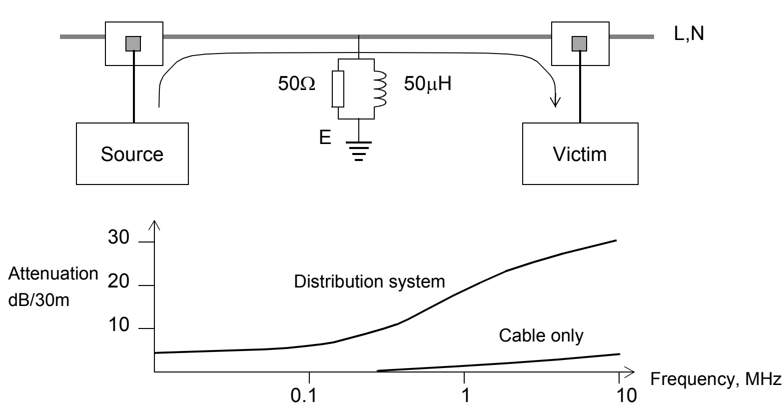


Figure 10.7 Coupling via the mains network

distribution system the disturbances and discontinuities introduced by load connections, cable junctions and distribution components will dominate the RF transmission characteristic. These all tend to increase the attenuation and introduce marked peaks and nulls in the frequency–impedance characteristic.

10.1.4 Radiated coupling

To understand how energy is coupled from a source to a victim at a distance with no intervening connecting path, you need to have a basic understanding of electromagnetic wave propagation. This section will do no more than introduce the necessary concepts. The theory of EM waves has been well covered in many other works [3][7][13].

10.1.4.1 Field generation

An electric field (E field) is generated between two conductors at different potentials. The field is measured in volts per metre and is proportional to the applied voltage divided by the distance between the conductors.

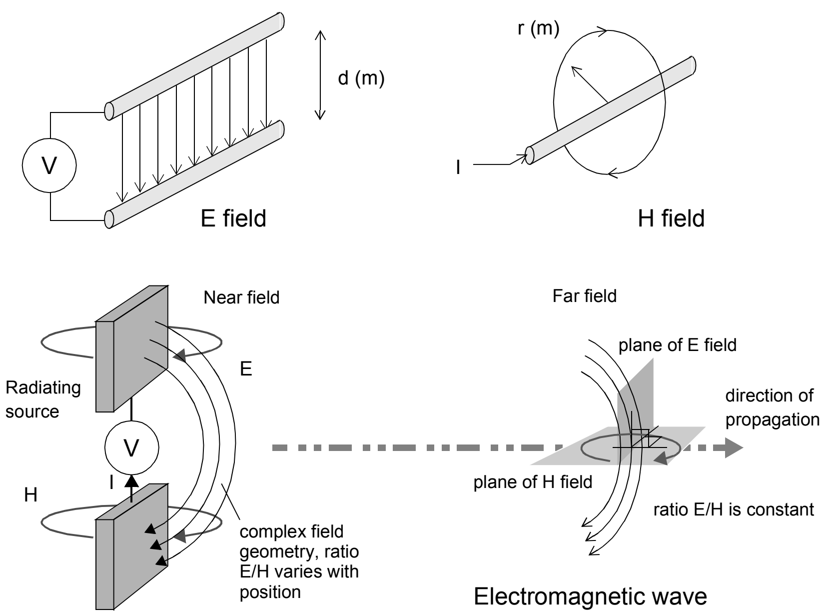


Figure 10.8 Electromagnetic fields

A magnetic field (H field) is generated around a conductor carrying a current, is measured in amps per metre and is proportional to the current divided by the distance from the conductor.

When an alternating voltage generates an alternating current through a network of conductors – a description which applies to any electronic circuit – an electromagnetic (EM) wave is generated which propagates as a combination of E and H fields. The speed of propagation is determined by the medium; in free space it is equal to the speed of light, $3 \cdot 10^8$m/s. Near to the radiating source the geometry and strength of the fields depend on the characteristics of the source. A conductor carrying a significant di/dt will generate mostly a magnetic field; a circuit node carrying a significant dv/dt will generate mostly an electric field. The structure of these fields will be determined by the physical layout of the source conductors, as well as by other conductors, dielectrics and permeable materials nearby. As you move around a typical electronic product, the fields vary in a highly complicated fashion and it is difficult to draw any hard and fast conclusions about their distribution and hence how to control them, as anyone who has used a near field probe for detailed diagnostics will know. But Maxwell's laws say that further away from the source, the complex three-dimensional field structure decays and only the components which are orthogonal to each other and to the direction of propagation remain. Figure 10.8 demonstrates these concepts graphically.

10.1.4.2 Wave impedance

The ratio of the electric to magnetic field strengths (E/H) is called the "wave impedance" (Figure 10.9). The wave impedance is a key parameter of any given wave as it determines the efficiency of coupling with another conducting structure, and also the effectiveness of any conducting screen which is used to block it. In the far field, that is for $d > \lambda/2\pi$, the wave is known as a plane wave and the E and H fields decay with distance at the same rate. Therefore its impedance is constant, and is equal to the impedance of free space given by equation (10.9):

$$Z_o = \sqrt{(\mu_o/\varepsilon_o)} = 120\pi = 377\Omega \tag{10.9}$$

where μ_o is $4\pi \cdot 10^{-7}$ H/m (the permeability of free space)
and ε_o is $8.85 \cdot 10^{-12}$ F/m (the permittivity of free space)

In the near field, $d < \lambda/2\pi$, the wave impedance is determined by the characteristics of the source. A low current, high-voltage radiator (such as a dipole) will generate mainly an electric field of high impedance, while a high current, low-voltage radiator (such as a loop) will generate mainly a magnetic field of low impedance. If (as a special case) the radiating structure happens to have an impedance around 377Ω, then a plane wave can in fact be generated in the near field, depending on geometry.

The region around $\lambda/2\pi$, or approximately one sixth of a wavelength, is the transition region between near and far fields. This is not a precise criterion, rather it indicates the region within which the field structure changes from complex to simple. Plane waves are always assumed to be in the far field, while for the near field it is necessary to consider individual electric or magnetic fields separately. Appendix D presents the formulae (Maxwell's field equations) that underpin this description.

10.1.4.3 The Rayleigh criterion

There is another definition of the transition between near and far fields, determined by the Rayleigh range. This has to do not with the field structure according to Maxwell's equations, but with the nature of the radiation pattern from any physical antenna (or equipment under test) which is too large to be a point source. To maintain the far field assumption, the phase difference between the field components radiated from the extremities of the antenna must be small, and therefore the path differences to these extremities must also be small in comparison to a wavelength. This produces a criterion that relates the wavelength and the maximum dimension of the antenna (or EUT) to the distance from it. Using the Rayleigh criterion, the far field distance is defined as:

$$d > 2D^2/\lambda \quad \text{where D is the maximum dimension of the antenna} \tag{10.10}$$

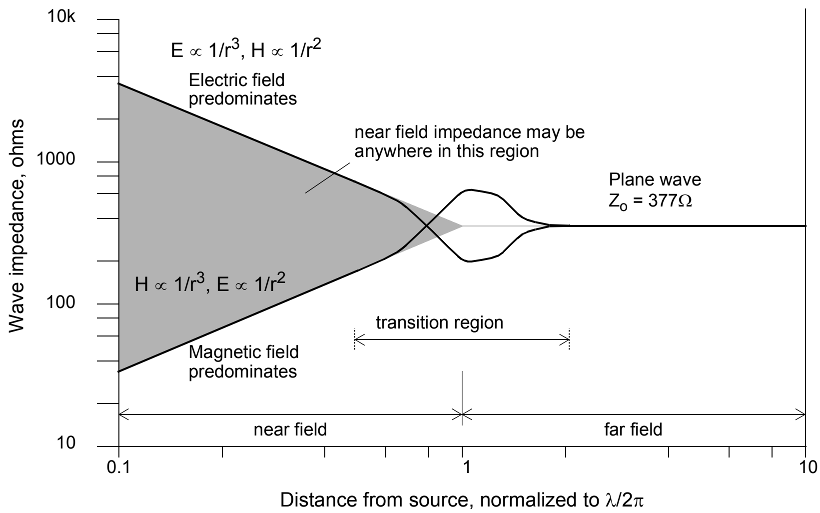


Figure 10.9 The wave impedance from Maxwell's laws

Table 10.1 shows a comparison of the distances for the two criteria for the near field/far field transition for various frequencies and EUT dimensions. Note how with typical EUT dimensions the Rayleigh range determines the far field condition for frequencies in excess of 100–200MHz.

Table 10.1 Rayleigh and Maxwell distances for transition to far field

Frequency	**Maximum dimension D (m)**	**Rayleigh** $d = 2D^2/\lambda$ (m)	**Maxwell** $d = \lambda/2\pi$ (m)
10MHz	1	0.067	4.77
	3	0.6	
30MHz	1	0.2	1.59
	3	0.6	
100MHz	0.3	0.06	0.477
	1	0.67	
	3	6.0	
300MHz	0.3	0.18	0.159
	1	2.0	
1GHz	0.3	0.6	0.0477
	1	6.67	

10.1.5 Coupling modes

The concepts of differential mode, common mode and antenna mode radiated field coupling are fundamental to an understanding of EMC and will crop up in a variety of guises throughout this book. They apply to coupling of both emissions and incoming interference.

10.1.5.1 Differential mode

Consider two items of equipment interconnected by a cable (Figure 10.10). The cable carries signal currents in ***differential*** mode (go and return) down the two wires in close proximity. A radiated field can couple to this system and induce differential mode interference between the two wires; similarly, the differential current will induce a radiated field of its own. The ground reference plane (which may be external to the equipment or may be formed by its supporting structure) plays no part in the coupling.

10.1.5.2 Common mode

The cable also carries currents in ***common*** mode, that is, all flowing in the same direction on each wire. These currents very often *have nothing at all to do with the signal currents*. They may be induced by an external field coupling to the loop formed by the cable, the ground plane and the various impedances connecting the equipment to ground, and may then create internal differential currents to which the equipment is susceptible. Alternatively, they may be generated by internal noise voltages between the ground reference point and the cable connection, and be responsible for radiated emissions. The existence of RF common mode currents means that *no* cable, whatever signal it may be intended to carry – even a single wire – can be viewed as safe from the EMC point of view.

Notice that the stray capacitances and inductances associated with the wiring and enclosure of each unit are an integral part of the common mode coupling circuit, and play a large part in determining the amplitude and spectral distribution of the common mode currents. These stray impedances are incidental rather than designed in to the equipment (or, let us say, they are "designed" by the mechanical designer rather than by the electronics designer). They don't appear on any circuit diagram, and are much harder to control or predict than those parameters such as cable spacing and filtering which determine differential mode coupling.

10.1.5.3 Antenna mode

Antenna mode currents are carried in the same direction by the cable and the ground reference plane. They should not arise as a result of internally-generated noise, but they will flow when the whole system, ground plane included, is exposed to an external field. An example would be when an aircraft flies through the beam of a radar transmission; the aircraft structure, which serves as the ground plane for its internal equipment, carries the same currents as the internal wiring. Antenna mode currents only become a problem for the radiated field susceptibility of self-contained systems when they are converted to differential or common mode by varying impedances in the different current paths.

10.1.5.4 Conversion between differential and common mode

Although it was said above that common mode currents may be unrelated to the intended signal currents, there may also be a component of common mode current

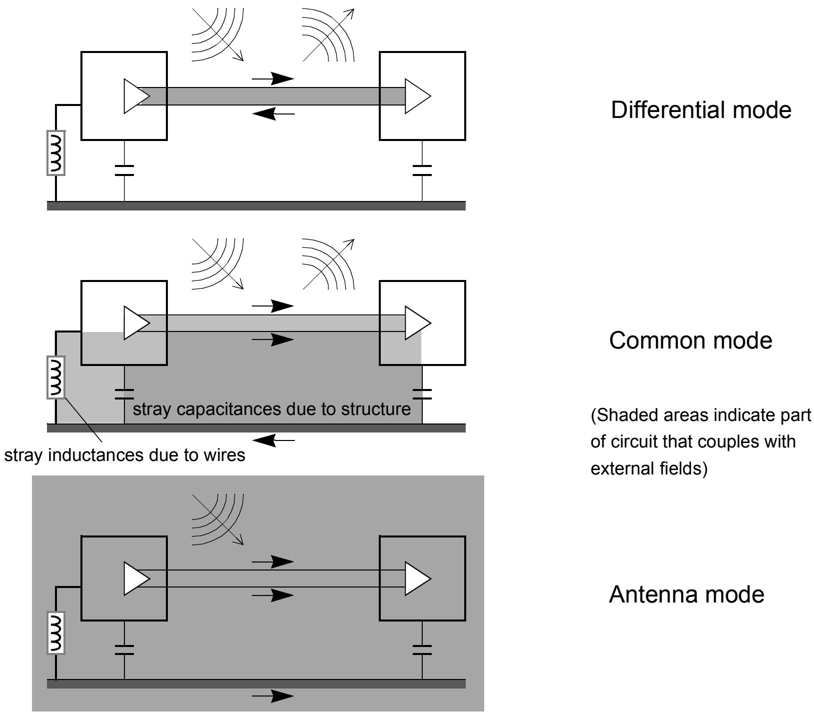


Figure 10.10 Radiated coupling modes

which *is* due to the signal current. Conversion occurs when the two signal conductors present differing impedances to their environment, represented by the external ground. These impedances are dominated at RF by stray capacitance and inductance related to physical layout, and are only under the circuit designer's control if that person is also responsible for physical layout.

In Figure 10.11 the differential mode current I_{DM} produces the desired signal voltage across load R_L. The common mode current I_{CM} does not flow through R_L directly but through impedances Z_A, Z_B and back via the external ground. Z_A, Z_B are not circuit components but distributed stray impedances, typically but not always capacitive, and are determined by such factors as surface area of PCB tracks and components and their proximity to chassis metalwork and other parts of the equipment. If $Z_A = Z_B$ then no voltage is developed across R_L by the common mode currents I_{CM}. But any inequality results in such a voltage, proportional to the differences in impedance:

$$V_{load(CM)} = I_{CM} \cdot Z_A - I_{CM} \cdot Z_B = I_{CM} \cdot (Z_A - Z_B) \tag{10.11}$$

For this reason, circuits which carry high frequency interfering signals (such as wideband data or video) or which could be susceptible to RF are best designed in such a way that the stray impedances of each conductor are balanced as nearly as possible. Alternatively, a common mode choke (section 13.2.4.1) is used which swamps the imbalance of the strays and reduces the magnitude of I_{CM}. As an example, both techniques are used to maximum effect in the standard interface circuit for high-speed Ethernet (100Mb/s) connections.

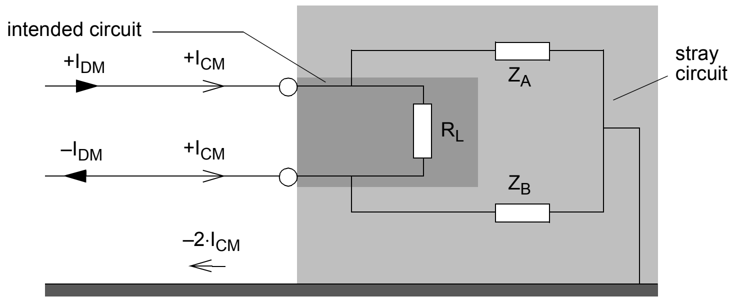


Figure 10.11 Differential to common mode conversion

The increasing popularity of wideband data transmission via unscreened cables within and between buildings has sharpened the problem of interference radiated from these cables. As well as the balance of the circuit at either end of the cable, the balance of the cable itself, as it passes near to other conducting structures in its environment, is an important factor. This is largely determined by the quality of the cable construction, and has resulted in a cable parameter known as "longitudinal conversion loss" (LCL) being defined. LCL is treated again in section 13.1.9.1.

10.1.5.5 Generalization

The principles demonstrated in the circuits of Figure 10.10 and Figure 10.11 are not limited to currents propagating down cables between modules. The circuits can be extended to include currents on interconnections between PCBs in an individual module, or even on tracks between parts of a PCB mounted on a chassis. Many EMC problems of products can be traced to the common mode currents flowing *within* them as well as outside them.

10.2 Emissions

When designing a product to a specification without knowledge of the system or environment in which it will be installed, you will normally separate the two aspects of emissions and susceptibility, and design to meet minimum requirements for each. Limits are laid down in various standards but individual customers or market sectors may have more specific requirements. In those standards which derive from CISPR (see Chapter 4), emissions are sub-divided into radiated emissions from the system as a whole, and conducted emissions present on the interface and power cables. Conventionally, the breakpoint between radiated (high frequency) and conducted (low frequency) is set at 30MHz, primarily for convenience of measurement. Radiated emissions can themselves be classified into those that derive from differential currents on internal PCBs or other wiring, and those from common mode currents on PCBs, conducting structures, or external cables that are connected to the equipment.

10.2.1 Radiated emissions

10.2.1.1 Radiation from the PCB

In most equipment, the primary sources are currents flowing in circuits (clocks, video and data drivers, and other oscillators or impulsive sources) that are mounted on printed circuit boards. Some of the energy is radiated directly from the PCB, which in the simplest instance can be modelled as a small loop antenna carrying the interference current (Figure 10.12). A small loop is one whose dimensions are smaller than a quarter

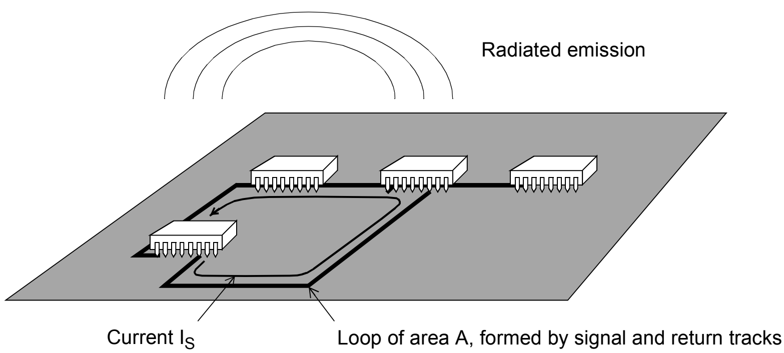


Figure 10.12 PCB radiated emissions

wavelength (λ/4) at the frequency of interest (e.g. 1m at 75MHz). Most PCB loops count as "small" at emission frequencies up to a few hundred MHz. When the dimensions approach λ/4 the currents at different points on the loop appear out of phase at a distance, so that the effect is to reduce the field strength at any given point. The maximum electric field strength from such a loop over a ground plane at 10m distance is proportional to the square of the frequency:

$$E = 263 \cdot 10^{-12} (f^2 \cdot A \cdot I_S) \text{ volts per metre [11]} \tag{10.12}$$

where A is the loop area in cm^2,
f (MHz) is the frequency of I_S the source current in mA.

In free space and in the far field, the field strength falls off proportionally to distance from the source. The figure of 10m is used as this is the standard measurement distance for the European radiated emissions standards. A factor of 2 times is allowed for worst-case field reinforcement due to reflection from the ground plane, which is also a required feature of testing to standards.

The loop whose area must be known is the overall path taken by the signal current and its return. Equation (10.12) assumes that I_S is at a single frequency. For square waves with many harmonics, the Fourier spectrum must be used for I_S. These points are taken up again in section 12.1.2.

Assessing PCB design

You can use equation (10.12) to indicate roughly whether a given PCB design will need extra screening. For example if A = 10cm^2, I_S = 20mA, f = 50MHz then the field strength E is 42dBμV/m, which is 12dB over the EN Class B limit. If the frequency and current are fixed, and the loop area cannot be reduced, screening will be necessary.

But the converse is not true. Differential mode radiation from small loops on PCBs is by no means the only contributor to radiated emissions; common mode currents flowing on the PCB and on attached cables can contribute much more. Paul [112] goes so far as to say:

> Predictions of radiated emissions based solely on differential-mode currents will generally bear no resemblance to measured levels of radiated emissions. Therefore, basing system EMC design on differential-mode currents and the associated prediction models that use them exclusively while neglecting to consider the (usually much larger) emissions due to common-mode currents can lead to a strong 'false sense of security'.

Common mode currents on the PCB itself are not at all easy to predict, in contrast with the differential mode currents which are governed by Kirchoff's current law. The return path for common mode currents is via stray capacitance to other nearby objects as well as via the stray inductance of intervening structures, and therefore a full prediction would have to take the detailed mechanical structure of the PCB and its case, as well as its proximity to ground and to other equipment, into account. Except for trivial cases this is to all intents and purposes impossible. It is for this reason more than any other that EMC design has earned itself the distinction of being a "black art".

10.2.1.2 The PCB as patch antenna

However, a PCB structure can be modelled in a different way. Consider that circuit operation creates a voltage and current distribution along the length of the transmission line formed between 0V and power planes, or 0V plane and signal tracks. Then at the edges of the transmission line the discontinuity creates radiating fields which propagate away from the PCB (Figure 10.13). In the case of planes, the radiation will occur from each pair of opposite edges. A radio antenna designer will recognize this as a description of a patch antenna, and in fact patch antenna theory can be used quite successfully to describe the radiation mechanism.

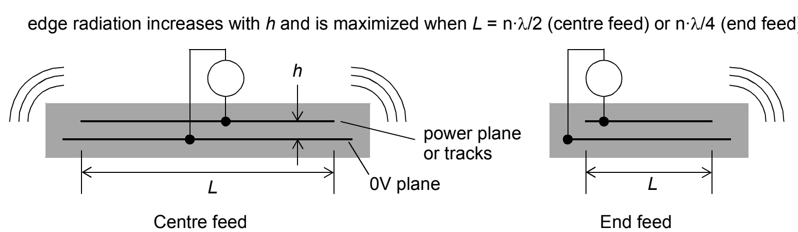


Figure 10.13 The patch antenna model

Radiation is increased by greater separation distance *h*, and is also maximized when there is 180º phase difference between two opposite edges, which means that the physical distance across the planes *L* is an electrical half wavelength (λ/2). Since a half wavelength within a PCB is $\sqrt{\varepsilon_r}$ shorter than in free space, for a typical fibreglass PCB (ε_r = 4.2) a plane size of, say, 15cm will be a half wavelength at around 490MHz; smaller planes will show correspondingly higher frequencies. At integer multiples of this frequency the PCB will show maxima in radiating efficiency.

The detailed emission pattern from a patch antenna depends on the edge geometry and on the feed point, particularly on where the noisiest components are located, which of course is a function of the PCB layout. The above analysis assumes a feed near the centre of the line; if it is near an edge then maximum feed current occurs when the line is n·λ/4 long. In two dimensions, different feed positions excite various x and y direction modes across the plane. In general, the analysis for typical complex circuit layouts with several layers, several potential sources and, of course, populated with components, is too complicated to allow you to evaluate alternatives for minimum radiation. Rather, it is best to make the relevant transmission line (power and 0V planes) the lowest possible impedance so as to create the minimum radiating voltage at the edges, which is handled by minimum separation distance *h* and by proper decoupling; and to design the plane layout so as to prevent the edges from creating sharp discontinuities which radiate effectively. This means avoiding long runs of straight edges with the two plane edges parallel and adjacent (see the 10·h rule described in section 11.2.2.7).

This description, of course, refers to radiation from a PCB in free space, or at least in an unshielded plastic enclosure. Placing PCBs in a complete or partial conductive enclosure will modify the model drastically, and if you wanted to analyse it, you would need a multi-level numerical electromagnetic modelling suite (and plenty of time!).

10.2.1.3 Radiation from cables

Fortunately (from some viewpoints) radiated coupling at VHF tends to be dominated by cable emissions, rather than by direct radiation from the PCB. This is for the simple reason that typical cables resonate in the 30–100MHz region, and their radiating efficiency is higher than PCB structures at these frequencies. The interference current is generated in common mode from ground noise developed across the PCB or elsewhere in the equipment and may flow along the conductors, or along the shield of a shielded cable. A model for this effect has been well described in [33]; simplistically, the capacitances between the noise-generating track and external ground, and between the PCB's ground reference and external ground, form a network for return of currents that are injected into cables connected to the PCB. A more detailed model can be generated which includes the inductance of the tracks and plane. Reference [33] describes how it is possible from this model to derive coupling parameters for a PCB which can then be applied to the large-scale system (enclosure and cables) of which the PCB is a part, to predict the common mode radiated emissions.

The model for cable radiation at lower frequencies (Figure 10.14) is a short ($L < \lambda/4$) monopole antenna over a ground plane. (When the cable length is resonant the model becomes invalid; see section D.3.6 in Appendix D for an equation describing emissions from resonant cables.) The maximum field strength, allowing +6dB for ground plane reflections at 10m, due to this radiation is directly proportional to frequency:

$$E = 1.26 \cdot 10^{-4} \cdot (f \cdot L \cdot I_{CM}) \text{ volts per metre [11]} \tag{10.13}$$

where L is the cable length in metres and
I_{CM} is the common mode current at f MHz in mA flowing in the cable.

For a 1m cable, I_{CM} must be less than 20µA for a 10m-distance field strength at 50MHz of 42dBµV/m – i.e., a thousand times less than the equivalent differential mode current! To meet the 30dBµV/m limit, the current needs to be 12dB or four times less, i.e. 5µA, under these conditions. And indeed, this value of 5µA when measured as a common mode cable current is regarded as a good indicator of likely compliance with radiated emissions limits [5] – any more than 5µA and the product has a good chance of failing the compliance test. I_{CM} can easily be measured in the development lab using a current probe or absorbing clamp (see section 6.2.2.2 and 6.2.2.3) and so this forms a useful diagnostic or pre-compliance check on the prospects for a given item.

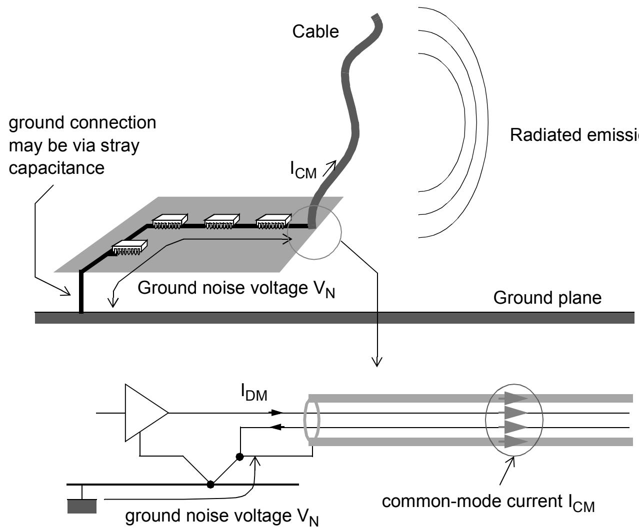


Figure 10.14 Cable radiated emissions

Common mode cable noise

At the risk of repetition, it is vital to appreciate the difference between common mode and differential mode cable currents. Differential mode current, I_{DM} in Figure 10.14, is the current which flows in one direction along one cable conductor and in the reverse direction along another. It is normally equal to the signal or power current, and is not present on the cable shield. It contributes little to the net radiation as long as the total loop area formed by the two conductors is small; the two currents tend to cancel each other. Common mode current I_{CM} flows equally in the same direction along all conductors in the cable, potentially including the shield, and is only related to the (differential) signal currents insofar as these are converted to common mode by unbalanced external impedances, and may be quite *un*related to them. For instance, an RS-232 interface cable will only be carrying data at a rate of perhaps 19.2kbaud, but can also unintentionally carry common mode noise from the circuit ground which is polluted by a processor clock and its harmonics at hundreds of MHz. It returns via the associated ground network and therefore the radiating loop area is large and uncontrolled. As a result, even a small I_{CM} can result in large emitted signals.

10.2.2 Conducted emissions

Interference sources within the equipment circuit or its power supply are coupled onto the power cable to the equipment. Interference may also be coupled either inductively or capacitively from another cable onto the power cable. Attention has focussed on the power cable as the prime source of conducted emissions since CISPR-based standards have historically only specified limits on this cable. The mains supply wiring is an efficient means of coupling interference around a building. However, signal and control cables can and do also act as coupling paths, especially those which travel long

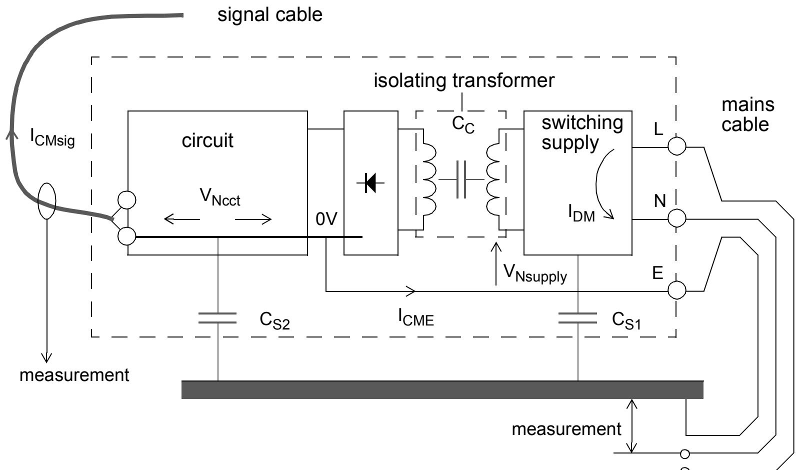


Figure 10.15 Coupling paths for conducted emissions

distances such as LAN and telecom lines, and later versions of the standards apply measurements to these cables as well.

The resulting interference may appear as differential mode (between live and neutral, or between signal wires) or as common mode (between live/neutral/signal and earth) or as a mixture of both. For signal and control lines, only common mode currents are of interest. For the mains port, the voltages between each phase/neutral and earth at the far end of the mains cable are measured. Differential mode emissions are normally associated with low frequency switching noise from the power supply, while common mode emissions can be due to the higher frequency switching components, internal circuit sources or inter-cable coupling.

10.2.2.1 Coupling paths

Figure 10.15, showing a typical product with a switched mode supply, gives an idea of the various paths these emissions can take. (Section 12.1.5 looks at SMPS emissions in more detail.) Differential mode current I_{DM} generated at the input of the switching supply is measured as an interference voltage across the load impedance of each line with respect to earth at the measurement point. Higher frequency switching noise components $V_{Nsupply}$ are coupled through C_C, the coupling capacitance between primary and secondary of the isolating transformer, to appear between L/N and E on the mains cable, and C_{S1} to appear with respect to the ground plane. Circuit ground noise V_{Ncct} (digital noise and clock harmonics) is referenced to ground by C_{S2} and coupled out via signal cables as I_{CMsig} or via the safety earth as I_{CME}.

The problem in a real situation is that all these mechanisms are operating simultaneously, and the stray capacitances C_S are widely distributed and unpredictable, depending heavily on proximity to other objects if the case is unscreened. A partially-screened enclosure may actually worsen the coupling because of its higher capacitance to the environment.

An extra consequence of coupling through C_S is that the interference measured at the mains connection can depend not only on the noise sources directly connected to the supply, but also on the connection of any signal lines and their loads. The circuit of

Figure 10.15 shows that I_{CMsig} flows both through the externally connected signal cable and the stray capacitances as well as the earth lead; if the earth lead is absent (as in safety class II apparatus) then C_S determines the whole path. If the common mode impedance to ground at the far end of the signal cable goes down, then I_{CMsig} will increase and therefore the voltage developed by this current across the impedance of C_S will also increase. This voltage is then passed on to the measurement at the mains port. From this analysis you can see that a true measurement of mains conducted emissions must be carried out with the correct and representative common mode load impedances connected to each of the relevant signal or other connections. Measuring conducted (or indeed radiated) emissions with signal ports disconnected will give false results.

10.2.2.2 Simplified equivalent circuits

The basic equivalent circuit for conducted emissions testing on the mains port is shown in Figure 10.16(a). The mains connection is represented by the AMN/LISN giving a defined RF impedance between live and earth, and between neutral and earth. The EUT contains both differential and common mode sources, generalized here as appearing in one case between live and neutral, and in the other case between both live and neutral with respect to earth. If the apparatus is safety class II there is no earth wire, but common mode signals can still return via the stray capacitance to the ground plane.

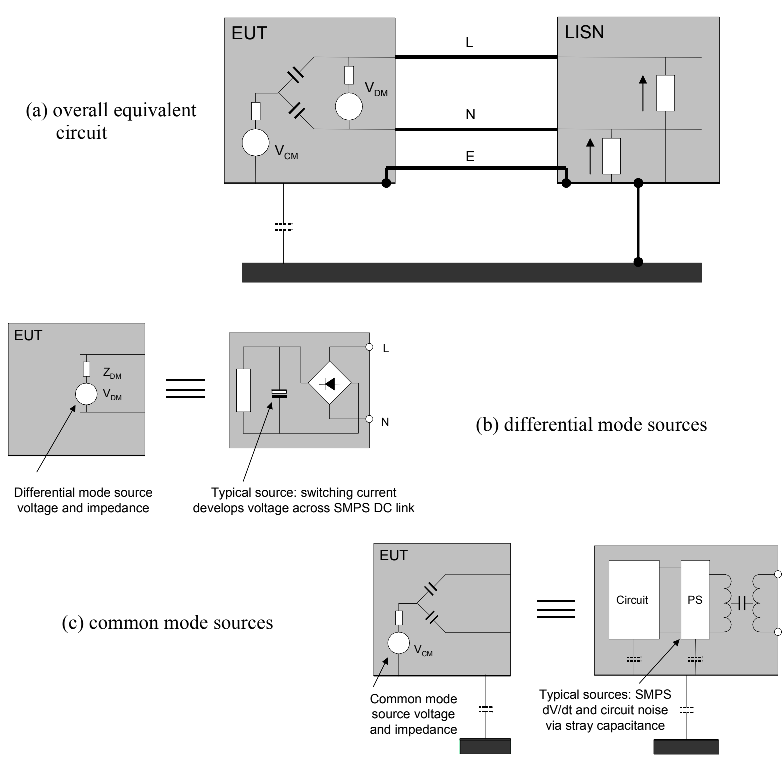


Figure 10.16 Equivalent circuits for conducted emissions tests

Differential mode sources appear between live and neutral connections without reference to the earth connection (Figure 10.16(b)). In circuits with switchmode power supplies or other power switching circuits the RF emissions are dominated by interference developed across the DC link to the switching devices. Although there will normally be a reservoir capacitor, the high di/dt through this capacitor will generate voltages at the harmonics of the switching frequency across its equivalent series impedance. Diode noise, if it is significant, will also appear in differential mode.

Common mode sources (Figure 10.16(c)) are more complex. The common mode voltage appears between both live and neutral with respect to earth. Since the mains input is normally isolated from earth, it is usual for common mode coupling to be capacitive.

The coupling is dominated by the inter-winding capacitance of the isolating transformer and the stray capacitances of noise sources, both in the power supply (e.g. from heatsinks) and the operating circuit. These capacitances are referred to earth, either directly or via the enclosure if this is conductive. A well-shielded enclosure will minimize “leakage” of this capacitive coupling and hence reduced conducted emissions. Other impedances may appear in the coupling path: for instance the leakage inductance of the isolating transformer is in series with its inter-winding capacitance and may give a series resonant peak in the MHz range.

10.3 Immunity

Electronic equipment will be susceptible to environmental electromagnetic fields and/ or to disturbances coupled into its ports via connected cables. An electrostatic discharge may be coupled in via the cables or the equipment case, or a nearby discharge can create a local field which couples directly with the equipment. The potential threats are:

- radiated RF fields;
- conducted transients;
- electrostatic discharge (ESD);
- magnetic fields;
- supply voltage disturbances.

Quite apart from legal requirements, equipment that is designed to be immune to these effects – especially ESD and transients – will save its manufacturer considerable expense through improved reliability and reduced field returns. Although many aspects of emission control are also relevant for immunity, in some cases the shielding and circuit suppression measures that are required for protection against ESD or RF interference may be more than you need for simple compliance with emission standards.

10.3.1 Radiated field

An external field can couple either directly with the internal circuitry and wiring in differential mode or with the cables to induce a common mode current (Figure 10.17). Coupling with internal wiring and PCB tracks is most efficient at frequencies above a few hundred MHz, since wiring lengths of a few inches approach resonance at these frequencies.

RF voltages or currents in analogue circuits can induce nonlinearity, overload or unintended DC bias and in digital circuits can corrupt data transfer [117]. Modulated

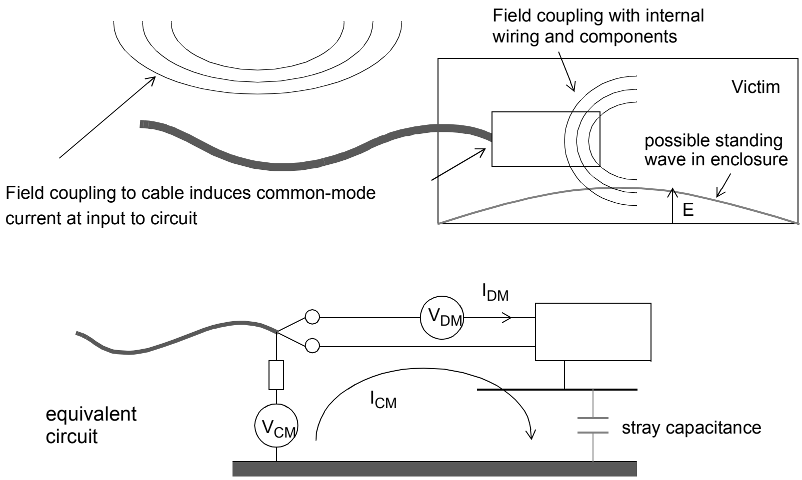


Figure 10.17 Radiated field coupling

fields can have greater effect than unmodulated ones. Likely sources of radiated fields are mobile transmitters, cellphones, high-power broadcast transmitters and radars. Field strengths between 1 and 10V/m from 20MHz to 1GHz are typical, and higher field strengths can occur in environments close to such sources.

Reciprocity

Because the coupling mechanisms for RF immunity are essentially the same as those for RF emissions, and because they are usually linear (unaffected by the amplitude of the coupled signal), the principle of *reciprocity* has evolved. In essence it states that:

> reductions in coupling which reduce emissions over a particular frequency range will also improve immunity over that frequency range

or, on the other hand,

> frequencies at which emissions are particularly troublesome will also be those at which immunity problems occur.

This concept relies on the observation that coupling mechanisms usually show resonant behaviour which maximizes the coupling at certain frequencies. The reciprocity principle should not be accorded too much weight: actual EMC performance depends on the operation of both source and victim circuits as well as on the coupling between them, and these are rarely either reciprocal or linear. Understanding the principle can help in dealing with many coupling-related problems, though.

10.3.1.1 Cable resonance

Cables are most efficient at coupling RF energy into equipment at the lower end of the vhf spectrum (30–100MHz). The external field induces a common mode current on the cable shield or on all the cable conductors together, if it is unshielded. The common mode cable current effects tend to dominate over the direct field interactions with the equipment as long as the equipment's dimensions are small compared with half the wavelength of the interfering signal.

A cable connected to a grounded victim equipment can be modelled as a single conductor over a ground plane, which appears as a transmission line (Figure 10.18, and compare this also to Figure 10.6). The current induced in such a transmission line by an external field increases steadily with frequency until the first resonance is reached, after which it exhibits a series of peaks and nulls at higher resonances [19]. The coupling mechanism is enhanced at the resonant frequency of the cable, which depends on its length and on the reactive loading of whatever equipment is attached to its end. A length of 2m is quarter-wave resonant at 37.5MHz, half-wave resonant at 75MHz.

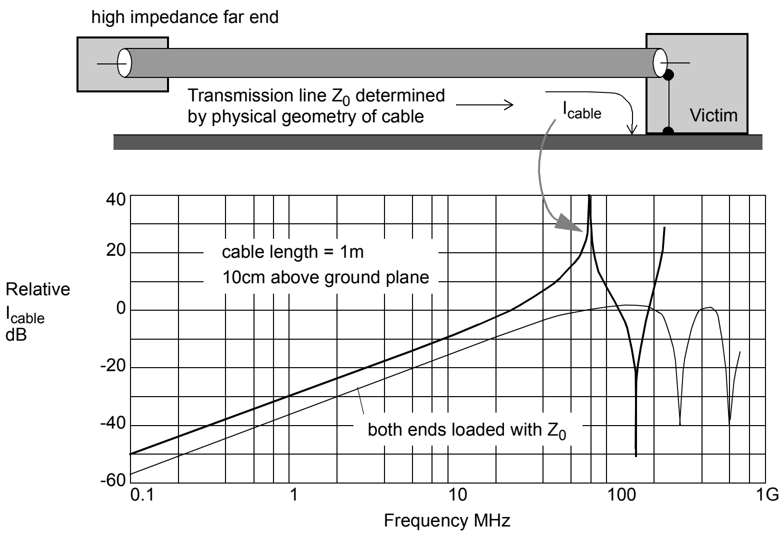


Figure 10.18 Cable coupling to radiated field

Cable loading

The dominant resonant mode depends on the RF impedance (high or low) at the distant end of the cable. If the cable is connected to an ungrounded object such as a hand controller it will have a high RF impedance, which will cause a high coupled current at quarter-wave resonance and high coupled voltage at half-wave. Extra capacitive loading such as body capacitance will lower its apparent resonant frequency.

Conversely, a cable connected to another grounded object such as a separately earthed peripheral will see a low impedance at the far end, which will generate high coupled current at half-wave and high coupled voltage at quarter-wave resonance. Extra inductive loading, such as the inductance of the earth connection, will again tend to lower the resonant frequency.

These effects are summarized in Figure 10.19. The RF common mode impedance of the cable varies from around 35Ω at quarter-wave resonance to several hundred ohms maximum. A convenient average figure (and one that is taken in many standards) is 150Ω. Because cable configuration, layout and proximity to grounded objects are outside the designer's control, attempts to predict resonances and impedances accurately are unrewarding.

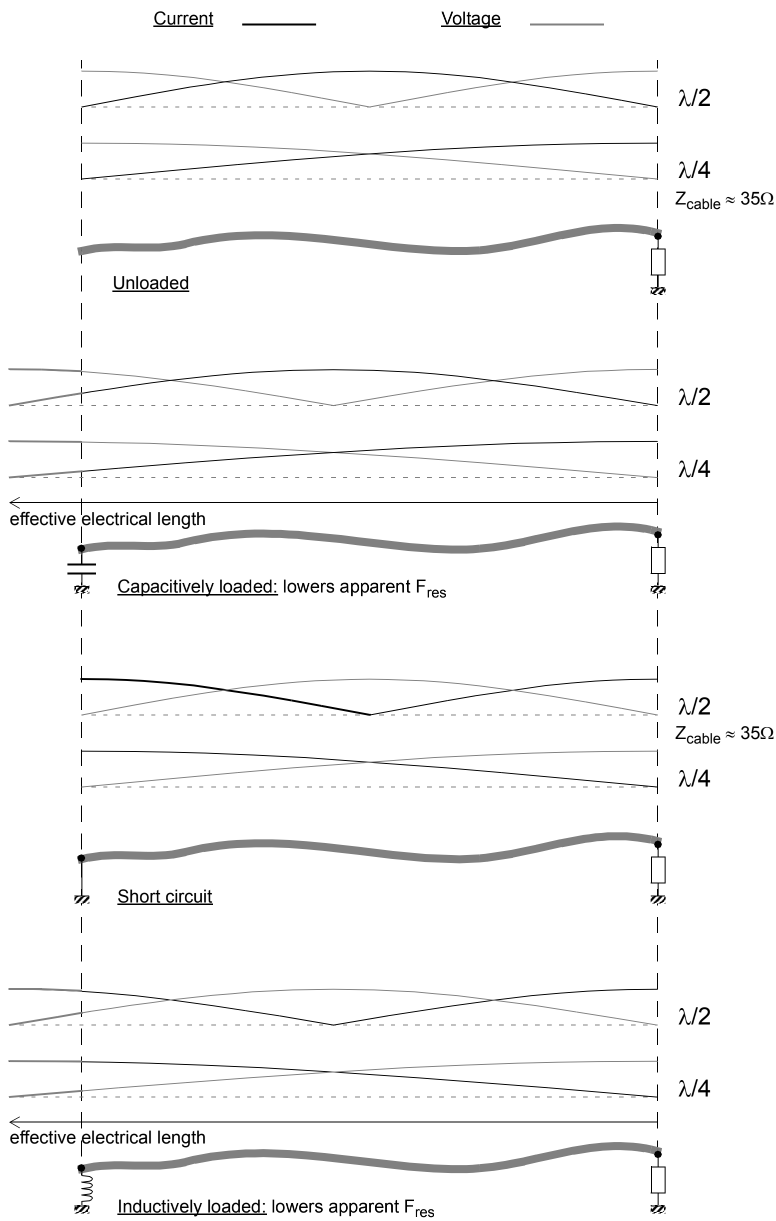


Figure 10.19 Current and voltage distribution along a resonant cable

10.3.1.2 Current injection

A convenient method for testing the RF susceptibility of equipment without reference to its cable configuration is to inject RF as a common mode current or voltage directly onto the cable port (see also section 7.1.4)[94]. This represents real-life coupling situations at lower frequencies well, until the equipment dimensions approach a half wavelength. It can also reproduce the fields (E_{RF} and H_{RF}) associated with radiated field coupling. The route taken by the interference currents, and hence their effect on the circuitry, depends on the various internal and external RF impedances to earth, as shown in Figure 10.20. Connecting other cables will modify the current flow to a marked extent, especially if the extra cables interface to a physically different location on the PCB or equipment. An applied voltage of 1V, or an injected current of 3–10mA, can be taken to correspond in typical cases to a radiated field strength of 1V/m.

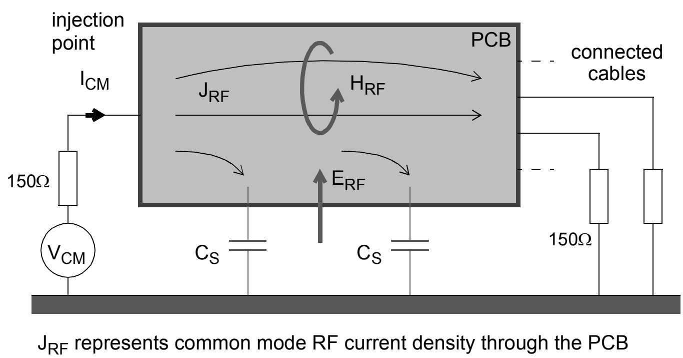


Figure 10.20 Common mode RF injection

However there is considerable disagreement over any single figure for conversion from radiated to injected, and it is generally accepted that conducted tests do not directly represent radiated tests at all [113], because of the variability attributable to multiple cable connections.

10.3.1.3 Cavity resonance

A screened enclosure can form a resonant cavity; standing waves in the field form between opposite sides when the dimension between the sides is a multiple of a half-wavelength. The electric field is enhanced in the middle of this cavity while the magnetic field is enhanced at the sides. In practice, real product enclosures show more complex behaviour since they are neither absolutely regular shapes nor are they empty. This effect is usually responsible for peaks in the susceptibility versus frequency profile in the UHF region, and is also a contributor to the reciprocal nature of susceptibility peaks corresponding with emission peaks. It is discussed further in section 14.1.3.

10.3.2 Transients

Transient overvoltages occur on the mains supply leads due to switching operations, fault clearance or lightning strikes elsewhere on the network. Transients over 1kV account for about 0.1% of the total number of transients observed. A study by the German ZVEI [74] made a statistical survey of 28,000 live-to-earth transients exceeding 100V, at 40 locations over a total measuring time of about 3,400 hours. Their results were analysed for peak amplitude, rate of rise and energy content. Table 10.2 shows the average rate of occurrence of transients for four classes of location, and Figure 10.21 shows the relative number of transients as a function of maximum transient amplitude. This shows that the number of transients varies roughly in inverse proportion to the cube of peak voltage.

Table 10.2 Average rate of occurrence of mains transients

Area Class	Average rate of occurrence (transients/hour)
Industrial	17.5
Business	2.8
Domestic	0.6
Laboratory	2.3

Sources:
Transients in Low Voltage Supply Networks, [74]

Characterization of Transient and CW Disturbances Induced in Telephone Subscriber Lines, [75]

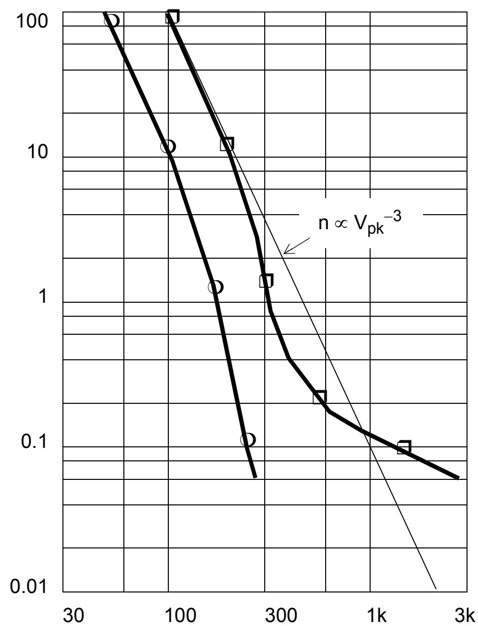


Figure 10.21 Relative number of transients (per cent) vs. maximum transient amplitude (volts)
❐ Mains lines (V_T 100V), ❍ Telecomm lines (V_T 50V)

High energy transients may destroy active devices in the equipment power supply. On the other hand fast-rising edges are more disruptive to circuit operation, since they are attenuated least by the coupling paths and they can generate large voltages in inductive ground and signal paths. The ZVEI study found that rate of rise increased roughly in proportion to the square root of peak voltage, being typically 3V/ns for 200V pulses and 10V/ns for 2kV pulses. Other field experience has shown that mechanical switching produces multiple transients (bursts) with risetimes as short as a few nanoseconds and peak amplitudes of several hundred volts. High frequency attenuation through the mains network (see section 10.1.3) restricts fast risetime pulses to those generated locally.

Analogue circuits are almost immune to isolated short transients, whereas digital circuits are easily corrupted by them. As a general guide, microprocessor equipment should be tested to withstand pulses at least up to 2kV peak amplitude. Thresholds below 1kV will give unacceptably frequent corruptions in nearly all environments, while between 1kV–2kV occasional corruption will occur. For high reliability equipment, a 4–6kV threshold is not too much.

10.3.2.1 Coupling mode

Mains transients may appear in differential mode (symmetrically between live and neutral) or common mode (asymmetrically between live/neutral and earth). Coupling between the conductors in a supply network tends to mix the two modes. Differential mode spikes are usually associated with relatively slow risetimes and high energy, and require suppression to prevent input circuit damage but do not, provided this suppression is incorporated, affect circuit operation significantly. Common mode transients are harder to suppress because they require connection of suppression components between live and earth, or in series with the earth lead, and because stray capacitances to earth are harder to control. Their coupling paths are very similar to those followed by common mode RF signals. Unfortunately, they are also more disruptive because they result in transient current flow in ground traces.

10.3.2.2 Spectral density and energy content

Transient interference is inherently broadband, and its frequency distribution is described by its amplitude spectral density: that is, the amplitude over a defined bandwidth versus frequency, expressed in volts per Hertz or volt-seconds. If the actual waveshape of a transient is known then the spectral density can be derived by taking the Fourier transform of this time domain waveform. In general of course the waveforms of real transients vary widely, but Figure 10.22 shows the spectral densities of the waveforms which have been standardized in the IEC 61000-4 series of immunity tests. If the frequency domain coupling transfer function is known even approximately, then the spectral density can be multiplied by this transfer function to work out the amplitude of an incoming transient at points of interest in the circuit [16].

The energy content of transients and surges is not simple to define. The actual energy available from the source is not all dissipated in the load. That proportion which is, depends on the ratio of the load and source impedances. In general, a load such as a surge suppressor will be non-linear and will also have a time or frequency dependence.

As an approximate indication of the energy content of a particular standard transient, the actual energy delivered by the generator into a defined resistive load can be calculated. For the ESD and EFT waveforms, these can be the calibration loads of 2Ω and 50Ω respectively. For the surge and ring waves, a load which matches the output impedance can be chosen, and the voltage or current waveform is delivered into this resistance with half the open circuit (or short circuit, for current) amplitude. In practice, of course, this doesn't happen since the load is not matched to the output impedance. In these cases the energy in Joules (watt seconds) is shown in Figure 10.22 and is given by

$$W = \frac{1}{R} \cdot \int_0^T \left(\frac{V(t)}{2}\right)^2 dt \qquad W = R \cdot \int_0^T \left(\frac{I(t)}{2}\right)^2 dt$$

where V(t) and I(t) are the open circuit voltage and short circuit current waveforms, respectively.

These graphs are for comparative purposes only – the real energy delivered to a particular EUT can only be calculated if the load impedance and characteristics, and the actual waveshape applied to this load, are known accurately.

10.3.2.3 Transients on signal lines

Fast transients can be coupled, usually capacitively, onto signal cables in common mode, especially if the cable passes close to or is routed alongside an impulsive

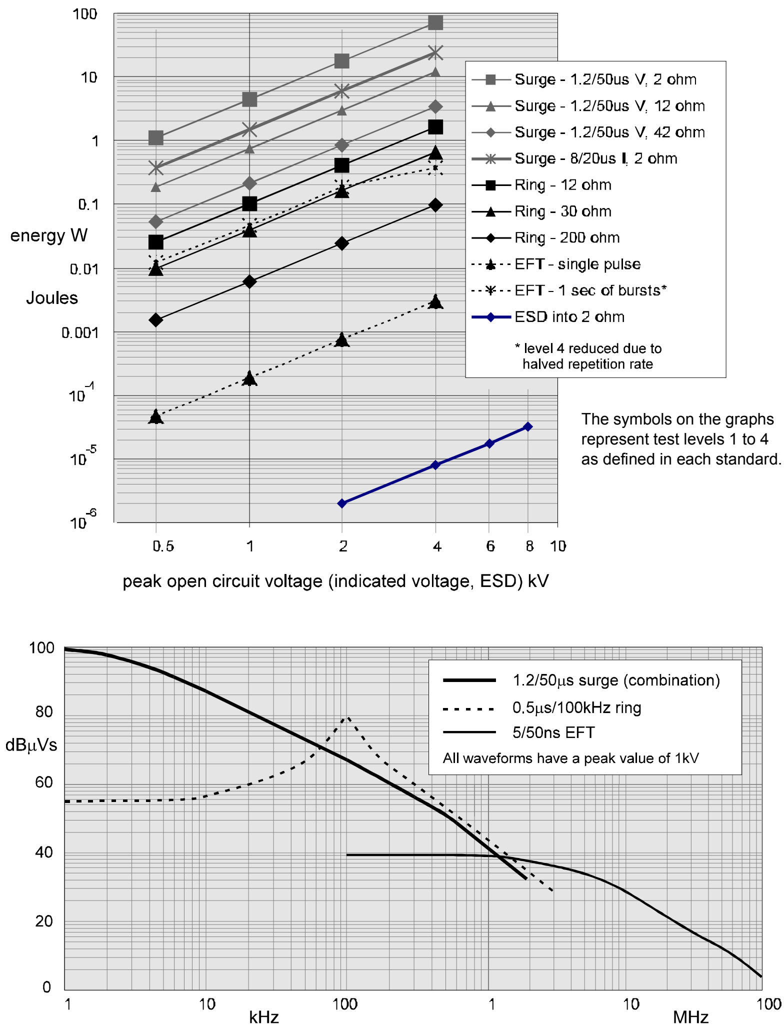


Figure 10.22 Amplitude spectral density and energy content

interference source. Although such transients are generally lower in amplitude than mains-borne ones, they are coupled directly into the I/O ports of the circuit and will therefore flow in the circuit ground traces, unless the cable is properly screened and terminated or the interface is properly filtered.

Other sources of conducted transients are telecommunication lines and the automotive 12V supply. The automotive environment can regularly experience transients that are many times the nominal supply range. The most serious automotive transients (Figure 10.23) are the load dump, which occurs when the alternator load is suddenly disconnected during heavy charging; switching of inductive loads, such as motors and solenoids; and alternator field decay, which generates a negative voltage spike when the ignition switch is turned off. ISO 7637 specifies transient testing in the automotive field.

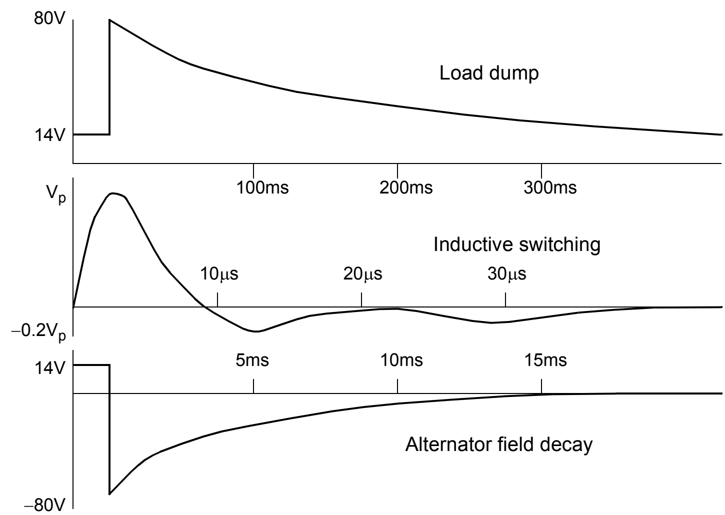


Figure 10.23 Automotive transients according to ISO 7637-2

Work on common mode transients on telephone subscriber lines [75] has shown that the amplitude versus rate of occurrence distribution also follows a roughly inverse cubic law as in Figure 10.21. Actual amplitudes were lower than those on the mains (peak amplitudes rarely exceeded 300V). A transient ringing frequency of 1MHz and rise times of 10–20ns were found to be typical. Telephone and data lines that enter a building from outside are particularly likely to be subject to lightning surges, and any ports that are connected to such lines should be designed to withstand such surges. Transient protection is discussed in more detail in section 13.2.5.

10.3.3 Electrostatic discharge

When two non-conductive materials are rubbed together or separated, electrons from one material are transferred to the other. This results in the accumulation of triboelectric charge on the surface of the material. The amount of the charge caused by movement of the materials is a function of the separation of the materials in the triboelectric series (Figure 10.24(a)) – positive materials give up electrons more readily, negative materials acquire them more readily. Additional factors are the closeness and area of contact, and rate of separation.

The voltage to which an object can be charged depends on its capacitance, following the law Q = CV. The human body can be charged by triboelectric induction to several kV. Because a perfect insulator does not allow movement of electrons, surface charges on an insulator remain in the area within which they were generated, but the human body is conductive and so a triboelectrically induced charge distributes itself over the body. The rate at which charge will bleed off a body to its surroundings, and so become neutralized, depends on the surface resistivity of the body and its surroundings. This in turn is a function of relative humidity: the more moisture there is in the air, the lower the surface resistivity of insulators and hence the quicker that charges bleed away. In practice, since movement is constantly generating charge, there is a balance between generation and dissipation which results in a typical level of charge voltage that can be found in a particular environment (Figure 10.24(b)).

When the body (in the worst case, holding a metal object such as a key) approaches a conductive object, the charge is transferred to that object normally via a spark, when the potential gradient across the narrowing air gap is high enough to cause breakdown. The energy involved in the charge transfer may be low enough to be imperceptible to the subject; at the other extreme it can be extremely painful. It is not essential that the target object is grounded. Charge transfer can occur between any two capacitive objects as long as there is a static potential difference between them, and a disruptive discharge current pulse will flow.

10.3.3.1 The ESD waveform

When an electrostatically charged object is brought close to a grounded target the resultant discharge current consists of a very fast (sub-nanosecond) edge followed by a comparatively slow bulk discharge curve. The characteristic of the hand/metal ESD current waveform is a function of the approach speed, the voltage, the geometry of the electrode and the relative humidity. The equivalent circuit for such a situation is shown in Figure 10.24(c). The capacitance C_D (typically 150pF for the human body) is charged via a high resistance up to the electrostatic voltage V. The actual value of V will vary as the charging and leakage paths change with the environmental circumstances and movements of the subject. When a discharge is initiated, the free space capacitance C_S, which is directly across the discharge point, produces an initial current peak the value of which is only limited by the local circuit stray impedance, while the main discharge current is limited by the body's bulk inductance and resistance Z_D.

The principal effects of an ESD in terms of equipment malfunction are produced by the discharge current pulse di/dt and its indirect effects. The rate of change of electric field dE/dt when the local static charge voltage collapses can also couple capacitively into high impedance circuits, and in some circumstances the high static electric field itself, before a discharge happens, may cause undesirable effects.

10.3.3.2 Coupling paths

The resultant sub-nanosecond transient equalizing current of several tens of amps follows a complex route to ground through the equipment, and is very likely to upset digital circuit operation if it passes through the circuit tracks. The paths are defined more by stray capacitance, case bonding and track or wiring inductance than by the designer's intended circuit. The high magnetic field associated with the current can induce transient voltages in nearby conductors that are not actually in the path of the current. Even if not discharged directly to the equipment, a nearby discharge such as to a metal desk or chair will generate an intense radiated field which will couple into unprotected equipment.

Critical areas which can act as sink points for the ESD are exposed metalwork, apertures, front panel components and connectors. Components and apertures can allow a discharge to take place via creepage across a surface to the circuits inside an enclosure, even if the enclosure itself is insulating. The breakdown voltage gradient in dry air is approximately 30kV per cm but this is reduced considerably across a surface, especially if the surface is contaminated with dirt or other substances.

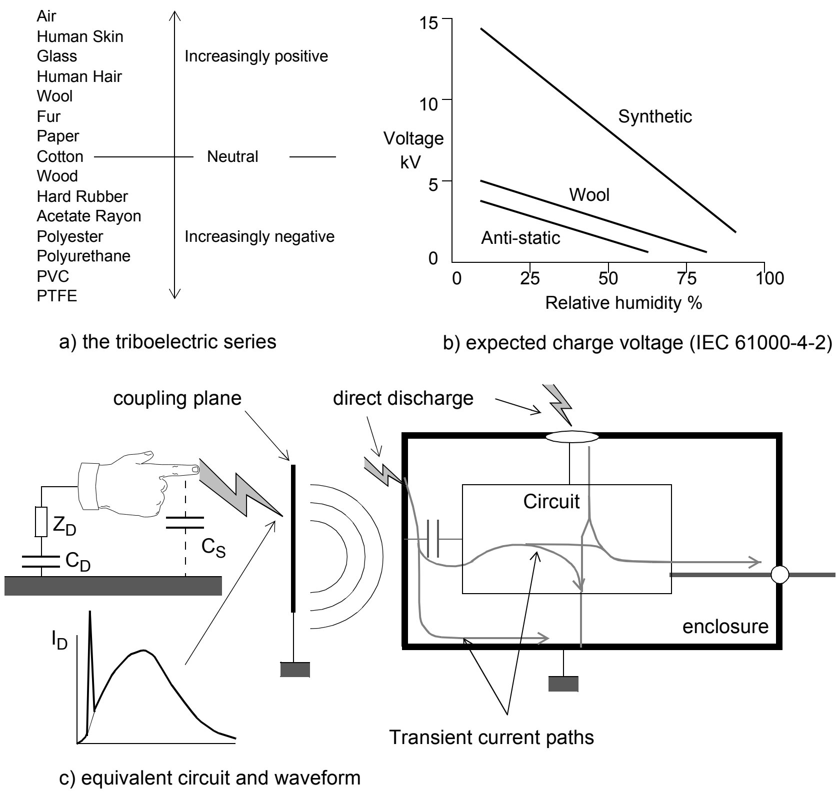


Figure 10.24 The electrostatic discharge

10.3.3.3 Secondary discharge

A common problem arises when a product enclosure is connected externally to ground at a different point and via a different route than the internal circuit. Because of the inductance of the various connections, a transient voltage will appear inside the enclosure, between the enclosure and the circuit (Figure 10.25). This voltage can then cause a secondary discharge to occur at unpredictable points inside the enclosure, which can be much more damaging and disruptive than the source discharge, since there is a lower impedance to limit the current, and also because a higher induced voltage occurs on a PCB track when an ESD occurs within a resonant structure. To prevent this, it is essential to bond the enclosure and the circuit board together at suitable points, typically at least at an interface ground (see section 11.2.3).

10.3.4 LF magnetic fields

Magnetic fields at low frequencies can induce interference voltages in closed wiring loops, their magnitude depending on the area that is intersected by the magnetic field. Non-toroidal mains transformers and switchmode supply transformers are prolific

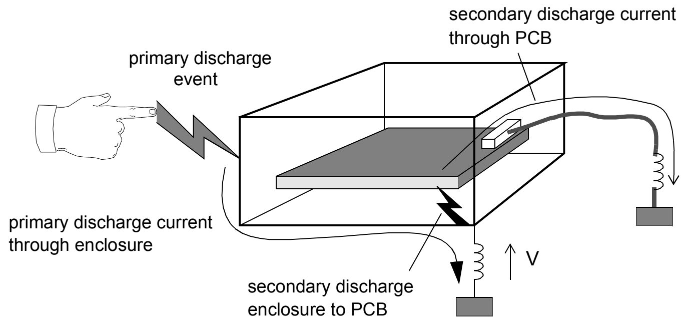


Figure 10.25 The cause of secondary discharge

sources of such fields and they will readily interfere with sensitive circuitry or components within the same equipment. Any other equipment needs to be immune to the proximity of such sources. Particular environments may result in high low-frequency or DC magnetic field strengths, such as electrolysis plant where very high currents are used, or certain medical apparatus.

50Hz currents in supply conductors are typical sources of magnetic fields. If the currents in a cable are balanced, i.e. the cable carries live and neutral together or all three phases together, then at a distance the magnetic fields from each conductor cancel and the net field is near zero. Close in to the cable though, the fields do not cancel perfectly since their source conductors are located at slightly different positions. The fields also will not cancel if the currents are not balanced, i.e. if there is some return path outside the cable – this is the same situation as was discussed earlier in section 10.1.5.2 for common mode current flow.

The flux density at a distance r metres at right angles to a single long conductor carrying a current *I* amps is given by:

$$B = (0.2 \cdot I)/r \quad \text{microtesla} \tag{10.14}$$

To find the field at a distance from a group of conductors, as in a cable, each carrying current of a particular phase and amplitude, it is only necessary to calculate the appropriate geometry and perform a vector sum of the field contributions from each conductor. A threshold value of 1μT is often recommended as a level below which an installation is acceptable [89].

The voltage developed by an external magnetic field in a single turn loop is:

$$V = A \cdot dB/dt \tag{10.15}$$

where A is the loop area in m^2 and
B is the component of flux density normal to the plane of the loop, in Tesla

It is rare for such fields to affect digital or large signal analogue circuits, but they can be troublesome with low-level circuits where the interference is within the operating bandwidth, such as audio or precision instrumentation. Specialized devices which are affected by magnetic fields, such as photomultiplier or cathode ray tubes, may also be susceptible.

10.3.4.1 Magnetic field screening

Conventional screening is ineffective against low frequency magnetic fields, because at such frequencies it relies on reflection rather than absorption of the field. Due to the low source impedance of magnetic fields reflection loss is low. Since it is only the component of flux normal to the loop which induces a voltage, changing the relative orientation of source and loop may be effective. LF magnetic shielding is only possible with materials which exhibit a high absorption loss such as steel, mu-metal or permalloy. As the frequency rises these materials lose their permeability and hence shielding efficiency, while non-magnetic materials such as copper or aluminium become more effective. Around 100kHz shielding efficiencies are about equal. Permeable metals are also saturated by high field strengths, and are prone to lose their permeability through handling. See section 14.1.2 for more discussion.

10.3.5 Supply voltage phenomena

Low frequency disturbances on the mains supply are covered in some detail in IEC61000 Part 2 sections 1 and 2. Section 1 [153] describes the environment, i.e. the nature of the disturbances that can be expected on public mains supplies, while section 2 [154] gives compatibility levels, i.e. the levels of disturbances that can be expected. The phenomena considered are:

- harmonics and inter-harmonics;
- voltage fluctuations, dips and short supply interruptions;
- voltage unbalance in three-phase supplies;
- mains signalling;
- power frequency variation.

Harmonics are considered further in section 10.4.

Brown-outs (voltage dips) and interruptions are a feature of all mains distribution networks, and are usually due to fault clearing or load switching elsewhere in the system (Figure 10.26). Such events will not be perceived by ordinary electronic equipment if its input reservoir hold-up time is sufficient, but if this is not the case then restarts and output transients can be experienced. Thyristor inverters may experience commutation failure and synchronous devices may lose synchronism. Typically, interruptions (as opposed to power cuts) can last for 10–500ms.

Load and line voltage fluctuations are maintained between +10% and −15% of the nominal line voltage in most industrialized countries. As a result of HD472/BS7697 [176], the EU countries have moved towards 230V ±10% at the point of connection to the consumer. Between 1995 and 1st January 2003, countries with a previously declared nominal voltage of 240V had a range of 230V +10% −6%, and those with a previous voltage of 220V had a range of 230V +6% −10%. Slow changes in the voltage within these limits occur on a diurnal pattern as the load on the power system varies. The declared voltage does not include voltage drops within the customer's premises, and so you should design stabilized power supplies to meet at least the −15% limit.

Dips exceeding 10% of nominal voltage occur up to four times per month for urban consumers, and more frequently in rural areas where the supply is via overhead lines [81][154]. Note that much wider voltage (and frequency) fluctuations and more frequent interruptions are common in those countries which do not have a well-developed supply network. They are also common on supplies which are derived from small generators.

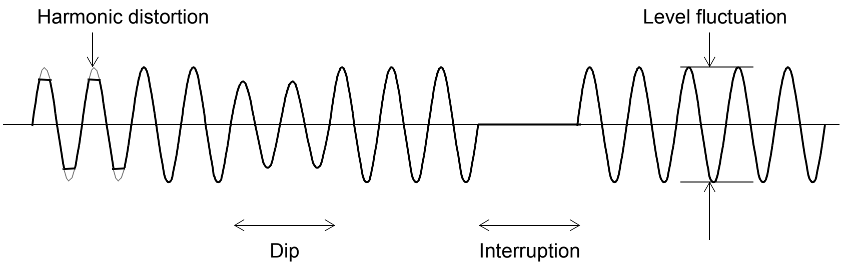


Figure 10.26 Mains supply fluctuations

Heavy industrial loads such as resistance and arc welding machines, large motors and arc furnaces can cause short-term step or random fluctuations and can affect many consumers fed from the same source. The main effect of these disturbances is flicker of lamp loads, which can cause physiological discomfort. Electronic power supply circuits can normally be designed to ignore them, although other circuits which use the 50Hz signal (e.g. for timing or phase reference) should have their operating bandwidth severely restricted by a 50Hz bandpass filter to ensure immunity from low-amplitude step changes.

10.4 Mains harmonics

One EMC phenomenon, which comes under the umbrella of the EMC Directive and is usually classified as an "emission", is the harmonic content of the mains input current. This is mildly confusing since the equipment is not actually "emitting" anything: it simply draws its power at harmonics of the line frequency as well as at the fundamental.

10.4.1 The supplier's problem

The problem of mains harmonics is principally one for the supply authorities, who are mandated to provide a high quality electricity supply. If the aggregate load at a particular mains distribution point has a high harmonic content, the non-zero distribution source impedance will cause distortion of the voltage waveform at this point, and possibly, due to supply network resonances, at other remote points. This in turn may cause problems for other users, and the currents themselves may also create problems (such as overheating of transformers and compensating components) for the supplier. The supplier does of course have the option of uprating the distribution components or installing special protection measures, but this is expensive and the supplier has room to argue that the users should bear some of the costs of the pollution they create.

Throughout the last few decades, harmonic pollution has been increasing and this increase has been principally due to low power electronic loads installed in large numbers. Between them, domestic TV sets and office information technology equipment account for about 80% of the problem. Other types of load which also take significant harmonic currents are not widely enough distributed to cause a serious problem yet, or are dealt with individually at the point of installation as in the case, for instance, of variable speed drives in industrial plant. The supply authorities are nevertheless sufficiently worried to want to extend harmonic emission limits to all classes of electronic products.

IEC 61000-2-2 defines the compatibility level for this phenomenon in terms of total harmonic distortion factor as 8% THD.

10.4.2 Non-linear loads

A plain resistive load across the mains draws current only at the fundamental frequency (50Hz in Europe). Most electronic circuits are anything but resistive. The universal rectifier-capacitor input draws a high current at the peak of the voltage waveform and zero current at other times; the well-known triac phase control method for power control (lights, motors, heaters, etc.) begins to draw current only partway through each half-cycle. These current waveforms can be represented as a Fourier series, and it is the harmonic amplitudes of the series that are subject to regulation.

The standard which covers mains harmonics is IEC 61000-3-2, first published in 1995 and revised with substantial changes in 2000. Its requirements are detailed in sections 4.5.1 and 8.1 where you will see that it applies either fixed limits to the harmonic content up to 2kHz (40th harmonic) or variable limits depending on the power drawn by the equipment; choice of limits depends on the class of product.

The limits are effectively an additional design constraint on the values of the input components, most notably the input series impedance (which is not usually considered as a desirable input component at all). Figure 10.27(a), which is a Fourier analysis of the current waveform calculated in the time domain, shows the harmonic content of input current for a rectifier-reservoir combination with a fairly high series resistance. This value of series resistance would not normally be found except with very inefficient transformer-input supplies. The fifth harmonic content just manages to meet Class D.

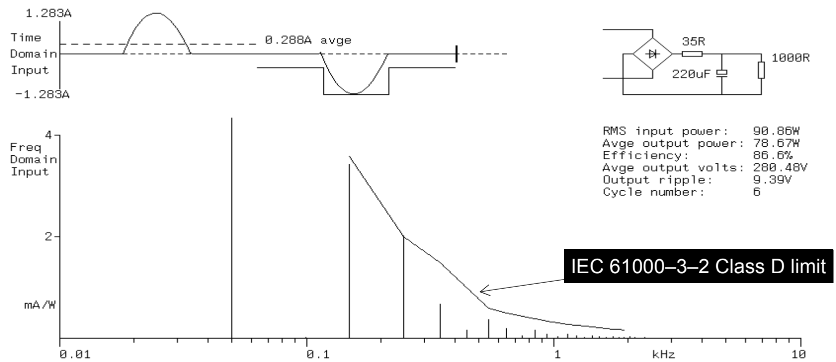
10.4.2.1 The effect of series resistance

Figure 10.27(b) illustrates the difference in input harmonics resulting from a tenfold reduction in input resistance. This level of input resistance would be typical for a direct-off-line switching supply and many highly efficient supplies could boast a lower R_S. The peak input current has increased markedly while its duty cycle has shrunk, leading to a much higher crest factor (ratio of peak to root mean square current) and thus higher levels of harmonics.

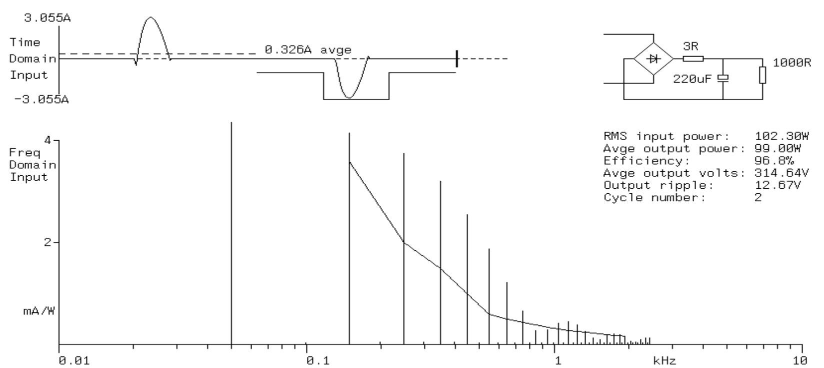
Increasing input series resistance to meet the harmonic limits is expensive in terms of power dissipation except at very low powers. In practice, deliberately dissipating between 10 and 20% of the input power rapidly becomes unreasonable above levels of 50–100W. In fact, the requirements of IEC 61000-3-2 do not apply to equipment having an active input power below 75W. Alternatives are to include a series input choke, which since it must operate down to 150Hz at the full input current is expensive in size and weight; or to include electronic power factor correction (PFC), which converts the current waveform to a near-sinusoid but is expensive in cost and complexity.

10.4.2.2 Power factor correction

PFC is essentially a switchmode converter on the front-end of the supply, and therefore is likely to contribute extra RF switching noise at the same time as it reduces input current harmonics. It is possible to combine PFC with the other features of a direct-off-line switching supply, so that if you are intending to use a SMPS anyway there will be little extra penalty. It also fits well with other contemporary design requirements such



(a)



(b)

Figure 10.27 Mains input current harmonics for rectifier-reservoir circuit

as the need for a "universal" (90–260V) input voltage range. Such power supplies can be bought off-the-shelf, or with extra design and development effort you can design a PFC-SMPS yourself. The availability of special-purpose control ICs makes the task easier.

Figure 10.28 shows the basis of operation of a power factor correction circuit. Instead of an input rectifier-reservoir combination, the rectified input feeds a switchmode boost converter circuit directly whose operational input voltage range extends from near-zero to the peak supply voltage. The pulse width of the switching circuit is regulated to give an average input current which approximates to the required sinusoidal waveshape. The effective distortion is very low, and therefore so is the harmonic content.

10.4.2.3 Phase angle control

Power control circuits which vary the switch-on point with the phase of the mains waveform are another major source of harmonic distortion on the input current.

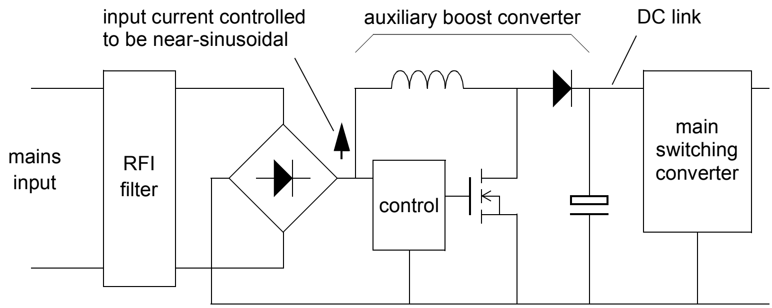


Figure 10.28 Schematic of a switchmode PFC circuit

Lighting dimmers and motor controllers are the leading examples of these. Figure 10.29 shows the harmonic content of such a waveform switched at 90° (the peak of the cycle, corresponding to half power). The maximum harmonic content occurs at this point, decreasing as the phase is varied either side of 90°. Lighting dimmers without input filtering or PFC of greater than about 5A rating are outlawed, since the limits are set at an absolute value.

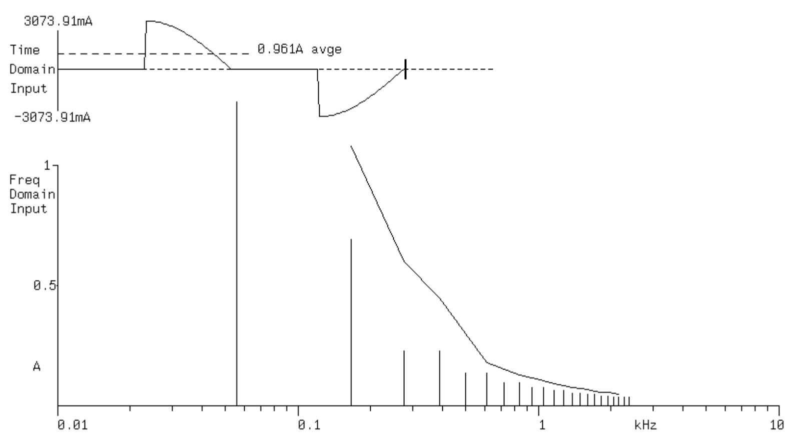


Figure 10.29 Mains input current harmonics for 500W phase control circuit at half power

Chapter 11

Layout and grounding

Designing for good EMC starts from the principle of controlling the flow of interference into and out of the equipment. You must assume that interference will occur to and will be generated by any product which includes active electronic devices. To improve the electromagnetic compatibility of the product you place barriers and route currents such that incoming interference is diverted or absorbed before it enters the circuit, and outgoing interference is diverted or absorbed before it leaves the circuit. A good analogy is to think of the circuit as a town, and the EMC control measures as bypasses or ring roads. The interference (traffic) is routed around the town rather than being forced to flow through it; the disruption to the town's operation is that much less.

In either case, you can conceive the control measures as applying at three levels, primary, secondary and tertiary, as shown in Figure 11.1.

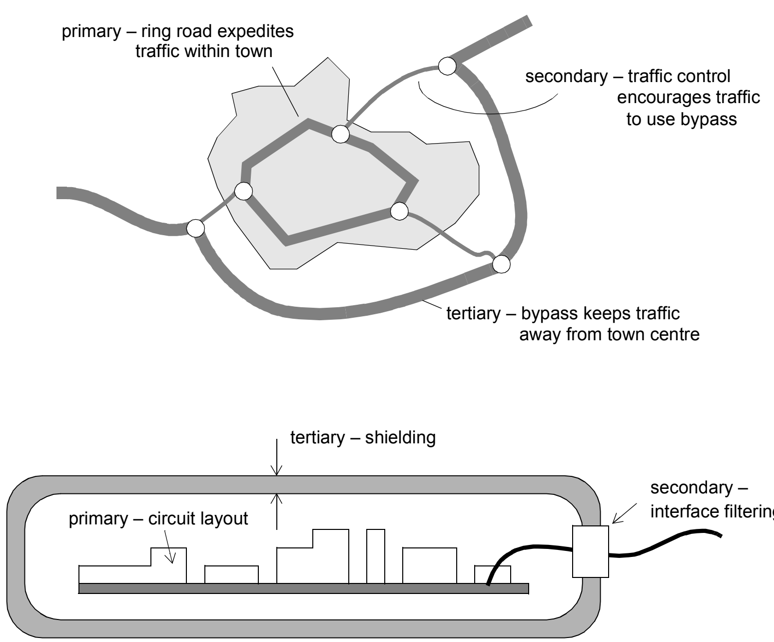


Figure 11.1 EMC control measures

Control at the primary level involves circuit design measures such as decoupling, balanced configurations, bandwidth and speed limitation, and especially board layout and grounding. For some low-performance circuits, and especially those which have no connecting cables, such measures may be sufficient in themselves. At the secondary level you must always consider the interface between the internal circuitry and external cables. This is invariably a major route for interference in both directions, and for some products (particularly where the circuit design has been frozen) all the control may have to be applied by filtering at these interfaces. Choice and mounting of connectors forms an important part of this exercise.

Full shielding (the tertiary level) is an expensive choice to make and you should only choose this option when all other measures have been applied. But since it is difficult or impossible to predict the effectiveness of primary measures in advance, it is wise to allow for the possibility of being forced to shield the enclosure. This means adapting the mechanical design so that a metal case could be used, or if a moulded enclosure is essential, you should ensure that apertures and joints in the mouldings can be adequately bonded at RF, that ground connections can be made at the appropriate places and that the moulding design allows for easy application of a conductive coating.

Chapter 12 looks at circuit design, and Chapters 13 and 14 consider the “classical” aspects of interfaces, filters and shielding. This chapter covers layout and grounding.

11.1 Equipment layout and grounding

The most cost-effective approach is to consider the equipment’s layout and ground regime at the beginning. No unit cost is added by a designed-in ground system. Ninety per cent of post-design EMC problems are due to inadequate layout or grounding: a well-designed layout and ground system can offer both improved immunity and protection against emissions, while a poorly designed one may well be a conduit for emissions and incoming interference. The most important principles are:

- partition the system to allow control of interference currents;
- consider ground as a path for current flow, both of interference into the equipment and conducted out from it;
- consider it also as a means of preventing interference currents from affecting signal circuits; this means careful placement of grounding points, and minimizing both the ground impedance itself and its transfer impedance to the circuit;
- design the conducting parts of the mechanical structure in the knowledge that they are unavoidably carrying interference currents, which you want to keep separate from your circuit.

11.1.1 System partitioning

The first design step is to partition the system. A poorly partitioned, or non-partitioned system (Figure 11.2) may have its component sub-systems separated into different areas of the board or enclosure, but the interfaces between them will be ill-defined and the external ports will be dispersed around the periphery. This makes it difficult to control the common mode currents that will exist at the various interfaces. Dispersal of the ports means that the distance between ports on opposite sides of the system is large, leading to high induced ground voltages as a result of incoming interference, and efficient coupling to the cables of internally generated emissions.

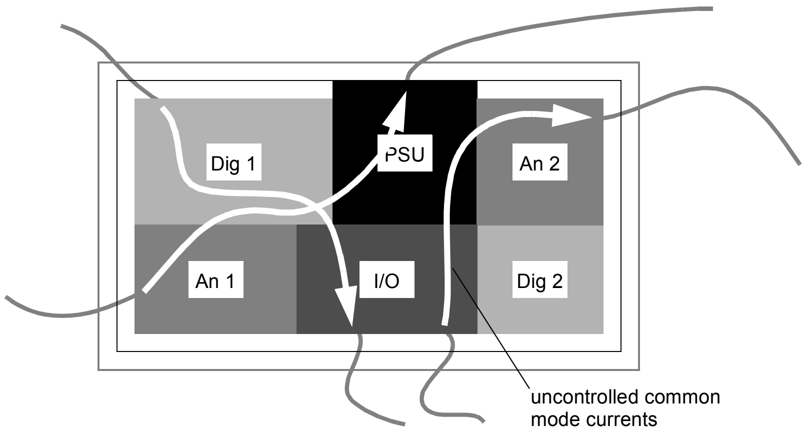


Figure 11.2 The haphazard system

Usually the only way to control emissions and immunity of such a system is by placing an overall shield around it and filtering each interface. In many cases it will be difficult or impossible to maintain integrity of the shield and still permit proper operation – the necessary apertures and access points will preclude effective attenuation through the shield.

11.1.1.1 The partitioned system

Partitioning separates the system into critical and non-critical sections from the point of view of EMC. Critical sections are those which contain radiating sources such as microprocessor logic or video circuitry, or which are particularly susceptible to imported interference: microprocessor circuitry and low-level analogue circuits. Non-critical sections are those whose signal levels, bandwidths and circuit functions are such that they are not susceptible to interference nor capable of causing it: non-clocked logic, linear power supplies and power amplifier stages are typical examples. Figure 11.3 shows this method of separation.

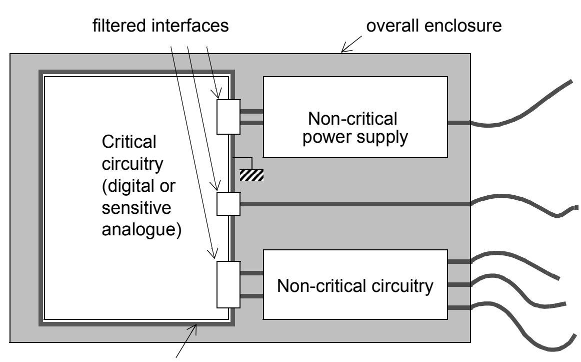


Figure 11.3 System partitioning

Control of critical sections

Critical sections can then be enclosed in a shielded enclosure into and out of which all external connections are carefully controlled. This enclosure may encase the whole product or only a portion of it, depending on the nature of the circuits: your major design goal should be to minimize the number of controlled interfaces, and to concentrate them physically close together. Each interface that needs to be filtered or requires screened cabling adds unit cost to the product. A product with no electrical interface ports – such as a pocket calculator or infra-red remote controller – represents an ideal case from the EMC point of view.

Note that the shield acts both as a barrier to radiated interference and as a reference point for ground return currents. In many cases, particularly where a full ground plane PCB construction is used, the latter is the more important function and it may be possible to do without an enclosing shield.

11.1.2 Grounding

Once the system has been properly partitioned, you can then ensure that it is properly grounded. Before continuing, it is as well to note that there is some confusion in terminology. In conventional electrical usage, "grounding" is American for the same function as is "earthing" in English; that is, a safety protective function. Since this is an English book and the EMC function is different, the words "ground" and "grounding" will be used here to distinguish the EMC function while "earth" and "earthing" will be used for the safety function.

The accepted purpose for grounding is to give a reference for external connections to the system. The classical definition of a ground is "an equipotential point or plane which serves as a reference for a circuit or system". Unfortunately this definition is meaningless in the presence of ground current flow. Even where signal currents are negligible, induced ground currents due to environmental magnetic or electric fields will cause shifts in ground potential. A good grounding system will minimize these potential differences by comparison with the circuit operating levels, but it cannot eliminate them. It has been suggested [5] that the term "ground" as conventionally used should be dropped in favour of "reference point" to make the purpose of the node clear.

An alternative definition for a ground is "a low impedance path by which current can return to its source" [105]. This emphasizes current flow and the consequent need for low impedance, and is more appropriate when high frequencies are involved. Ground currents always circulate as part of a loop. The task is to design the loop in such a way that induced voltages remain low enough at critical places. You can only do this by designing the ground circuit to be as compact and as local as possible.

The most important EMC function of a ground system is to *minimize interference voltages at critical points compared to the desired signal*. To do this, it must present a low *transfer* impedance path at these critical locations. The concept of transfer impedance is introduced in section 11.1.2.2.

11.1.2.1 Current through the ground impedance

When designing a ground layout you must know the actual path of the ground return current. The amplifier example in Figure 11.4 illustrates this. The high-current output ΔI returns to the power supply from the load; if it is returned through the path Z1–Z2–Z3 (Figure 11.4(a)) then an unwanted voltage component is developed across Z2 which is in series with the input V_S, and depending on its magnitude and phase the circuit will oscillate. This is an instance of common impedance coupling, as per section 10.1.1.

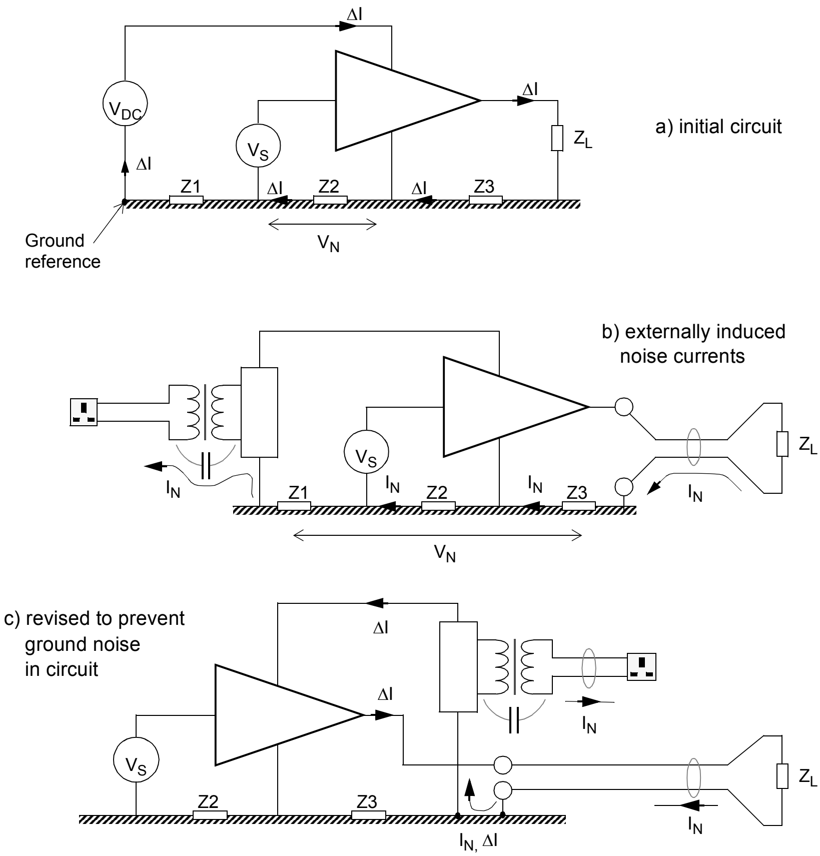


Figure 11.4 Ground current paths in an example circuit

For EMC purposes, instability is not usually the problem; rather it is the interference voltages V_N which are developed across the impedances that create emission or susceptibility problems. At high frequencies (above a few kHz) or high rates of change of current the impedance of any linear connection is primarily inductive and increases with frequency ($V = -L \cdot di/dt$), hence ground noise increases in seriousness as the frequency rises. The effect of adding external connections is illustrated in Figure 11.4(b). Interference current I_N induced in, say, the output lead, flows through the ground system, passing through Z2 again and therefore inducing a voltage in series with the input, before exiting via stray capacitance to the mains supply connection. The same route could be taken in reverse by incoming mains-borne noise.

To deal with the problem, it is simply necessary to ensure that the interfering currents are not allowed to flow through the sensitive part of the ground network, as is shown by the rearrangement of Figure 11.4(c). On a circuit diagram these circuits would appear identical; in practical realization, when laid out on a PCB, they are different.

11.1.2.2 Transfer impedance

For EMC purposes, grounding provides a set of interconnected current paths, designed to have a low transfer impedance Z_T, in order to minimize interfering voltages at sensitive interfaces which may or may not be ground-referred. Z_T determines the strength of the unwanted source in the signal circuit due to interference current flowing in the common mode circuit. Depending on the interface under consideration, we tackle the transfer impedance of an appropriate part of the interconnected grounding paths. The result is a grounding structure, whose three-dimensional shape is designed for low Z_T. The structure can be realized as a "shielding" enclosure, a chassis plate, a plane layer on a PCB, or a cable conduit, whatever is required by the application.

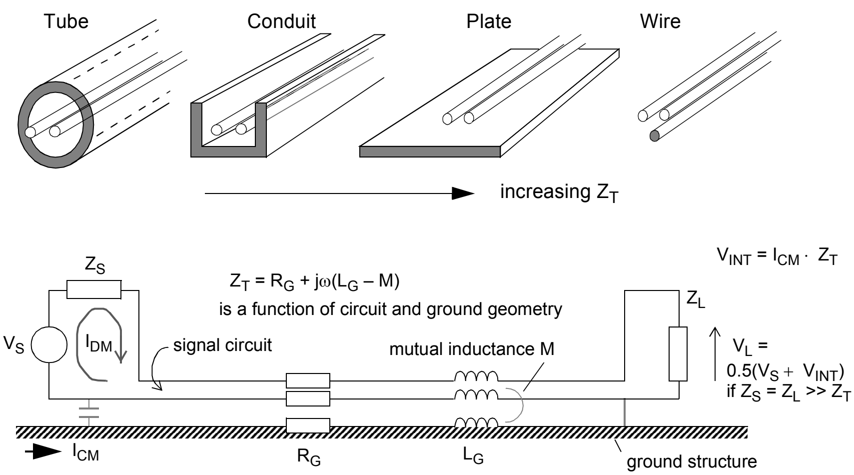


Figure 11.5 Transfer impedance of the ground structure

The important quality of such a structure (Figure 11.5) is that when a disturbance current flows in it, only a small interference voltage V_{INT} is generated differentially in the signal circuit. The dominant contribution to Z_T for external interference currents flowing in the ground is mutual inductance between common and differential mode circuits. Z_T is reduced in proportion to the mutual inductance linking the structure and the circuit. The grounding structure should have minimum resistance R_G and self-inductance L_G. Z_T is minimized by a solid enclosing tube – you may recognize this as a coaxial shielded cable – and is worst if the structure is non-existent, since all interference current then flows in the circuit. A haphazard ground structure with appreciable series impedance – caused by apertures and discontinuities, see section 14.1.3 – is little better, and the first level of usefulness is offered by a parallel wire conductor. Practical compromise structures are the conduit or the flat plate. The closer the differential mode circuit is physically to such a grounding structure, the less the Z_T, provided that the structure is unbroken in the direction of current flow.

This discussion has concentrated on ground currents, since in general the low inherent impedance of ground structures allows relatively high interference currents to flow; EMC consists in recognizing this and taking the appropriate precautions. It is important, though, that noise voltages are not allowed to build up on the grounding structure relative to the circuit, since the close coupling recommended above for low Z_T purposes also implies a high capacitance, which in turn would allow such voltages to couple more readily into the circuit. This suggests that the circuit 0V reference should itself be connected to the ground structure at least at one point, normally close to the most sensitive part of the circuit; but this will *encourage* the distribution of current between the ground structure and the circuit! This paradox can only be resolved by physically designing the ground structure for the least possible Z_T.

11.1.3 Ground systems

Ignoring for now the need for a safety earth, a grounding system as intended for a circuit reference can be configured as single point, multi-point or as a hybrid of these two.

11.1.3.1 Single point

The single point grounding system (Figure 11.6(a)) is conceptually the simplest, and it eliminates common impedance ground coupling and low frequency ground loops. Each circuit module has its own connection to a single ground, and each rack or sub-unit has one bond to the chassis. Any currents flowing in the rest of the ground network do not couple into the circuit. This system works well up to frequencies in the MHz region, but the distances involved in each ground connection mean that common mode potentials between circuits begin to develop as the frequency is increased. At distances of odd multiples of a quarter wavelength, the circuits are effectively isolated from each other. At the same time, stray capacitance starts to contribute to “sneak” current paths that bypass the single point wiring.

A modification of the single point system (Figure 11.6(b)) ties together those circuit modules with similar characteristics and takes each common point to the single ground. This allows a degree of common impedance coupling between those circuits where it won’t be a problem, and at the same time allows grounding of high frequency circuits to remain local. The noisiest circuits are closest to the common point in order to minimize the effect of the common impedance. When a single module has more than one ground, these can be tied together with back-to-back diodes to prevent damage when the circuit is disconnected.

11.1.3.2 Multi-point

Hybrid and multi-point grounding (Figure 11.6(c)) can overcome the RF problems associated with pure single point systems. Multi-point grounding is necessary for digital and large signal high frequency systems. Modules and circuits are bonded together with many short ($< 0.1\lambda$) links to minimize ground-impedance-induced common mode voltages. Alternatively, many short links to a chassis, ground plane or other low impedance conductive body are made. This may not be suitable for sensitive audio circuits, if the loops that are introduced then create magnetic field pick-up. It is difficult to keep 50/60Hz interference out of such circuits, but this is only a problem if the circuits themselves can be affected by it, and if its amplitude is large because the loop areas are large. Circuits which operate at higher frequencies or levels are generally not susceptible to this interference. The multi-point sub-system can be brought down to a single point ground in the overall system.

11.1.3.3 Hybrid

Hybrid grounding uses reactive components (capacitors or inductors) to make the grounding system act differently at low frequencies and at RF. This may be necessary in sensitive wideband circuits. In the example of Figure 11.6(c), the sheath of a relatively long cable is grounded directly to chassis via capacitors to prevent RF standing waves from forming. The capacitors block DC and low frequencies and therefore prevent the formation of an undesired extra ground loop between the two modules.

When using such reactive components as part of the ground system, you need to take care that spurious resonances (which could enhance interference currents) are not introduced into it. For example if 0.1μF capacitors are used to decouple a cable whose self-inductance is 0.1μH, the resonant frequency of the combination is 1.6MHz. Around this frequency the cable screen will appear to be anything but grounded!

When you are using separate DC grounds and an RF ground plane (such as is offered by the chassis or frame structure), reference each sub-system's DC ground to the frame by a 10–100nF capacitor. The two should be tied together by a low impedance link at a single point where the highest di/dt signals occur, such as the processor motherboard or the card cage backplane.

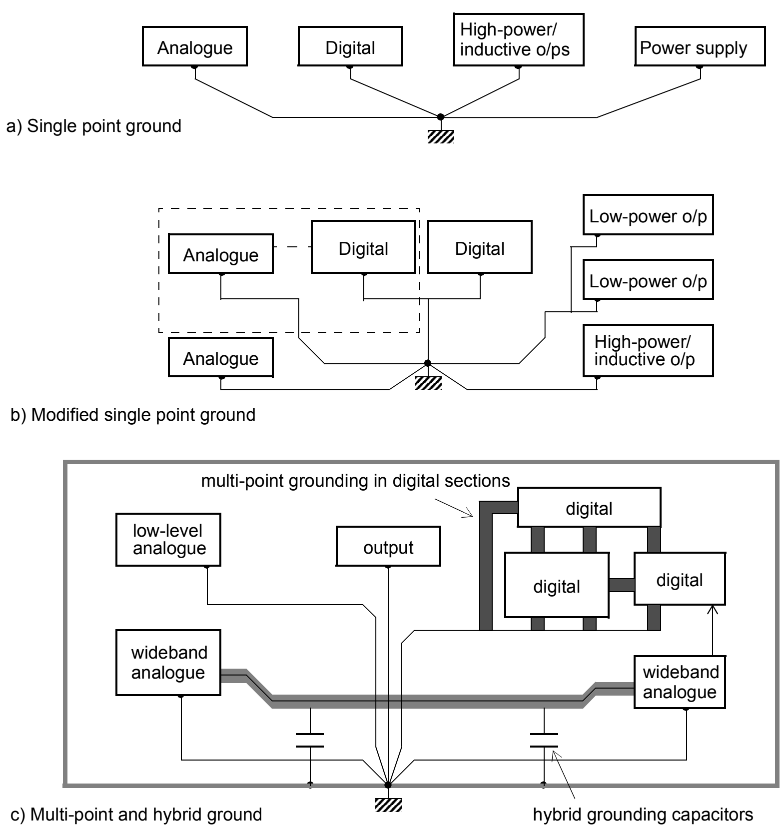


Figure 11.6 Grounding systems

11.1.3.4 The impedance of ground wires

When a grounding wire runs for some distance alongside a ground plane or chassis before being connected to it, it appears as a transmission line. This can be modelled as an LCR network with the L and C components determining the characteristic impedance Z_0 of the line (Figure 11.7).

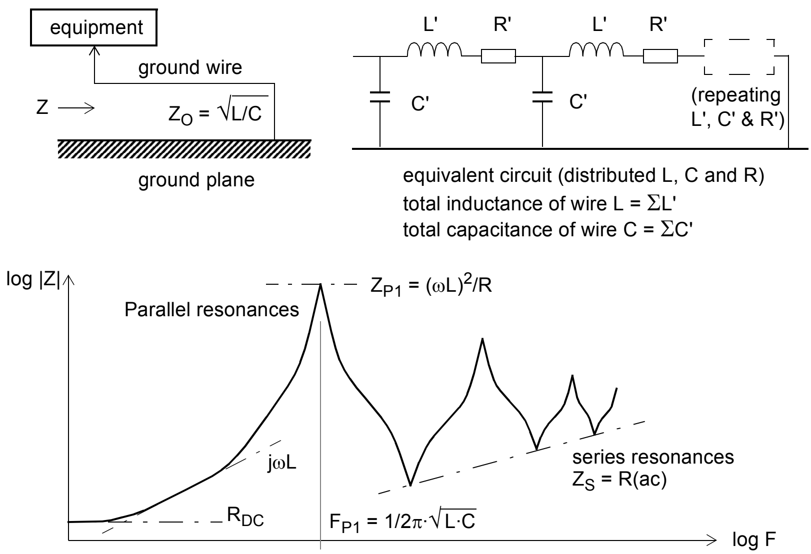


Figure 11.7 The impedance of long ground wires

As the operating frequency rises, the inductive reactance exceeds the resistance of the wire and the impedance increases up to the first parallel resonant point. At this point the impedance seen at the end of the wire is high, typically hundreds of ohms (determined by the resistive loss in the circuit). After first resonance, the impedance for a lossless circuit follows the law:

$$Z = Z_O \cdot \tan [\omega \cdot x \cdot \sqrt{(L/C)}] \tag{11.1}$$

where x is the distance along the wire to the short

and successive series (low impedance) and parallel (high impedance) resonant frequencies are found. As the losses rise due to skin effect, so the resonant peaks and nulls become less pronounced. To stay well below the first resonance and hence remain an effective conductor, the ground wire should be less than 1/20th of the shortest operating wavelength.

11.1.3.5 The safety earth

From the foregoing discussion, you can see that the safety earth (the green and yellow wire) is not an RF ground at all. Many designers may argue that everything is connected to earth via the green and yellow wire, without appreciating that this wire has a high and variable impedance at RF. In general the safety earth connection is not necessary for EMC purposes: after all, battery powered apparatus can function quite successfully without it. A good low impedance connection to an RF reference provided locally by a chassis, frame or plate *is* necessary and in many cases must be provided *in parallel with* the safety earth. On the other hand, if we are considering mains emissions or incoming mains-borne disturbances, which are propagating in common mode with respect to the safety earth, then a safety earth connection is required for EMC purposes, so that filter capacitors can be tied to it. This is not because it is an earth, but because it is the return path for the disturbance currents.

It may even be necessary for you to take the safety earth *out* of the RF circuit deliberately, by inserting a choke of the appropriate current rating in series with it (see section 13.2.3.6). This is because it can offer an alternative path for interference currents to invade or circulate within the system, and interrupting this path is a simple way of improving EMC.

11.1.3.6 Ground map

A fundamental tool for use throughout the equipment design is a ground map. This is a diagram which shows all the ground reference points and grounding paths (via structures, cable screens, etc. as well as tracks and wiring) for the whole equipment. It concentrates on grounding only; all other circuit functions are omitted or shown in block form. Its creation, maintenance and enforcement throughout the project design should be the responsibility of the EMC design authority.

11.1.3.7 Summary

In the context of product-level grounding, and so for dimensions up to say 1m, at frequencies below 1MHz single point grounding is possible and in some circumstances preferable. Above 10MHz a single point ground system is not feasible because wire and track inductance raises the ground impedance unacceptably, and stray capacitance allows unintended ground return paths to exist. For high frequencies multi-point grounding to a low inductance ground plane or shield is essential. This creates ground loops which may be susceptible to magnetic field induction, so should be avoided or specially treated in a hybrid manner when used with very sensitive circuits.

For EMC purposes, *even a circuit which is only intended to operate at low frequencies must exhibit good immunity from RF interference*. This means that those aspects of its ground layout which are exposed to the interference – essentially all external interfaces – must be designed for multi-point grounding. At the bare minimum, some low inductance ground plate or plane must be provided at the interfaces.

Grounding principles

- All conductors have a finite impedance which increases with frequency
- Two physically separate ground points are not at the same potential unless no current flows between them
- At high frequencies there is no such thing as a single point ground

11.2 PCB layout

The way in which you design a printed circuit board makes a big difference to the overall EMC performance of the product which incorporates it. The principles outlined above must be carried through onto the PCB, particularly with regard to partitioning, interface layout and ground layout. This means that the circuit designer must exert tight control over the layout draughtsman, especially when CAD artwork is being produced. Conventional CAD layout software works on a node-by-node basis, which if allowed to will treat the entire ground system as one node. This is fine if you are designing with a single ground plane for the whole product, but will cause problems with multiple ground planes if left uncorrected.

Design of a PCB with multilayer technology is discussed shortly. But if you are using older technology, i.e. double- or single-sided, the safest way to lay out a PCB is to start with the ground traces, manually if necessary, then to incorporate critical signals such as HF clocks or sensitive nodes which must be routed near to their ground returns, and then to track the rest of the circuitry at will. As much information should be provided with the circuit diagram as possible, to give the layout draughtsman the necessary guidance at the beginning. These notes should include:

- physical partitioning of the functional sub-modules on the board;
- positioning requirements of sensitive components and I/O ports;
- marked up on the circuit diagram, the various ground nodes that are being used, together with which connections to them must be regarded as critical;
- where the various ground nodes may be commoned together, and where they must *not* be;
- which signal tracks must be routed close to the ground tracks, plus any other constraints on the signal track routing.

One consequence of the universal single ground plane approach, evaluated in section 11.2.3.3, is that it eases the dilemmas that flow from the third and fourth points above.

11.2.1 Ground layout without a ground plane

11.2.1.1 Track impedance

Careful placement of ground connections goes a long way towards reducing the noise voltages that are developed across ground impedances. But on any non-trivial printed circuit board it is impractical to eliminate circulating ground currents entirely. The other aspect of ground design is to minimize the value of the ground impedance itself.

Track impedance is dominated by inductance at frequencies higher than a few kHz (Figure 11.8). You can reduce the inductance of a connection in two ways:

- minimizing the length of the conductor, and if possible increasing its width;
- running its return path parallel and close to it.

The inductance of a PCB track is primarily a function of its length, and only secondarily a function of its width. For a single conductor of diameter d and length l inches, equation (11.2) gives the self-inductance (further equations are given in Appendix D):

$$L = 0.0051 \cdot l \cdot (\ln(4l/d) - 0.75) \text{ microhenries} \quad (11.2)$$

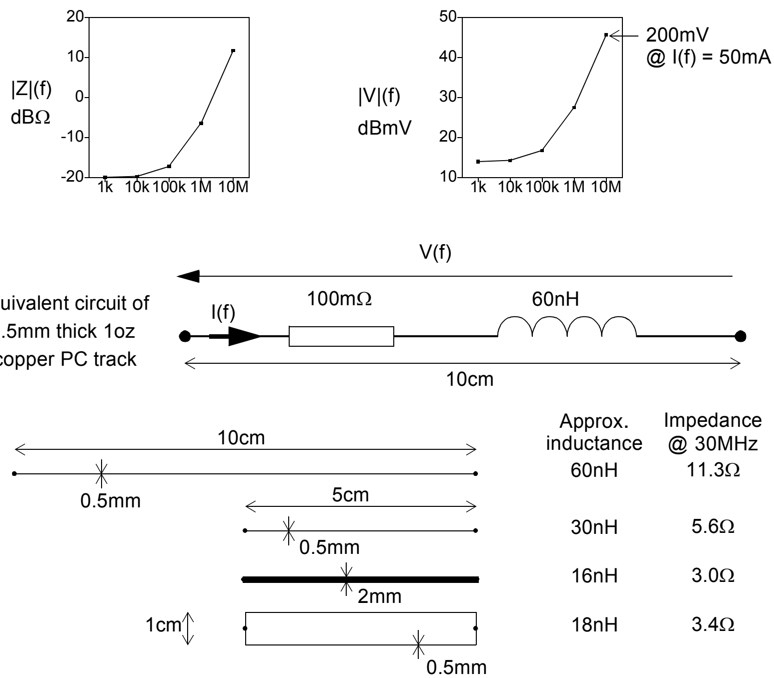


Figure 11.8 Impedance of printed circuit tracks

Because of the logarithmic relationship of inductance and diameter, doubling this dimension does not produce a 50% decrease in inductance. Paralleling tracks will reduce the inductance *pro rata* provided that they are separated by enough distance to neutralize the effect of mutual inductance (see Figure 10.4 on page 226). For narrow conductors spaced more than 1cm apart, mutual inductance effects are negligible.

11.2.1.2 Gridded ground

The logical extension to paralleling ground tracks is to form the ground layout in a grid structure (Figure 11.9). This maximizes the number of different paths that ground return current can take, and therefore minimizes the ground inductance for any given signal route. Such a structure is well suited to low cost digital double-sided PCB layout with multiple packages, when individual signal/return paths are too complex to define easily [71]. The grid structure doesn't have to be regular, and indeed in most such layouts the sizes of the apertures in the grid are anything but uniform.

A wide ground track is preferred to narrow for minimum inductance, but even a narrow track linking two otherwise widely-separated points is better than none. The grid layout is best achieved by putting the grid structure down first, before the signal or power tracks are laid out. You can follow the X-Y routing system for double-sided boards, where the X-direction tracks are all laid on one side and the Y-direction tracks all on the other. Offensive (high di/dt) signal tracks can then be laid close to the ground tracks to keep the overall loop area small; this may call for extra ground tracking, which should be regarded as an acceptable overhead.

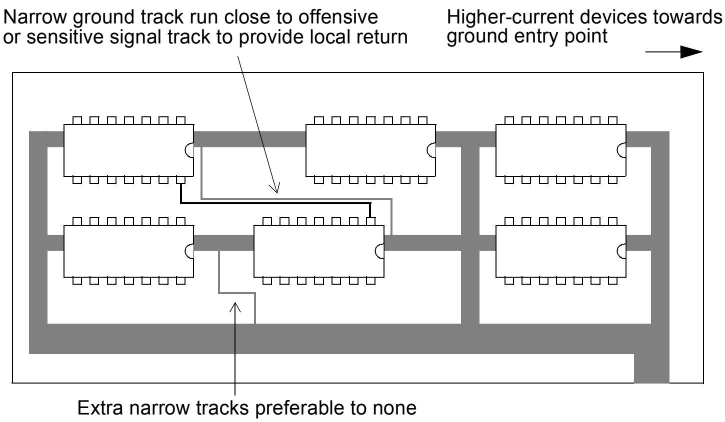


Figure 11.9 The gridded ground structure

Undesirable ground style

The one type of ground configuration that you should not use for any class of circuit is the "comb" style in which several ground spurs are run from one side of the board (Figure 11.10). Such a layout forces return currents to flow in a wide loop even if the signal track is short and direct, and contributes both to increased radiation coupling and to increased ground noise generation. The significant common ground impedance introduced between packages on the board may also cause circuit malfunction. The comb can easily be converted to a proper grid by adding bridging tracks at intervals across the spurs.

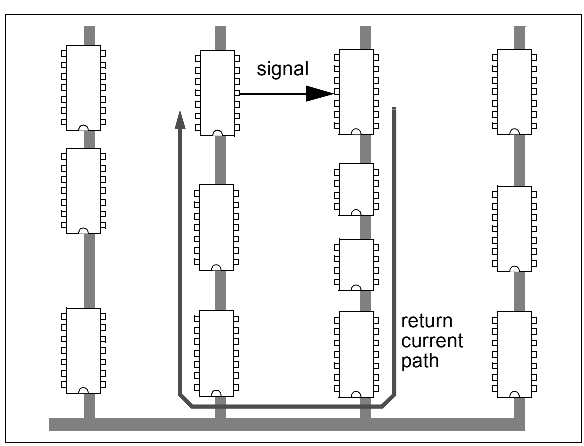


Figure 11.10 Undesirable: the comb ground structure

11.2.2 Using a ground plane

The limiting case of a gridded ground is when an infinite number of parallel paths are provided and the ground conductor is continuous, and it is then known as a ground plane. This is easy to realize with a multilayer board and offers the lowest possible ground path inductance. It is essential for RF circuits and digital circuits with high clock speeds, and offers the extra advantages of greater packing density and a defined characteristic impedance for all signal tracks [100] (see section 11.2.2.6). A common four-layer configuration includes the power supply rail as a separate plane, which gives a low power-ground impedance at high frequencies.

The main EMC purpose of a ground plane is to provide a low-impedance ground and power return path to minimize induced ground noise. Shielding effects on signal tracks are secondary and are in any case nullified by the components, which stand proud of the board. There is little to be gained in general from having power and ground planes outside the signal planes on four-layer boards, especially considering the extra aggravation involved in testing, diagnostics and rework. The exception will be where there is significant E-field (dv/dt) coupling to or from the tracks which exceeds the coupling due to the components. In this case putting the power and ground planes to the outside of the board will provide an E-field shield to these tracks, but it is rare that such coupling is the dominant factor (backplanes are the most typical exception), and EMC performance is usually best served by keeping the power and ground plane layers adjacent and very close.

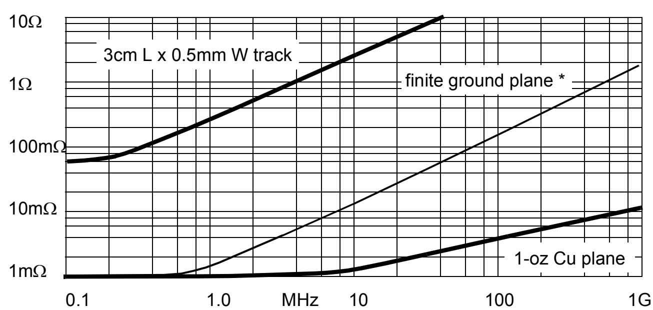


Figure 11.11 Impedance of ground plane versus track

* 5cm wide ground plane at 0.8mm under 3cm length of track, according to equation 11.3

Figure 11.11 compares the impedance between any two points (independent of spacing) on a ground plane *of infinite extent* with the equivalent impedance of a short length of track. The impedance starts to rise at higher frequencies because the skin effect increases the effective resistance of the plane, but this effect follows the square root of frequency (10dB/decade) rather than the inductive wire impedance which is directly proportional to frequency (20dB/decade). For a *finite* ground plane, the geometrical centre should see the ideal impedance while points near the outside will see much higher values as the impedance becomes markedly inductive. The corollary of this is that you should never place critical tracks or devices near the outside edge of the ground plane. An approximation for the inductance of the return path per unit length for a given signal track in the bulk of the PCB has been given by [56] as:

$$L \approx 5\ (h/w)\ \mathrm{nH/cm} \qquad (11.3)$$

where w is the width of the plane, h is the distance between the signal track of interest and the return plane, and the length of the return path is much greater than h.

11.2.2.1 Ground plane on double-sided PCBs

A partial ground plane is also possible on double-sided PCBs. This is not achieved merely by filling all unused space with copper and connecting it to ground – since the purpose of the ground plane is to provide a low inductance ground path, it must be positioned under (or over) the tracks which need this low inductance return. At high frequencies, return current does not take the geometrically shortest return path but will flow preferentially in that part of the plane which is in the neighbourhood of its signal trace. This is because such a route encloses the least area and hence has the lowest overall inductance. Thus the use of an overall ground plane ensures that the optimum return path is always available, allowing the circuit to achieve minimum inductive loop area by its own devices [131].

A partial ground plane

Figure 11.12 illustrates the development of the ground plane concept from the limiting case of two parallel identical tracks. To appreciate the factors which control inductance, remember that the total loop inductance of two parallel tracks which are carrying current in *opposite* directions (signal and return) is given by equation (11.4):

$$L_{tot} = L_1 + L_2 - 2M \qquad (11.4)$$

where L_1, L_2 are the inductances of each track and
M is the mutual inductance between them

M is inversely proportional to the spacing of the tracks; if they were co-located it would be equal to L and the loop inductance would be zero. In contrast, the inductance of two identical tracks carrying current in the *same* direction is given by:

$$L_{tot} = (L + M)/2 \qquad (11.5)$$

so that a closer spacing of tracks *increases* the total inductance, towards that of one track on its own. Since the ground plane is carrying the return current for signal tracks above it, it should be kept as close as possible to the tracks to keep the loop inductance to a minimum. For a continuous ground plane this is set only by the thickness of the intervening board laminate.

11.2.2.2 Ground planes for low cost boards

The ground plane approach is well established for high-value, high-speed digital circuits for which multilayer PCBs are the norm. What is sometimes less appreciated is that it is directly relevant to low-cost analogue circuits as well. Adding a ground plane to a cheap single-sided analogue board is often the single most cost-effective change for improving EMC.

Despite the encroachment of digital technology, analogue circuits are still widely used for simple, cost-sensitive applications, particularly in the consumer market. Examples are security sensors, timers and residual current detectors. For these products, cost is the driving factor and both component count and assembly time must be kept to a minimum. Volumes may be high and assembly is often done in low labour cost areas, so that through hole insertion may still be economic by comparison with surface mounting. PCB technology may be single sided on PRBF, if double-sided PTH on fibreglass represents too high a cost. These products typically are unconcerned with RF emissions, and historically EMC has never been considered in their design. Under the EMC Directive though, RF and transient immunity is important, even as it is for reliability of operation. This immunity can frequently best be achieved by implementing a ground plane on the board.

Adding a grounded copper layer

A plane can be implemented retrospectively by simply adding a top copper layer to a single-sided board design, without changing the original layout (Figure 11.13). The top copper etch pattern is limited to clearance holes around each component lead – the original pad pattern in reverse. The top layer is connected by a through wire or eyelet to the 0V track on the underside.

This construction does not make the plane behave in the optimum fashion, where current flowing within it minimizes magnetic coupling to the circuits – the plane ensuring minimum possible loop area for each circuit path. It cannot do this as long as the 0V returns for each and every circuit are not connected to it. It does, though, act as

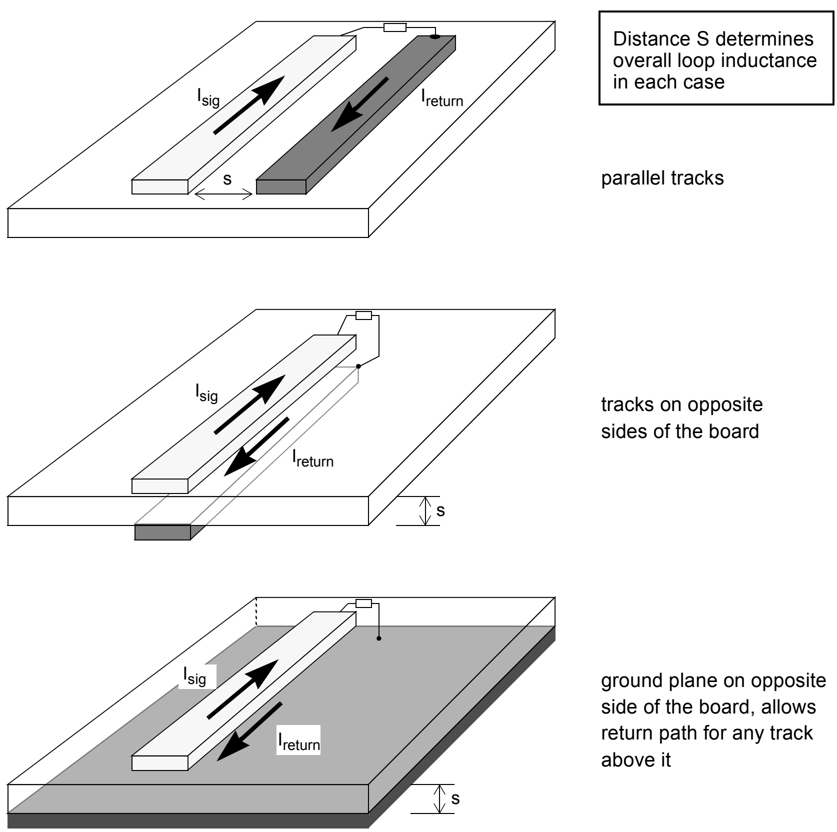


Figure 11.12 Return current paths

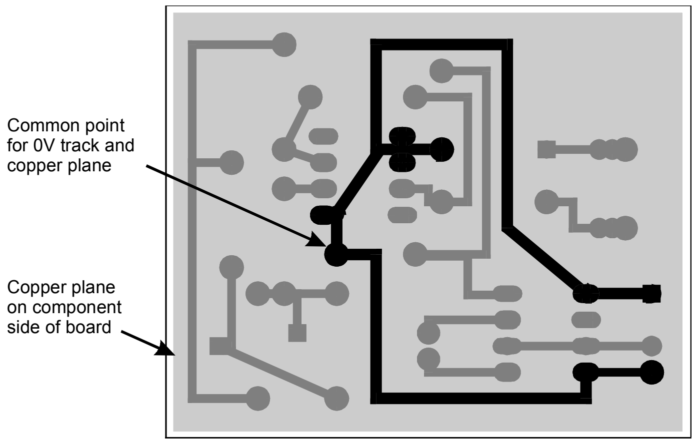


Figure 11.13 Simple copper layer added to a single-sided PCB

a partial electric field screen, reducing the effect of capacitive coupling to the tracks. This is most effective for higher impedance circuits where capacitive coupling is the prime threat. It will also reduce differential voltages across the length and width of the PCB, induced either magnetically or electrically. You can think of it as creating a "quasi-equipotential" area for RF interference over the region of the circuit. The "image plane" discussed in section 14.1.5 works in a similar fashion.

Its effect is maximized if the 0V connection is made in the vicinity of the most sensitive part of the circuit, and if this part of the circuit is located away from the edges of the plane. The purpose of this is to ensure that the remaining interference-induced voltage differences between the plane and the circuit tracks are lowest in the most critical area.

Making the copper layer into a ground plane

If some cost increase is permissible and a double-sided PTH (plated through hole) board can be used, then it is preferable to transfer all 0V current onto the plane layer, which then becomes a true ground plane. All circuit 0V connections are returned to the ground plane via PTH pads and the original 0V track is redundant. In a revised or new layout, the 0V track is omitted entirely, possibly allowing a tighter component and tracking density. Figure 11.14 shows this transition.

This will have the effect not only of reducing capacitive coupling to the circuit, as discussed above, but also magnetic coupling, since all circuit paths now inherently have the least loop area. As long as sensitive circuits are located well inboard of the edges of the plane they will enjoy the greatest protection. Fringing fields at the edge of a ground plane increase its effective inductance, and magnetic coupling to the circuit tracks worsens.

It is not essential that the ground plane area covers the whole of the board, or even all of the tracks, as long as it does cover the tracks and circuit areas which are critical for immunity.

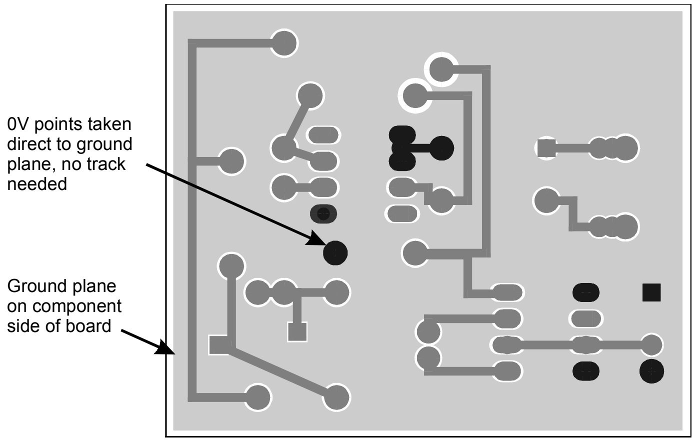


Figure 11.14 Taking all 0V connections to the ground plane

11.2.2.3 Breaks in the ground plane

What *is* essential is that the plane remains unbroken in the direction of current flow. Any deviations from an unbroken plane effectively increase the loop area and hence the inductance. If breaks are necessary it is preferable to include a small bridging track next to a critical signal track to link two adjacent areas of plane (Figure 11.15) – this then develops the layout into a ground grid. A slot in the ground plane will nullify the beneficial effect of the plane if it interrupts the current, however narrow it is. This is why a multilayer construction with an unbroken internal ground plane is the easiest to design, especially for fast logic which requires a closely controlled track characteristic impedance. If you use double-sided board with a partial ground plane, bridging tracks as shown in Figure 11.15 should accompany all critical tracks, especially those which carry sensitive signals a significant way across the board.

The most typical source of ground plane breaks is a slot due to a row of through hole connections, such as required by a SIL or DIL IC package (Figure 11.16). The effect of

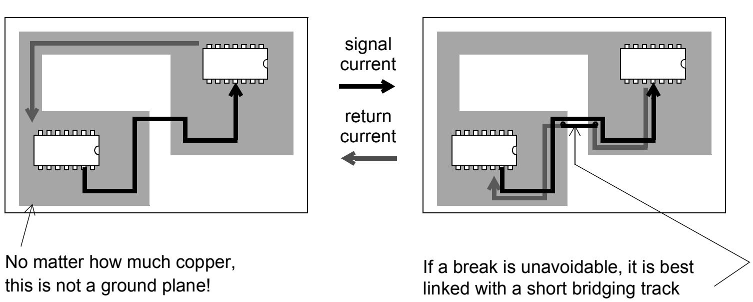


Figure 11.15 A broken ground plane

such a slot is worst for the tracks in its centre, since the return current is diverted all the way to the edge of the slot. This translates to greater ground inductance and an unnecessary increase in induced interference voltage in this area of the board's ground network. If the slot straddles a particularly sensitive circuit then worsened immunity will result, even though you have gone to the trouble of putting a ground plane onto the board. It is a simple matter to ensure that each pair of holes is bridged by a thin copper trace – most etching technology is capable of this degree of resolution – it adds nothing to the production cost and can give several dB improvement in immunity in the best case.

The inductance of a slot in the plane can be modelled as a length of short circuited transmission line. [56] gives an approximation for this inductance as shown in Figure 11.16. It is instructive to compare the order of magnitude of values this gives versus that obtained from equation (11.3): a 3cm track over an unbroken ground plane of width 5cm, 0.8mm above it, gives 0.24nH return path inductance. A 3cm long slot under this track of width 1mm, with 4cm of board either side (i.e. board is 8cm x 5cm), will increase this inductance to 2.6nH.

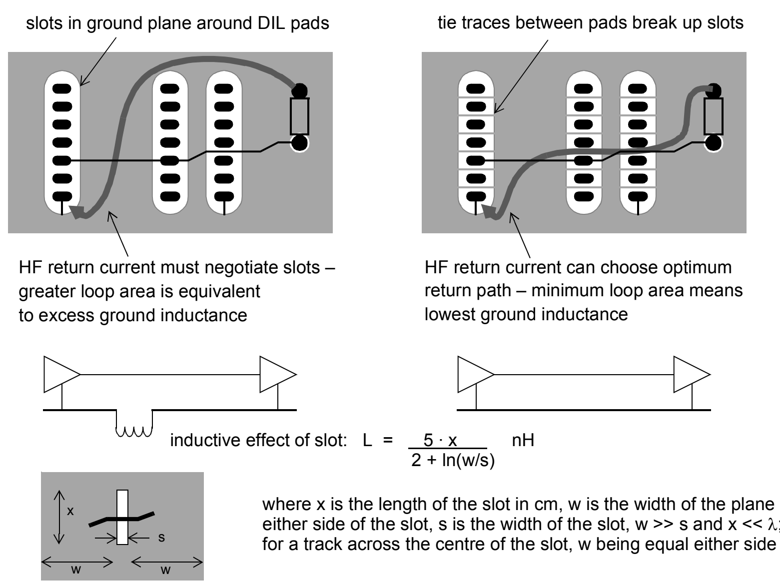


Figure 11.16 Dealing with slots at DIL pinouts

11.2.2.4 Multilayer boards

All of the foregoing, with respect to ground plane construction in the plane of the board, applies to multilayer technology. Additionally, the order of the layers becomes significant when high-speed circuits are implemented on multilayer boards.

In these configurations, the most important aspect is that every critical signal layer should be adjacent to a ground or power plane layer. Also, power and ground planes should be on adjacent layers to take advantage of the inter-layer capacitance for high frequency decoupling. Critical tracks should be routed adjacent to ground rather than power planes, for preference. Such tracks (typically carrying high di/dt signals such as clocks) should also not jump through vias from one ground reference layer to another, unless the ground layers are tied together with vias at that point. In general, using multiple ground (0V) layers is perfectly acceptable and sometimes essential, but it should be accompanied by via stitching widely distributed around the planes (see also sections 11.2.2.6 and 11.2.2.7 shortly).

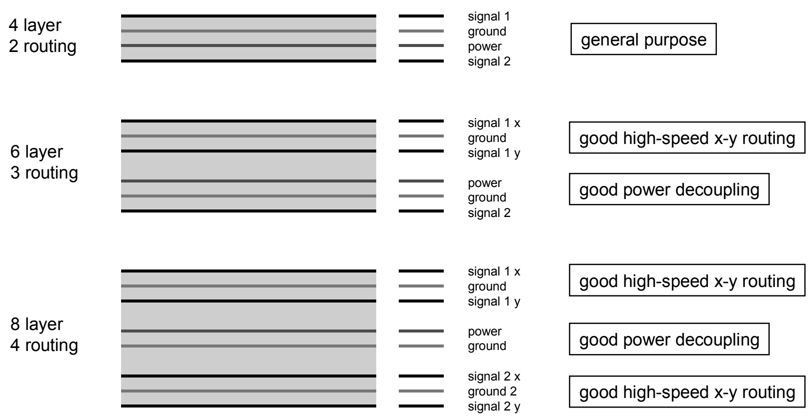


Figure 11.17 Layer stacking on a multilayer board

Figure 11.17 illustrates possible (not by any means mandatory) layer configurations for various layer multiples. If x- and y-orientation layers share a common ground plane this allows both constant impedance and minimum separation from the plane layer for high-speed routing. High-density designs can use more than one set of x-plane-y layers as shown in the 8-layer example above. If high-speed tracks can be distinguished from non-critical ones, the non-critical tracks can be run on a separate layer adjacent to a second plane (marked signal 2 in the six-layer example). In all cases, putting power and ground planes on adjacent inner layers gives maximum distributed decoupling capacitance between them.

The main EMC purpose of a ground plane is to provide a low impedance return path to minimize induced ground noise. Shielding effects on signal tracks are secondary. Putting power and ground planes outside the signal layers on multilayer boards will only be a significant advantage if E-field shielding of the tracks is necessary, and this will in any case be compromised by the unavoidable fields from the components. But an extra ground plane on the outside layer of a board can be helpful when capacitive coupling, such as from large-area components carrying HF voltages, needs to be dealt with by extra shielding measures. [114] describes a situation in which a microprocessor heatsink was electrically bonded to a ground plane on the board (see section 11.2.2.8 later), and the inductance of the bond connection was noticeably more significant when the ground plane was on an inside layer than when it was outside. This resulted in several dB increase in emissions in the frequency range around 1GHz. If your circuit has very wideband EMC aspects then such an approach is worthwhile, but for less demanding applications the physical disadvantages associated with outside plane layers tend to make them unattractive.

11.2.2.5 Crosstalk

A ground plane is a useful tool to combat crosstalk, which is strictly speaking an internal EMC phenomenon. Crosstalk coupling between two tracks is mediated via inductive, capacitive and common ground impedance routes, usually a combination of all three (Figure 11.18). The effect of the ground plane is to significantly reduce the common ground impedance Z_G, by between 40–70dB, in the case of an infinite ground plane compared to a narrow track.

The ground plane may also reduce mutual inductance coupling (M_{12}) by ensuring that the coupled current loops are not co-planar. Capacitive coupling between the tracks will not be directly affected by the ground plane, but the lowered impedance of the line (equivalent to saying that C_{1G} and C_{2G} have been increased) will reduce capacitive crosstalk amplitude.

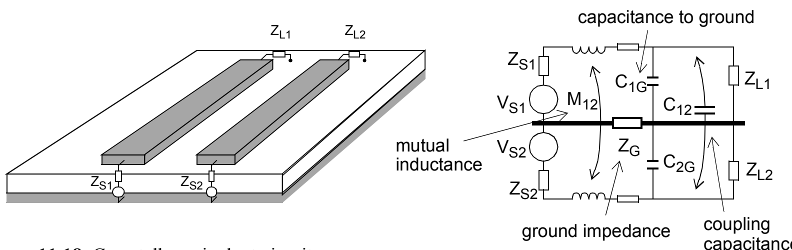


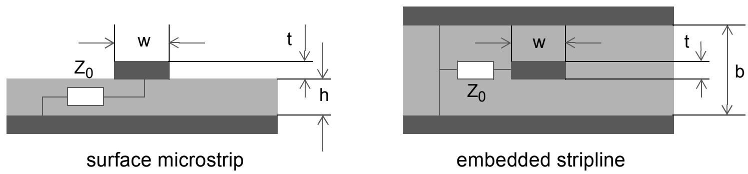
Figure 11.18 Crosstalk equivalent circuit

Crosstalk problems, or internal interference coupling, are by no means limited to digital circuits, although these tend to make the problem most visible. A common threat to the immunity of analogue circuits is crosstalk from other circuits carrying high interference currents or voltages. Mains-borne interference is the most widespread, but RF or transient interference induced onto signal cables can also be significant. You should always be on the lookout to minimize crosstalk from these sources by ensuring the maximum separation (hence minimum M_{12} and C_{12}) between the circuits; or by interposing electric field screening between them, if C_{12} is dominant; or by implementing a ground plane, to minimize Z_G and maximize C_{1G} and C_{2G}.

11.2.2.6 Constant impedance track layers

High frequency circuits need matched impedances – see section 12.1.2.4. When an HF signal has to be passed across a board without distortion due to variations in the geometry of the tracks carrying it, these have to be designed as transmission lines with a known, constant impedance. This invariably means that they should be placed on a layer next to a reference ground plane and should remain on that layer for the whole of their run: any vias which might take them to another layer represent a discontinuity in impedance, as would any jump to another reference plane. The characteristic impedance looking along the line is determined entirely by the cross-sectional geometry, and so the same geometry has to be maintained for the total length. No variations in width (w) or separation (h) from the ground plane are permissible.

Various equations have been published which offer a simple analytical way to calculate the characteristic impedance of a given geometry, but these are always approximations, valid over restricted dimensional ranges (typically worse for larger w/h), and are subject to varying degrees of error. Complicated equations are usually more accurate, and numerical modelling rather than analytical calculation gives the



Approximate widths for given impedances: t = 35µm, ε_r = 4.2

	w mm for surface microstrip		w mm for embedded stripline	
Z_0 Ω	h = 0.25 mm	h = 0.5 mm	b = 0.3 mm	b = 0.6 mm
30	0.77		0.21	0.47
50	0.43	0.9	0.09	0.21
70	0.23	0.5		0.09
90	0.12	0.28		
100		0.2		

Figure 11.19 Track geometries and impedances (see Annex D section D.5.3 for equations)

most accurate result [42]. Several proprietary software packages are available to perform this modelling, using the geometry to determine the per-unit-length inductance and capacitance L_0 and C_0 and then calculating Z_0 from $Z_0 = \sqrt{(L_0/C_0)}$. Some equations for the more common configurations are given in Appendix D (section D.5.3), and Figure 11.19 shows some figures for typical applications using these equations.

You can see from this that surface microstrip lines are more suited to higher impedances than embedded striplines; the latter are more suitable for high energy clocks simply because the use of two reference planes encloses the magnetic field radiation more effectively. Single-ended line configurations are shown above; a differential transmission line formed from two striplines or microstrips against a common ground plane is harder to model because of the coupling between the two lines. If they are widely separated such that the coupling between them is low, the differential characteristic impedance $Z_{0(diff)}$ can be approximated by twice the Z_0 of each half of the pair. For closer pairs with greater mutual coupling then $Z_{0(diff)} = 2 \cdot Z_0 \cdot (1 - k)$ where k is the coupling factor between them.

The impedances depend on practical dimensions of w and h and also on the exact relative permittivity ε_r of the board material. Each of these is subject to some error, for instance ε_r is generally given at 1MHz by default whereas a track's Z_0 is of most interest in the range above 500MHz; at these frequencies ε_r can fall by typically 5%. The value depends on the glass-to-resin content of fibreglass materials and different materials of the same generic type can vary by several percent. Also the specified width of a track is subject to deviations in the PCB manufacturing process, as is the thickness of the layers. Even with the narrowest production trace width, striplines and microstrip cannot achieve high impedances: a practical range of impedances is between 50 and 100 ohms. In all cases, high impedances require large separations and hence very thick dielectrics, making PCBs excessively thick. Lower values are easier to achieve but the track width with thin separations becomes unwieldy in practical layouts.

For all these reasons, and because PCB manufacturers are themselves normally in the best position to determine the actual Z_0 of tracks on their boards, a usual approach is to specify on the PCB drawing the actual intended impedance of tracks of a given width on a specified common impedance layer.

11.2.2.7 Power planes

The subject of power planes raises two issues:

- should they be board-wide or segmented into islands?
- how should they be coupled to the ground plane?

The assumption is that the ground (0V) plane is the reference for the whole circuit. Then each supply (there may be more than one supply voltage) needs to be decoupled to this 0V. Such decoupling is most effective if the power plane is on the next layer to a 0V plane, and if the separation distance is minimized. This gives maximum capacitance between the two, or to put it another way, it creates a transmission line structure with the lowest possible Z_0. So given a single 0V plane on an inner layer, this allows a maximum of two power plane layers within the same area (one each side of the 0V plane, see Figure 11.20).

Doing this will of course deny the 0V plane to any signal tracks that might need it; so in practice, if you have more than one supply voltage to a VLSI IC, you will need at least two 0V plane layers, one for the power supplies and one for the signals. Yet more supply voltages will need further 0V planes, or will compromise the decoupling of some of the power planes. Multiple 0V planes need many stitching vias in order to maintain the lowest high frequency voltage differential between them. If at any point you choose to allow tracks to jump from a layer next to one plane to a layer next to another, at least one via next to the jump is mandatory to keep integrity of the return current path.

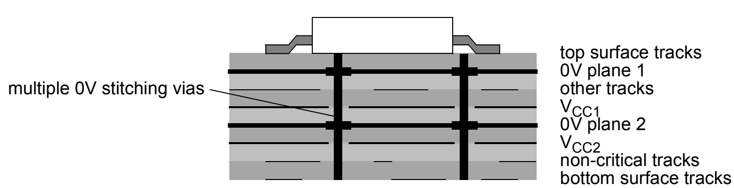


Figure 11.20 Layer stack-up for two supply planes

Segmentation

The mention of transmission lines above gives a hint of the problem that can be caused by a single large power plane – particularly one (say 3V3 or 5V) that spreads across the whole PCB and mirrors the 0V plane. This creates a transmission line "sandwich" that will have principal resonances determined by its x and y dimensions. At the resonant frequencies two undesirable things happen: the impedance between V_{CC} and 0V becomes high at certain positions around the board, and the radiating efficiency of the edges of the sandwich is maximized. The resonant frequencies are affected by the dielectric between the planes. For FR4 with $\varepsilon_r = 4.2$, the frequency is reduced by $\sqrt{4.2}$ from the free-space value, so a board with dimensions 30cm x 12cm will have a bare-board resonance in the y direction of 610MHz and in the x direction of 244MHz. Clock energy at these frequencies will radiate effectively, and RF susceptibilities can be expected to be worst at the same frequencies.

The solution to this problem is, firstly, not to allow the power plane to exactly mirror the 0V plane; if the power plane edge is inset from the ground plane edge by at least 10 · h (Figure 11.21) then the radiating efficiency of the edges of the resulting transmission line is noticeably reduced. But beyond this, consider breaking a large, homogeneous area of power plane into islands. Separating a large plane into individual smaller planes, each with its own decoupling capacitors and supplied via an isolating series choke, pushes the transmission line resonances typically into the GHz region where they are likely to be less harmful. It also reduces cross-coupling via the power plane of different circuits on the board, and reduces the spread of V_{CC} noise across the board. The series chokes – for instance, surface mount ferrites – need only have a few tens of ohms impedance in the VHF range to be effective at segregating the individual planes. The segmentation boundaries should follow natural divisions in the circuit structure, particularly with respect to clock distribution, as far as possible.

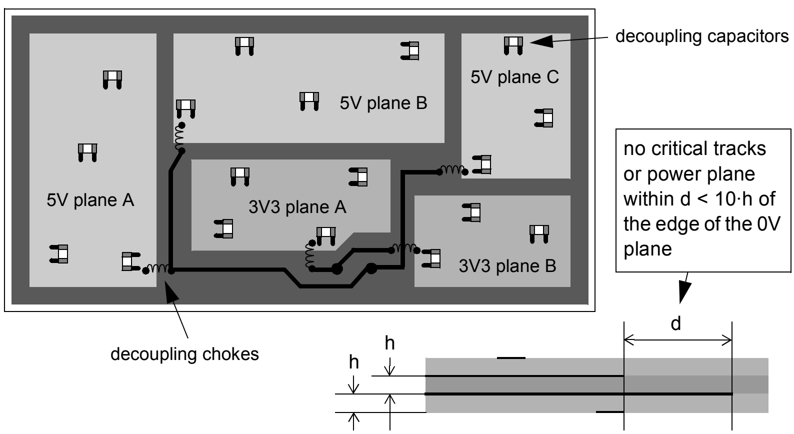


Figure 11.21 Power plane segmentation and the 10 · h rule

This segmentation approach is not appropriate for ground planes, which should always be continuous over the whole PCB.

As we have seen, several aspects of the EMC performance depend on the layer separation distance, so that the detail of the vertical sandwich construction of the PCB can be as important for its EMC as is the layout. The separation is determined by the thickness of the pre-preg and core materials used in the build of the bare board. This is often left to the PCB manufacturer and is not specified by the circuit designer, or anyone else. As a result you end up with the default thicknesses used by a particular board supplier, which might be perfectly adequate for the EMC performance of the product and so is never questioned; but if another board supplier is chosen during the product life cycle, it is entirely possible that a different set of thicknesses could be used which result in changed performance. To guard against this, make sure that you specify layer thicknesses in the PCB drawing, if necessary checking with your preferred supplier what their defaults will be.

11.2.2.8 Heatsinks on PCBs

Heatsinks can be a particular cause of EMC difficulties on PCBs. The subject is touched on again in section 12.1.5.2 on SMPS design. Here, the concern is the effect of stray capacitance from heatsinks on digital devices carrying RF voltages, such as microprocessors and ASICs.

Consider a simplified equivalent circuit of a PCB with a 0V plane, a processor, a heatsink on the processor, and an enclosure around the whole assembly (Figure 11.22).

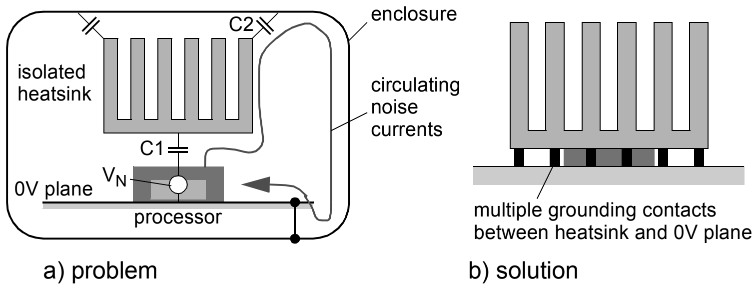


Figure 11.22 Heatsinks on processors

The processor is the source of the RF noise: voltages V_N on the silicon at clock frequencies appear with respect to the 0V plane and are capacitively coupled directly to the metal of the heatsink. The heatsink must have good thermal contact to the chip, and a by-product of this is a high coupling capacitance C1. In its turn, the heatsink is a large chunk of metal, and it has a high capacitance C2 to its surroundings, particularly the enclosure.

The effect of this is to energize the entire enclosure with circulating currents at the clock frequency and its harmonics. Any weakness in the enclosure will allow these frequencies to radiate. To control this, the most effective measure is to prevent the heatsink from carrying high levels of clock noise. This means it must be grounded to the 0V plane on the PCB, and cannot be allowed to float. Capacitance C1 and noise source V_N cannot be removed, but if the heatsink is grounded then the circulating noise currents remain in its locality and the voltage on the heatsink is minimized. The enclosure itself is not stressed, since C2 is not fed from a noise source.

The catch with this is that any grounding inductance resonates in parallel with the capacitance of the heatsink so that at the resonant frequency, even a supposedly grounded heatsink will carry high levels of noise and couple them out to the enclosure. A heatsink self-capacitance of 10pF and an inductance of 3nH (no more than a few mm of grounding contact) results in a resonant frequency of 900MHz, which could well be troublesome for many types of processor. If you take the above advice, ground the heatsink at one point and the emissions at some frequencies go up, this is most probably what is happening. The inductance of the grounding connection needs to be well below 1nH to be effective, and this means that many grounding contacts around the outside of the heatsink must be provided, not just one or two. The most effective designs use a continuous conductive gasket around the periphery of the heatsink (which must be conductively finished) to a ground plane contact strip on the surface of the PCB.

11.2.3 Configuring I/O and circuit grounds

11.2.3.1 The interface ground structure

Decoupling and shielding techniques to deal with common mode currents appearing on cables both require a "clean" ground area, not contaminated by internally generated noise, and to which external disturbances can be filtered to prevent them passing into the circuit. This forms the low transfer impedance grounding structure, as per section 11.1.2.2, for the cable interface. *Filtering at high frequencies is next to useless without such a ground*. Unless you consider this as part of the layout specification early in the design phase, such a ground will not be available. Provide an interface ground by grouping all I/O leads in one area, and connecting their shields and decoupling capacitors to a separate ground plane in this area. This interface ground plane can be on a separate part of the main circuit PCB [106] as shown in Figure 11.23, or it can be a metal plate on which the connectors are mounted. The external ground (which may be only the mains safety earth) and the metal or metallized case, if one is used, are connected here as well, via a low inductance link. It is also possible for the interface to be an entire PCB in itself, with an integral ground plane to which the filter components and cable screens are connected and which is then connected to an external metal or metallized chassis via suitable gaskets or other forms of bonding. The filtered signals and power are then passed through to internal PCBs via a board-to-board connection. Figure 11.24 shows a typical arrangement for a product with digital, analogue and interface sections.

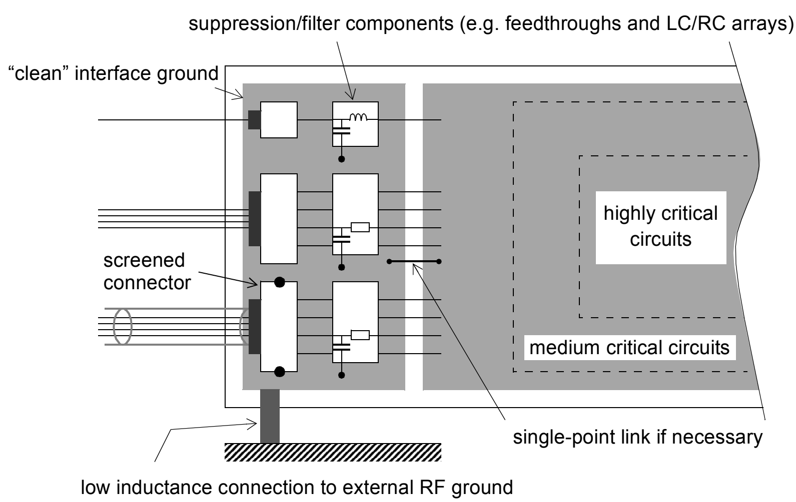


Figure 11.23 The interface ground on a PCB

This interface ground should only connect to the internal circuit ground, if at all, at one point. This prevents noise currents flowing through the interface ground plane and "contaminating" it. No other connections to the interface ground are allowed. As well as preventing common mode emissions, this layout also shunts incoming interference currents (transient or RF) to the interface ground and prevents them flowing through susceptible circuitry. If for reasons of enclosure design it is essential to have cables interfacing with the unit or PCB at different places around its periphery, you should still arrange to couple them all to a separate ground structure through which no circuit currents are flowing. In this case a chassis plate is mandatory.

These requirements can be ensured by devoting an area along one edge of the PCB to the interface ground plane, including the I/O resistive, capacitive and inductive filtering and suppression elements on it (but no other circuits), and separating the circuit 0V from it either absolutely, or by a single connecting link. RF-critical circuitry should be widely separated from the I/O area.

For ESD protection the circuit ground *must* be referenced to the chassis ground; this prevents transient high voltages appearing between the two and creating a secondary discharge (*cf* sections 10.3.3.3 and 12.2.1.2). This can easily be done by using multiple plated-through holes to the ground plane and metallic standoff spacers. If there has to be DC isolation between the two grounds at this point, use a 10–100nF RF ceramic capacitor at each location. You can provide a clean I/O ground on plug-in rack mounting cards by using wiping finger style contacts to connect this ground track directly to the chassis.

11.2.3.2 Separate circuit grounds

There are two schools of thought as to maintaining separate 0V references for different operational parts of a circuit. The first (see Figure 11.24) says that you should never extend a digital ground plane over an analogue section of the PCB as this will couple digital noise into the analogue circuitry. A single point connection between digital and analogue grounds can be made at the system's analogue-to-digital converter. It is very important in this arrangement *not* to connect the digital circuitry separately to an external ground [46]. If you do this, extra current paths are set up which allow digital circuit noise to circulate in the clean ground.

Interfaces directly to the digital circuitry (such as a port input or output) should be buffered so that they do not need to be referenced to the digital 0V. The best interface is an opto-isolator or relay, but this is of course expensive. When you can't afford isolation, a separate buffer IC which can be referenced to the I/O ground is preferable;

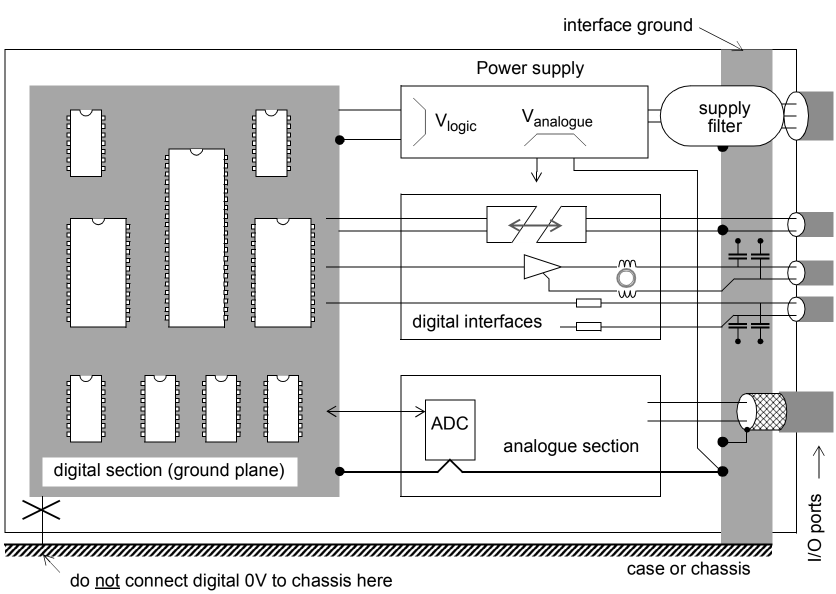


Figure 11.24 Multiple ground areas

otherwise, buffer the port with a series resistor or choke, and decouple the line *at the board interface* (not somewhere in-board) with a capacitor and/or a transient suppressor to the clean ground. More is said about I/O filtering in section 13.2.4.

Notice how the system partitioning, discussed in section 11.1.1, is essential to allow you to group the I/O leads together and away from the noisy or susceptible sections. Also the mains cable, as far as EMC is concerned, is another I/O cable. Assuming that you are using a block mains filter, fit this to the "clean" ground reference plate directly.

11.2.3.3 Solid universal circuit ground

The second school of thought, diametrically opposed to the first, says that all parts of the circuit should be referenced to a single, unbroken 0V reference plane. If this approach is chosen, it is crucial that a high quality, low impedance plane is used; any 0V tracks would be unacceptable. The effect of this construction is that all signal and power return currents flow in the plane, so that common impedance coupling between different parts of the circuit is unavoidable, and it is only careful layout and the low impedance of the plane which mitigates this coupling to an acceptable extent. Mixing sensitive analogue with high-speed digital circuits on the same board is risky, although with careful partitioning it can still be successful.

But in an unscreened product, the PCB is exposed directly to the RF environment and multiple planes will act as effective dipole antennas, concentrating induced interference at their junctions, or radiating fields due to voltage differences between them. The advantage of a single plane is that there is no opportunity for voltage differences to arise, as a result of interference or for any other reason, between different 0V nodes, since there are no different nodes. The tricky question of where exactly to place the separation boundary is avoided. So is the hazard of 0V differentials, potentially a problem with the first approach. If the board is mounted on a metal chassis, the ground plane can be tied to this chassis at multiple points, and this improves the RF "solidity" of the whole assembly, so that induced interference currents will inherently cause a lower level of internal disturbance.

In the end, you are free to use either approach, as long as the implications of the decision are appreciated. Segregation of grounds removes the threat of cross-interference between different parts of the circuit but requires you to think carefully about placing the boundaries and defining the circuit's relationship to the chassis; with complex multi-sectioned circuits it can be difficult to administer. Unifying the ground makes it easy to control the ground design but increases the risk of one part of the circuit interfering with another.

11.2.3.4 Cable screen connection

Input/output decoupling is of critical importance with either circuit grounding regime because it is vital to keep cable common mode interference currents to a minimum. As an example, consider Figure 11.25. If the cable screen or return is taken to the wrong point with respect to the output driver decoupling capacitor, the high-speed current transitions on the driver supply (which flow through the decoupling capacitor traces) generate common mode voltage noise V_N which is delivered to the cable and which appears as a radiated emission. Cable screens must always be taken to a point at which there is the minimum noise with respect to the system's ground reference. This means that the dedicated interface ground area must always be implemented and all cable connections either taken directly to it (screens) or filtered to it (signal and power lines).

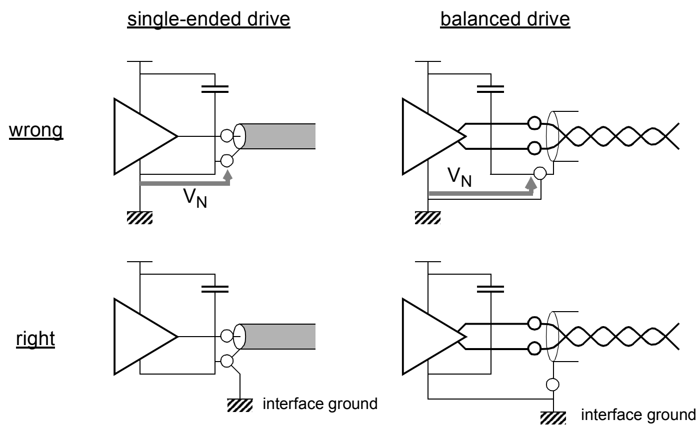


Figure 11.25 The point of connection of I/O cable screens

11.2.4 Rules for PCB layout

Because it is impractical to optimize the PCB layout for all individual signal circuits, you have to concentrate on those which are the greatest threat. These are the ones which carry the highest di/dt most frequently, especially clock lines and data bus lines, and square-wave oscillators at high powers, especially in switching power supplies. From the point of view of susceptibility, sensitive circuits – particularly edge-triggered inputs, clocked systems, and precision analogue amplifiers – must be similarly treated. Once these circuits have been identified and partitioned you can concentrate on dealing with their loop inductance and ground coupling. The aim should be to ensure that circulating ground noise currents do not get the opportunity to infect the system.

A structured approach to PCB layout for EMC

These rules can at least partly form the basis for an in-house EMC design policy and design review checklist; some can be implemented in a CAD design rule checker. They assume that you will use at least one 0V (ground) plane.

Before routing begins,

- identify and label high di/dt circuits
- identify and label sensitive circuits
- identify 0V plane(s) including separate areas
- identify power plane segments
- identify the interface ground plane
- decide on layer stack-up:
  - identify 0V plane layers
  - identify constant impedance layer(s)
  - identify power plane layer(s), next to 0V plane: some of these may also carry tracks in areas where power planes aren't needed

- check component placement to ensure:
  - no unnecessarily long track routes
  - no proximity of noisy components to sensitive ones
  - critical circuits away from ground plane edge
  - maximum partitioning of interfaces and functional sections: no unprotected interface routes to pass close to operational circuits
  - all filter components at the interface they are protecting
- board-to-board connections: ground pins should be distributed along multi-way connectors close to high-speed or sensitive signals
- identify points for bonding the ground plane(s) to chassis

During routing,

- ensure no tracks cross any breaks in a 0V plane
  - if they must, then make sure series buffer resistors are placed appropriately at the break
- flag any breaks or gaps in a 0V plane and decide whether they are necessary or can be minimized
- check that critical and constant-impedance tracks do not swap layers
  - if they must, they should be routed above or below a single 0V plane, not jump to a different 0V plane
- check adequate placement of decoupling capacitors
  - near device power pins
  - minimum inductance track/via layout
- ensure that interface filtering and transient protection is tracked with low inductance to the interface ground plane
- identify and control common impedance current paths for power switching circuits and sensitive wideband circuits
- check implementation of $10 \cdot h$ rule for power planes and critical tracks versus the ground plane edge
- for balanced differential signal track pairs, confirm that adequate balance is maintained along the entire run – separation of at least $3 \cdot h$ from other tracks is usually enough
- minimize surface areas of nodes with high dv/dt
- if empty areas of any layer are flood filled with copper, ensure that each such area is connected to 0V, not floating

If you have to design a PCB without a 0V plane,

- identify critical (high-current, high di/dt or sensitive) circuit loops, including the appropriate segment of the 0V track
- minimize enclosed loop areas in these loops
- flood-fill and mesh the 0V tracks as much as possible

Chapter 12

Digital and analogue circuit design

Circuit design is an essential contributor to EMC. In this area we can distinguish the techniques used to control radio frequency emissions from an operating circuit, and those used to control radio frequency and transient immunity of an operating circuit. There are some common points between these two, but there are also some differences which mean that to completely address the compatibility of a product you need to deal with both. Low frequency (mains supply) EMC is also a function of circuit design, principally of the power supply, but does not suffer from the mystery surrounding RF effects and is not considered in this chapter.

12.1 Design for emissions control

Digital circuits are prolific generators of electromagnetic interference. High frequency square waves, rich in harmonics, are distributed throughout the system. The harmonic frequency components reach into the part of the spectrum where cable and enclosure resonance effects are important. Analogue circuits are in general much quieter because high frequency square waves are not normally a feature. A major exception is wide bandwidth video circuits, which transmit broadband signals up to several tens of MHz, or over a hundred MHz for high resolution video. Any analogue design which includes a high frequency oscillator or other high di/dt circuits must follow HF design principles, especially with regard to ground layout.

Some low frequency amplifier circuits can oscillate in the MHz range, especially when driving a capacitive load, and this can cause unexpected emissions. The switching power supply is a serious cause of interference at low to medium frequencies since it is essentially a high-power square wave oscillator.

12.1.1 The Fourier spectrum

12.1.1.1 The time domain and the frequency domain

Basic to an understanding of why switching circuits cause interference is the concept of the time domain/frequency domain transform. Most circuit designers are used to working with waveforms in the time domain, as viewed on an oscilloscope, but any repeating waveform can also be represented in the frequency domain, for which the basic measuring and display instrument is the spectrum analyser (section 6.1.2). Whereas the oscilloscope shows a segment of the waveform displayed against time, the spectrum analyser will show the same signal displayed against frequency. Thus the relative amplitudes of different frequency components of the signal are instantly seen.

The mathematical tool which allows you to analyse a known time domain waveform in the frequency domain is called the Fourier transform. The necessary equations for the Fourier transform are covered in Appendix D, section D.7. Figure 12.2 shows the spectral amplitude compositions of various types of waveform (phase relationships are rarely of any interest for EMC purposes). The sinewave has only a single component at its fundamental frequency. A square wave with infinitely fast rise and fall times has a series of odd harmonics (3, 5, 7, etc. multiples of the fundamental frequency) extending to infinity. A sawtooth contains both even and odd harmonics.

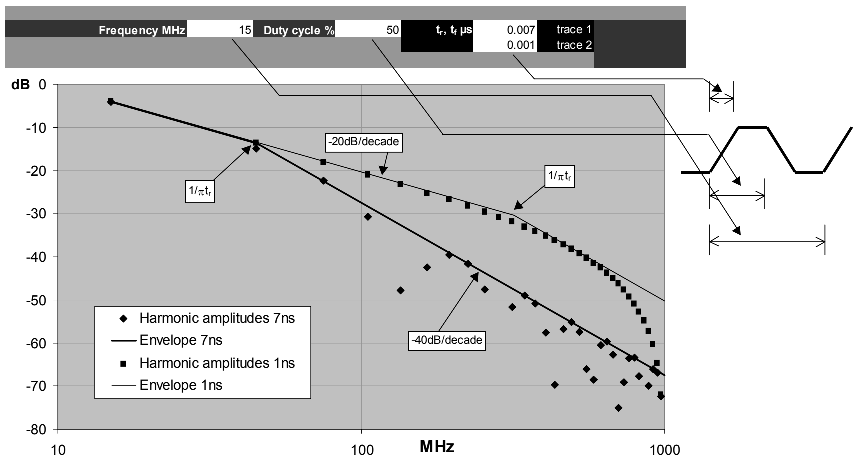


Figure 12.1 Harmonic envelope of a 50% duty cycle trapezoid

Switching waveforms can be represented as trapezoidal; a digital clock waveform is normally a square wave with finite rise and fall times. The harmonic amplitude content of a trapezoid decreases from the fundamental at a rate of 20dB per decade until a breakpoint is reached at $1/\pi t_r$, after which it decreases at 40dB/decade (Figure 12.1). Of related interest is the differentiated trapezoid, which is an impulse with finite rise and fall times. This has the same spectrum as the trapezoid at higher frequencies, but the amplitude of the fundamental and lower order harmonics is reduced and flat with frequency. (This property is intuitively obvious as a differentiator has a rising frequency response of +20dB/decade.) Reducing the trapezoid's duty cycle from 50% has the same effect of decreasing the fundamental and low frequency harmonic content.

Asymmetrical slew rates and duty cycles other than 50% generate even (multiples of 2, 4, 6, etc.) as well as odd harmonics. This feature is important, since differences between high- and low-level output drive and load currents mean that most logic circuits exhibit different rise and fall times, and it explains the presence and often preponderance of even harmonics at the higher frequencies. Changing the duty cycle even slightly away from 50% will dramatically increase the amplitude of even harmonics with little change to the odd components; conversely a duty cycle of 33.3% causes a null in the odd harmonics.

12.1.1.2 Choice of logic family

The damage as far as emissions are concerned is done by switching edges which have a fast rise or fall time (note that this is not the same as propagation delay and is rarely specified in data sheets; where it is, it is usually a maximum figure). Using the slowest

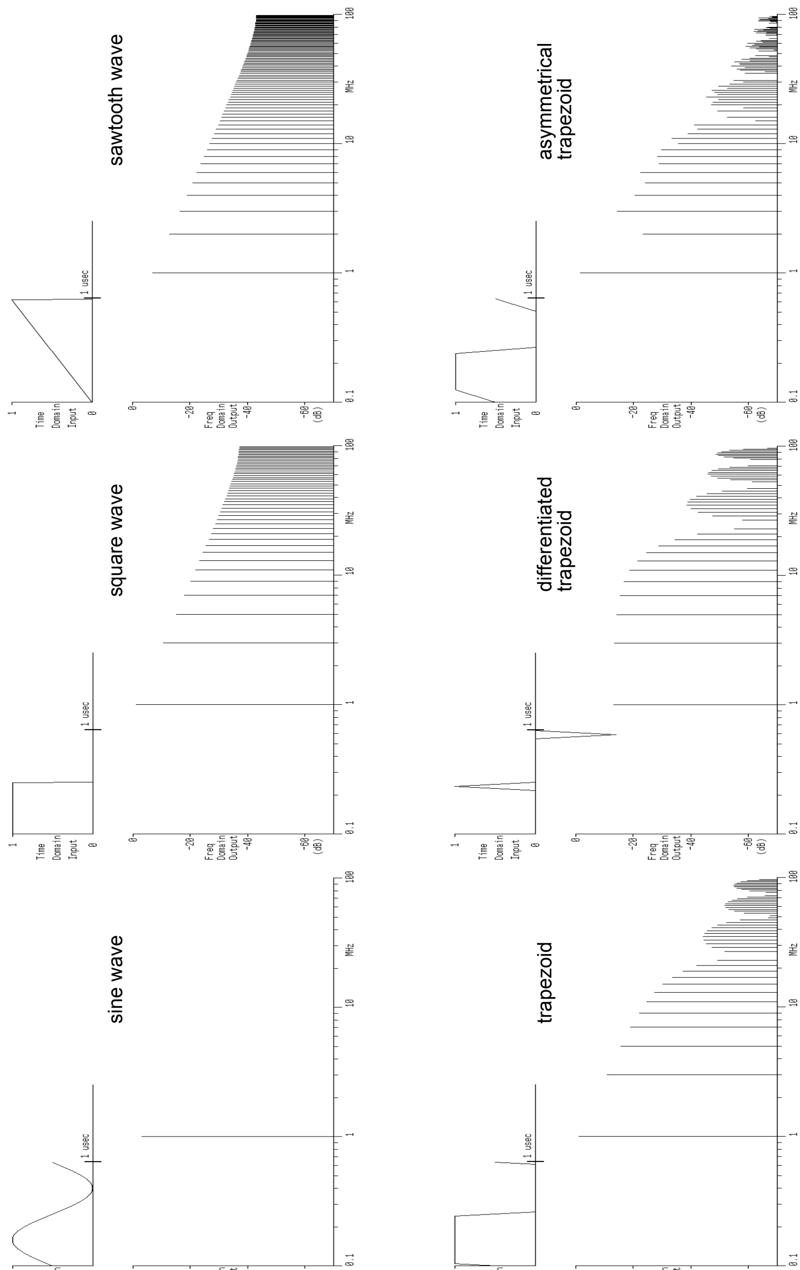


Figure 12.2 Frequency spectra for various waveforms

risetime compatible with reliable operation will minimize the amplitude of the higher-order harmonics where radiation is more efficient. Figure 12.1 shows the calculated harmonic amplitudes for a 15MHz clock with risetimes of 1ns and 7ns. An improvement approaching 20dB is possible at frequencies around 400MHz by slowing the risetime.

The advice based on this premise must be, use the slowest logic family that will do the job; don't use fast logic when it is unnecessary. Treat with caution any proposal to substitute devices from a faster logic family, such as replacing 74HC parts with 74AC. Where parts of the circuit must operate at high speed, use fast logic only for those parts and keep the clock signals local. This preference for slow logic is unfortunately in direct opposition to the demands of software engineers for ever-greater processing speeds.

The graph in Figure 12.3 shows the measured harmonic envelope of a 10MHz 50% duty cycle waveform for three devices of different logic families in the same circuit. Note the emphasis in the harmonics above 200MHz for the 74AC and 74F types. From the point of view of immunity, a slow logic family will respond less readily to fast transient interference (see section 12.2.2.1).

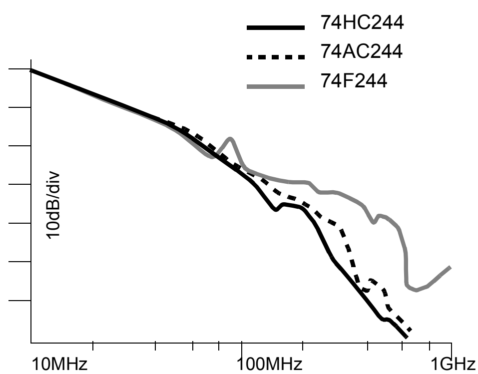


Figure 12.3 Comparison of harmonic envelopes of different logic families

Some IC manufacturers are addressing the problem of RF emissions at the chip level. By careful attention to the internal switching regime of VLSI devices, noise currents appearing at the pins can be minimized. The transition times can be optimized rather than minimized for a given application [73]. Alternatively, the shape of the transition can be "softened" to reduce higher order harmonic content without substantially reducing the logic switching speed. Revised package design and smaller packages can allow the decoupling capacitor (section 12.1.3) to be placed as close as possible to the chip, without the internal leadframe's inductance negating its effect; also, the reduction in operating silicon area gained from shrinking silicon design rules can be used to put a respectable-sized decoupling capacitor (say 1nF) actually on the silicon.

12.1.1.3 I_{CC} at switching

The Fourier components of the output waveform are not the only reason for emissions of logic circuits. The current taken from the V_{CC} pin or pins related to driving the outputs has two components (Figure 12.4):

- the current fed through the output pin(s) and used for charging or discharging the capacitance of each output node;
- the current taken through the totem-pole output stage transistors as they switch, which is not passed through the output pin(s).

Both of these components appear at the V_{CC} pin and have to be decoupled, as discussed shortly in section 12.1.3. The second component does not involve the signal tracks and is not affected by the quality of their routing or layout, but it is significantly affected by the nature of the decoupling regime and of the use of ground and power planes. Logic families which control this unwanted current (sometimes known as "delta-I" noise) are widely touted as helping to reduce emissions overall.

As well as the current associated with output drivers, there are similar supply currents associated with switching each of the internal nodes of the device. Individually, these are orders of magnitude less than the output drive currents. But in a typical VLSI device there are many more of them: thousands at least, and if as is usual the device is synchronously clocked then all of the current transitions occur simultaneously, at the clock edge. Therefore the supply current of any VLSI device tends to be dominated by short but potentially high amplitude current spikes at each transition of the internal clock.

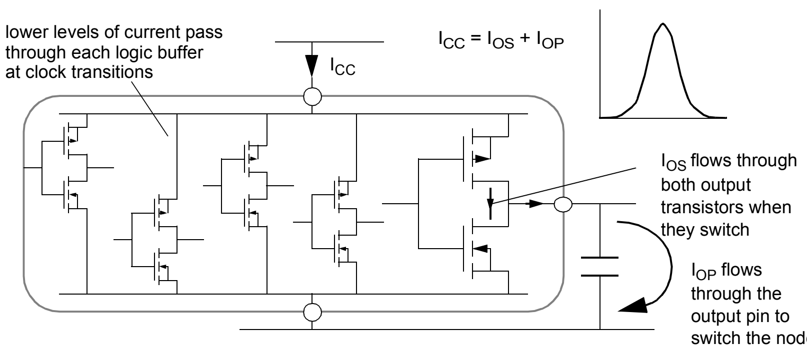


Figure 12.4 Current through the supply pin

Since these current transients are almost entirely a function of the internal design of the chip, it is for this reason as much as any other that different manufacturers' implementations of a particular device may be noisier or quieter in EMC terms. In any case, their existence ensures that high quality supply decoupling (section 12.1.3) in conjunction with carefully laid out power and 0V planes is essential for VLSI devices.

12.1.2 Radiation from logic circuits

The following section discusses differential and common mode radiation mechanisms from circuits on a PCB, assuming a single source frequency. The equipment is taken to have a plastic enclosure, so that no screening effect is offered and the PCB radiates as if it is a bare board. Conductive enclosures complicate the issue.

12.1.2.1 Differential mode radiation

As is shown in section 10.2.1.1, the radiation efficiency of a small loop is proportional to the square of the frequency (+40dB/decade). This relationship holds good until the periphery of the loop approaches a quarter wavelength, for instance about 15cm in epoxy-glass pcb† at 250MHz, at which point the efficiency peaks. Superimposing this characteristic onto the harmonic envelope of a trapezoidal waveform shows that differential mode emissions (primarily due to current loops) will be roughly constant with frequency (Figure 12.5(a)) above a breakpoint determined by the risetime [11].

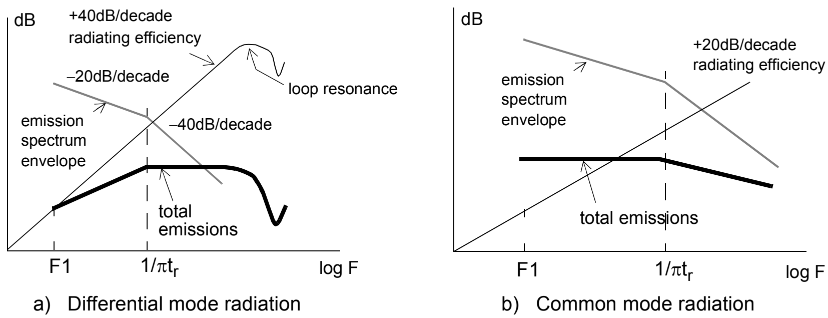


Figure 12.5 Emissions from digital trapezoid waves via different paths

The actual radiated emission envelope at 10m can be derived from equation (10.12) on page 235 provided the peak-to-peak current, risetime and fundamental frequency are known. The Fourier coefficient at the fundamental frequency F1 is 0.64, therefore the emission at F1 will be:

$$E = 20\log_{10} . [119 . 10^{-6} (f^2 . A . I_{pk})] \text{ dB}\mu\text{V/m} \qquad (12.1)$$

from which the +20dB/decade line to the breakpoint at $1/\pi t_r$ is drawn. In fact, by combining the known rise and fall times and transient output current capability for a given logic family with the trapezoid's Fourier spectrum at various fundamental frequencies, the maximum radiated emission can be calculated for different loop areas. If these figures are compared with the EN Class B radiated emissions limit (30dBμV/m at 10m up to 230MHz), a table of maximum allowable loop area for various logic families and clock frequencies can be derived. Table 12.1 shows such a table. The working for an example from this table (74ALS at 30MHz) is shown underneath.

The ΔI figure in Table 12.1 is the dynamic switching current that can be supplied by the device to charge or discharge the node capacitance. It is not the same as the steady-state output high or low load current capability. Some manufacturers' data will give these figures as part of the overall family characteristics. The same applies to t_r and t_f, which are typical figures, and are not the same as maximum propagation delay specifications.

Layout and construction implications

The implication of these figures is that for clock frequencies above 30MHz, or for fast logic families (AC, AS or F), a ground plane layout is essential as the loop area

† The effect of the epoxy-glass dielectric is to slow the wave propagation in the PCB and hence reduce the effective wavelength, by a factor proportional to $\sqrt{\varepsilon_r}$ if the track is on an inner layer; see Appendix D section D.5.3

Table 12.1 Differential mode emission: allowable loop area

Logic family	t_r/t_f ns	ΔI mA	Loop area cm^2 at clock frequency 4MHz	10MHz	30MHz	100MHz
4000B CMOS @ 5V	40	6	1000	400	–	–
74HC	6	20	45	18	6	–
74LS	6	50	18	7.2	2.4	–
74ALS	3.5	50	10	4	1.4	0.4
74AC	3	80	5.5	2.2	0.75	0.25
74F	3	80	5.5	2.2	0.75	0.25
74AS	1.4	120	2	0.8	0.3	0.15

Loop area for 30dBμV/m 30MHz–230MHz, 37dBμV/m 230–1000MHz at 10m

Working: take 74ALS family with F_{clk} = 30MHz as example. Worst case is at 150MHz (5th harmonic)

Fourier analysis of the source current using section D.7 on page 467 with $(t + t_r)/T = 0.5$, T = 33.3ns, t_r = 3.5ns and I = 50mA gives $I_{(5)}$, the current at the 5th harmonic, as 3.83mA.

From equation (10.12), for a field strength E of 30dBμV/m and $I_{(5)}$ at 150MHz as above, the allowable loop area A is **1.395cm^2** (rounded to 1.4 in the table).

restrictions cannot be met in any other way. Even this is insufficient if you are using fast logic at clock frequencies above 30MHz. The loop area introduced by the device package dimensions exceeds the allowed limit and extra measures (some form of shielding) are unavoidable. This information can be useful in the system definition stages of a design, when the extra costs inherent in choosing a higher clock speed (versus, for example, multiple processors running at lower speeds) can be estimated.

Table 12.1 applies to a single radiating loop. Usually, only a few tracks dominate the radiated emissions profile. Radiation from these loops can be added on a root mean square basis, so that for n similar loops the emission is proportional to $\sqrt{n}$. If the loops carry signals at different frequencies then their emissions will not add.

Do not make the mistake of thinking that if your circuit layout satisfies the conditions in Table 12.1 then your radiated emissions will be below the limit. Total radiation is frequently dominated by common mode emissions, as we are about to discover, and Table 12.1 only relates to differential mode emissions. But if the circuit does *not* satisfy Table 12.1, then extra precautions will definitely be needed.

12.1.2.2 Common mode radiation

Common mode radiation which is due mainly to cables and large metallic structures increases at a rate linearly proportional to frequency (ignoring resonances), as shown in equation (10.13) on page 237. There are two factors which make common mode coupling the major source of radiated emissions:

- cable radiation is much more effective than from a small loop, and so a smaller common mode current (of the order of microamps) is needed for the same field strength;
- cable resonance usually falls within the range 30–100MHz, and radiation is enhanced over that of the short cable model.

A similar calculation to that done for differential mode can be done for cable radiation on the basis of the model shown in Figure 12.6. This assumes that the cable is driven by a common mode voltage developed across a ground track which forms part of a logic circuit. The ground track carries the current ΔI which is separated into its frequency components by Fourier analysis, and this current then generates a noise voltage differential V_N of $\Delta I \cdot j\omega \cdot L$ between the ground reference (assumed to be at one end of the track) and the cable connection (assumed to be at the other). A factor of –20dB is allowed for lossy coupling to the ground reference; such coupling is often due to stray capacitance and structural resonances and is very difficult to quantify. At some frequencies the loss will be much greater, but at others (for instance if a series resonance is involved) it could be much less, approaching 0dB. The cable impedance is assumed to be a resistive 150Ω and constant with frequency – this is a value which represents the average of typical cables between their resonant extremes. PCB dimensions are assumed to be negligible compared to the cable dimensions.

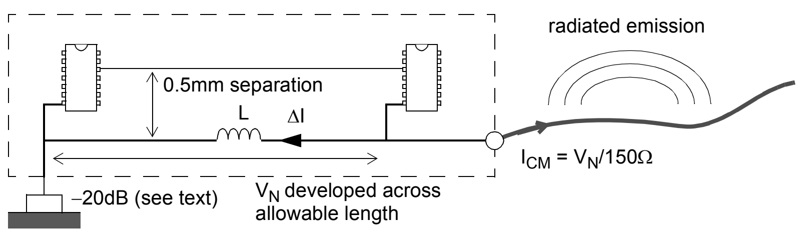


Figure 12.6 Common mode emission model

Track length implications

The inductance L is crucial to the level of noise that is emitted. In the model, it is calculated from the length of a 0.5mm wide track separated from its signal track by 0.5mm, so that mutual inductance cancellation reduces the overall inductance. Table 12.2 tabulates the resulting allowable track lengths versus clock frequency and logic family as before, for a radiated field strength corresponding to the EN Class B limits.

This model should not be taken too seriously for prediction purposes. Too many factors have been simplified: cable resonance and impedance variations with frequency and layout, track and circuit resonance and self-capacitance, and resonance and variability of the coupling path to ground have all been omitted. The purpose of the model is to demonstrate that logic circuit emissions are normally dominated by common mode factors. Common mode currents can be combated by:

- ensuring that logic currents do not flow between the ground reference point and the point of connection to external cables;
- filtering all cable interfaces to a "clean" ground as discussed in section 11.2.3.1;
- screening cables with the screen connection made to a "clean" ground;
- minimizing ground noise voltages by using low inductance ground layout, preferably involving a ground plane.

Table 12.2 shows that the maximum allowable track length for the higher frequencies and faster logic families is impracticable (fractions of a millimetre!). Therefore one or a combination of the above techniques will be essential to bring such circuits into

Table 12.2 Common mode emission: allowable track length

Logic family	t_r/t_f ns	ΔI mA	Track length cm at clock frequency 4MHz	10MHz	30MHz	100MHz
4000B CMOS @ 5V	40	6	180	75	–	–
74HC	6	20	8.5	3.2	1	–
74LS	6	50	3.25	1.3	0.45	–
74ALS	3.5	50	1.9	0.75	0.25	0.08
74AC	3	80	1.0	0.4	0.14	0.05
74F	3	80	1.0	0.4	0.14	0.05
74AS	1.4	120	0.4	0.15	0.05	–

Allowable track length for 30dBμV/m 30MHz–230MHz, 37dBμV/m 230–1000MHz at 10m; cable length = 1m; layout: parallel 0.5mm tracks 0.5mm apart (2.8nH/cm)

Working: take 74HC family with F_{clk} = 10MHz as an example. Worst case is at 90MHz (9th harmonic).

From equation (10.13), for a field strength E of 30dBμV/m and 1m cable length, I_{CM} must be 2.8μA.

From $V_N = I_{CM} \cdot 150$ and including 20dB coupling attenuation, V_N = 4.18mV.

Fourier analysis of the source current using section D.7 on page 467 with $(t + t_r)/T = 0.5$, T = 100ns, t_r = 6ns and I = 20mA gives $I_{(9)}$, the current at the 9th harmonic, as 0.826mA.

Now from $L = V_N/2 \cdot \pi \cdot f \cdot I_{(9)}$, the allowable inductance across which V_N will be developed at $I_{(9)}$ and 90MHz is 8.95nH which at 2.8nH/cm gives **3.19cm** allowed.

compliance. Clearly, if the ground reference point in Figure 12.6 is moved to be adjacent to the cable interface, no driving voltage is developed and the cable becomes benign. This is the purpose of the interface clean ground structure discussed in section 11.2.3.1. With this approach the common mode emissions are due only to common mode currents flowing directly in the PCB tracks. This is not to say that the effects of such currents are negligible – with high clock frequencies and unshielded products PCB common mode radiation can certainly present a real problem. The model can be modified to remove the cable and instead treat the PCB itself as the radiating structure. This has been explored in section 10.2.1.2.

12.1.2.3 Dealing with clock emissions

The main source of radiation in digital circuits is the processor clock (or clocks) and its harmonics. All the energy in these signals is concentrated at a few specific frequencies, with the result that clock signal levels are 10–20dB higher than the rest of the digital circuit radiation (see for example Figure 12.7, which shows emissions dominated by a 40MHz clock oscillator). Since the commercial radiated emissions standards do not distinguish between narrowband and broadband, these narrowband emissions should be minimized first, by proper layout, grounding and buffering of clock lines. Then pay attention to other broadband sources, especially data/address buses and backplanes, and video or high-speed data links.

It is common to find circuits being driven from clock signals whose transition times are substantially faster than is necessary. Certainly there are many high-performance circuits where clock timing is critical and transitions must be as fast as possible, but this is not universal. Where circuit constraints allow it, you should deliberately slow clock edges to minimize harmonic generation. This can be done in three ways: series

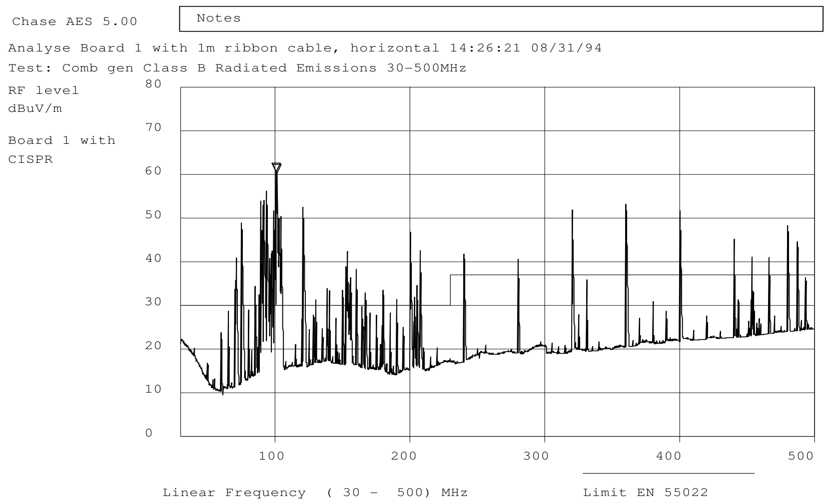


Figure 12.7 Typical emissions plot showing clock harmonics

impedance, parallel capacitance or by using a low-performance buffer. Figure 12.8 shows the first two of these. Generally, slugging the clock output with a parallel capacitor is undesirable because although it has the desired effect of reducing the dv/dt feeding into the clock line, it increases the capacitive loading on the driver and hence increases the di/dt drawn from its supply pins; the overall effect may be to worsen the emissions rather than improve them.

Stopper resistors

It is preferable to increase the series impedance of the driver output at the harmonic frequencies, and this can best be done with a low-value resistor or a small ferrite impeder in series with the output (Figure 12.8(b)). Low-loss inductors are less helpful as they tend to introduce ringing.

In fact, series resistors can be used all over a digital board, not just in clock lines; they are small enough (0603 or 0402 size, or in multiple arrays) to be used in every line of an entire bus. The effect of a series resistor of 22–47Ω at the output of each pin of a bus driver or parallel interface on an ASIC or FPGA, in concert with the distributed capacitance of the bus line being driven, is to damp the bus edge transitions as well as

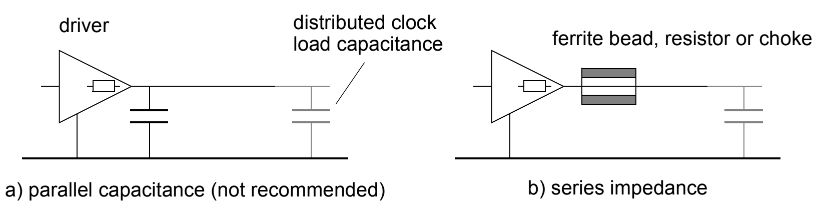


Figure 12.8 Controlling the clock edges

to reduce the V_{CC} spikes by isolating the driver output from the capacitance. All that is necessary is to ensure that the extra resistance does not cause a violation of timing constraints. You needn't restrict use of these resistors just to the expected high frequency lines; the effect of internal bond wire inductance is to infect the chip's own power rails with clock noise (see the section on ground bounce, 12.2.3) and this means that even low frequency output port drivers can be carriers of this noise away from the device. Series stopper resistors near the device output pins will prevent it from propagating far across the board.

Spread-spectrum clock generation

One possible alternative to damping the clock edges is a technique more often associated with radio transmission, known as spread-spectrum clock generation. In this technique the clock frequency itself is modulated by 0.5–2% by a waveform selected for the most even spectral spreading, which was patented in 1996 by Lexmark [79]. This results in a wider distribution of the spectral energy associated with each clock harmonic so that the level measured in a constant 120kHz bandwidth falls, it is claimed by between 10 and 20dB [96]. This is achieved without any extra effort in layout and without slowing the clock risetimes.

The clock frequency will exhibit some jitter and therefore the technique may be restricted in applications which need accurate clock timing such as telecoms or video products, but there will undoubtedly be many applications which can use it, and it may present itself as an attractive "quick fix" for remedial work. One effect of spreading both up and down in frequency is potentially to violate clock set-up and hold constraints – a higher instantaneous frequency giving tighter and perhaps unacceptable timing margins. This can be avoided by "down-spreading", that is, by modulating only downwards in frequency [57]. Another problem is caused by PLL (phase-lock-loop) frequency multiplication, which may not track the desired modulation waveform accurately, so that the expected best-case reduction in emissions does not materialize.

The technique does not reduce the total amount of radiated energy, and it has been compared to "getting rid of a cow-pat by stamping on it". To an EMC purist, of course, it should not be regarded as a substitute for proper design and layout in the first place.

Backplanes and daughter board connectors

Buses that drive several devices or backplanes which themselves drive several boards, carry much higher switching currents (because of the extra load capacitance) than circuits which are compact and/or lightly loaded. Products which incorporate a backplane are more prone to high radiated emissions. A high-speed backplane should always use a multilayer board with a ground plane, and daughter board connectors should include a ground pin adjacent to every high-speed clock, data or address pin (Figure 12.9). If this is impractical or too expensive, multiple distributed ground returns can be used to minimize loop areas, although this is at best a second-rate compromise and would be unacceptable if transmission line distribution were involved. The least significant data/address bit usually has the highest frequency component of a bus and should be run closest to its ground return. Clock distribution tracks must *always* have an adjacent ground return. The capacitive load on the clock signal at each daughter board should be minimized, by having a buffer on the board for local clock distribution.

Low-voltage logic families

The move from 5V V_{CC} to 3.3V and below has been driven by the need to reduce the speed-power product of digital circuits, so that more performance can be squeezed out

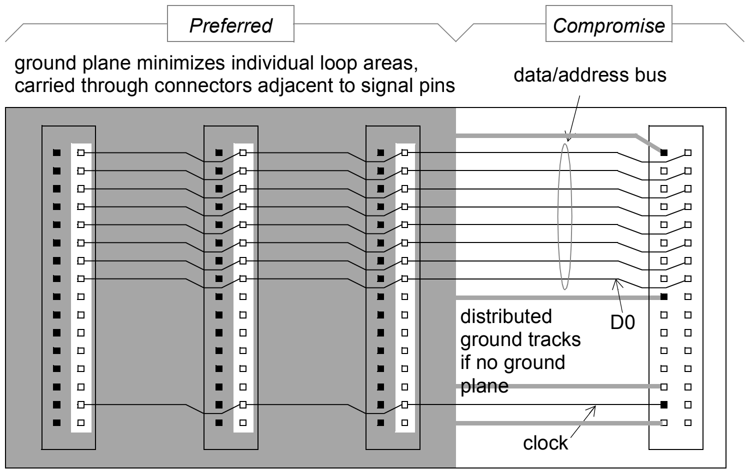


Figure 12.9 Backplane layout

of a lower power consumption. The consequences for EMC performance are equivocal. On the one hand, reducing the voltage swing of external clocks and data makes these signals in themselves less aggressive for RF emissions. But on the other, lower logic voltage thresholds mean potentially lower RF and transient immunity levels.

In fact, the trend to lower supply voltages has merely meant that circuits have become faster and more complex for the same power consumption. Current drain from the supply rails has not reduced and can sometimes be greater, and this is associated with increased supply ΔI noise which is a serious RF emissions hazard itself. In addition, the need for multiple low-voltage supplies (often 3V3, 2V5 and 1V8 are all needed for a single VLSI device) causes headaches for PCB layout when several power planes need to be referenced to the same 0V plane. All in all, moving to low-voltage logic has not made the EMC designer's life any easier.

12.1.2.4 Ringing on transmission lines

If you transmit data or clocks down long lines, these must be terminated to prevent ringing. Ringing is generated on the transitions of digital signals when a portion of the signal is reflected back down the line due to a mismatch between the line impedance and the terminating impedance. A similar mismatch at the driving end will re-reflect a further portion towards the receiver, which gets reflected back again; and so on. Severe ringing will affect the data transfer, by causing spurious transitions, if it exceeds the device's input noise margin.

Aside from its effect on noise margins, ringing may also be a source of radiated interference in its own right. The amplitude of the ringing depends on the degree of mismatch at either end of the line while its frequency depends on the electrical length of the line (Figure 12.10). The question is, at what length does a PCB track or cable need to be treated as a transmission line? This can be determined by whether the logic transition has been completed before the reflected part has made the round trip. A digital driver-receiver combination should be analysed in terms of its transmission line behaviour if:

$2 \times t_{PD} \times \text{line length} > \quad \text{transition time}$ (12.2)

where t_{PD} is the line propagation delay in ns per unit length [100], which depends on the dielectric constant of the board material and can be calculated from section D.5.3 (Appendix D)

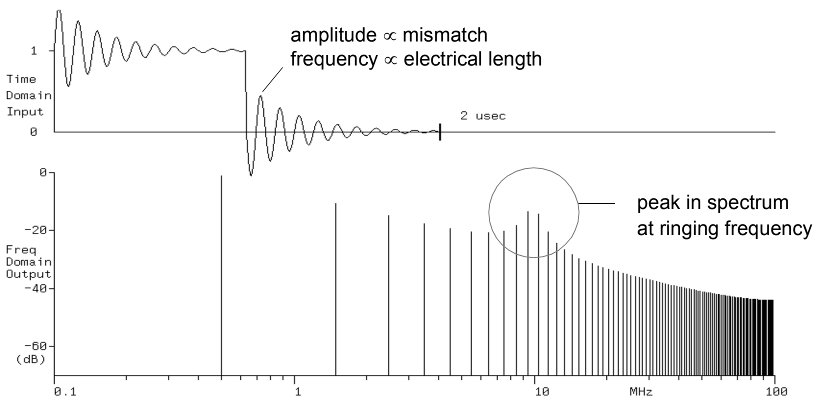


Figure 12.10 Ringing due to a mismatched transmission line

This means matching the track's characteristic impedance to the source and load impedances, and may require extra components to terminate the line at either end. Most digital circuit data and application handbooks (e.g. [10],[15]) include advice and formulae for designing transmission line systems in fast logic. Table 12.3 is included as an aid to deciding whether the particular circuit you are concerned with should incorporate transmission line principles.

Table 12.3 Critical transmission line length

Logic family	t_r/t_f ns	Critical line length
4000B CMOS @ 5V	40	3 m
74HC	6	45 cm
74LS	6	45 cm
74ALS	3.5	26 cm
74AC	3	22.5 cm
74F	3	22.5 cm
74AS	1.4	10.5 cm

Line length calculated for dielectric constant = 4.5 (FR4 epoxy glass), t_{PD} = 0.067ns/cm

12.1.3 Digital circuit decoupling

No matter how good the V_{CC} and ground connections are, they will introduce an impedance which will create switching noise from the transient switching currents taken from the V_{CC} pins, as shown in Figure 12.4. The purpose of a decoupling capacitor is to maintain a low dynamic impedance from the individual IC supply voltage to ground. This minimizes the local supply voltage droop when a fast current pulse is taken from it, and more importantly it minimizes the lengths of track which carry high di/dt currents.

12.1.3.1 Component placement

Placement is critical; the capacitor must be tracked close to the circuit it is decoupling (see Figure 12.11). "Close" in this context means less than half an inch for fast logic such as AS-TTL, AC or ECL, especially when high-current devices such as bus drivers are involved; if all outputs on an octal bus buffer are heavily loaded and the output changes from $\#FF_H$ to $\#00_H$ or vice-versa, a current pulse which may exceed an amp passes through the supply pins. For low-current, slow devices such as 4000B-series CMOS the requirement is more relaxed; but a typical ASIC or FPGA clocked at hundreds of MHz needs a capacitor adjacent to every pin, most conveniently mounted on the opposite side of the board under the package.

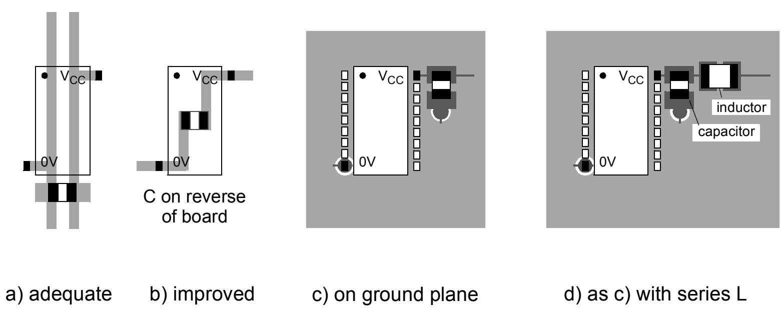


Figure 12.11 Decoupling capacitor positioning

When decoupling capacitors are used exclusively with power and ground planes, as is usual for high-performance multilayer designs, the positioning of individual capacitors is less important, although considerations of current distribution still have an effect as discussed later (page 305). It is still normal to use at least one capacitor next to each power pin, but the effect of the planes is to act as a transmission line that interconnects all such decouplers in parallel, with the minimum of inductance. Therefore rather than primarily decoupling individual devices, capacitors on a power/ground plane network are effectively reducing the impedance between the planes over as wide a bandwidth as possible. Thus the injected high frequency currents shown in Figure 12.4 create the lowest possible noise voltages.

To achieve this each decoupler should present a low impedance especially at high frequency where the planes' dimensions become a significant fraction of a wavelength. Since this is normally in the region of a few hundred MHz for typical sized boards, the capacitor is operating mainly above self resonance, and its actual contribution is to add its small package inductance in parallel with all the others on the plane. So, as we are about to see, the component's inductance is its most important characteristic. It will have lowest impedance at its self resonant frequency, which will be determined by its value and its mounted self-inductance: $F_{res} = 1/(2\pi \cdot \sqrt{L_m \cdot C})$, where L_m is the sum of the package inductance and the PCB mounting inductance due to pads and vias to the planes. For instance, a 2200pF 0402 capacitor with about 1.5nH of total mounted inductance will have a SRF of 88MHz. At this frequency, it should be closer to the device it is decoupling than roughly one tenth of a quarter wavelength [25] in order that the propagation delay across the planes in the x-y direction should not effectively detach the capacitor from the device pin: in FR4 glass fibre PCB material, this example corresponds to $(300/88)/(40\cdot\sqrt{4.5}) = 4$cm. Generally, then, capacitors can be placed around the periphery of a device rather than directly underneath it with little compromise. Larger value capacitors with lower SRFs are even less critical.

12.1.3.2 Component selection

This leads to the view that the crucial factor when selecting capacitor type for high-speed logic decoupling is package or lead inductance rather than absolute value. Minimum inductance offers a low impedance to fast pulses. Multilayer chip, or if leaded is necessary, small disk ceramics are preferred. The overall inductance of each connection is the sum of both lead and track inductances.

Capacitance value for low frequencies

The minimum required value for the total parallel capacitance on a particular power rail depends on the specification for tolerable supply voltage deviation, and can be calculated from equation (12.3):

$$C_{min} = \Delta I \cdot \Delta t / \Delta V \qquad (12.3)$$

ΔI and Δt can to a first order be taken from the figures in Table 12.1 and Table 12.2 while ΔV depends on your judgement of permissible supply voltage droop at the capacitor. Typically on a 5V supply a power rail droop of 0.25V is reasonable; lower voltage supplies can tolerate less droop. For an octal buffer taking 50mA per output and switching in 6ns, the required capacitance for $\Delta V = 0.25$V is 9.6nF. For smaller devices and faster switching times, less capacitance is required, and often the optimum capacitance value is as low as 1000pF. Then, the question becomes how many discrete capacitors of what value do you need to provide this total with the least overall inductance; invariably, many small packages are better than a few large ones.

Small tantalum capacitors are to be preferred for bulk decoupling because due to their non-wound construction their self-inductance is very much less than for an aluminium electrolytic of the same value. Their equivalent series resistance is fairly high and this has the beneficial effect of minimizing impedance peaks from their self-inductance resonating with the lower value capacitors in parallel on the power rail.

PCB layout for high frequencies

Assuming you will be using small chip multilayer ceramics, with a package size of 0402, 0603 or at a pinch 0805, the determining factor for the inductance of the component is its PCB layout. This is illustrated in Figure 12.12. The two pads of the capacitor are taken through vias to the relevant power and ground planes; the separation of the vias, the dimensions of the capacitor and the height of each via between the surface pad and the buried plane determine the overall loop area, and hence the inductance contributed by the layout. If the board costs are to stay low then the vias should not be placed in the pads, so layout 'X' is the best option.

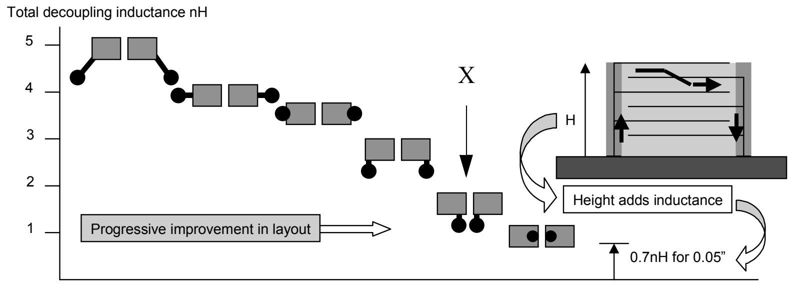


Figure 12.12 Decoupling capacitor track layout for minimum inductance

12.1.3.3 Use of series inductors

Liberally sprinkling a board with decoupling capacitors of itself is not necessarily the optimum decoupling regime. This is particularly the case where one or a few ICs contribute the majority of the offending di/dt through its or their power supply pins. A common example of this situation is the use of a single-chip microprocessor with no other digital components; the main emissions source is the clock harmonic current drawn through the V_{CC} pin. When one decoupling capacitor is positioned close to the processor package and others are positioned elsewhere on the board, intuitively it would seem that the noise currents would flow through and be decoupled by the adjacent capacitor. But in practice this is not so at all frequencies; the inductance of the interconnecting tracks forms a series tuned circuit with the remote decoupling capacitors, and at the resonant frequencies the noise currents flowing towards the remote capacitors are actually greater than if these capacitors were not present [58]. This in turn causes worse emissions at these frequencies when the capacitors are added.

This effect can be seen through analysis of the equivalent circuit (with arbitrary but representative circuit values, see Figure 12.13). The interference source is modelled as a frequency-dependent voltage generator feeding the decoupling and power rail network through a source impedance of 100Ω. The "far end" of the power supply is modelled as a 10μF capacitor in parallel with 1Ω. C1 is the local decoupling capacitor, C2 is the remote one placed across another IC, represented in the model by 100Ω. (This value of resistance may or may not represent the RF impedance of an IC; it has little effect on the analysis.) Track and parasitic inductances are included as shown. The two results graphs show the decoupling effect represented as a ratio of I_{IN} to I_1 or I_2 in the appropriate length of track, l_1 or l_2. The left-hand graph shows the decoupling effect in the track between C1 and C2; the right-hand graph shows the effect downstream of C2. The solid line shows the decoupling due to the equivalent circuit as drawn, with the dotted lines representing various changes.

Analysing the results of the model

Firstly, the resonant peaks in both graphs' solid lines giving an *increase* in track current (dB > 0) at the lower frequencies are evident. This is due to resonances of L_{+1} and L_{+2} with C1 and C2. Above 100MHz all the curves are flat, the actual decoupling being determined by the ratio of inductances since all capacitors are operating above self-resonance. The action happens mostly in the 2–50MHz frequency range.

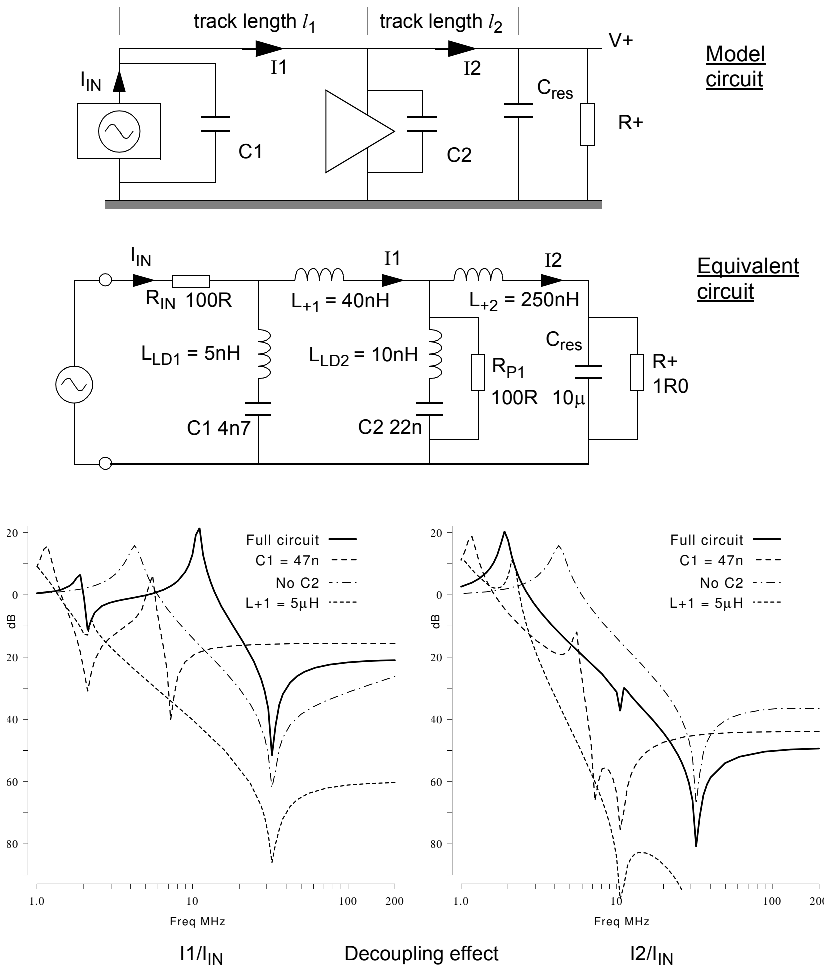


Figure 12.13 Analysing the decoupling equivalent circuit

Increasing C1 from 4n7 to 47n improves the decoupling in the mid-frequencies at the expense of shifting the resonances lower – which is not necessarily undesirable. Because it is a larger capacitor, the self-inductance L_{LD1} has been increased to 10nH and this worsens the decoupling above 20MHz.

Removing C2, although worsening the decoupling of track l_2, in fact improves the decoupling of l_1 above 7MHz, because the impedance looking into l_1 has increased and therefore less noise current flows into it. The impact of this modification in real life will depend on the relative significance of l_1 and l_2 in actually coupling interference out of the PCB.

The most substantial improvement occurs when the impedance looking into l_1 is significantly increased. This can only be achieved by inserting a discrete inductor; in the analysis the inductor value was chosen to be 5µH, but this is fairly arbitrary. You can see that the improvement over most of the frequency range is in the order of 40dB. This technique is shown schematically in Figure 12.11(d). As a general design rule, you should plan to include such series inductors (available as chip ferrite components, taking up negligible extra space at low cost) in the $+V_{CC}$ feed to every IC which is expected to contribute substantially to the noise pollution of the supply lines.

Power planes

The addition of series inductors is advisable even where both power and ground planes are used. In fact segmentation of power planes, as discussed in section 11.2.2.7, requires their use to partition one plane segment from another.

If the board has four or more layers and includes a power plane, the inductance of the coupled planes is so low that resonances are not analysed according to the simple model above. However, the placement of decoupling capacitors on a ground-and-power plane board is still significant. Imagine the situation shown in Figure 12.14: an IC is mounted somewhere on a four-layer board and it injects I_{CC} noise into the power and ground planes (A). Elsewhere on the board is a decoupling capacitor.

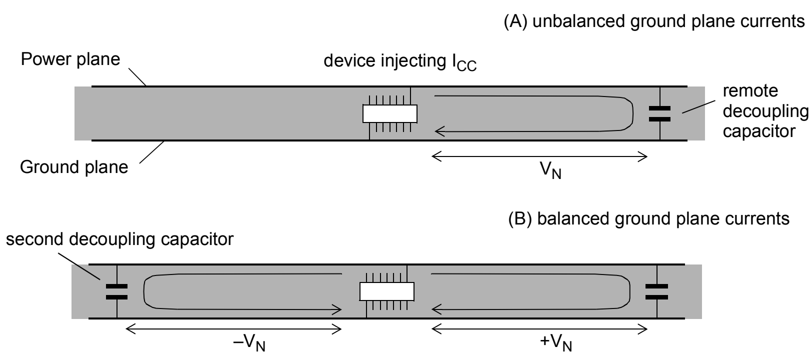


Figure 12.14 Unbalanced power/ground plane currents

The current that flows between the two will create a noise voltage across the impedance of the plane and this noise voltage will create common mode emissions, as discussed in section 10.2.1, either by direct radiation from the board or by coupling to connected cables. If, now, another capacitor is placed in a position on the board exactly opposite from the first (B) with respect to the device, then a counterbalancing noise voltage $-V_N$ is produced. The effect of this will be to cancel the first $+V_N$ and to reduce the board's emissions.

Emissions from this cause can either be controlled by reducing the I_{CC} noise developed by a device, or by careful layout of decoupling capacitors to balance the board. The latter is impractical for all but the simplest of circuits, but distributing many capacitors across the board, particularly close to high-noise devices, will control the resulting currents as far as possible. As discussed earlier, keeping the smallest capacitors within one tenth of a quarter wavelength (at their self-resonant frequency) from the pin they are decoupling will minimize the distance that the decoupling currents must travel through the power planes [25]. The I_{CC} noise from individual devices can be reduced by applying ferrite impeders in each supply connection as shown in Figure

12.11(d). It is also helpful to group together devices that generate high I_{CC} noise so that their interconnections are short, they can all use a small segmented power plane, and to keep them away from the edge of the board.

12.1.4 Analogue circuits: emissions

In general analogue circuits do not exhibit the high di/dt and fast risetimes that characterize digital circuits, and are therefore less responsible for excessive emissions. Analogue circuits which deliberately generate high frequency signals (remembering that the emissions regulatory regime currently begins at 150kHz, and in some cases lower than this) need to follow the same layout, decoupling and grounding rules as already outlined. It is also possible for low frequency analogue circuits to operate unintentionally outside their design bandwidth.

12.1.4.1 Instability

Analogue amplifier circuits may oscillate in the MHz region and thereby cause interference for a number of reasons:

- feedback-loop instability;
- poor decoupling;
- output stage instability.

Capacitive coupling due to poor layout and common-impedance coupling are also sources of oscillation. Any prototype amplifier circuit should be checked for HF instability, whatever its nominal bandwidth, in its final configuration and layout. Feedback instability is due to too much feedback near the unity-gain frequency, where the amplifier's phase margin is approaching a critical value. It may be linked with incorrect compensation of an uncompensated op-amp.

Even if the circuit doesn't actually oscillate, near-instability is a potential problem for RF immunity as well: if the amplifier circuit is configured so that its stability margin is reduced at frequencies of a few MHz, then RF coupled into the circuit at around these frequencies will be much more likely to cause susceptibility effects (see section 12.2.6.4).

12.1.4.2 Decoupling

Power supply rejection ratio falls with increasing frequency, and power supply coupling to the input at high frequencies can be significant in wideband circuits. The higher the impedance of the power rails, the more unwanted coupling exists.

This is cured by good decoupling at the power supply pins, but typical 0.01–0.1μF decoupling capacitors may resonate with the parasitic inductance of long power leads in the MHz region, so decoupling-related instability problems may show up in this range. Paralleling a low-value capacitor with a 1–10μF electrolytic capacitor will drop the resonant frequency and stray circuit Q to a manageable level. The electrolytic's series inductance could resonate with the ceramic capacitor and actually worsen the situation, although the losses of a few ohms in series with the capacitor usually prevent this, which is a good argument for not using low-loss 1–10μF ceramic capacitors in this case. The input stages of multi-stage high gain amplifiers may need additional resistance or a ferrite bead in series with each stage's supply to improve decoupling from the power rails.

12.1.4.3 Output stage instability

Capacitive loads cause a phase lag in the output voltage by acting in combination with the op-amp's open-loop output resistance (Figure 12.15). This increased phase shift reduces the phase margin of a feedback circuit, possibly by enough to cause oscillation. A typical capacitive load, often invisible to the designer because it is not treated as a component, is a length of unterminated coaxial cable. Until the length starts to approach a quarter-wavelength at the frequency of interest, coax looks like a capacitor: for instance, 10m of the popular RG58C/U 50Ω type will be about 1000pF. Other capacitive loads may be found within a system, such as a sample-and-hold circuit, but the cable load is of special interest for EMC since it couples any RF oscillations both into and out of the circuit. To cure output instability, decouple the capacitance from the output with a low-value series resistor, and add high frequency feedback with a small direct feedback capacitor C_F which compensates for the phase lag caused by C_L [31]. When the critical frequency is high a ferrite chip is an acceptable substitute for R_S.

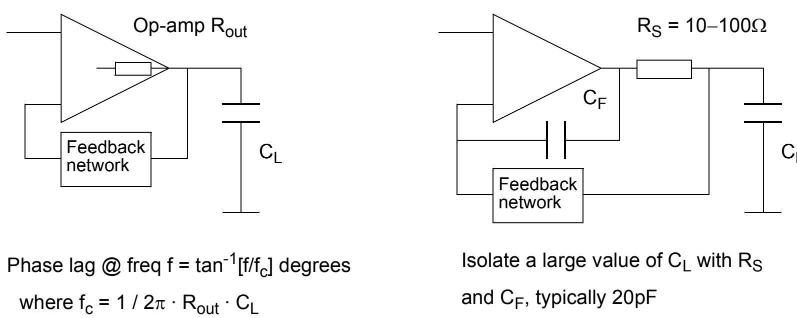


Figure 12.15 Instability due to capacitive loads

12.1.5 The switching power supply

Power switching circuits of all sorts present extreme difficulties in containing generated interference [144]. Typical switching frequencies of 50–500kHz and their harmonics can be emitted by both differential and common mode conducted and radiated mechanisms. Switching waveform asymmetry normally ensures that, except at the lowest orders, both odd and even harmonics are present. A further cause of broadband noise may be due to reverse recovery switching of the input rectifier diodes.

If the fundamental frequency is stable and well controlled then a spectrum of narrowband emissions is produced which can extend beyond 30MHz when the waveform transition times are fast. A measurement bandwidth of 9kHz means that individual harmonics can be distinguished if the fundamental frequency is greater than about 20kHz. Self-oscillating converters will normally show drift or modulation due to input or output ripple, which has the effect of broadening individual harmonic lines so that an emission "envelope", rather than narrowband harmonics, is measured.

Figure 12.16 shows a typical direct-off-line switching supply with the major emission paths marked; topologies may differ, or the transformer may be replaced by an inductor but the fundamental interference mechanisms are common to all designs.

12.1.5.1 Radiation from a high di/dt loop

Magnetic field radiation from a loop which is carrying a high di/dt can be minimized by reducing the loop area or by reducing di/dt. With low output voltages, the output rectifier and smoothing circuit may be a greater culprit in this respect than the input circuit. Loop area is a function of layout and physical component dimensions (see section 11.2.1). di/dt is a trade-off against switching frequency and power losses in the switch. It can to some extent be controlled by slowing the rate-of-rise of the drive waveform to the switch.

Unfortunately, the trend towards minimizing power losses and increasing frequencies goes directly against the requirements for low EMI. The lowest di/dt for a given frequency is given by a sinewave: sinusoidal converters (such as the series resonant converter, see [108]) or other variants with reduced risetimes have reduced EM emissions.

Magnetic component construction and layout

As explained in section 10.3.4.1, screening will have little effect on the magnetic field radiation due to this current loop, although it will reduce the associated electric field. The transformer (or inductor) core should be in the form of a closed magnetic circuit in order to restrict magnetic radiation from this source. A toroid is the optimum from this point of view, but may not be practical because of winding difficulties or power losses; if you use a gapped core such as the popular E-core type, the gap should be directly underneath the windings since the greatest magnetic leakage flux is to be found around the gap.

Fast risetimes together with high currents can extend direct radiation both from the switching magnetics and from the circuit structure to above 30MHz, making it a threat for the radiated emissions test. Additionally, such radiation can couple into the output

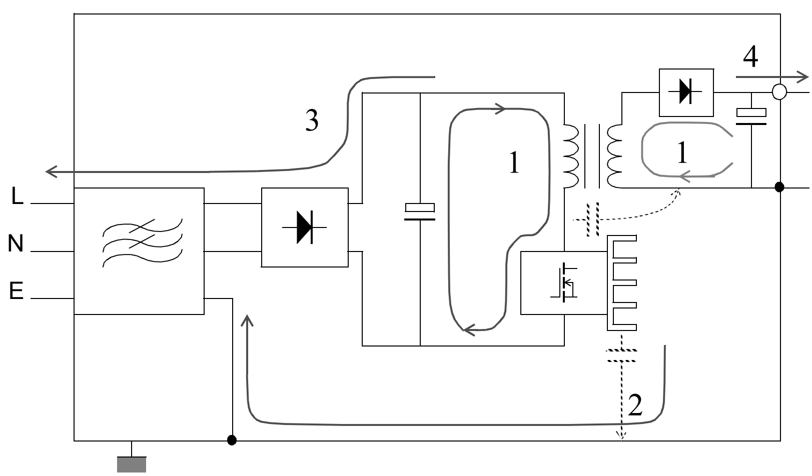


Figure 12.16 Switching supply emission paths

or mains leads and be responsible for conducted interference (over 150kHz to 30MHz) if the overall power supply layout is poor. You should always keep any wiring which leaves the enclosure well away from the transformer or inductor. In the same vein, coupling between the various inductors in the supply needs to be avoided. Stray leakage flux from the transformer could couple with the core of the input choke, which would dramatically reduce the attenuation of the input filter of which it is a part. Both magnetic and electric field coupling should be considered.

12.1.5.2 Capacitive coupling to ground

High dv/dt at the switching point (the collector or drain of the switching transistor) will couple capacitively to ground and create common mode interference currents. The solution is to minimize dv/dt, and minimize coupling capacitance or provide a preferential route for the capacitive currents through appropriate screening (Figure 12.17).

Dv/dt is reduced by a snubber and by keeping a low transformer leakage inductance and di/dt. These objectives are also desirable, if not essential, to minimize stress on the switching device, although they increase power losses. The snubber capacitor is calculated to allow a defined dv/dt with the maximum load as reflected through the transformer; the series resistor must be included to limit the discharge current through the switching device when it switches on. You can if necessary include a diode in parallel with the resistor to allow a higher resistor value and hence lower switching device ratings.

Capacitive screening

Capacitive coupling is reduced by providing appropriate electrostatic screens, particularly in the transformer and on the device heatsink; in mains SMPSs these are the two points which carry the highest dv/dt. Note the proper connection of the screen: to either supply rail, which allows circulating currents to return to their source, not to the safety earth. On the transformer, an external foil screen connected to 0V will also reduce coupling of the high dv/dt on the outside of the winding to other parts of the circuit.

Even if the transformer is not screened, its construction can aid or hinder capacitive coupling from primary to secondary, and from the whole component to others around it (Figure 12.17(b)). Separating the windings onto different bobbins reduces their coupling capacitance but increases leakage inductance. Coupling is greatest between nodes of high dv/dt; so the end of the winding which is connected to V_{CC} or ground can partially screen the rest of the winding in a multi-layer design. Conversely, if you don't take care with this aspect, the transformer can be the major contributor to the circuit's emissions.

Switching device heatsinks can also cause problems. Particularly with high-voltage transistors, the switching node (drain of a MOSFET or collector of an IGBT) carries the highest dv/dt but is closely coupled to the heatsink tab: indeed with a non-isolated device it *is* the heatsink tab. Even with iso-tab devices there is a significant capacitance within the package between the semiconductor and the tab. This means that the one structure that you should never bolt the heatsink to is the equipment chassis or case, since this references the switching noise directly to the ground node. Unfortunately, that is usually the most thermally desirable place for it.

If at all possible, the metalwork of the heatsink should be directly connected to the 0V rail of the circuit that uses it. When this is impossible, for thermal or safety reasons, there should be an electrostatic screen between the device and the heatsink, connected

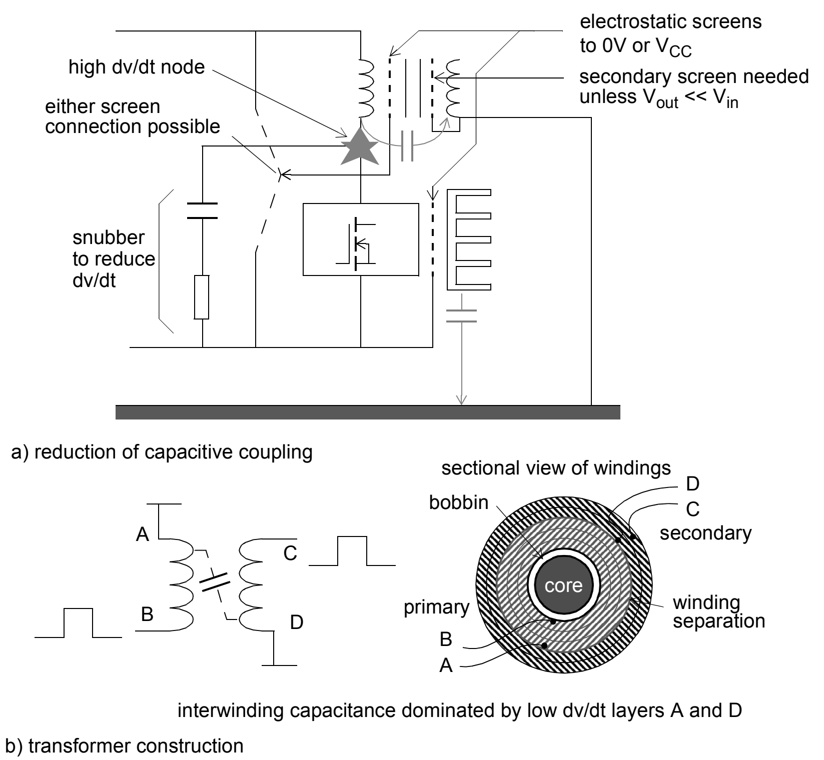


Figure 12.17 Common mode capacitive coupling

to this 0V; thermal pads with integral foil screens are available for this purpose. Where this is also impossible, aim for the minimum achievable coupling capacitance between the switching node and the heatsink and expect to have to work harder at (and pay more for) your filter design.

Physical separation of parts carrying high dv/dt is desirable [144] although hard to arrange in compact products. Extra electric field screening to 0V of the offending component(s) with small formed tinplate parts is an alternative.

The 20MHz peak

When you compare conducted emissions plots from many different types of product using switchmode mains power supplies, a striking similarity emerges – the existence of a peak in the emissions profile in the region of 5–25MHz, usually due to high order harmonics of the switching circuit [141]. If the level of the peak is below the limit then no further effort is needed, but if it is higher there usually follows a protracted period of remedial work, much of which appears to have no useful effect. Why is there this hump, and why is it so hard to get rid of?

The answer is best illustrated by discussing the overall testing equivalent circuit. There are four significant parts to this circuit (Figure 12.18(a)): the equipment under test including a mains filter, its mains lead, the LISN, and the ground plane. The EUT internal circuit is represented by a common mode noise source V_N referred to the chassis and capacitively feeding the common mode equivalent of the mains filter. This capacitance C_{CM} as discussed above is the main coupling path in most cases. Good circuit design and layout seeks to minimize its value, since it forms a potential divider with C_Y, the combined phase-to-earth capacitance of the mains filter. In most portable apparatus C_Y's maximum value is set by safety earth leakage current limits.

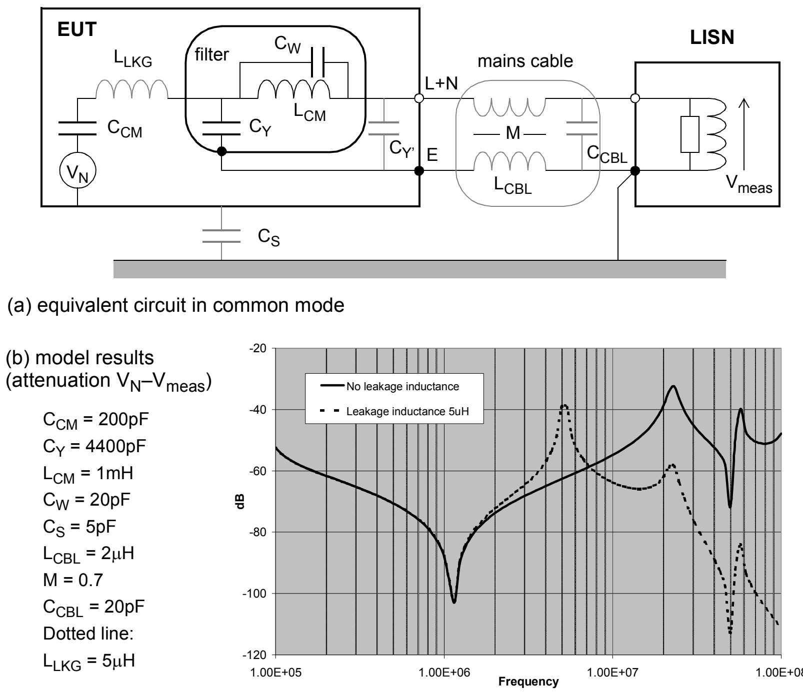


Figure 12.18 Analysis of the high frequency emissions peak

All but the simplest filters will include a series common mode choke L_{CM}. In this equivalent circuit a single winding is shown, since live and neutral are regarded as identical and with the conventional double-wound compensating choke the two windings reinforce each other. An important extra parasitic component is the winding capacitance C_W which appears across the windings and limits the HF performance of the choke.

Provided that it is electrically short (less than a tenth of a wavelength at the highest frequency, i.e. 1m for 30MHz) the transmission line effects of the mains cable can be ignored and the cable is represented by its lumped equivalent. This is dominated by the self-inductance due to the cable length, modified by mutual inductance and capacitance between the conductors – shown by capacitor C_{CBL} and a transformer of self-inductance L_{CBL} and mutual coupling M.

The EUT capacitance to the ground plane C_S is also part of the overall circuit. For EUTs with no safety earth it is a crucial part of the common mode coupling path. Any safety earth connection appears in parallel with this capacitance.

With this description of the equivalent circuit, it is easy to see that there are several possibilities for resonances to occur, and the main problem is determining which components are critical. The easiest way to do this is to model the circuit. The results of the modelling (Figure 12.18(b), solid line) show exactly the sort of hump in the transfer function from V_N to the LISN measurement point that is expected from the measurements. The self-capacitance of the choke is in fact resonating with the inductance of the mains cable, which of course is affected by cable length and the mutual coupling of the wires within it. The effect of this resonance can be substantially reduced by placing a subsidiary Y-capacitor of 100–470pF on the mains cable side of the common mode choke, $C_{Y'}$ in Figure 12.18(a). This lowers the frequency of the resonance and reduces its amplitude.

Another more difficult problem occurs when the common mode coupling path *within* the EUT includes some inductance. This is again quite usual with switched-mode power supplies when the inductance in question is the leakage inductance of the switching transformer. The leakage inductance L_{LKG} appears in series with C_{CM} as shown in Figure 12.18(a) and the resonant frequency of this combination is typically in the 5–10MHz range. Figure 12.18(b), dotted line, shows the effect of this leakage inductance resonance. Depending on the relative values of the stray capacitances and inductances, the resonant peaks may be separate as shown or may overlap and reinforce each other.

Because C_{CM} is also largely determined by the interwinding capacitance of the transformer, the resonance is impossible to shift without a complete redesign of the transformer – usually not an option by the time EMC compliance testing of an established design is being carried out. Deliberately increasing the leakage inductance – although undesirable for other reasons – may have a beneficial effect on the amplitude of the coupling peak. An alternative is to include a small extra C-M choke before C_Y in the mains filter.

12.1.5.3 Differential mode interference

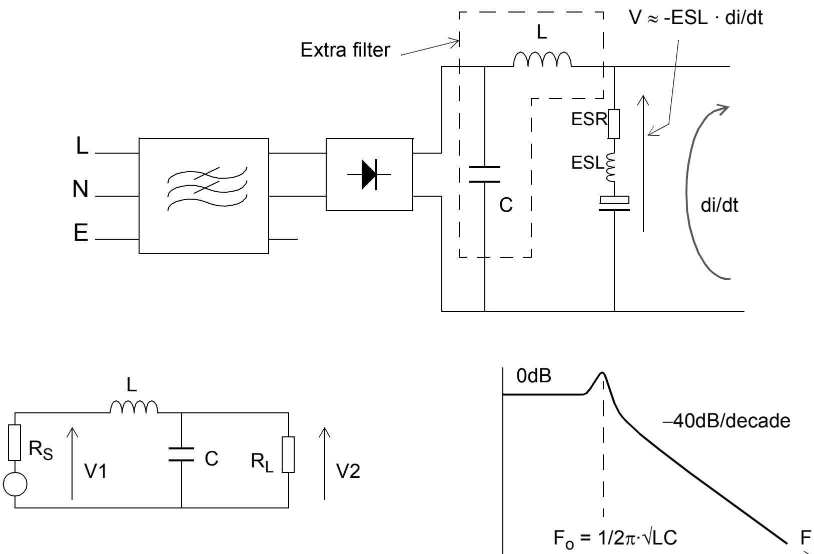
Differential mode interference is caused by the voltage developed across the finite impedance of the input reservoir capacitor with high di/dt due to the switched power. It is often the dominant interference source for the lower switching harmonics. Choosing a capacitor with low equivalent series impedance (ESL and ESR) will improve matters, but it is impossible to obtain a low enough impedance in a practical capacitor to make generated noise negligible.

Extra series inductance and parallel capacitance on the input side will attenuate the voltage passed to the input terminals. A capacitor on its own will be ineffective at low frequencies because of the low source impedance. Series inductors of more than a few tens of microhenries are difficult to realize at high DC currents (remembering that the inductor must not saturate at the peak ripple current, which is much higher than the DC average current), and multiple sections with smaller inductors may be more effective than a single section. When several parallel reservoir capacitors are used, one of these may be separated from the others by the series inductor; this will have little effect on the overall reservoir but will offer a large attenuation to the higher frequency harmonics at little extra cost.

Figure 12.19 demonstrates filtering arrangements. The LC network may also be placed on the input side of the rectifier. This will have the advantage of attenuating broadband noise caused by the rectifier diodes switching at the line frequency. The mains input filter itself (see section 13.2.3) will offer some differential mode rejection.

It is also possible to choose switching converter topologies with input inductors such as the Cúk circuit [66] which reduce, but do not eliminate (because of stray capacitance across the inductor), fast di/dt transitions in the input and/or output waveforms.

When you are testing the performance of a differential mode filter, be sure always to check it at the maximum operating input power. Not only do the higher switching currents generate more noise, but the peak mains input current may drive the filter inductor(s) into saturation and make it ineffective.



Insertion loss of simple LC filter:

$$|V1/V2| = |1 - \omega^2 LC\cdot(R_L/(R_S + R_L)) + j\omega\cdot((CR_LR_S + L)/(R_S + R_L))|$$

For low R_S and high R_L this reduces to

$$|V1/V2| = |1 - \omega^2 LC|$$

Standard mains impedance R_L is approximated by 100Ω//100μH differentially, which approaches 100Ω above 300kHz. Note resonance of LC at low frequency gives insertion gain

Figure 12.19 Differential mode filtering

12.1.5.4 Output noise

Switching spikes are a feature of the DC output of all switching supplies, partly because of the finite impedance of the output reservoir and partly due to coupling across the isolating transformer interwinding capacitance. Such spikes are conducted out of the unit on the output lines in both differential and common mode, and may re-radiate onto other leads or be coupled to the ground connection. A low-ESL reservoir capacitor is preferable, but good differential mode suppression can be obtained, as with the input, with a high frequency L-section filter. 20–40dB is obtainable with a ferrite bead and 0.1μF capacitor above 1MHz. Common mode spikes will be unaffected by adding a filter, and this is a good way to diagnose which mode of interference is being generated. To remove common mode interference, place a capacitor of 1–10nF between the primary and secondary 0V nodes: this acts as a voltage divider against the transformer's interwinding capacitance, but its applicability is limited by the isolation voltage and leakage capacitance requirements you face from input to output.

The abrupt reverse recovery characteristic of the output rectifier diode(s) can create extra high frequency ringing and transients. When the current is switched from conduction to blocking through a conventional diode there is a short period during which the forward current continues to flow, which causes undesirable extra power dissipation in the diode. Fast recovery diodes are especially designed to minimize the period during which this happens, to allow higher switching speeds. Unfortunately, although these devices don't take long to recover, when they do, there is a very fast switch-off of the remanent current – in other words, a high di/dt. This excites the resonant circuits created by the transformer secondary leakage inductance along with stray capacitance, leading to high levels of emission at the resonant frequencies, modulated by the switching frequency. These can be attenuated by choosing soft recovery diodes or by paralleling the diodes with an RC snubber.

12.1.6 Other power switching circuits

The SMPS is not the only source of RF emissions, although because of its ubiquity it is the one most frequently encountered in the mains conducted emissions test. Any circuit that switches significant power at frequencies from tens to several hundreds of kHz is also capable of serious RF emissions problems, such as:

- variable speed motor drives;
- Class D audio power amplifiers;
- high-efficiency lighting converters;
- phase angle lighting and heating controllers.

The underlying purpose of all these circuits is to switch power as fast as possible so that minimum power is lost in the switching device, usually a power MOSFET, IGBT or triac. As we have seen, the faster the switching the greater the content of high frequency harmonics in the switched current and voltage waveforms.

Invariably in the design of these products you must take similar precautions to those outlined above for power supplies. Both the differential and common mode coupling paths must be addressed; nodes of high dv/dt must be identified and designed to minimize capacitive coupling, and loops of high di/dt must be identified and designed to minimize inductive coupling. PCB track layout must take into account the threat of common impedance coupling and you will often need to implement single point grounding schemes to deal with this. Both differential and common mode filtering will always be needed and such filters are often a significant cost, size and weight overhead. A generalized equivalent circuit for any of these products is as shown in Figure 12.20, from which you can see that although the main switching circuit may be coupled to the output of the device through filters, there is still potential for a substantial interference signal to be radiated from the output leads or conducted back through the mains.

12.2 Design for immunity

Because the microprocessor is a state machine, processor-based circuits are prone to corruption by fast transients which can cause the execution of false states. Great care is necessary to prevent any clocked circuit (not just microprocessor-based) from being

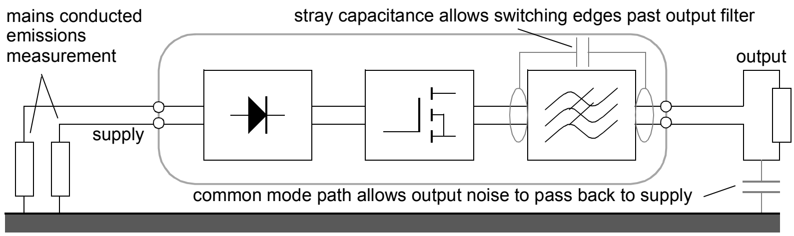


Figure 12.20 General power switching circuit

susceptible to incoming interference. Analogue signals are more affected by continuous interference, which is rectified by non-linear circuit elements and causes bias or signal level shifts. The immunity of analogue circuits is improved by minimizing amplifier bandwidth, maximizing the signal level, using balanced configurations and electrically isolating I/O that will be connected to "dirty" external circuits.

12.2.1 Digital circuits: interference paths

Most of the critical interference in microprocessor circuits is ground-borne, whether it is common mode RF or transients, and the damage is done by conversion of common mode ground noise to differential mode noise at susceptible signal nodes. This occurs because of a high transfer impedance from common to differential mode due to poor PCB layout. Differential mode interference will not propagate far into the circuit from the external interfaces. Therefore, firstly you should concentrate on the best PCB tracking to minimize mode conversion – that is, appropriate use of a ground plane; and secondly, lay out the mechanical design of the product to keep ground interference currents away from the logic circuits. Filter the I/O leads or isolate them, to define a preferential safe current path for interference. Radiated RF fields that generate differential mode voltages internally are dealt with in the same way as differential RF emissions, by minimizing circuit loop area with a ground plane, and by restricting the bandwidth of susceptible circuits where this is feasible.

12.2.1.1 Interference paths – transients

A typical microprocessor-based product, including power supply, operator interface, processor board, enclosure and external connections can be represented at high frequency [29] by the layout shown in Figure 12.21. The 0V rail will appear as a network of inductances with associated stray capacitances to the enclosure. If it's a 0V plane, then the inductances will be lower but the total capacitance will be higher: the principle is the same. An incoming common mode transient current on the mains can travel through the circuit's 0V rail, generating ground differential spikes as it goes, through any or all of several paths as shown (observe the influential effect that stray capacitance has on these paths):

1: primary-to-secondary capacitance through the power supply to 0V, through the equipment and then to case;

2: as above, but then out via an external connection;

3: direct to case, then via stray capacitance to 0V and out via an external connection.

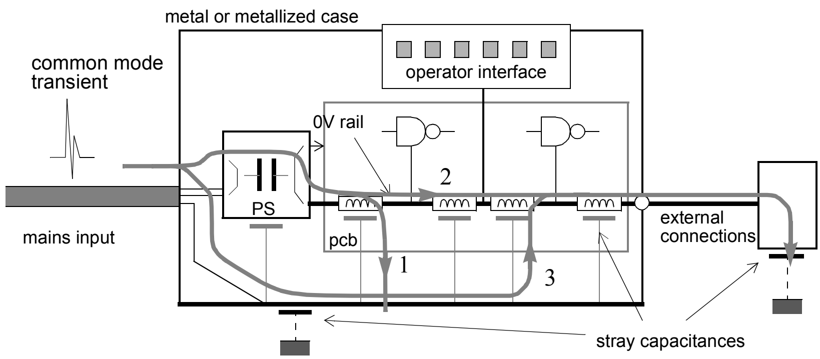


Figure 12.21 Representative high frequency equivalent circuit: transients

If there are no external connections, (1) is the only problem and can be cured by a mains filter and/or by an electrostatic screen to the case in the mains transformer. (2) arises because an external connection can provide a lower impedance route to ground than case capacitance. You cannot control the impedance to ground of external connections, so you have to accept that this route will exist and ensure that the transient current has a preferential path via the case to the interface which does not take in the circuit. This is achieved by ensuring that the case structure is well bonded together, i.e. it presents a low impedance path to the transient, and by filtering interfaces to the case at the point of entry/exit (see section 13.2.4 and section 11.2.3).

If the enclosure is non-conductive then transient currents have no choice but to flow through the circuit. Local grouping of interfaces is essential, as well as liberal use of common mode chokes to increase the impedance seen looking from the circuit into the interface cable; but you must expect some interference current to flow through the PCB itself, so a low transfer impedance for minimum mode conversion, as provided by a good 0V plane, is also needed. Effectively, removing the protection offered by a metal case places much greater stress on the PCB design.

With external connections, route (3) can actually be *worsened* by a mains filter, since at high frequencies parts of the enclosure can float with respect to true ground; and in the transient immunity test the safety earth line is tested as much as the live and neutral. The case, which is connected directly to the safety earth and/or indirectly at high frequencies to live and neutral through the mains filter, will have transients injected straight into it. Even large conducting structures may exhibit high impedances, and hence voltage differentials when subjected to fast transients. The safety earth – the green and yellow wire – is *not* a reference point at HF (refer back to section 11.1.3.4 for ground wire impedances) and if this is the only "earth" connection, the voltage on the enclosure is defined by a complex network of inductances (connected cables) and stray capacitances which are impossible to predict. To deal with this problem when the case is inadequately grounded, the external interfaces must be filtered to the case, just as is the power supply input. This provides a preferential path through the metalwork, and not through the circuit, for the interference currents. This will also deal with transient currents injected via the interface.

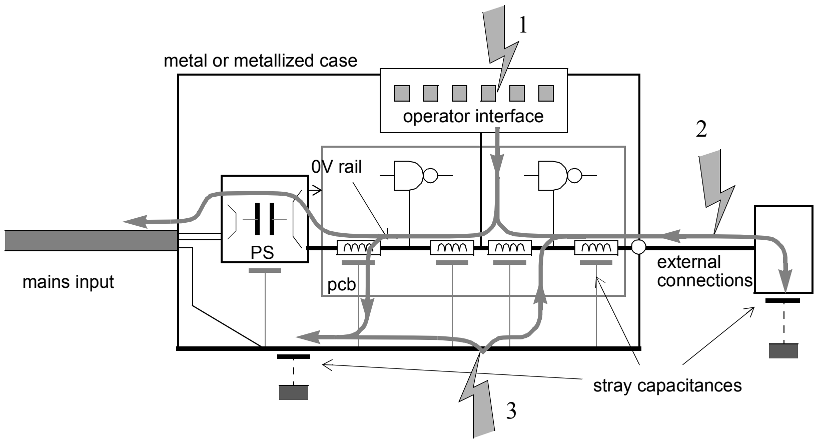


Figure 12.22 Representative high frequency equivalent circuit: ESD

In all cases, grouping all I/O leads together with the supply lead (see Figure 11.24 on page 284) will offer low-inductance paths that bypass the circuit and prevent transient currents from flowing through the PCB tracks.

12.2.1.2 Interference paths – ESD

An electrostatic discharge can occur to any exposed part of the equipment. Common trouble spots as shown in Figure 12.22 are keyboards and controls (1), external cables (2) and accessible metalwork (3). A discharge to a nearby conductive object (which could be an ungrounded metal panel on the equipment itself) causes high local transient currents which will then also induce currents within the equipment by inductive or common impedance coupling.

Because there are many potential points of discharge, the possible routes to ground that the discharge current can take are widespread. Many of them will include part of the PCB ground layout, via stray capacitance, external equipment or exposed circuitry, and the induced transient ground differentials will cause maloperation in the same way as described above for injected common mode transient bursts. The discharge current will take the route (or routes) of least inductance. If the enclosure is well bonded to ground then this will be the natural sink point. If it is not, or if it is non-conductive, then the routes of least inductance will be via the connecting cables.

In any case, it will be hard to prevent the PCB from being subject to some of the ESD currents, and the di/dt of an ESD stress is the highest of all EMC phenomena (10^9 or 10^{10} A/s), so this again puts the emphasis on good layout practice to minimize mode conversion transfer impedance. If the edge of the PCB may be exposed, as in card frames, then a useful trick is to run a “guard trace” around it, unconnected to any circuitry, and separately bond this to ground to act as a sacrificial path.

When the enclosure consists of several conductive panels then these must all be well bonded together, following the rules described in section 14.1 for shielded enclosures. All metallic covers and panels must be bonded together with a low impedance connection (<2.5mΩ at DC) in at least two places. If this is not done then the edges of the panels will create very high transient fields as the discharge current attempts to cross them. If they are interconnected by lengths of wire, the current through the wire will cause a high magnetic field around it which will couple effectively with nearby PCB tracks.

The discharge edge has an extremely fast risetime (sub-nanosecond, see section 10.3.3.1) and so stray capacitive coupling is essentially transparent to it, whilst even short ground connectors of a few nH will present a high impedance. For this reason the presence or absence of a safety ground wire (which has a high inductance) will often make little difference to the system response to ESD.

12.2.1.3 Transient and ESD protection

Techniques to guard against corruption by transients and ESD are generally similar to those used to prevent RF emissions, and the same components will serve both purposes. Specific strategies aim to prevent incoming transient and ESD currents from flowing through the circuit, and instead to absorb or divert them harmlessly and directly to ground (Figure 12.23). To achieve this for ESD the simplest approach is not to allow an ESD event to happen at all, which means that all around the EUT enclosure there are no accessible conductive surfaces. "Accessible" in this context means accessible to an air discharge from a finger charged to typically 8kV (15kV for certain more rigorous standards), so that creepage and clearance paths from the outside to circuits on the inside must also be considered. As a general rule, you should aim to maintain at least a 2mm clearance gap between tracks on the edges of PCBs and the inside surfaces of enclosures near apertures, and/or 6mm of creepage path from accessible outer surfaces to inner conducting parts, including conductive coatings on the plastic itself. For small products this can be quite a challenge.

Further than this,

- keep all external interfaces physically near each other;

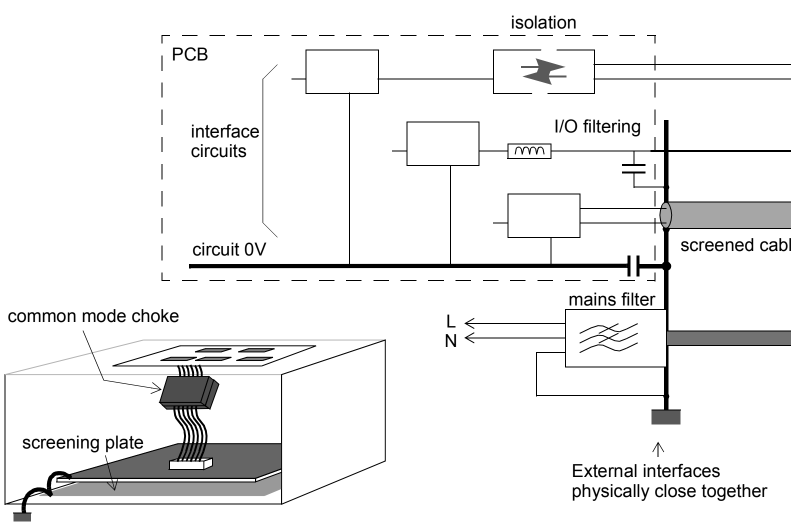


Figure 12.23 Transient and ESD protection

- filter all interfaces to ground at their point of entry;
- if this is not possible, isolate susceptible interfaces with a common mode ferrite choke or opto-couplers;
- use screened cable with the screen connected directly to the interface ground;
- screen PCBs from gaps in the enclosure or external discharge points with extra internally grounded plates.

For reduced ESD susceptibility, the circuit ground needs to remain stable during the ESD event. A low inductance ground network is essential, but this must also be coupled (by capacitors or directly) to a master reference ground structure to prevent significant transient potential differences between the structure and the circuit during an ESD event.

I/O cables and internal wiring may provide coupling paths for the current, just as they are routes into and out of the equipment for common mode RF interference. The best way to eliminate susceptibility of internal harnesses and cables is not to have any, through economical design of the board interconnections. If you must have internal wiring, make sure that its routing is controlled to avoid close coupling with any particularly susceptible (or emissive) circuits. External cables must have their shields well decoupled to the ground structure, following the rules in section 13.1.7, that is 360° bonding of cable screens to connector backshells and no pigtails [125].

Insulated enclosures make the control of ESD currents harder to achieve, and a well-designed and low inductance circuit ground is essential. But, if the enclosure can be designed to have no apertures which provide air gap or creepage paths to the interior then no direct discharge will be able to occur, provided the material's dielectric strength is high enough. You will still need to protect against the field of an indirect discharge, though.

The operator interface

Keyboards present an operator interface which is frequently exposed to ESD. Keyboard cables should be foil-and-braid shielded which is 360° grounded at both ends to the low-inductance chassis metalwork. Plastic key caps will call for internal metal or foil shielding between the keys and the base PCB which is connected directly to the cable shield, to divert transients away from the circuitry. The shield ground should be coupled to the circuit ground at the cable entry point via a 10–100nF capacitor to prevent ground potential separation during an ESD event. A membrane keypad with a polyester surface material has an inherently high dielectric strength and is therefore resistant to ESD, but for optimum protection it should incorporate a ground plane to provide a bleed path for the accumulated charge and to improve RF immunity.

This ground plane must be "hidden" from possible discharges by sealing it behind the membrane surface. The edges of ground plane layers are a potential problem with air discharge if there is only a short distance between the conductive plane and the edge of the membrane assembly. Such ground layers should have a border around the edge of preferably 1cm or more to increase the creepage distance. The ground layer should be taken to chassis directly via a separate short, wide tab, not carried with the keypad signal connections (Figure 12.24). Membranes without such a ground plane must be regarded as unprotected interfaces, and all lines that connect to the driver on the PCB must be heavily RC filtered and preferably provided with ESD surge protection.

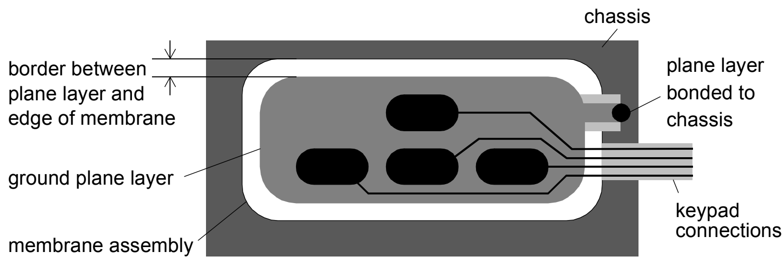


Figure 12.24 Ground plane on a membrane keypad

Transient protection devices for ESD

Interfaces may be particularly susceptible to an ESD event, if they are easily accessible to personnel, or can be approached by the connector of another device carrying a different charge level. In these circumstances transient protection of the interface pins is necessary. Section 13.2.5 discusses the general approach of transient suppressors; for ESD purposes the devices used must be very fast but do not have to carry or dissipate high energy levels. The peak current of an 8kV discharge event is only 30A for a nanosecond or so. The purpose of the suppressor is to divert current from a susceptible device to a suitable grounding structure, which will normally be the chassis but could be the interface power planes if these can carry the current without disrupting the circuit operation. This can be achieved either with a transzorb-type (zener or varistor) device, surface mounted and in a small (0805 or 1206) package for minimum lead inductance, or with steering diodes which dump the applied current into the 0V or supply rails. Figure 12.25 shows these methods.

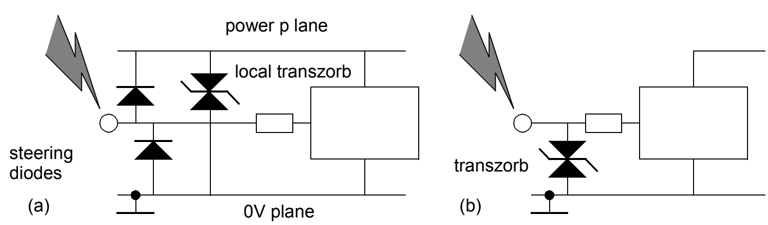


Figure 12.25 ESD interface protection

The problem facing designers who have to deal with high-speed digital interfaces such as the USB or DVI specifications, is that although the individual pins at these interfaces may need ESD protection there is precious little leeway for adding extra parallel components. Any such component will add capacitance, and for interfaces which employ matched transmission lines such capacitance is undesirable since it degrades the matching. This has led to the development of some very low-capacitance diodes specifically for the purpose; below 1pF is possible, which should satisfy most specifications. Transzorbs cannot be made with such low capacitances, but steering diodes can, since these are inherently operated with reverse bias under normal conditions. Packages with several diode pairs are available for protecting several lines at once.

In contrast to a transzorb, a steering diode does not absorb the ESD energy itself but merely diverts it to a local sink which can cope with the energy without harm. This means that such a sink must be explicitly provided: if it is not within the diode package itself then a separate transzorb and/or decoupling capacitor is necessary across the power and 0V planes near to the interface. This safely dissipates the current passed through the steering diodes without allowing a significant power voltage rise across the protected IC. Bearing in mind the fast rise time of the current pulse, there should be minimum inductance in the path created by the diodes, planes and transzorb.

12.2.1.4 Vulnerable circuit points

Power supply

CMOS devices are fairly tolerant of slow V_{CC} changes, but this is less true of other technologies. You should consider the possible impact of voltage dips, interruptions and variations, and check the circuit operation under a wide variety of conditions. Surges can cause both short-term overvoltages and undervoltages on the supply rail. Supply voltages exceeding the devices' absolute maximum ratings (typically 7V for 5V supplies, or 4.6V for 3V3) can cause CMOS latchup, in which the device behaves like a triggered thyristor between its supply pins, shorting out the positive supply; if some current limiting is not provided, this can easily be destructive. In many cases, such current limiting could be as simple as a resistor of a few ohms in series with the supply – suitably decoupled, of course.

Oscillator, reset and interrupt inputs

The clock oscillator input pin is high impedance and particularly likely to be affected. A transient coupled onto this pin will generate a shortened processor clock period, which is liable to cause a spurious instruction or data access, thus corrupting the program counter or memory. The reset and edge-triggered interrupt inputs are also sensitive and glitches here have widespread effects on the processor operation. On many microcontrollers, the reset pin is specified for a minimum active duration of several clock cycles. If it is pulsed for a shorter period, an incomplete reset occurs, with unpredictable results.

Always treat these pins on a microcontroller with the greatest care (Figure 12.26): never run long lines to them, put in buffering resistors where possible, keep the tracks closely coupled to the ground plane.

I/O ports

Input/output ports connected to long tracks or external connections are vulnerable on three counts. Small amounts of noise can give spurious data. Higher levels can corrupt the contents of I/O control or data registers, while direct ESD can cause CMOS latch-up.

12.2.2 Logic noise immunity

The ability of a logic element to operate correctly in a noisy environment involves more than the commonly quoted static noise margins. To create a problem an externally generated transient must cause a change of state in an element which then propagates through the system. Systems with clocked storage elements or those operating fast enough for the transient to appear as a signal are more susceptible than slow systems or those without storage elements (combinational logic only).

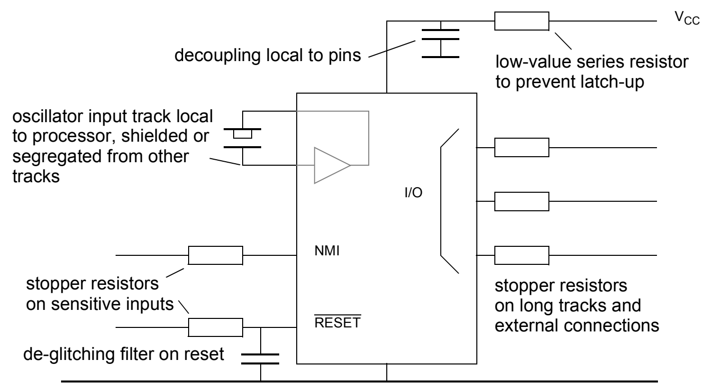


Figure 12.26 Immunity precautions around microcontrollers

12.2.2.1 Dynamic noise margin

The effect of a fast transient will depend on the peak voltage coupled into the logic input, and also on the speed of response of the element. Any pulse positive-going from 0V but below the logic switching threshold (typically 1.4V for 74HCT, 50% V_{CC} for 74HC circuits) will not cause the element input to switch from 0 to 1 and will not be propagated into the system. Conversely a pulse above the threshold will cause the element to switch. But a pulse which is shorter than the element's response time will need a higher voltage to cause switchover, and therefore the graph shown in Figure 12.27 can be constructed [10][15], which illustrates the susceptibility of different logic families versus pulse width and amplitude. Bear in mind that switching and ESD transients may lie within the 1–5ns range. Here is another argument for slow logic!

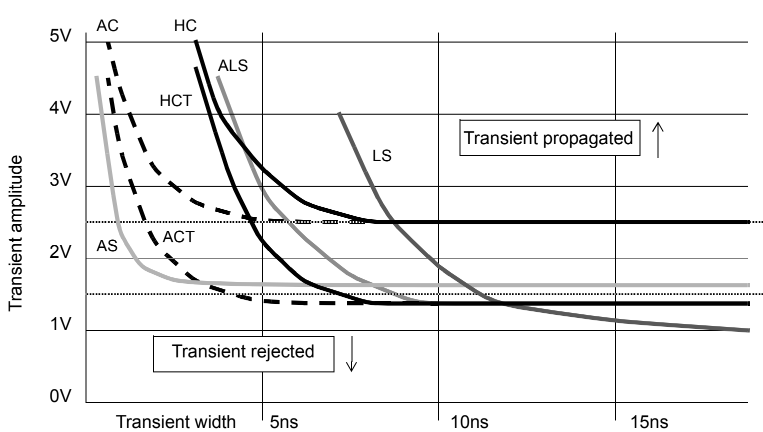


Figure 12.27 Dynamic noise margins

With synchronous logic, the time of arrival of the transient with respect to the system clock (assuming it corrupts the data line rather than the clock line, due to the former's usually greater area) is important. If the transient does not coincide with the active clock edge then an incorrect value on the data line will not propagate through the system. Thus you can expand the graphs of Figure 12.27 to incorporate another dimension of elapsed time from the clock edge. Tront [134] has simulated a combinational logic circuit with a flip-flop in 3-micron CMOS technology, and generated a series of "upset windows" in this way to describe the susceptibility of that particular circuit to interference. Such a simulation process, using the simulation package SPICE3, can pinpoint those parts of a circuit which have a high degree of susceptibility.

12.2.2.2 Transient coupling

The amplitude of any pulse coupled differentially into a logic input will depend on the loop area of the differential coupling path which is subjected to the transient field $H_{transient}$ due to transient ground currents $I_{transient}$, and also on the impedance of the driving circuit – less voltage is coupled into a lower impedance. If sensitive signal tracks are run close to their ground returns as recommended for emissions, then the resulting loop area is small and little interference is coupled differentially into the sensitive input (Figure 12.28). This is another way of stating that the transfer impedance is lowest for tightly-coupled signal and return tracks.

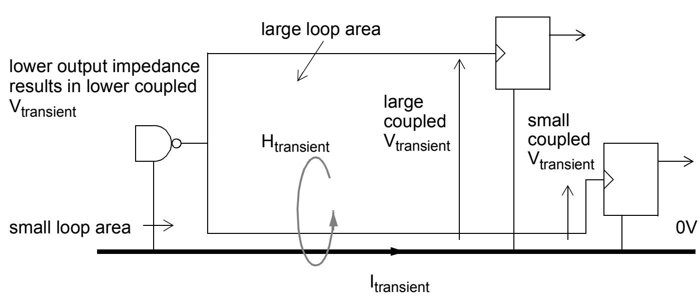


Figure 12.28 Transient coupling via signal/return current loops

12.2.2.3 Susceptibility to RF

Susceptibility of digital circuits to RF, as opposed to transient interference, tends to be most marked in the 20–200MHz region. Susceptibility at the component level is broadband, although there are normally peaks at various frequencies due to resonances in the coupling path. As the frequency increases into the microwave region, component response drops as parasitic capacitances offer alternative shunt paths for the RF energy, and the coupling becomes less efficient. Prediction of the level of RF susceptibility of digital circuits using simulation is possible for small-scale integrated circuits [139] but the modelling of the RF circuit parameters of VLSI devices requires considerable effort, and the resources needed to develop such models for microprocessors and their associated peripherals is overwhelmed by the rate of introduction of new devices.

The first effect of RFI to be noticed on dynamically active logic circuits is timing jitter at the transitions (Figure 12.29). The RF signal is added to the input voltage as it switches through the non-linear active transition region of the device, which causes either an advance or a delay on the transition edge as it is seen by the input. If the circuit timing is critical then this itself can cause maloperation. Thus, to increase RF immunity the first design requirement is to keep the system timing as relaxed as possible throughout.

As the RF level increases then actual spurious logic level transitions occur and these will propagate through the circuit if the logic response speed is high enough. Different thresholds can be found for logic 0 and logic 1 effects because of the different driver output impedances in the two states. Active high transition clock inputs are far more sensitive when held at logic 0 than when at logic 1.

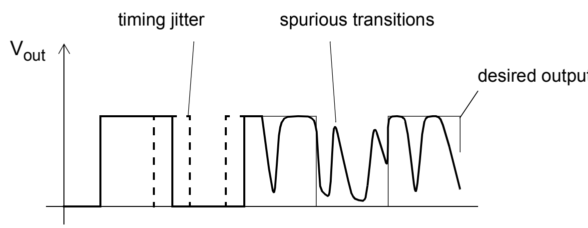


Figure 12.29 RF effects on logic circuits

12.2.3 Signal integrity and ground bounce

Many of the aspects of PCB layout for digital circuits already discussed fall easily under another heading, that of signal integrity. Simply put, this refers to design of the interconnections within digital circuits such that they don't compromise the integrity of signal transmission between them. This is not principally a matter of external EMC, more it is a case of ensuring that the circuit actually works: that is, it doesn't interfere with itself. But good signal integrity has the beneficial side effect of reducing the circuit's exposure to external disturbances, particularly transients and RF.

Design aspects that are relevant for signal integrity in high-speed systems include:

- transmission line matching of interconnecting traces (see section 12.1.2.4);
- control of ground bounce at internal IC nodes and external pins.

We have already looked at transmission lines. Ground bounce is a threat in several ways. Consider the equivalent circuit of an IC including its internal bond wires as shown in Figure 12.30. This representation assumes perfect (zero inductance) ground and power planes, but takes into account the inductance of the bond wires between the chip and its connecting pins, and the inductance of the connecting pins to the planes.

At each output transition, a current spike is drawn through the GND pin (see section 12.1.1.3) partly to charge/discharge the output node capacitance and partly because of ΔI currents. This current passes through the ground inductance (the same thing happens with the opposite polarity to the supply pins) and creates a voltage pulse at the internal 0V node of the IC. Thus, compared to the (assumed perfect) 0V rail of the ground plane, the IC's internal 0V experiences a "bounce" at each current transition. To put it in context, if the total ground inductance is 10nH and di/dt is 50mA/ns the amplitude of the bounce is 0.5V.

This voltage disturbance is shared with all the circuits on the chip. In a complex device this can mean that one part of the IC is capable of affecting other parts, since the

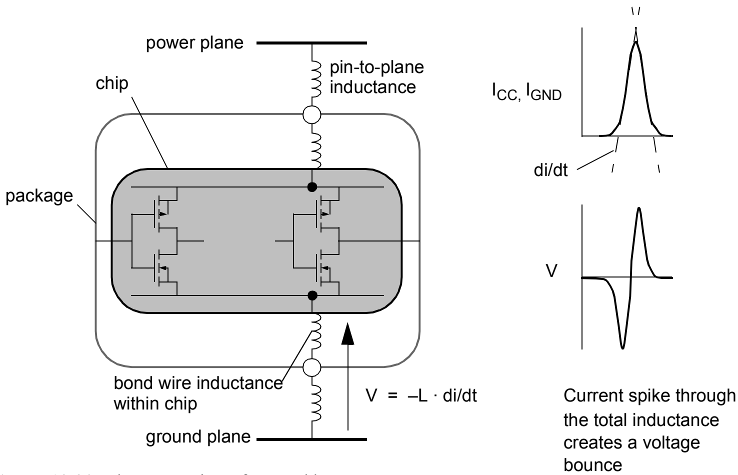


Figure 12.30 The generation of ground bounce

rogue voltage appears in series with other inputs and therefore affects the dynamic logic zero threshold, possibly leading to mis-clocking or mis-reading of logic levels. The chip manufacturer can only mitigate the problem by offering several 0V pins so that the bond wire inductances are paralleled, and/or by providing several distinct 0V nodes within the chip to remove the common impedance coupling. The only cures available to the designer are not to overload the outputs with capacitance, and to make sure that every 0V pin is solidly bonded to the ground plane by local, minimum inductance vias.

Figure 12.30 assumes that the ground and power planes do not themselves suffer from voltage bounce. But this is only true if they have zero impedance, and practical power distribution networks will not meet this ideal. If there is too much inductance in the supply traces between ICs, then the ground bounce problem is exported to create voltage differentials between the 0V pins of these ICs with a comparable result, that of different ICs in the same system interfering with each other (Figure 12.31).

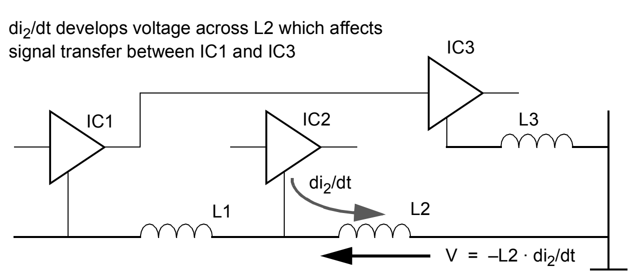


Figure 12.31 System-developed ground bounce

Excessive ground bounce has a number of consequences:

- in the extreme, it will prevent the circuit working;
- more usually, it will create errors and strange effects which may be attributed to other causes such as marginal timing;

- the degradation in thresholds may not show effects under normal conditions but will make the circuit more susceptible to external interference;
- the voltage developed along the supply rails makes the circuit inherently more noisy and contributes to higher RF emissions (*cf* section 12.1.2.2).

12.2.4 The microprocessor watchdog

Circuit techniques to minimize the amplitude and control the path of disruptive interference go a long way towards "hardening" a microprocessor circuit against corruption. But they cannot *eliminate* the risk. The coincidence of a sufficiently high-amplitude transient with a vulnerable point in the data transfer is an entirely statistical affair. The most cost-effective way to ensure the reliability of a microprocessor-based product is to accept that the program *will* occasionally be corrupted, and to provide a means whereby the program flow can be automatically recovered, preferably transparently to the user. This is the function of the microprocessor watchdog [21].

Some micros on the market include built-in watchdog devices, which may take the form of an illegal-opcode trap, or a timer which is repetitively reset by accessing a specific register address. If such a watchdog is available, it should be used, because it will be well matched to the processor's operation; otherwise, one must be designed-in to the circuit.

12.2.4.1 Basic operation

The most serious result of a transient corruption is that the processor program counter or address register is upset, so that it starts interpreting data or empty memory as valid instructions. This causes the processor to enter an endless loop, either doing nothing or performing a few meaningless or, in the worst case, dangerous instructions. A similar effect can happen if the stack register or memory is corrupted. Either way, the processor will appear to be catatonic, in a state of "dynamic halt".

A watchdog guards against this eventuality by requiring the processor to execute a specific simple operation regularly, regardless of what else it is doing, on pain of consequent reset. The watchdog is actually a timer whose output is linked to the $\overline{\text{RESET}}$ input, and which itself is being constantly re-triggered by the operation the processor performs, normally writing to a spare output port. This operation is shown schematically in Figure 12.32.

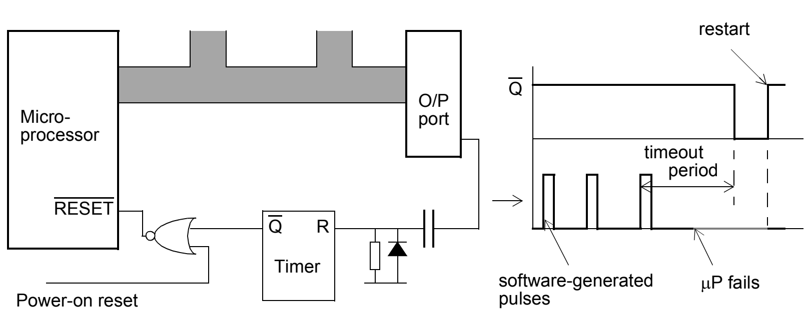


Figure 12.32 Watchdog operation

12.2.4.2 Timeout period

If the timer does not receive a "kick" from the output port for more than its timeout period, its output goes low ("barks") and forces the microprocessor into reset. The timeout period must be long enough so that the processor does not have to interrupt time-critical tasks to service the watchdog, and so that there is time for the processor to start the servicing routine when it comes out of reset (otherwise it would be continually barking and the system would never restart properly). On the other hand, it must not be so long that the operation of the equipment could be corrupted for a dangerous period. There is no one timeout period which is right for all applications, but usually it is somewhere between 10ms and 1s.

12.2.4.3 Timer hardware

The watchdog circuit has to exceed the reliability of the rest of the circuit and so the simpler it is, the better. A standard timer IC is quite adequate, but the timeout period may have an unacceptably wide variation in tolerance, besides needing extra discrete components. A digital divider such as the 4060B fed from a high frequency clock and periodically reset by the report pulses is a more attractive option, since no other components are needed. The divider logic could instead be incorporated into an ASIC if this is present for other purposes. The clock has to have an assured reliability in the presence of transient interference, but such a clock may well already be present or could be derived from the unsmoothed AC input at 50/60Hz.

An extra advantage of the digital divider approach is that its output in the absence of re-triggering is a stream of pulses rather than a one-shot. Thus if the micro fails to be reset after the first pulse, or more probably is derailed by another burst of interference before it can re-trigger the watchdog, the watchdog will continue to bark until it achieves success (Figure 12.33). This is far more reliable than a monostable watchdog that only barks once and then shuts up.

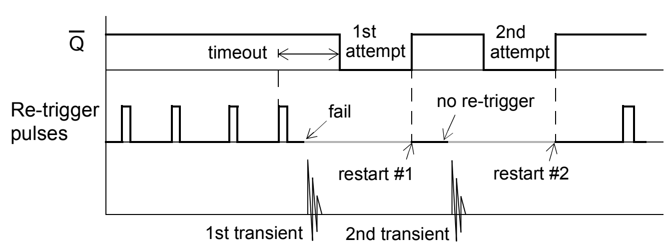


Figure 12.33 The advantage of an astable watchdog

A programmable timer must not be used to fulfil the watchdog function, however attractive it may be in terms of component count. It is quite possible that the transient corruption could result in the timer being programmed off, thereby completely silencing the watchdog. Similarly, it is unsafe to disable the watchdog from the program while performing long operations; corruption during this period will not be recoverable. It is better to insert extra watchdog "kicks" during such long sequences.

12.2.4.4 Connection to the microprocessor

Figure 12.32 shows the watchdog's $\overline{Q}$ output being fed directly to the $\overline{RESET}$ input along with the power-on reset (POR) signal. In many cases it will be possible and preferable to trigger the timer's output from the POR signal, in order to assure a defined reset pulse width at the micro on power-up.

It is essential to use the $\overline{\text{RESET}}$ input and not some other signal to the micro such as an interrupt, even a non-maskable one. The processor may be in any conceivable state when the watchdog barks, and it must be returned to a fully characterized state. The only state which can guarantee a proper restart is reset. If the software must know that it was the watchdog that was responsible for the reset, this should be achieved by reading a separate latched input port during initialization.

12.2.4.5 Source of the re-trigger pulse

Equally important is that the micro should not be able to carry on kicking the watchdog when it is catatonic. This demands AC coupling to the timer's re-trigger input, as shown by the R–C–D network in Figure 12.32. This ensures that only an edge will re-trigger the watchdog, and prevents an output which is stuck high or low from holding the timer off. The same effect is achieved with a timer whose re-trigger input is edge- rather than level-sensitive.

Using a programmable port output in conjunction with AC coupling is attractive for two reasons. It needs two separate instructions to set and clear it, making it very much less likely to be toggled by the processor executing an endless loop; this is in contrast to designs which use an address decoder to produce a pulse whenever a given address is accessed, which practice is susceptible to the processor rampaging uncontrolled through the address space. Secondly, if the programmable port device is itself corrupted but processor operation otherwise continues properly, then the re-trigger pulses may cease even though the processor is attempting to write to the port. The ensuing reset will ensure that the port is fully re-initialized. As a matter of software policy, programmable peripheral devices should be periodically re-initialized anyway.

12.2.4.6 Generation of the re-trigger pulses in software

If possible, two independent software modules should be used to generate the two edges of the report pulse (Figure 12.34). With a port output as described above, both edges are necessary to keep the watchdog held off. This minimizes the chance of a rogue

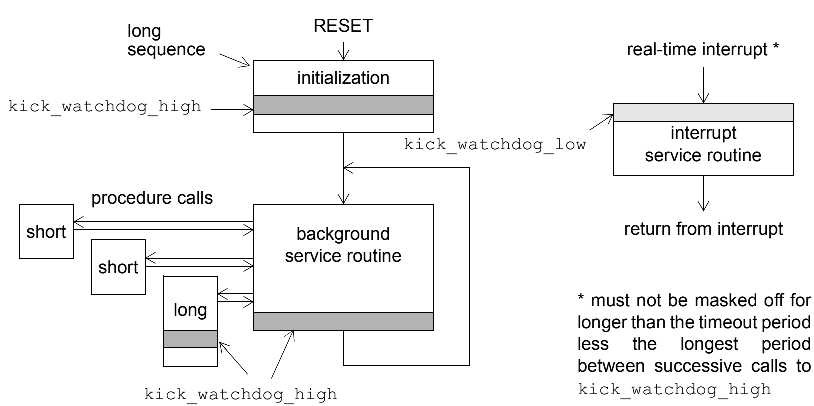


Figure 12.34 Software routine for watchdog re-trigger

software loop generating a valid re-trigger pulse. At least one edge should only be generated at one place in the code; if a real-time "tick" interrupt is used, this edge could be conveniently placed at the entry to the interrupt service routine, whilst the other is placed in the background service module. This has the added advantage of guarding against the interrupt being accidentally masked off.

Placing the watchdog re-trigger pulse(s) in software is the most critical part of watchdog design and repays careful analysis. On the one hand, too many calls in different modules to the pulse generating routine will degrade the security and efficiency of the watchdog; but on the other hand, any non-trivial application software will have execution times that vary and will use different modules at different times, so that pulses will have to be generated from several different places. Two frequent critical points are on initialization, and when writing to non-volatile (EEPROM) memory. These processes may take several tens of milliseconds. Analysing the optimum placement of re-trigger pulses, and ensuring that under all correct operating conditions they are generated within the timeout period, is not a small task.

Ideally, the microprocessor should only refresh the watchdog if it is certain that *all* its software routines are running normally. You may be able to identify the most critical routines and define a byte of corresponding task flags. When each task runs it sets its own flag; a watchdog refresh task running in the main loop keeps monitoring the flags, and when all have been set it toggles the watchdog refresh and clears the byte to zero.

12.2.4.7 Testing the watchdog

This is not at all simple, since the whole of the rest of the circuit design is bent towards making sure the watchdog never barks. Creating artificial conditions in the software is unsatisfactory because the tested system is then unrepresentative. An adequate procedure for most purposes is to subject the equipment to repeated transient pulses which are of a sufficient level to corrupt the processor's operation predictably, if necessary using specially "weakened" hardware. For safety critical systems you may have to perform a statistical analysis to determine the pulse repetition rate and duration of test that will establish acceptable performance. An LED on the watchdog output is useful to detect its barks. A particularly vulnerable condition is the application of a burst of spikes, so that the processor is hit again just as it is recovering from the last one. This is unhappily a common occurrence in practice.

As well as testing the reliability of the watchdog, a link to disable it must be included in order to test new versions of software.

12.2.5 Defensive programming

Some precautions against interference can be taken in software. Standard techniques of data validation and error correction should be widely used. Hardware performance can also be improved by well-thought-out software. Some means of disabling software error-checking is useful when optimizing the equipment hardware against interference, as otherwise weak points in the hardware will be masked by the software's recovery capabilities. For example, software which does not recognize digital inputs until three polls have given the same result will be impervious to transients which are shorter than this. If your testing uses only short bursts or single transients the equipment will appear to be immune, but longer bursts will cause maloperation which might have been prevented by improving the hardware immunity.

Not all microprocessor faults are due to interference. Other sources are intermittent connections, marginal hardware design, software bugs, meta-stability of asynchronous circuits, etc. Nevertheless it is necessary to prepare for interference-induced errors, and typical software techniques to enhance immunity are to:

- type-check and range-check all input data;
- sample input data several times and either subject it to rate-of-change checks, for analogue data, or validate it, for digital data;
- incorporate parity checking and data checksums in all data transmission;
- protect data blocks in volatile memory with error-detecting and correcting algorithms;
- wherever possible rely on level- rather than edge-triggered interrupts;
- periodically re-initialize programmable interface chips (PIAs, ACIAs, etc.).

When a failure occurs on testing, it is often difficult to find out what actually happened. Connecting an in-circuit emulator while undergoing ESD or EFT testing is not recommended since it drastically worsens the interference coupling. Usually, you have to resort to deducing the internal state after failure from the state of the I/O and bus lines. If you can program spare I/O pins to give a diagnostic status indication this is made easier. Alternatively or in addition, unused non-volatile memory can be used to store a diagnostic trace, which is recovered after failure [53].

12.2.5.1 Input data validation and averaging

If you can set known limits on the figures that enter as digital input to the software then you can reject data which are outside those limits. When, as in most control or monitoring applications, each sensor inputs a continuous stream of data, this is simply a question of taking no action on false data. Since the most likely reason for false data is corruption by a noise burst or transient, subsequent data in the stream will probably be correct and nothing is lost by ignoring the bad item. Data-logging applications might require a flag on the bad data rather than merely to ignore it.

This technique can be extended if there is a known limit to the maximum rate-of-change of the data. An input which exceeds this limit can be ignored even though it may be still within the range limits. Software averaging on a stream of data to smooth out process noise fluctuations can also help remove or mitigate the effect of invalid data.

You should take care when using sophisticated software for error detection not to lock out genuine errors which need flagging or corrective action, such as a sensor failure. The more complex the software algorithm is, the more it needs to be tested to ensure that these abnormal conditions are properly handled.

12.2.5.2 Digital inputs

A similar checking process should be applied to digital inputs. In this case, there are only two states to check so range testing is inappropriate. Instead, given that the input ports are being polled at a sufficiently high rate, compare successive input values with each other and take no action until two or three consecutive values agree. This way, the processor will be “blind” to occasional transients which may coincide with the polling time slot. (This method is, of course, nothing more than a variant of the standard technique of “de-bouncing” switch inputs.) It does mean that the polling rate must be two or three times faster than the minimum required for the specified response time, which in turn may require a faster microprocessor than originally envisaged.

12.2.5.3 Interrupts

For similar reasons to those outlined above, it is preferable not to rely on edge-sensitive interrupt inputs. Such an interrupt can be set by a noise spike as readily as by its proper signal. Undoubtedly edge-sensitive interrupts are necessary in some applications, but in these cases you should treat them in the same way as clock inputs to latches or flip-flops, and take extra precautions in layout and drive impedance to minimize their noise susceptibility. If there is a choice in the design implementation, then favour a level-sensitive interrupt input.

If an interrupt is non-maskable, excessive noise on its input pin can cause a burst of consecutive interrupts which will quickly result in a stack overflow. This should be dealt with by monitoring the value of the stack pointer at the beginning of the interrupt routine, and forcing a hardware reset if it gets too large.

12.2.5.4 Token passing

Software structures tend to mimic business management structures: the big decisions are made at the top, but enacted by those at the bottom [53]. The danger comes when those at the bottom execute tasks without authorization from the top. Interference could force a microcontroller into a low-level sub-routine, perhaps by corrupting its program counter, accidentally initiating an action which has serious consequences.

Token passing (Figure 12.35) provides a means whereby each sub-routine can check the authority of the calling routine. The top-level (decision maker) routine sends a specific byte value (token) to the first sub-routine, which checks it against a list of valid tokens and if necessary passes it on. Normally, the tokens match and the action is executed successfully, but if not, a reset is forced. There is of course a price to pay in terms of code size and execution time, for which reason you may want to use token passing selectively rather than globally.

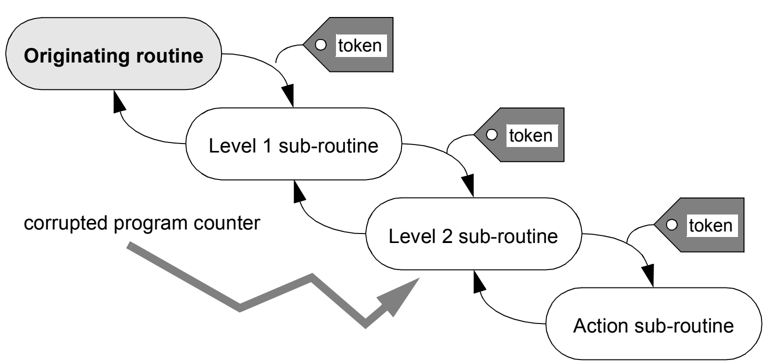


Figure 12.35 Token passing through sub-routines

12.2.5.5 Data and memory protection

Volatile memory (RAM, as distinct from ROM; EEPROM and flash memory could be categorized as volatile since its data may be corrupted while being written) is susceptible to various forms of data corruption. These can be detected by placing critical data in tables in RAM. Each table is then protected by a checksum, which is stored with the table. Checksum-checking diagnostics can be run by the background routine automatically at whatever interval is deemed necessary to catch RAM corruption, and an error can be flagged or a software reset can be generated as required. The absolute values of RAM data do not need to be known provided that the checksum is recalculated every time a table is modified. Beware that the diagnostic routine is not interrupted by a genuine table modification or vice versa, or errors will start appearing from nowhere! Of course, the actual partitioning of data into tables is a critical system design decision, as it will affect the overall robustness of the system.

Some non-volatile memory devices include a software locking mechanism, which disables the write facility. Full use should be made of this feature, unlocking only for those infrequent times when fresh data has to be written. If the power fails or the processor crashes in the middle of writing to non-volatile memory, the stored data will be unavoidably corrupted. To manage the problem, use a COPY–MODIFY–STORE–ERASE sequence to prevent critical data from being part overwritten.

12.2.5.6 Unused program memory

One of the threats discussed in the section on watchdogs above was the possibility of the microprocessor accessing unused memory space due to corruption of its program counter. If it does this, it will interpret whatever data it finds as a program instruction. In such circumstances it would be useful if this action had a predictable outcome.

An 8-bit bus access to a non-existent address returns the data $\#FF_H$, provided there is a passive pull-up on the bus. Nothing can be done about this. However, un-programmed ROM also returns $\#FF_H$ and this can be changed. A good approach is to convert all unused $\#FF_H$ locations to the processor's one-byte NOP (no operation) instruction (Figure 12.36). The last few locations in ROM can be programmed with a JMP RESET instruction, normally three bytes, which will have the effect of resetting the processor. Then, if the processor is corrupted and accesses anywhere in unused memory, it finds a string of NOP instructions and executes these (safely) until it reaches the JMP RESET, at which point it restarts. An alternative is to block-fill the unused locations with the STOP instruction rather than the NOP, which then relies on the hardware watchdog restarting the system.

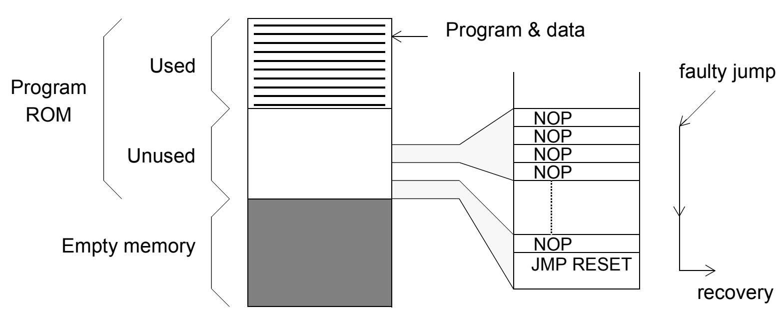


Figure 12.36 Protecting unused program memory with NOPs

The effectiveness of this technique depends on how much of the total possible memory space is filled with NOPs or STOPs, since the processor can be corrupted to a random address. (In practice, the corruption is often less than random, due perhaps to particularly susceptible states or a particular data/address line being the weakest point.) If the processor accesses an empty bus, its action will depend on the meaning of the $\#FF_H$ instruction. The relative cheapness of large ROMs and EPROMs means that you could consider using these, and filling the entire memory map with ROM, even if your program requirements are small.

12.2.5.7 Re-initialization

As well as RAM data, you must remember to guard against corruption of the set-up conditions of programmable devices such as I/O ports or UARTs. Many programmers assume erroneously that once an internal device control register has been set up (usually in the initialization routine) it will stay that way forever. Experience shows that control registers can change their contents, even though they are not directly connected to an external bus, as a result of interference. This may have consequences that are not obvious to the processor: for instance if an output port is re-programmed as an input, the processor will happily continue writing data to it oblivious of its ineffectiveness.

The safest course is to periodically re-initialize all critical registers, perhaps in the main idling routine if one exists. Timers, of course, cannot be protected in this way. The period between successive re-initializations depends on how long the software can tolerate a corrupt register, versus the software overhead associated with the re-initialization.

12.2.6 Transient and RF immunity – analogue circuits

Analogue circuits in general are not as susceptible to transient upset as digital, but may be more susceptible to demodulation of RF energy. This can show itself as a DC bias shift which results in measurement non-linearities or non-operation, or as detection of modulation, which is particularly noticeable in audio and video circuits. Such bias shift does not affect digital circuit operation until the bias is enough to corrupt logic levels, at which point operation ceases completely. Improvements in immunity result from attention to the four areas as set out below. The greatest RF signal levels are those coupled in via external interface cables and so interface circuits should receive the first attention.

Analogue immunity principles

- minimize circuit bandwidth
- maximize signal levels
- ensure a good circuit stability margin
- use balanced signal configurations
- isolate particularly susceptible paths

12.2.6.1 Audio rectification

This is a term used rather loosely to describe the detection of RF signals by low frequency circuits. It is responsible for most of the ill effects of RF susceptibility of both analogue and digital products.

When a circuit is fed an RF signal that is well outside its normal bandwidth, the circuit can respond either linearly or non-linearly (Figure 12.37). If the signal level is

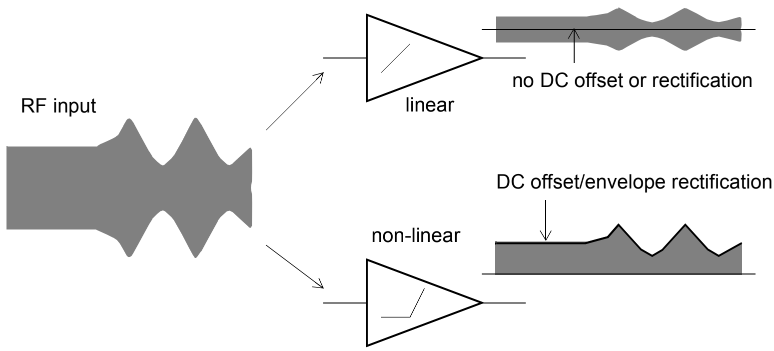


Figure 12.37 RF demodulation by non-linear circuits

low enough for it to stay linear, it will pass from input to output without affecting the wanted signals or the circuit's operation. If the level drives the circuit into non-linearity, then the envelope of the signal (perhaps severely distorted) will appear on the circuit's output. At this point it will be inseparable from the wanted signal, and indeed the wanted signal will itself be affected by the circuit's forced non-linearity.

The response of the circuit depends on its linear dynamic range and on the level of the interfering signal at the point in the circuit which has the most non-linearity. Usually this is its input. All other factors being equal, a circuit which has a wide dynamic range will be more immune to RF than one which has not.

12.2.6.2 Bandwidth, level and balance

The level of the interfering signal can be reduced by restricting the operating bandwidth to the minimum acceptable. You can achieve this (referring to Figure 12.38) by input RC or LC filtering (1), feedback RC filtering (2), and low-value (10–33pF) capacitors (3) or resistors (4) directly at the input terminals. RC filters may degrade stability (see section 12.2.6.4) or worsen the circuit's common mode rejection (CMR) properties, and the value of C must be kept low to avoid this, but an improvement in RF rejection of between 10 to 35dB over the range 0.15 to 150MHz has been reported [130] by including a 27pF feedback capacitor on an ordinary inverting op-amp circuit. High frequency CMR is determined by the imbalance between capacitances on balanced inputs, and the typical low-value tolerance of 10% may make capacitive filtering unacceptable for this reason alone. If an increased input resistance would be too high and might affect circuit DC conditions, a lossy ferrite-cored choke or bead is an alternative series element.

Localized low-value series resistance (4) is a useful approach at the inputs of op-amps and comparators – their high input impedance allows a series resistor of up to a few hundred ohms to be applied at each input, literally at the pins. In conjunction with the input capacitance, this attenuates any induced RF voltages that may be induced along the input tracks. Since surface mount resistors are extremely cheap, it is worth implementing this as a standard technique for all analogue circuits.

You should design for signal level to be as high as possible throughout, consistent with other circuit constraints, but at the same time impedances should also be maintained as low as possible to minimize capacitive coupling and these requirements may conflict. The decision will be influenced by whether inductive coupling is

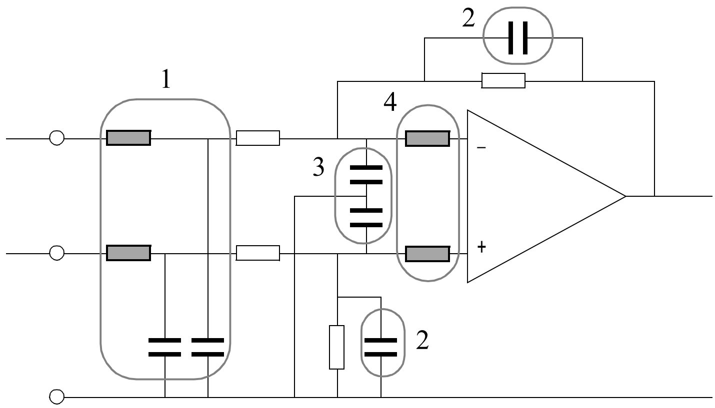


Figure 12.38 Bandwidth limitation in a differential amplifier

expected to be a major interference contributor. If it is (because circuit loop areas cannot be made acceptably small), then higher impedances will result in lower coupled interference levels. Refer to the discussion on inductive and capacitive coupling (section 10.1.1).

Balanced circuit configurations allow maximum advantage to be taken of the inherent common mode rejection (CMR) of op-amp circuits. But note that CMR is poorer at high frequencies and is affected by capacitive and layout imbalances, so it is unwise to trust too much in balanced circuits for good RF and transient immunity.

A common fix for improving the immunity of a discrete transistor circuit is to incorporate either or both of a resistor or ferrite chip in series with the base and a low-value (10–33pF) capacitor directly across the base-emitter junction (Figure 12.39). The effect of this is to reduce the RF applied across the junction, where the non-linearity occurs, but the components need to be mounted right next to the transistor connections. As a matter of standard design practice, you should put such components in place wherever there is a circuit input that could be exposed to incoming RF.

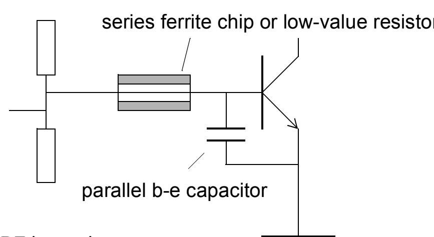


Figure 12.39 Increasing discrete transistor RF immunity

12.2.6.3 The effect of op-amp type and operating conditions

It has been observed [139] that in the frequency range 1–20MHz, mean values of demodulated RFI are 10 to 20dB lower for BiFET op-amps than for bipolar op-amps. This may well have to do with the inherent non-linearity of the bipolar input base-emitter junction. A research project carried out by the author [142] investigated the contributions to uncertainty of the immunity tests, and one aspect was the non-linearity of response of op-amp circuits with injected interference. It was clear that:

- different op-amp technologies have different characteristic transfer functions of injected RF to DC offset, with CMOS types showing less effect than bipolar;
- for any given type, different responses can be observed at different frequencies;
- the DC operating conditions often have a marked effect on the response, and this is particularly noticeable when single-supply devices are operated close to their lower common mode voltage limit;
- only one of the devices that were investigated, the LM301A which has the oldest design, showed a response that was approaching linearity over the range tested. All others showed non-linearity to a greater or lesser extent.

A sample of the results illustrating these points is shown in Figure 12.40. Notice particularly that for the bipolar LM358 (the National part's response is shown, but Philips and Harris parts performed in the same way) in single supply mode with the input near the 0V rail, as soon as the RF voltage is enough to force the input outside its declared common mode range the output swings immediately to the positive supply rail and stays there. This doesn't happen when the same part is used in dual supply mode with the input at mid-rail, and it doesn't happen with the CMOS TLC272, although it does happen with the BiCMOS CA3240. A general observation is that you should avoid operating op-amp inputs very close to one end or the other of their common mode range, as this makes them much more susceptible to RF.

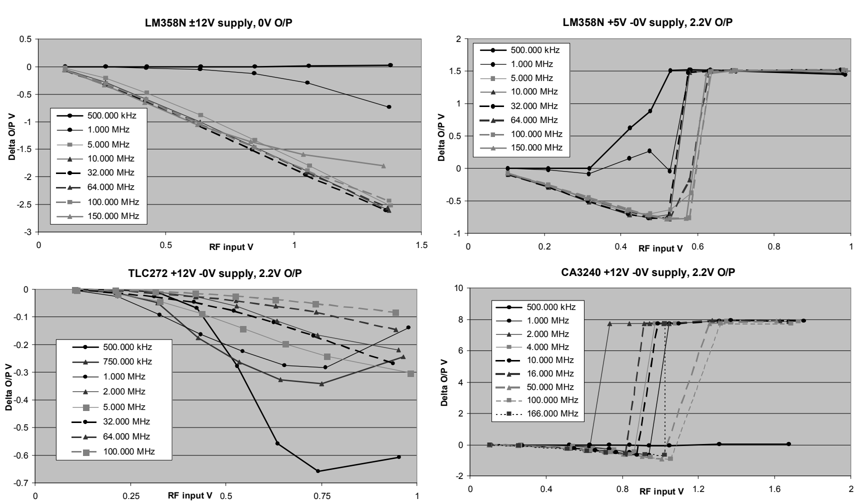


Figure 12.40 Op-amp DC response versus RF input voltage (from [142] Annex F)
NB the Y axis in each case is the deviation from the undisturbed DC output level

12.2.6.4 Stability

Op-amps with gain bandwidths of hundreds of MHz or even GHz are routinely used in many applications. The stability of wideband amplifiers was briefly mentioned in section 12.1.4.1 in the context of emissions, but a quasi-stable amplifier is also a threat to immunity. If the circuit is close to oscillation but not actually unstable at a particular radio frequency, this is equivalent to saying it has a peak in its frequency response at that frequency (Figure 12.41). If an interfering RF signal is applied at this peak the amplifier will happily respond to it, most probably saturating and corrupting its desired signal. This is not an uncommon occurrence particularly on the conducted immunity test in the range 5–30MHz, where peaks in susceptibility cannot be explained by the usual structural resonances of higher frequencies.

Apart from inadequate design of feedback circuitry, amplifier instability is usually a result of bad circuit layout or poor supply decoupling. You need to pay particular attention to these aspects with wideband devices. Also, low-value capacitors at the input as recommended above in Figure 12.38 can affect the feedback to worsen the stability of the circuit, so they should be used with caution; the input pin series resistor method is preferable. Stability problems can be diagnosed without access to RF immunity test facilities by feeding a square wave with a fast rising edge into the circuit, and looking for ringing on the output.

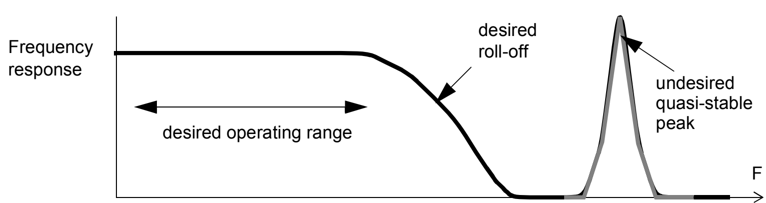


Figure 12.41 Amplifier quasi-stability

12.2.6.5 Isolation

Signals may be isolated at input or output with either an opto-coupler or a transformer (Figure 12.42). The ultimate expression of the former is fibre optic data transmission, which with the falling costs of fibre optic components is becoming steadily more attractive in a wide range of applications. Given that the major interference coupling route is via the connected cables, using optical fibre instead of wire completely removes this route. This leaves only direct coupling to the enclosure, and coupling via the power cable, each of which is easier to deal with than electrical signal interfaces. Signal processing techniques will be needed to ensure accurate transmission of precision AC or DC signals, which increases the overall cost and board area.

Coupling capacitance

Isolation breaks the electrical ground connection and therefore substantially removes common mode noise injection, as well as allowing a DC or low frequency AC potential difference to exist. However there is still a residual coupling capacitance between primary and secondary which will compromise the isolation at high frequencies or high rates of common mode dv/dt. This capacitance is typically 2–3pF per device for an opto-coupler; where several channels are isolated the overall coupling capacitance (from one ground to the other) rises to several tens of pF. This common mode

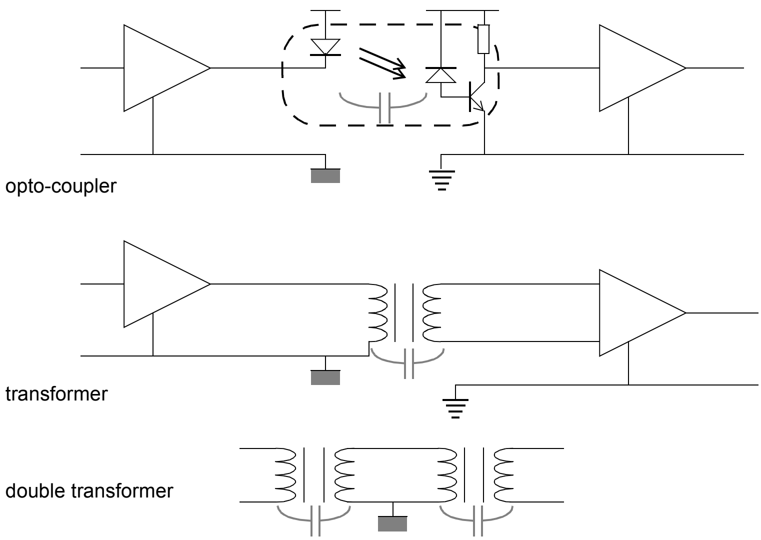


Figure 12.42 Signal isolation

impedance is a few tens of ohms at 100MHz, which is not much of a barrier!

Electrostatically screened transformers and opto-couplers are available where the screen reduces the coupling of common mode signals into the receiving circuit, and hence improves the common mode transient immunity *of that (local) circuit*. This improvement is gained at the expense of increasing the overall capacitance across the isolation barrier, and hence reducing the impedance of the transient or RF coupling path to the rest of the unit. A somewhat expensive solution to this problem is to use two unscreened transformers in series, with the intervening coupled circuit separately grounded.

A particular problem with opto-couplers, not suffered by transformers, is that high-level but short transients can couple into the base of the receiving photo-transistor and cause its saturation. This then effectively "blinds" the coupler for however long it takes for the device to come out of saturation, which may be several microseconds; the overall impact on the circuit is to apparently stretch the duration of the transient to a potentially unacceptable extent. The same effect happens with RF interference, and is due to capacitive coupling directly to the base (Figure 12.43), especially when this is brought out to a separate termination on the package. The problem is not as bad with photodiode couplers, and can be mitigated with a phototransistor by connecting its base to emitter with a low-value resistor, at the expense of sensitivity.

Brøndegaard Nielsen [102] investigated a number of widely-available opto-couplers in the presence of fast transient and RF interference. He concluded that

> 2kV transients always caused some short wrong signals on the optocoupler output ... you have to accept a wrong signal of up to 1µs on each single transient of the burst, if you want the optocoupler to insulate fast transient voltages ... if you cannot accept a short time wrong signal from an optocoupler, you have to limit the dV/dt value on the galvanic barrier by common mode

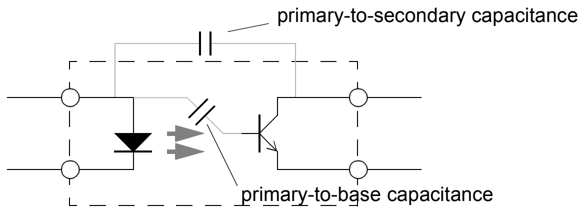


Figure 12.43 Parasitic coupling in the opto-coupler

> filters, or you have to limit the signal bandwidth of the transferred signal by a differential mode filter after the output of the optocoupler ... the most transient susceptible optocouplers were found among the types with an external base pin.

It is best to minimize the number of channels by using serial rather than parallel data transmission. Do not compromise the isolation further by running tracks from one circuit near to tracks from the other. Figure 10.4 on page 226 will give some idea of the degree of mutual coupling versus track spacing. Isolated sections must be laid out to prevent any coupling to other parts of the circuit, including ground planes: the main (non-isolated) 0V plane should stop halfway across the isolator, and no part of the isolated circuit should overlap it.

Chapter 13

Interfaces and filtering

Figure 11.1 on page 258 divided EMC control measures into three levels. This chapter deals with the secondary (interfaces and filtering) level. If your product has no interfaces – let's say it is a hand-held, battery operated device – then you already have a head start when it comes to EMC since only the enclosure port is relevant to you. But for the rest of us, power, signal and data interfaces are a fact of life and they offer a ready route for disturbances into and out of a product. The first thing that gets connected to an interface is a cable, so we need to consider how cables couple these disturbances, and if they are to be screened, how the screen connection needs to be dealt with in practice; and then if the interface is unscreened, how to specify and implement the necessary filtering.

13.1 Cables and connectors

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

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

13.1.1 The mode of propagation

Although this subject was covered in section 10.1.5, it is particularly important to appreciate the difference between common mode and differential mode currents on cables since this difference directly affects many of the principles that apply to cable design for EMC.

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

Common mode current, I_{CM}, flows equally in the same direction along all conductors in the cable, including the shield if this is present, and may or may not be related to the signal currents. That part of the signal current which does not return via the cable but leaks out through stray coupling, does appear as a common mode component; this aspect is related to the longitudinal conversion loss of the cable, which is discussed later (section 13.1.9.1). The other major source is the noise voltage developed within the circuit and referred between the point of connection of the cable, and the circuit's ground reference. This is why it is good practice to couple the circuit to ground at the interfaces, which will minimize this noise voltage.

I_{CM} returns via the associated ground network and therefore the loop area involved in the coupling is large and uncontrolled (Figure 13.1). As a result, even a small I_{CM} can result in large emitted signals, or an impinging field causes large levels of I_{CM}. On the other hand, the total differential mode currents I_{DM} sum to zero in the cable and hence they create no net magnetic field around the whole cable.

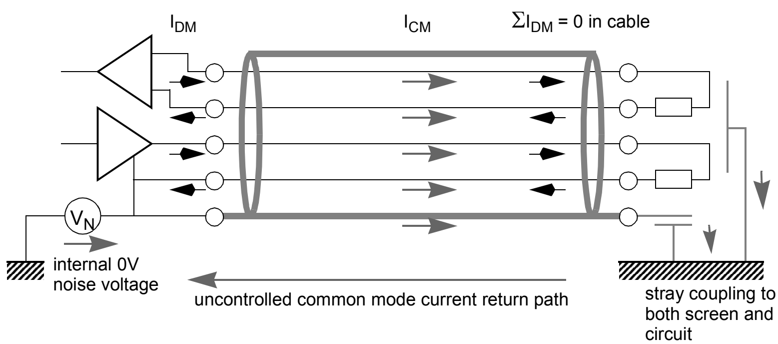


Figure 13.1 The distinction between differential and common mode cable currents

13.1.2 Cable return currents

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

It is not intuitively obvious that return currents will necessarily flow in the conductor which is local to the signal wires, when there may be several alternative return paths for them to take – for instance, an external ground connection at each end of the cable. At DC, the return currents are indeed shared between paths only by the ratio of conductor resistances. But as the frequency increases, the mutual inductance of the coupled pair (twisted, coaxial or simply adjacent in the cable bundle) tends to reduce the impedance presented to the return current by its local return compared to other paths, because the enclosed loop area is smallest for this path (Figure 13.2). The effect relies on the fact that the return current is equal in magnitude but opposite in sign to the signal current. Therefore their respective magnetic fields will tend to cancel, the cancellation being complete if the two wires are co-located. If the magnetic fields have cancelled, then by definition the inductance of the path is zero. In practice, co-location of conductors is naturally impossible, but it can be closely approximated by twisting the pair or by adopting a coaxial geometry.

This effect is also responsible therefore for the magnetic shielding property of

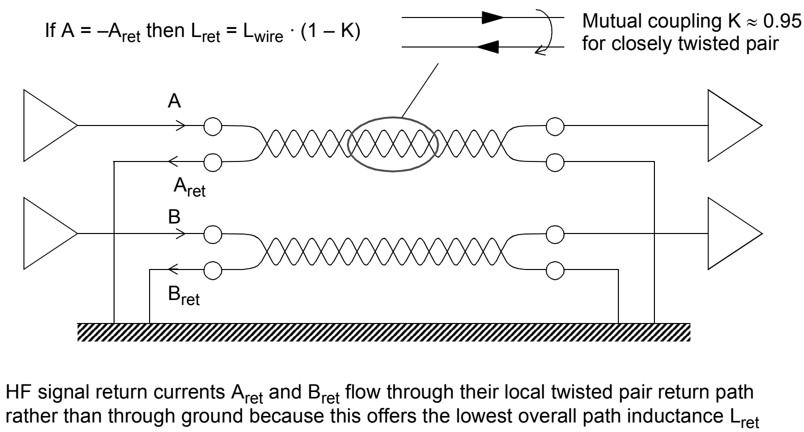


Figure 13.2 Signal return current paths

coaxial cable, and is the reason why current in a ground plane remains local to its signal track (compare section 11.2.2).

13.1.3 Crosstalk

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

Visualize two circuit pairs in a single cable (Figure 13.3). Along the length of the cable there is distributed capacitance between every conductor and each of the other three conductors. Similarly, there is mutual inductance linking every conductor to each of the others. At frequencies where the cable length is much shorter than a wavelength, the L and C can be simplified to a matrix of discrete reactances, but at higher frequencies it is necessary to assign elemental reactances to an infinitesimally short length and then integrate these over the length of the cable.

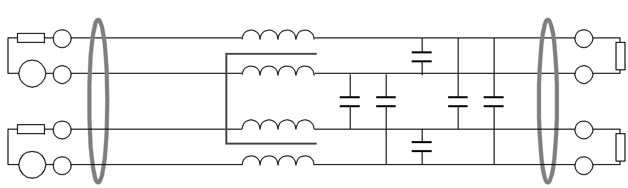


Figure 13.3 Intra-cable crosstalk

The mutual capacitance and inductance between the two conductors that form each circuit pair are benign, and determine the characteristic impedance of that pair (from Z_0

= √L/C, assuming no losses). In conjunction with the circuit driving and load impedances they will determine the bandwidth capability of the cable/equipment system. But the mutual impedances *between the pairs* are undesirable. These result in crosstalk interference between the two circuits.

13.1.3.1 Capacitive crosstalk

The voltages appearing on + and – of pair 1 are coupled by the mutual capacitances onto + and – of pair 2. The amplitude of the induced voltage is determined by the values of the capacitances and the circuit impedances, and the rate of change of source voltage (dv/dt). Balanced circuits, and a balanced cable construction which equalizes the capacitances, will minimize the effective crosstalk since voltages induced on or from the + conductor will be nearly cancelled by those induced on or from the – conductor. Unbalanced circuits with high dv/dt and high impedances will be the most susceptible to capacitive crosstalk. Screening each pair individually (Figure 13.4) will remove the capacitive crosstalk almost entirely, since the mutual capacitances between pairs are eliminated, to be replaced by mutual capacitance from each pair to its screen and mutual capacitance between screens. (Screens without 100% optical coverage, such as braids, will still allow a small amount of capacitance directly between conductors, through the gaps in the screen.) The screens must of course be connected to a fixed potential (which may be system earth, or sometimes circuit 0V); voltages will still be developed longitudinally along the screens as a result of their resistance, and this along with other screen imperfections is then the limiting factor in capacitive crosstalk suppression.

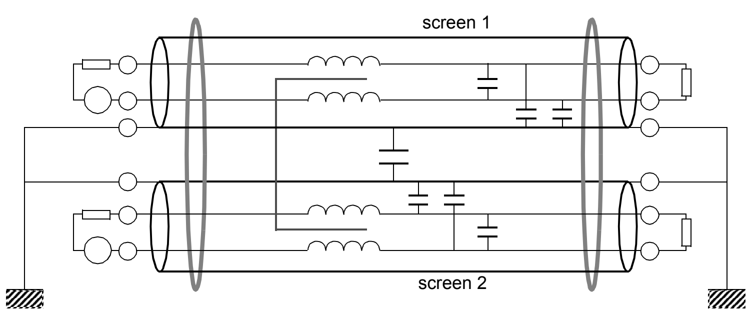


Figure 13.4 Screening against capacitive crosstalk

13.1.3.2 Inductive crosstalk

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

13.1.3.3 Distributed crosstalk

At high frequencies the cable must be considered as a distributed structure and the mutual impedances of elemental lengths have to be integrated over the whole length. The phase differences along the length become significant, and the contributions of inductive and capacitive crosstalk result in constructive interference at each end at some frequencies, and destructive interference (nulls in the coupling) at each end at others. What is more, the sense of the wave travelling down the cable becomes significant, and it is necessary to talk of "near-end" crosstalk (NEXT) as distinct and different from "far-end" crosstalk (FEXT). This book does not attempt to go into the detail of distributed crosstalk analysis – see Chapter 10 of Paul [12] or section 4.3 of Tsaliovich [17] for an in-depth treatment.

13.1.4 Cable screens at low frequencies

Optimum screening requires different connection regimes for interference at low frequencies (audio to a few hundred kHz) and at radio frequencies. These two regimes may well be mutually exclusive – the best example being the oft-quoted advice to ground the screen at one end only, which may be correct for low frequencies but is definitely incorrect for high frequencies. Now that you have to meet RF emissions and immunity requirements for the EMC Directive, let alone for practical performance in the field, this advice is obsolete. If grounding the screen at both ends causes problems, then there is most probably a deficiency in the system grounding scheme. To see why this is so, consider the circuit of Figure 13.5.

13.1.4.1 Screen currents and magnetic shielding

An overall screen, grounded only at one end, provides good shielding from capacitively coupled interference (Figure 13.5(a)) but none at all from magnetic fields, which induce a noise voltage in the loop that is formed when both source and load are grounded. (Beware: different principles apply when either source or load is not grounded!) To shield against a magnetic field, *both* ends of the screen must be grounded. This allows an induced current (I_S in Figure 13.5(b)) to flow in the screen which will negate the induction effect in the centre conductor, as described in section 13.1.2. The effect of this current begins to become apparent only above the cable cut-off frequency F_C, which is a function of the screen inductance and resistance and is around 1–2kHz for braided screens or 7–10kHz for aluminium foil screens. Above about five times the cut-off frequency, the voltage induced in the centre conductor remains constant with frequency (Figure 13.5(c)) and therefore the shielding effect continues to increase with frequency.

The same principle applies when you are shielding a conductor to prevent magnetic field *emission*. The return current must flow through the screen, and this will only occur (for a circuit which is grounded at both ends) at frequencies substantially above the shield cut-off frequency. Hence the difficulty in shielding against low frequency magnetic fields, which is explored further in section 14.1.2.

We can note at this point that, if we are only interested in shielding against low frequency *capacitively* induced interference, there is nothing wrong with the single-ended grounding method of Figure 13.5(a) and it is, of course, widely used in instrumentation systems for precisely this purpose.

13.1.4.2 Where to ground the cable screen

There are two problems with grounding the screen at both ends in the circuit of Figure 13.5(b). One is that it becomes a circuit conductor and any voltage dropped across the

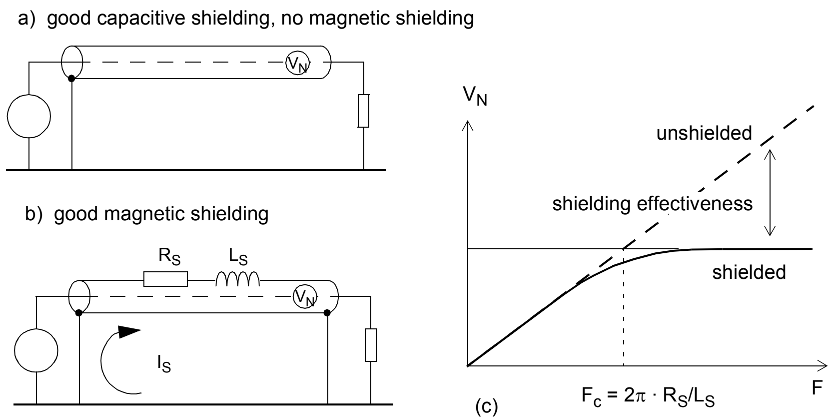


Figure 13.5 Magnetic shielding effectiveness versus screen grounding

screen impedance will be injected in series with the signal. Whenever a *circuit* is grounded at both ends, only a limited amount of magnetic shielding is possible because of the large interference currents induced in the screen-ground loop, which develop an interference voltage along the screen. To minimize low frequency magnetic field pick-up, one end of the circuit should be isolated from ground, the circuit loop area should be small, and the screen should not form part of the circuit. You can best achieve this by using shielded twisted pair cable with the screen grounded at only one end. The screen then takes care of capacitive coupling while the twisting minimizes magnetic coupling.

For a circuit with an ungrounded source the screen should be grounded at the input common, whereas if the input is floating and the source is grounded then the screen should be grounded to the source common. These arrangements (Figure 13.6) minimize capacitive noise coupling from the screen to the inner conductor(s), since they ensure the minimum voltage differential between the two. Notice though that as the frequency increases, stray capacitance at the nominally ungrounded end reduces the efficiency of either arrangement by allowing undesired ground and screen currents to flow.

Undesired screen currents

The second problem is that if there is a significant voltage difference between the grounds at the two ends of the screen, connecting both ends will encourage the screen to carry a resulting current. This will be limited only by the impedance of the screen and the source impedances of the voltage differential; both of these could be very low, and the current that flows could be enough to damage the cable. Such voltage differentials are not unusual on large sites or between buildings. This is often the practical reason cited for the installation engineer's mantra, don't connect the screen at both ends.

The preferred solution here is to bite the bullet and connect a parallel earth conductor (as discussed in Chapter 15, and [22] and [156]) along the length of the installed cable so that it takes the brunt of the earth currents that may flow. This solution has historically been resisted by installation engineers on grounds of cost, but if in fact the PEC is a suitably bonded metallic cable tray or conduit, the actual cost can be a lot

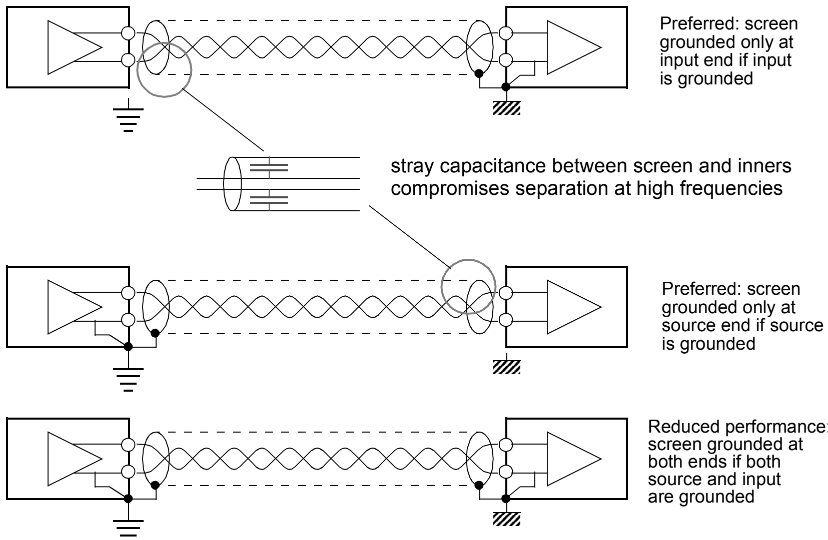


Figure 13.6 Screen grounding arrangements versus circuit configuration

less than is imagined. This allows the cable screen itself to be bonded at both ends, or not, as EMC circumstances dictate.

But as we are about to see, if you use a screened cable with no other precautions and leave the screen open at one end, your RF shielding at the unconnected end is non-existent. The above discussion applies only for *low frequency* shielding. Not everybody appreciates this state of affairs: even standards writers can be caught out, as the following quote from the NMEA 2000 specification for marine electronic data communications [101] illustrates:

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

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

13.1.5 Cable screens at RF

Once the cable length approaches a quarter wavelength at the frequency of interest†, screen currents due to external fields become a fact of life. An open circuit at one end of the cable becomes transformed into a short circuit a quarter wavelength away, and screen currents flow in a standing wave pattern whether or not there is an external

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

connection (Figure 13.7). The magnitude of the current is related to the characteristic impedance of the transmission line formed by the cable and the ground reference (this behaviour is discussed in section 10.3.1.1). Even below resonant frequencies, stray capacitance can allow screen currents to flow.

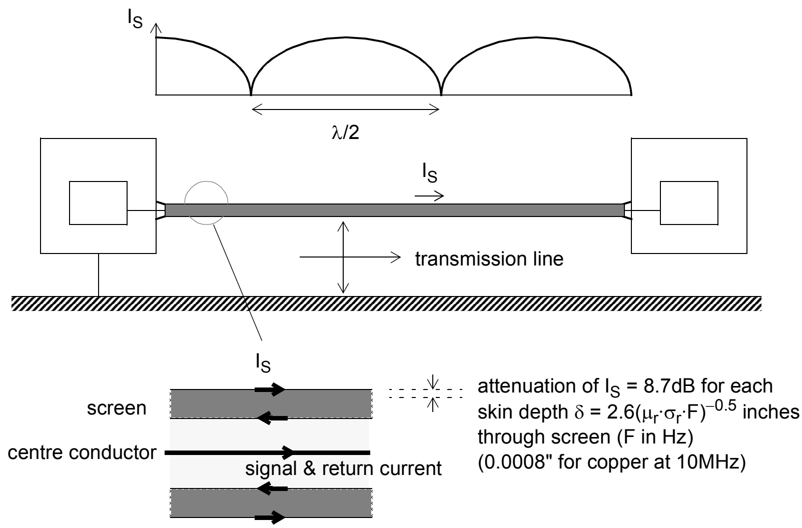


Figure 13.7 The cable screen at RF

Separation of inner and outer screen currents

However, at high frequencies the inner and outer of the screen are isolated by skin effect, which forces currents to remain on the surface of the conductor. Signal currents on the inside of the screen do not couple with interference currents on the outside. The skin depth as shown in Figure 13.7 is the depth at which surface currents are attenuated by 8.7dB and depends on the reciprocal of the square root of the material properties and frequency. At 10MHz in copper it is 0.8 thou or 0.021mm. So a solid screen thickness of 0.25mm would give an attenuation from outer to inner surface of 103dB. Thus multiple grounding of the screen, or grounding at both ends, although it deliberately allows screen currents to flow, does not introduce interference voltages on the inside to the same extent as at low frequencies. This effect is compromised by a braided screen due to its incomplete optical coverage – that is, there are small gaps between braid strands – and because the strands are continuously woven from inside to out and back again. It is also more seriously compromised by the quality of the screen ground connection at either end, as is discussed in section 13.1.7.

13.1.6 Types of cable screen

The performance of cable screens depends on their construction. Figure 13.8 shows some of the more common types of screen available commercially at reasonable cost; for more demanding applications specialized screen constructions such as optimized or multiple braids are available at a premium. Of course, you can also run unscreened cable in shielded conduit, in a separate braided screen or wrap it with screening or

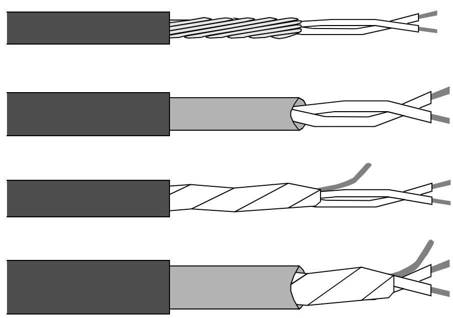


Figure 13.8 Common screen types

permeable material. These options are most useful for systems or installation engineers.

- *Lapped wire* screens consist of wires helically wound onto the cable. They are very flexible, but have poor screening effectiveness and are noticeably inductive at HF, so are restricted to audio use.
- *Single braid* screens consist of wire woven into a braid to provide a metallic frame covering the cable, offering 80–95% coverage and reasonable HF performance. The braid adds significantly to cable weight and stiffness.
- *Laminated tape or foil* with drain wire provides a full cover but at a fairly high resistance and hence only moderate screening efficiency. Light weight, flexibility, small diameter and low cost are retained. Making a proper termination to this type of screen is difficult; screen currents will tend to flow mainly in the drain wire, making it unsuitable for magnetic screening since the geometry doesn't allow for proper field cancellation, although its capacitive screening is excellent.
- *Composite foil and braid* combines the advantages of both laminated foil and single braid to optimize coverage and high frequency performance.
- *Multiple braid* screens improve the performance of single braids by deliberately separating the inner and outer current flows, and allowing the screens to be dedicated to different (low and high frequency) purposes.

13.1.6.1 Surface transfer impedance

The screening performance of shielded cables is best expressed in terms of surface transfer impedance (STI). This is denoted by Z_T and is a measure of the voltage induced per unit length on the inner conductor(s) of the cable by an interference current flowing down the cable outer shield, which will vary with frequency and is normally expressed in milliohms per metre. (The same parameter was introduced in section 11.1.2.2 to describe the performance of grounding structures in general.) A perfect screen would not allow any voltage to be induced on the inner conductors and would have an STI of zero, but practical screens will couple some energy onto the inner via the screen impedance. At low frequencies it is equal to the DC resistance of the screen.

Figure 13.9 compares STI versus frequency for various types of cable screen construction. The initial decrease in STI with frequency for the better performance

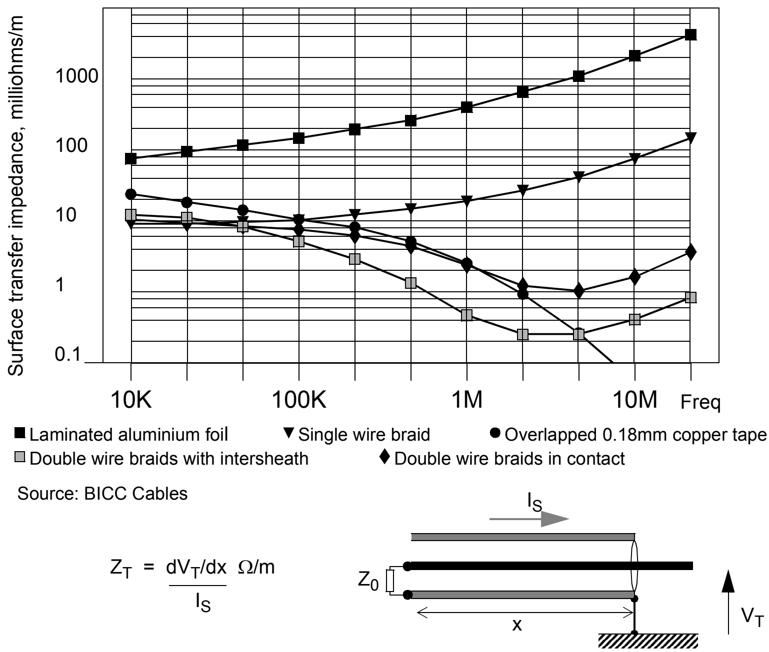


Figure 13.9 Surface transfer impedance of various screen types

screens is due to the skin effect separating signal currents on the inside of the screen from noise currents on the outside. The subsequent increase is due to field distortion by the holes and weave of the braid. Once the frequency approaches cable resonance then STI figures become inapplicable to real situations. This is because the condition of constant screen current I_S per unit length cannot in general be met, because of standing waves. Only when the cable *and its external ground reference* form a transmission line of known characteristic impedance, and this transmission line is properly terminated at each end, will the RF current be constant along the length. A special jig which ensures this condition is one method for measuring Z_T; standardized methods have been published in IEC 62153-4 [159] and EN 50289-1-6 [151], and an introduction to the subject can be found in IEC TR 61917 [158].

Note that the inexpensive types have a worsening STI with increasing frequency, and that the laminated foil screen is approximately 20dB worse than a single braid, due to its higher resistance and to the field distortion introduced by the drain wire, which carries the major part of the longitudinal screen current. A solid copper screen, on the other hand, just gets better and better with increasing frequency because its skin effect is not compromised.

13.1.7 Screened cable connections

13.1.7.1 How to ground the cable shield

The over-riding requirement for terminating a cable screen is a connection direct to the metal chassis or enclosure ground which exhibits the lowest possible impedance. This ensures that interference currents on the shield are routed to ground without passing through or coupling to other circuits. The best connection in this respect is one in which the shield is extended up to and makes a solid 360° connection with the ground plane or chassis (Figure 13.10). This is best achieved with a hard-wired cable termination using a conductive gland and ferrule which clamps over the cable screen. A connector will always compromise the quality of the screen-to-chassis bond, but some connectors are very much better than others.

Connector types

Military-style connectors allow for this construction, as do the standard ranges of RF coaxial connectors such as N type or BNC. Of the readily available commercial multi-way connectors, only those with a connector shell that is designed to make positive 360° contact with its mate are suitable. Examples are the subminiature D range with dimpled tin-plated shells. Connector manufacturers are now introducing properly designed conductive shells for other ranges of mass-termination connector as well, and the new generation of high-speed data connectors such as USB and DVI are designed from the outset with this in mind.

The importance of the backshell

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

13.1.7.2 The effect of the pigtail

A pigtail connection is one where the screen is brought down to a single wire and extended through a connector pin to the ground point. Because of its ease of assembly it is very commonly used for connecting the screens of data cables. Indeed, some older data interface standards explicitly provide a pin in the connector for this purpose. Unfortunately, it may be almost as bad as no connection at high frequencies because of the pigtail inductance [87][110]. This can be visualized as being a few tens of nanohenries in series with the cable screen connection (Figure 13.11), which develops a common mode voltage on the screen at the interface as a result of the interference current I_S flowing down the screen. This voltage then couples readily onto the inner conductors, or vice versa, noise voltages on the inner conductors couple readily out onto the screen.

The equivalent surface transfer impedance of such a connection rises rapidly with increasing frequency until it is dominated by the pigtail inductance, and effectively negates the value of a good HF screened cable. At higher frequencies resonances with the stray capacitances around the interface limit the impedance, but they also make the actual performance of the connection unpredictable and very dependent on construction

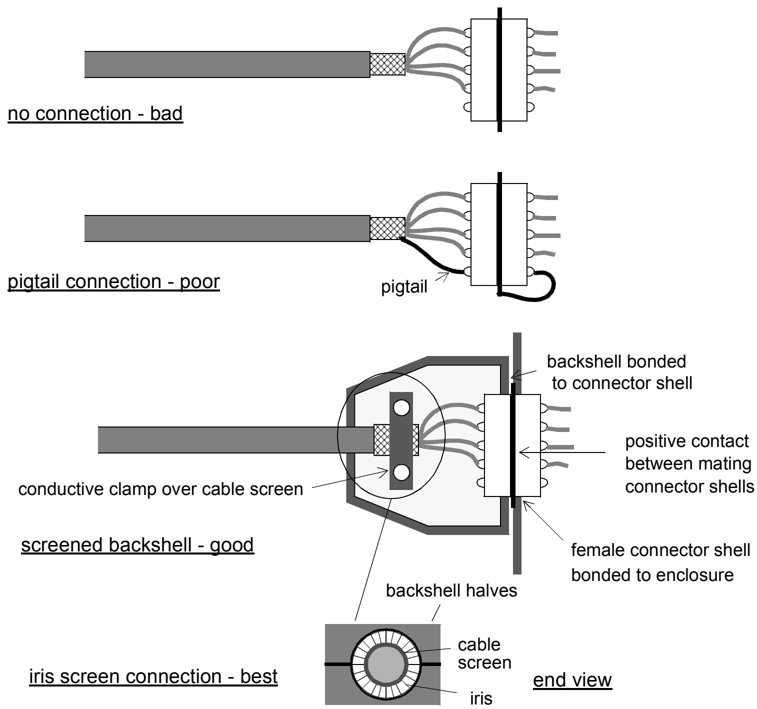


Figure 13.10 Cable screen connection methods at RF

and movement. If a pigtail connection is unavoidable then it must be as short as possible, and preferably doubled and taken through two pins on opposite ends of the connector so that its inductance is halved. Even with this precaution, it is best to regard interfaces at which you know that a pigtail is involved as essentially unscreened, and to apply good practice filtering as well.

The connector shell to chassis link

The effective length of the pigtail extends from the end of the cable screen through the connector and up to the point of the ground plane or chassis connection. The once-common practice of mounting screened connectors on a PCB with the screening shell taken to ground via a length of track – which in the worst case travels the length of the board – is equivalent to deliberate insertion of a pigtail on the opposite side of the connection. Screened connectors must always be mounted so that their shells are bonded directly to chassis, sometimes even needing a conductive gasket to ensure the bond. A lower performance compromise that is sometimes acceptable is to take the shell to a “quiet ground” (not circuit 0V) on the PCB, which is then taken via pillars to the chassis (see section 11.2.3).

13.1.7.3 Terminating shielded cables without connectors

Systems builders are frequently faced with the requirement to bring many field cables into a cabinet and wire these to terminal blocks, whence the signals are routed to the

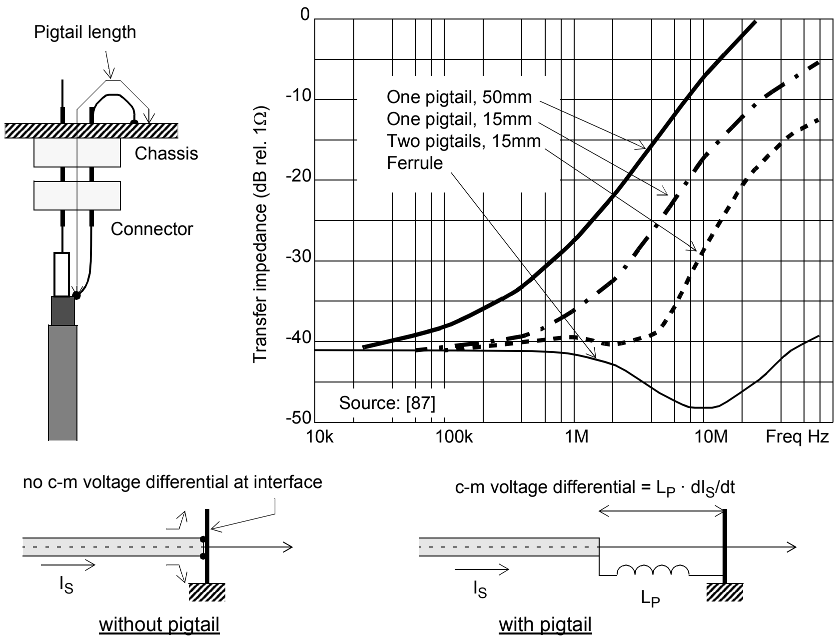


Figure 13.11 The bane of the pigtail

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

13.1.8 Unscreened cables

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

13.1.8.1 Twisted pair

Twisted pair is a particularly effective and simple way of reducing both magnetic and capacitive interference pick-up. Twisting the wires tends to ensure a uniform distribution of capacitances to structures outside the cable. Both capacitance to ground and to extraneous sources are balanced. This means that common mode capacitive coupling is also balanced, allowing high common mode rejection provided that the rest of the circuit is also balanced.

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

The crosstalk or external coupling to any unscreened cable, whether or not it is twisted pair, has both capacitive and inductive components. The equivalent circuit (Figure 13.12) depicts the capacitive coupling as a current source I_C from external sources onto each conductor half-twist, while the inductive coupling is a voltage source V_C in series with each conductor.

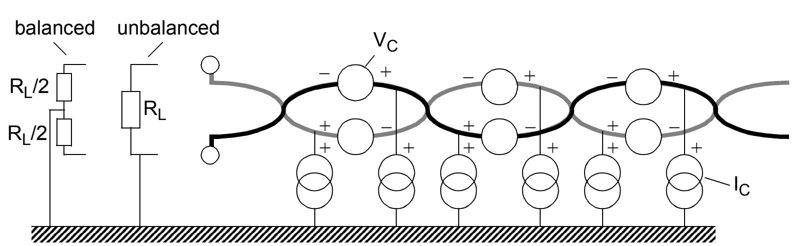


Figure 13.12 Equivalent circuit for coupling to twisted pair

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

Balancing the circuit eliminates (to a first order) the effect of capacitive coupling, since the current sources sum equally in common mode in both halves of the differential circuit, and allows the crosstalk to be determined purely by residual inductive coupling. This will be sensitive to the uniformity of the twists and of the field through which the cable passes, but is unaffected by whether the circuit is or is not balanced.

13.1.8.2 Ribbon cable

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

Ribbon should be shielded if it carries high frequency signals and is extended outside a screened enclosure, but you will find that proper termination of the shield is usually incompatible with the use of a mass-termination connector. Cable designs can be obtained with an integral ground plane underneath the conductors, or with full coverage screening [107]. However, the shielding performance figures for ground plane and shielded cables assume a low-inductance termination, which is difficult to achieve in practice; typical terminations via drain wires will worsen this performance, more so at high frequencies.

Ground configuration in ribbon

The performance of a ribbon cable carrying high frequency data is very susceptible to the configuration of its ground return(s). The cheapest configuration is to use one ground conductor for the whole cable (Figure 13.13(a)). This creates a large inductive loop area for the signals on the opposite side of the cable, and crosstalk and ground impedance coupling between signal circuits. It is highly undesirable, but if for other reasons you must have the minimum number of conductors, then at least place the ground conductor in the middle of the cable and place the most aggressive or most sensitive signals adjacent to it.

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

13.1.8.3 Ground plane flexi

A particularly effective way to carry high frequency signals between boards within a product, and which provides better performance than a ground plane ribbon while being cheaper and easier to use, is the ground plane flexi connector assembly (Figure 13.14). The double-sided flexi has one side dedicated to a ground plane while the other side carries the signal tracks. Alternate pins on the surface-mount connector at each end take the 0V and are via'd through to the ground plane. This low-impedance ground return, very close to the signal tracks, ensures that the minimum of ground noise is developed between the two ends of the circuit.

This construction gives a simple, cheap and effective mass-termination connection system. Flexis can be designed for any physical configuration; some suppliers will do

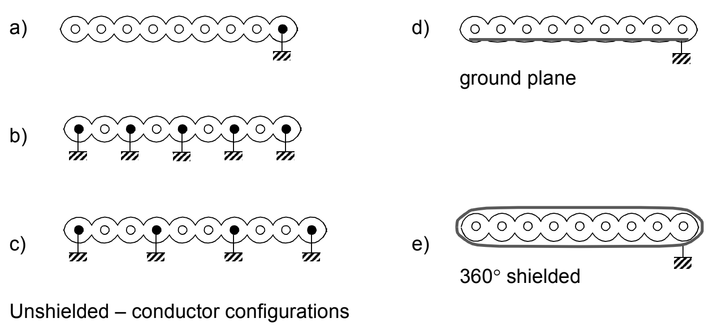


Figure 13.13 Ribbon cable configurations

simple designs as standard parts. The method of construction and assembly is exactly like that of a standard PCB except that the substrate is thin and flexible rather than being rigid fibreglass.

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

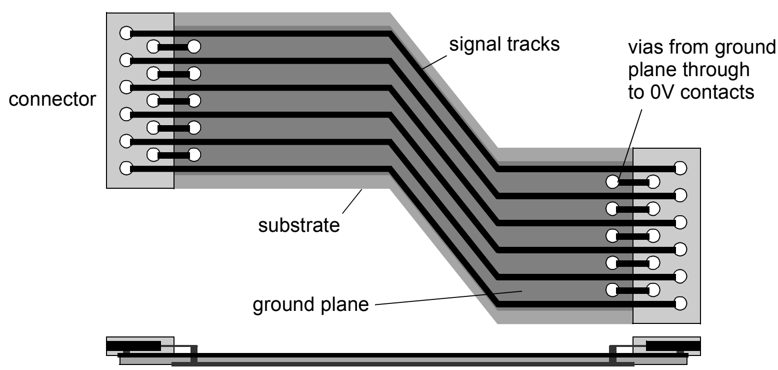


Figure 13.14 The ground plane flexi

13.1.8.4 Ferrite loaded cable

Common mode currents in cable screens are responsible for a large proportion of overall radiated emission. A popular technique to reduce these currents is to include a common mode ferrite choke around the cable, typically just before its exit from the enclosure – see section 13.2.2.1. Such a choke effectively increases the HF impedance of the cable to common mode currents without affecting differential mode (signal) currents.

An alternative to discrete chokes is to surround the screen with a continuous coating of flexible ferrite material. This has the advantage of eliminating the need for an extra component or components, and since it is absorptive rather than reflective it reduces discontinuities and hence possible standing waves at high frequencies. This is particularly useful for minimizing the effect of the antenna cable when making radiated field measurements (section 6.5.2.3). It can also be applied to unscreened cables such as mains leads. Such "ferrite-loaded" cable is unfortunately expensive, not widely available and, like other ferrite applications, is only really effective at very high frequencies. Its use is more suited to one-off or ad hoc applications than as a production item. It can be especially useful when transients or ESD conducted along the cable are troublesome in particular situations.

13.1.9 Structured cabling: UTP versus STP

A common issue for equipment with data communication ports is how it will interface with cabling that has already been installed within a building, or "structured cabling". International and European specifications for such cabling have been published as ISO/IEC 11801 and EN 50173 [146]. These documents classify cables by their performance, as Category 3, 5, or 6, with enhanced and extended specifications in preparation. The most important specifications from the operational perspective are near-end crosstalk, attenuation and characteristic impedance, and these are laid down in the standards. Unshielded twisted pair (UTP) is in principle capable of meeting these requirements, and there is now a large installed base of such cabling in commercial premises to allow the widespread adoption of local area networks.

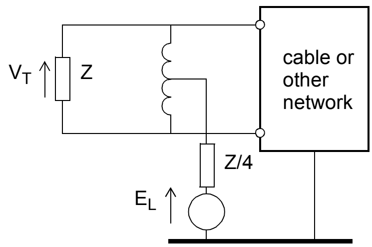
There is no clear preference as to whether the data cable should be shielded or unshielded. A good quality shielded cable can ensure minimum coupling with its environment, but only if it is properly terminated (see section 13.1.7) and if the shield is maintained unbroken along its length – not an easy matter if it includes joints or patch panels. If this is not the case, then a well-specified UTP cable is likely to be better overall, if the equipment to which it is connected has good common mode rejection.

If you are designing a product with a LAN or telecom data interface then you will need to decide which type of cable to use. If shielded, then you must provide for a correctly terminated shielded connector, and ensure that the installation uses this connector with the proper cable in the right way. If unshielded, then the connector is less important, but the interface must be actively designed for good common mode rejection, which will mean ensuring that the physical layout is balanced and, usually, incorporating a wideband common mode choke (section 13.2.4). It will also be necessary to specify the Longitudinal Conversion Loss (LCL) of allowable connected cables – typically by restricting them to one or other of the IEC 11801 categories.

13.1.9.1 Longitudinal Conversion Loss

The LCL of a balanced cable system – or indeed any one- or two-port network – is a measure of the mode conversion exhibited by the system, that is the degree to which an inadequately balanced termination will develop an unwanted transverse (differential) signal when excited by a longitudinal (common mode) signal. It is measured as shown in Figure 13.15 [194]. Although this diagram shows a differential mode signal generated by a common mode input, the principle is reciprocal and can be used to describe unwanted common mode signals developed by intended differential signals.

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



$$\mathrm{LCL} = 20 \log_{10} \left| \frac{E_L}{V_T} \right| \ \mathrm{dB}$$

Figure 13.15 Longitudinal conversion loss

$$I_{CM}\,(\mathrm{dB\mu A}) = U_T(\mathrm{dB\mu V}) - \mathrm{LCL}\,(\mathrm{dB}) - 20 \log_{10} \left|\{2Z_0 \cdot (Z_{cm} + Z_{ct})/(Z_0 + 4Z_{cm})\}\right| \quad (13.1)$$

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

So for a Class B current limit of 30dBµA, as per CISPR 22 third edition, an LCL for category 3 cable of 50dB, a Z_0 of 100Ω, and $Z_{cm} = Z_{ct} = 25\Omega$, the maximum permissible signal level at any given frequency is 114dBµV or 0.5V. The higher the common mode impedances Z_{cm} and Z_{ct}, the more differential signal can be allowed; but more importantly, the lower the LCL, the greater the level of interference that is created. This has important implications both for testing conducted emissions from telecom ports, as discussed in section 6.2.2.4, and also for using poorly specified cables for passing broadband data – such as for broadband over power lines (section 1.3.5).

13.2 Filtering and suppression

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

13.2.1 Filter configuration

In EMC work, “filtering” almost always means low-pass filtering. The purpose is normally to attenuate high frequency components, which are assumed to be interference, while passing low frequency ones which are assumed to be wanted. Various simple low-pass configurations are shown in Figure 13.16, and filter circuits are normally made up from a combination of these. The effectiveness of the filter configuration depends on the impedances seen at either end of the filter network.

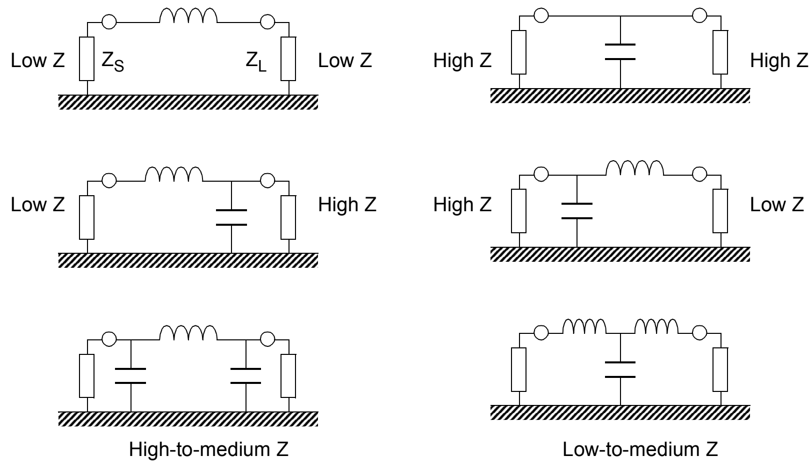


Figure 13.16 Filter configuration versus impedance

13.2.1.1 Source and load impedances

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

Conventionally, filters are specified for terminating impedances of 50Ω at each end because this is convenient for measurement and is an accepted RF standard. If you are sourcing filters specified in this way, it is a convenient way to compare the performance of different units but it won't help you much in knowing how much attenuation to expect in the actual circuit. In the real application, Z_S and Z_L are complex and perhaps unknown at the frequencies of interest for suppression. If either or both has a substantial reactive component then resonances are created which may convert an insertion loss into an insertion gain at some frequencies. Differential mode impedances may be predictable if the components which make up the source and load are well characterized at RF, but common mode impedances such as are presented by cables or the stray reactances of mechanical structures are essentially unpredictable. Practically, cables have been found to have common mode impedances in the region of 100 to 400Ω except at resonance, and a figure of 150Ω is commonly taken as a rule of thumb (see also section 10.3.1.1).

13.2.1.2 Parasitic reactances

Filter components, like all others, are imperfect. Inductors have self-capacitance, capacitors have self-inductance. This complicates the equivalent circuit at high frequencies, and means that a typical filter using discrete components will start to lose its performance above a break frequency determined by its parasitics. The larger the components are physically, the lower will be the break frequency. For capacitors, as the frequency increases beyond capacitor self-resonance the impedance of the capacitors in the circuit actually rises, so that the insertion loss begins to fall. This can be countered by using special construction for the capacitors (see the next section). Similarly, inductors have a self-resonant frequency beyond which their impedance starts to fall. Filter circuits using a single choke are normally limited in their performance by the self-resonance of the choke (see Figure 13.17) to 40 or 50dB. Better performance than this requires multiple filter sections.

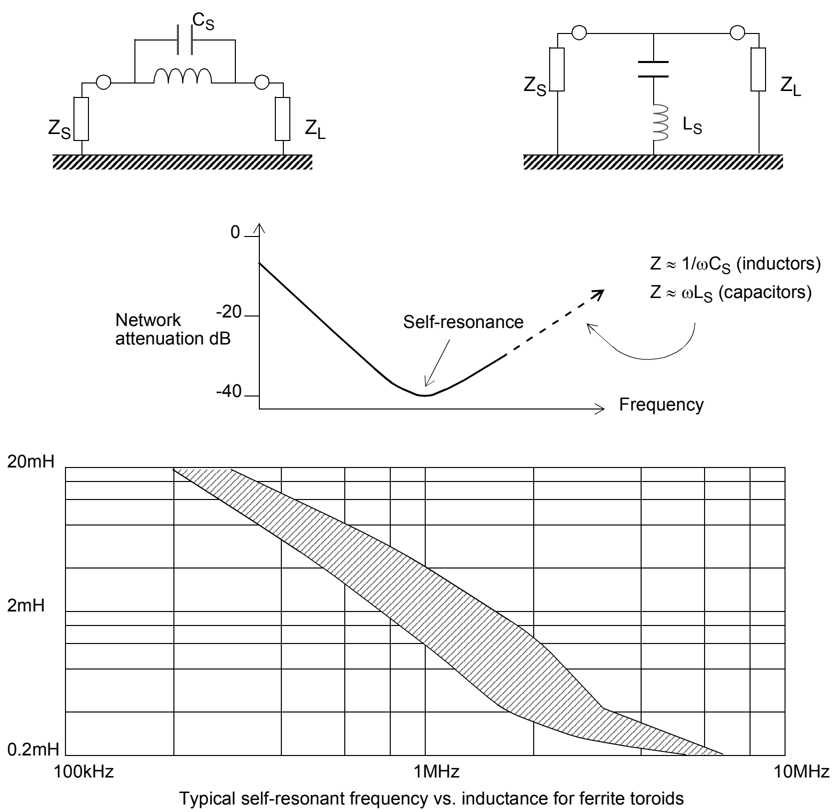


Figure 13.17 Self-resonant effects due to parasitic reactances [55]

Capacitors

Ceramic capacitors are usually regarded as the best for RF purposes, and they are of course easy to source. Small capacitors with short leads (preferably a chip component) will have the lowest self-inductance. The inductance is made up of a combination of that due to the attached leads and that due to the package; a chip component only exhibits the latter, and their self-resonant frequency can therefore be double that of a leaded capacitor of the same value. Some extra inductance is added by the tracking and vias, and tracks to capacitors used for filtering and decoupling should be short and direct, in order not to lose this advantage through additional track inductance. As we have seen in the discussion on decoupling (section 12.1.3), smaller packages give better results: for instance widely-available chip ceramic data gives a self-resonant frequency for a 10nF component as 65MHz in 0402, 58MHz in 0603, 56MHz in 0805 and 45MHz in 1206. Unless you need a particularly high voltage rating, stay with the smaller parts.

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

Inductors

The more turns N an inductor has, the higher will be its inductance (proportional to N^2) but also the higher its self-capacitance. The number of turns for a given inductance can be reduced by using a high permeability core, but these also exhibit a high dielectric constant which tends to increase the capacitance again, for which reason you should generally use a bobbin on a high-permeability core rather than winding directly onto the core. For minimum self-capacitance the start and finish of a winding should be widely separated; winding in sections on a multi-section bobbin is one way to achieve this. A single layer winding exhibits the lowest self-capacitance. If you have to use more turns than can be accommodated in a single layer, progressive rather than layer winding (see Figure 13.18) will minimize the capacitance.

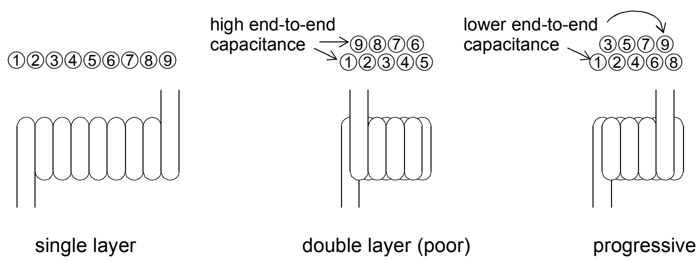


Figure 13.18 Inductor winding techniques

13.2.1.3 Component layout

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

A poor ground offers a common impedance which rises with frequency (section D.5.2) and couples through HF interference via the filter's local ground path. Common input-output wiring does the same thing through mutual capacitance or inductance, and it is also possible for the "clean" wiring to couple with the unfiltered side through inappropriate routing. The cures (Figure 13.19) are to directly couple the filter's ground terminal to the lowest inductance ground of the equipment, preferably the chassis, and to keep the I/O leads separate, preferably screened from each other. It is best to position the filter so that it straddles the equipment shielding, where this exists.

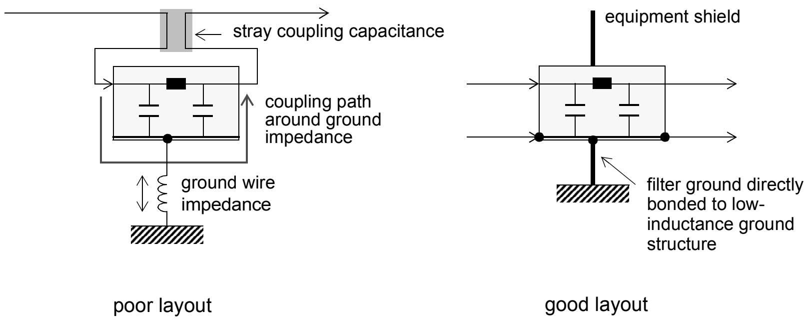


Figure 13.19 The effect of filter layout

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

13.2.2 Components

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

13.2.2.1 Ferrites

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

As described in section 10.1.1.2, current flowing through a conductor creates a magnetic field around it. Transfer of energy between the current and the magnetic field is effected through the "inductance" of the conductor – for a straight wire the self-inductance is typically 20nH per inch. Placing a magnetically permeable material around the conductor increases the flux density for a given field strength and therefore increases the inductance.

Ferrite is such a material; its permeability is controlled by the exact composition of the different oxides that make it up (ferric, with typically nickel and zinc) and depends heavily on frequency. Also the permeability is complex and has both real and imaginary parts, which translate into both inductive and resistive components of the impedance "inserted" into the line passed through the ferrite (Figure 13.20). The ratio of these components varies with frequency – at the higher frequencies the resistive part dominates (the ferrite can be viewed as a frequency-dependent resistor) and the assembly becomes lossy, so that RF energy is dissipated in the bulk of the material and resonances with stray capacitances are avoided or damped.

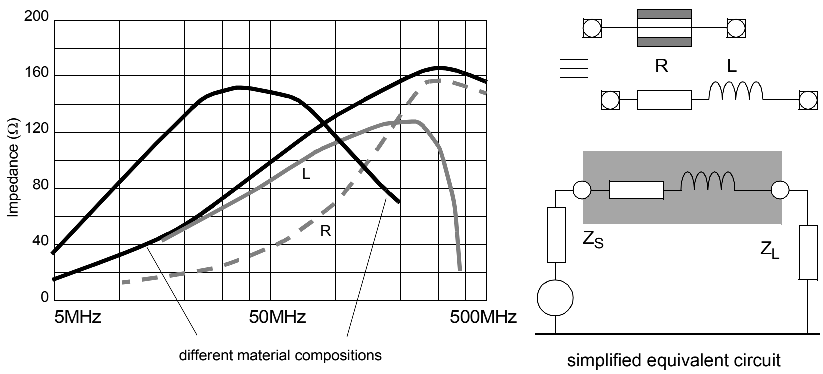


Figure 13.20 Ferrite impedance versus frequency and equivalent circuit

Cable currents

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

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

The effect of impedance

As with any other component, when a ferrite is placed in circuit it operates between source and load impedances. A quick glance at the equivalent circuit in Figure 13.20 shows that maximum attenuation due to the simple impedance divider will occur when Z_S and Z_L are low. For example, if Z_S and Z_L are 10 ohms and the ferrite impedance at a given frequency is 100 ohms, the total attenuation (with versus without ferrite) is:

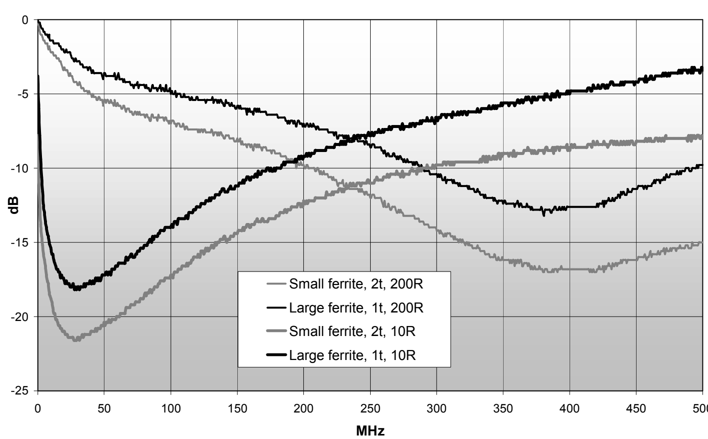
$$A = 20\log_{10} [(10+10)/(10+100+10)] = -15.6\text{dB} \tag{13.2}$$

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

$$A = 20\log_{10}[(150+150)/(150+100+150)] = -2.5\text{dB} \quad (13.3)$$

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

Clip-on ferrite impedances rarely exceed 200–300 ohms, and consequently the attenuation that can be expected from placing a ferrite on an open cable is typically no more than 6–10dB, with 20 dB being achievable at certain frequencies where the cable shows a low impedance. The plots in Figure 13.21 show the actual attenuation for two core sizes at two different circuit impedances.



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

Figure 13.21 Ferrite attenuation in different circuit impedances

13.2.2.2 Choosing and using ferrite components

Size and shape

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

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

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

Number of turns

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

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

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

Capacitance

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

Resistance

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

Saturation

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

13.2.2.3 Ferrite chip components

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

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

13.2.2.4 Three-terminal capacitors

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

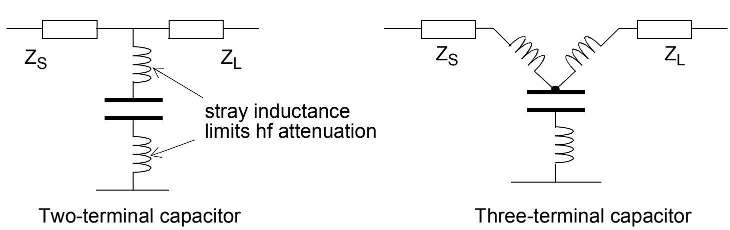


Figure 13.22 The three-terminal capacitor

This lead inductance can be put to some use if the capacitor is given a three-terminal construction (Figure 13.22), separating the input and output connections. The lead inductance now forms a T-filter with the capacitor, greatly improving its high frequency performance. A ferrite bead on each of the upper leads will further enhance the lead inductance and increase the effectiveness of the filter when it is used with a relatively low-impedance source or load. The three-terminal configuration can extend the range of a small ceramic capacitor from below 50MHz to beyond 200MHz, which is particularly useful for interference in the vhf band. To fully benefit from this approach, you must terminate the middle (ground) lead directly to a low-inductance ground such as a ground plane, otherwise the inductance remaining in this connection will defeat the capacitor's purpose.

Surface mount capacitors are also available in a quasi-feedthrough or three-terminal form, in which the device has an elongated shape, and the end terminals carry the signal through the capacitor while the middle body terminal is bonded directly to a ground plane on the PCB. The proper PCB layout for this is shown in Figure 13.23.

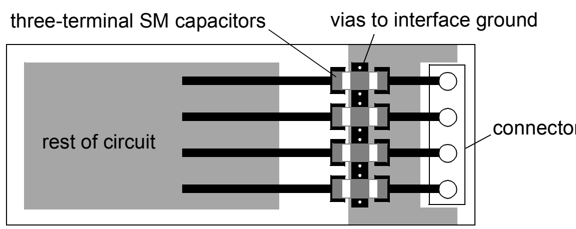


Figure 13.23 Using three-terminal surface mount capacitors

13.2.2.5 Feedthrough capacitors

Any leaded capacitor is still limited in effectiveness by the inductance of the connection to the ground point. For the ultimate performance, and especially where penetration of a screened enclosure must be protected at UHF and above then a feedthrough (or leadthrough) construction (Figure 13.24) is essential. Here, the ground connection is made by screwing or soldering the outer body of the capacitor directly to the metal screening or bulkhead. Because the current to ground can spread out for 360° around the central conductor, there is effectively no inductance associated with this terminal and the capacitor performance is maintained well into the GHz region. This performance is compromised if a 360° connection is not made or if the bulkhead is limited in extent. To create a π-section filter, the inductance of the through lead can be increased by separating the ceramic metallization into two parts and incorporating a ferrite bead within the construction. Feedthrough capacitors are available in a wide range of voltage and capacitance ratings but their cost increases with size.

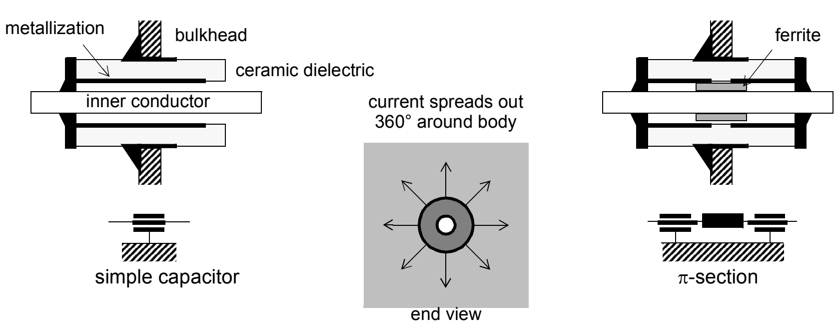


Figure 13.24 The feedthrough capacitor

13.2.3 Mains filters

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

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

In fact, the market for mains filters really took off with the introduction of regulations on conducted mains emissions, compounded by the rising popularity of the switchmode power supply. With a switching supply, a mains filter is essential to meet these regulations. EMC has historically tended to be seen as an afterthought on commercial equipment, and there have been many occasions on which retro-fitting a single component mains filter has brought a product into compliance, and this has also encouraged the development of the mains filter market. A real benefit is that safety approvals needed for all components on the mains side of the equipment have been already dealt with by the filter manufacturer if a single-unit filter is used.

13.2.3.1 Application of mains filters

Merely adding a block filter to a mains input will improve low frequency emissions such as the low harmonics of a switching power supply. But HF emissions (above 1MHz) require attention to the layout of the circuitry around the filter (see section 13.2.1.3). Treating it like any other power supply component will not give good HF attenuation and may actually worsen the coupling, through the addition of spurious resonances and coupling paths. Combined filter and CEE22 inlet connector modules are a good method of ensuring correct layout, providing they are properly bonded to a specified clean ground.

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

13.2.3.2 Typical mains filter

A typical filter (Figure 13.25) includes components to block both common mode and differential mode components. The common mode choke L consists of two identical windings on a single high-permeability core, usually toroidal but sometimes of E-I construction, configured so that differential (line-to-neutral) currents cancel each other. This allows high inductance values, typically 1–10mH, in a small volume without fear of choke saturation caused by the mains frequency supply current.

The sense of the windings (Figure 13.26) is such that differential currents, in which the “go” current in one wire is equal and opposite to the “return” current in the other,

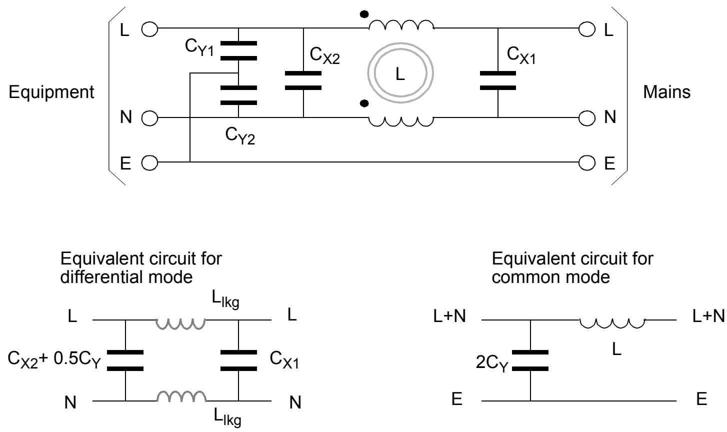


Figure 13.25 Typical mains filter and its equivalent circuit

each create a magnetic flux in the core, but because they are equal and opposite the two fluxes cancel, leaving no net magnetic flux. Thus since the core is invisible the differential mode inductance is very small, being dominated by the residual difference between the windings, known as the leakage inductance.

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

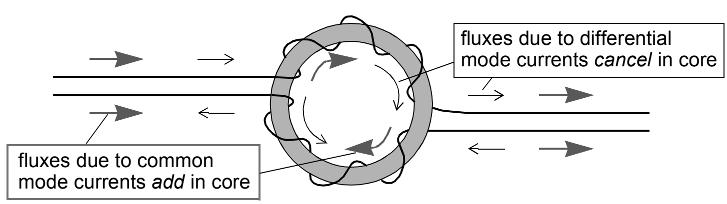


Figure 13.26 The common mode choke

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

Common mode capacitors

Capacitors C_{Y1} and C_{Y2} attenuate common mode interference and if C_{X2} is large, have no significant effect on differential mode. The effectiveness of the C_Y capacitors depends very much on the common mode source impedance of the equipment (Figure

13.27). This is usually a function of stray capacitance coupling to earth, which depends critically on the mechanical layout of the circuit and the primary-to-secondary capacitance of the mains transformer, and can easily exceed 1000pF. The attenuation offered by the potential divider effect of C_Y may be no more than 15–20dB. The common mode choke is the more effective component, and in cases where C_Y is very severely limited more than one common mode choke may be needed. Calculation of appropriate component values is covered in Appendix D (section D.6).

Differential mode capacitors

Capacitors C_{X1} and C_{X2} attenuate differential mode only but can have fairly high values, 0.1 to 0.47µF being typical. Either may be omitted depending on the detailed performance required, remembering that the source and load impedances may be too low for the capacitor to be useful. For example a 0.1µF capacitor has an impedance of about 10Ω at 150kHz, and the differential mode source impedance seen by C_{X2} may be considerably less than this for a power supply in the hundreds of watts range, so that a C_{X2} of this value would have no effect at the lower end of the frequency range where it may be most needed.

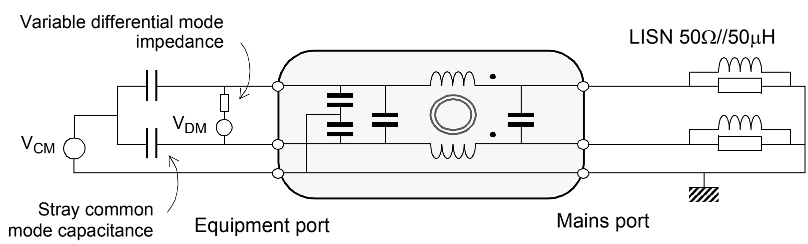


Figure 13.27 Impedances seen by the mains filter

13.2.3.3 Safety considerations

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

$$I_{LKG} = V \cdot 2\pi F \cdot C \cdot 1.2 \quad \mu A \tag{13.4}$$

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

Values for this current range from 0.25mA to 3.5mA depending on the applicable standard, safety class and use of the apparatus (Table 13.1). Special installation conditions apply where the leakage exceeds 3.5mA. Medical equipment has an even lower leakage requirement, typically 0.1mA. Note that this is the *total* leakage current due to the apparatus; if there are other components (such as transient suppressors) which also form a leakage path to earth, the current due to them must be added to that due to C_Y, putting a further constraint on the value of C_Y.

BS 613, which specifies EMI filters in the UK, allows a maximum value for C_Y of 5000pF with a tolerance of ±20% for safety Class I or II apparatus. The frequently

Table 13.1 Allowed earth leakage limits in common safety standards

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

specified value of 0.75mA leakage current gives a maximum capacitance of around 4.7nF on each phase for a voltage of 250V at 50Hz, so this value is typical in general-purpose filter units.

Component ratings

Both C_X and C_Y carry mains voltages continuously and must be specifically rated to do this. Failure of C_X will result in a fire hazard, while failure of C_Y will result in both a fire hazard and a potential shock hazard. “X” and “Y” class components to EN 132400 (similar to IEC 60384-14) are designed and marketed specifically for these positions; safety standards mandate their use. EN 132400 has various requirements (Table 13.2) including peak impulse voltage, voltage endurance and flammability.

Table 13.2 EN132400 impulse voltage and endurance ratings

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

13.2.3.4 Insertion loss versus impedance

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

The 50Ω termination does not reflect the real situation. The mains port HF impedance can be generalized for both common and differential mode by a 50Ω//50μH network as provided by a CISPR 16 LISN (section 6.2.2.1); when the product is tested for compliance, this network will be used anyway. The equipment port impedance will vary substantially depending on load and on the HF characteristics of the input components such as the mains transformer, diodes and reservoir. Differential mode impedance is typically a few ohms for small electronic products, while common mode impedance as discussed above can normally be approximated by a capacitive reactance of 100–1000pF. Because these load impedances differ from 50Ω they may enhance resonances within the filter and thus achieve insertion *gain* at some frequencies – typically below the 150kHz lower cut-off frequency of the conducted emissions limits.

One alternative method for measuring filter insertion loss is to use terminating impedances of 0.1Ω/100Ω and reverse. This method is defined in CISPR publication 17, and results in more realistic performance data under some circuit conditions, but not all filter manufacturers provide figures using this test method. If you can assume pure resistive terminating impedances, you can derive the expected insertion loss performance from the published 50Ω data using the equations in section D.6.

The circuit of Figure 13.27 is easy to set up with a circuit analysis program. Provided that you remember to include estimated parasitic reactances (capacitance across the choke coils, inductance in series with capacitors) you can make a fair attempt at designing the mains filter circuit for the optimum required attenuation from known emission levels.

13.2.3.5 Core saturation

Filters are specified for a maximum working RMS current, which is mainly determined by allowable heating in the common mode choke. Capacitor input power supplies have a distinctly non-sinusoidal input current waveform (see the discussion on mains harmonics in section 10.4, and Figure 13.28(a)), with a peak current of between three and ten times the RMS, which represents a high "crest factor". But, for reasons of space efficiency, the ferrite core of a mains filter is typically designed so that the common mode inductance has dropped to 80% of its nominal value at the filter's rated current. To protect the common mode inductance against saturation, this 80% point should occur at the maximum value of the current waveform: that is, the filter should be rated not for the RMS but for the peak current. The magnetic field due to the supply current is compensated inside a common mode ferrite core, but the stray inductance (around 1% of nominal) still leads to saturation at the peak current. The effect of saturation is mainly felt at the bottom end of the spectrum near 150kHz, where a loss of attenuation of more than 10dB can occur.

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

The core will also saturate when it is presented with a high-voltage, high energy common mode surge, such as a switching transient on the mains (Figure 13.28(b)). The surge voltage will be let through delayed and with a slower risetime but only slightly attenuated with attendant ringing on the trailing edge. Standard mains filters designed

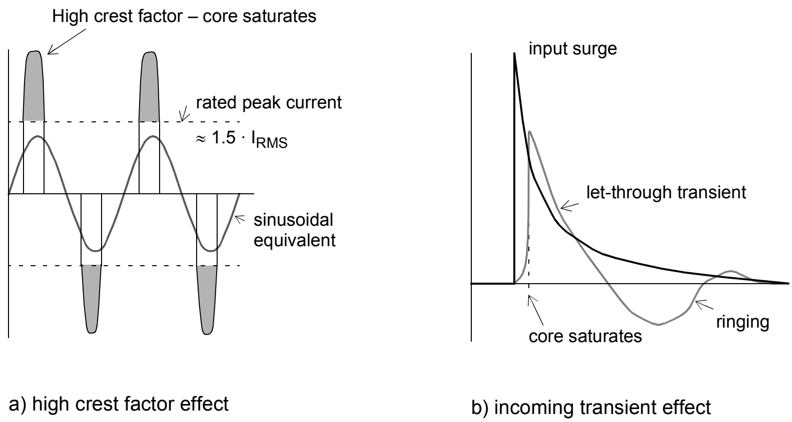


Figure 13.28 Core saturation effects

only for attenuating frequency-domain emissions are inadequate to cope with large incoming common mode transients, though some are better than others. Differential mode transients require considerably more energy to saturate the core and these are more satisfactorily suppressed.

13.2.3.6 Extended performance

In some cases the insertion loss offered by the typical configuration won't be adequate. This may be the case when for example a high-power switching supply must meet the most stringent emission limits, or there is excessive coupling of common mode interference, or greater incoming transient immunity is needed. The basic filter design can be extended in a number of ways (Figure 13.29):

- *extra differential line chokes*: these are separate chokes in L and N lines which are not cross-coupled and therefore present a higher impedance to differential mode signals, giving better attenuation in conjunction with C_X. Because they must not saturate at the full AC line current they are much larger and heavier for a given inductance.
- *an earth line choke*: this increases the impedance to common mode currents flowing in the safety earth and may be the only way of dealing with common

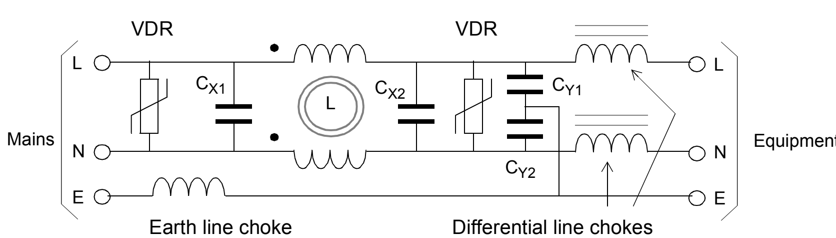


Figure 13.29 Higher-performance mains filter

mode interference, both incoming and outgoing, when C_Y is already at its maximum limit and nothing can be done about the interference at source. Because it is in series with the safety earth its fault current carrying capability must satisfy safety standards. Ensure that it is not short-circuited by any extra earth connection: this is most often provided by a cable to another item of equipment, and makes the earth line choke unusable except in apparatus without such other connections.

- *transient suppressors*: a device such as a voltage-dependent resistor (VDR) across L and N will clip incoming differential mode surges (see also section 13.2.5). If it is placed at the mains port then it must be rated for the full expected transient energy, but it will prevent the choke from saturating and protect the filter's C_X; if it is placed on the equipment side then it can be substantially downrated since it is protected by the impedance of the filter. Of course, it has no effect on common mode transients.

In addition to these extra techniques the basic filter π-section can be cascaded with further similar sections, perhaps with inter-section screens and feedthroughs to obtain much higher insertion loss. For these levels of performance the filter must be used in conjunction with a well-screened enclosure to prevent high frequency coupling around it. Large values of C_X should be protected with a bleeder resistor in parallel, to prevent a hazardous charge remaining between L and N when the power is removed (detailed requirements can be found in safety specifications such as IEC 60335/EN 60335).

13.2.4 I/O filtering

If I/O connections carry only low bandwidth signals and low current it is possible to filter them using simple RC low-pass networks (Figure 13.30(a)). The decoupling capacitor must be connected to the clean I/O ground (see section 11.2.3) which may not be the same as circuit 0V.

13.2.4.1 Common mode choke

Adequately high values of R and C may not be possible with high-speed data links or interfaces carrying significant current, but it is possible to attenuate common mode currents entering or leaving the equipment without affecting the signal frequencies by using a discrete common mode choke arrangement. The choke has several identical windings on the same core such that the fields from differential currents cancel each other whereas common mode currents add, in the same fashion as the mains common mode choke described in section 13.2.3.2. Such units are available commercially (sometimes described as "data line filters") or can be custom designed. It is important when designing in a signal line common mode choke that you make sure the differential currents really do cancel in the core: this means that all lines handled by the interface, *including the 0V line* if it is used as a return, must pass through individual windings on a single choke, as shown in Figure 13.30(b).

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

13.2.4.2 Filtered connectors

A convenient way to incorporate both the capacitors and to a lesser extent the inductive components of Figure 13.30 is within the external connector itself. Each pin can be

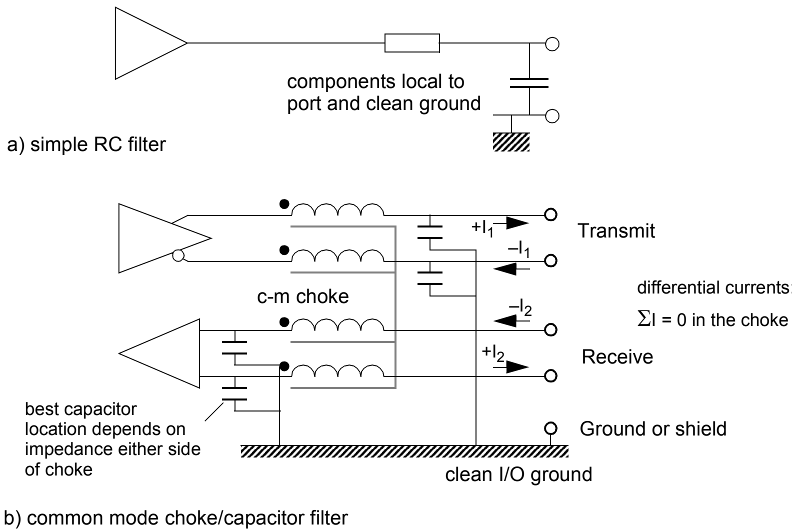


Figure 13.30 I/O filtering techniques

configured as a feedthrough capacitor with a ceramic jacket, its outside metallization connected to a matrix which is grounded directly to the connector shell (Figure 13.31(a)). Thus the inductance of the ground connection is minimized, provided that the connector shell itself is correctly bonded to a clean ground, normally the metal backplate or chassis of the unit. Any series impedance in the ground path not only degrades the filtering but will also couple signals from one line into another, leading to designed-in crosstalk.

The advantage of this construction is that the insertion loss can extend to over 1GHz, the low frequency loss depending entirely on the actual capacitance (typically 50–2000pF) inserted in parallel with each contact. With some ferrite incorporated as part of the construction, a π-filter can be formed as with the conventional feedthrough (section 13.2.2.5). No extra space for filtering needs to be provided. The filtered connector has obvious attractions for retro-fit purposes, and may frequently solve interface problems at a stroke. You can also obtain ferrite blocks tailored to the pinout dimensions of common multiway connectors, which effectively offer individual choking for each line with a single component.

Disadvantages are the significant extra cost over an unfiltered connector; if not all contacts are filtered, or different contacts need different capacitor values, you will need a custom part. Voltage ratings may be barely adequate and reliability may be worsened. A small "piggy-back" or "mezzanine" board of chip capacitors mounted immediately next to the connector (Figure 13.31(b)), with their ground connection made via the lowest possible inductance, can be equally effective up to hundreds of MHz, is cheaper, and is inherently customized.

13.2.4.3 Circuit effects of filtering

When you use any form of capacitive filtering, the circuit must be able to handle the extra capacitance to ground, particularly when filtering an isolated circuit at radio

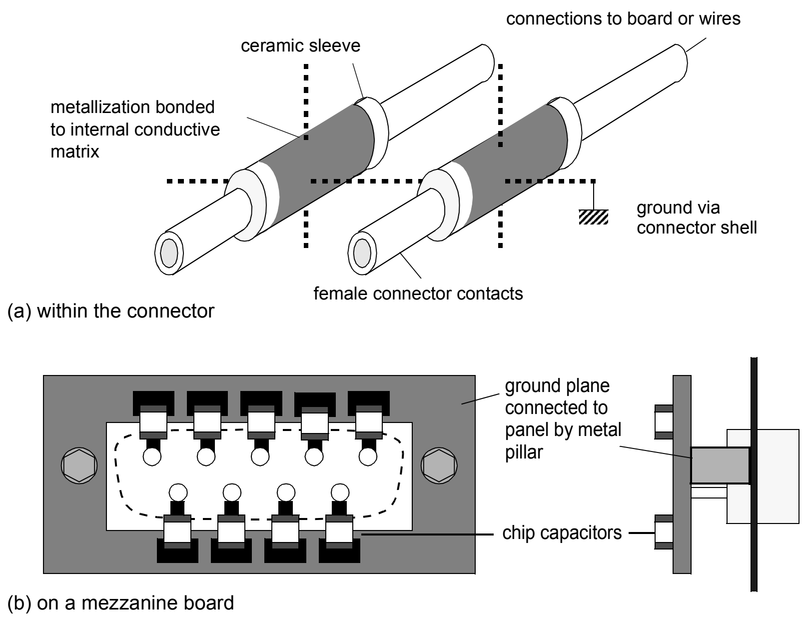


Figure 13.31 Filtered connector pins

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

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

As a further hazard, capacitive loading of low frequency analogue amplifier outputs may also push the output stage into instability (see section 12.1.4.3).

13.2.5 Transient suppression

Incoming transients on either mains or signal lines are reduced by non-linear devices: the most common are varistors (voltage-dependent resistors, or VDRs), zeners and spark gaps (gas discharge tubes). The device is placed in parallel with the line to be protected (Figure 13.32) and to normal signal or power levels it appears as a high impedance – essentially determined by its self-capacitance and leakage specifications. When a transient which exceeds its breakdown voltage appears, the device changes to a low impedance which diverts current from the transient source away from the

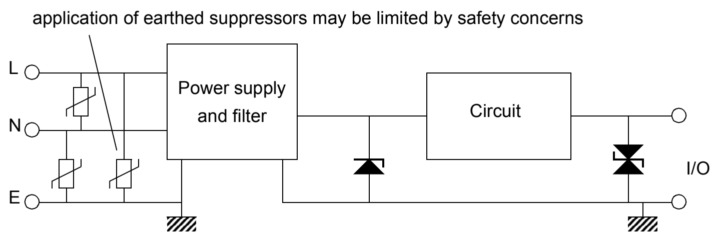


Figure 13.32 Typical locations for transient suppressors

protected circuit, limiting the transient voltage (Figure 13.33). It must be sized to withstand the continuous operating voltage of the circuit, with a safety margin, and to be able to absorb the energy from any expected transient.

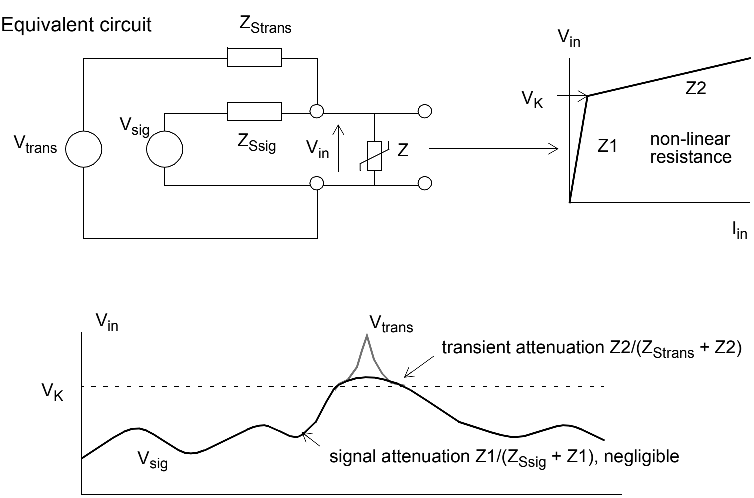


Figure 13.33 The operation of a transient suppressor

The first requirement is fairly simple to design to, although it means that the transient clamping voltage is usually 1.5–2 times the continuous voltage, and circuits that are protected by the suppressor must be able to withstand this. The second requirement calls for a knowledge of the source impedance Z_{Strans} and probable amplitude of the transients, which is often difficult to predict accurately especially for external connections. This determines the amount of energy which the suppressor will have to absorb. Reference [48] gives details of how to determine the required suppressor characteristics from a knowledge of the circuit parameters, and also suggests design values for the energy requirement for suppressors on AC power supplies. These are summarized in Table 13.3. IEEE C62.41 [208] gives further details of expected transient sources and amplitudes, as well as detailing the test methods recommended to check transient immunity.

Table 13.3 Suggested transient suppressor design parameters

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

Table 13.4 Comparison of transient suppressor types

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

Table 13.4 compares the characteristics of the most common varieties of transient suppressor. Variations on the three basic types are available; for instance the ZnO varistor is available in monolithic multilayer form which allows clamp voltages down to 5V, and can also be fabricated with a specified capacitance to offer a varistor-cum-capacitor in a single component.

Combining types

You may sometimes have to parallel different types of suppressor in order to achieve a performance which could not be given by one type alone. For example, telecoms applications require signal line protection not just from straightforward transients, but also from local lightning strikes and shorting between the signal line and AC power lines [38]. These can only be dealt with by providing both primary and secondary protection; primary protection, offered by gas discharge tubes (GDTs) or carbon spark gaps, will remove the major part of the incoming energy but leaves an initial spike due to its slow response time. Secondary protection, provided by a semiconductor device, is faster and needs only to deal with the residual energy.

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

13.2.5.1 Layout of transient suppressors

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

The component leads must be short (suppressors are available in SM chip form) and they must be connected locally to the circuit that is to be clamped (Figure 13.34). Any common impedance coupling, via ground or otherwise, must be avoided. ESD spike suppression can be improved by raising the RF impedance of the input circuit with a lossy ferrite component. Where suppressors are to be combined with I/O filtering you may be able to use the three-terminal varistor/capacitor devices that are now available.

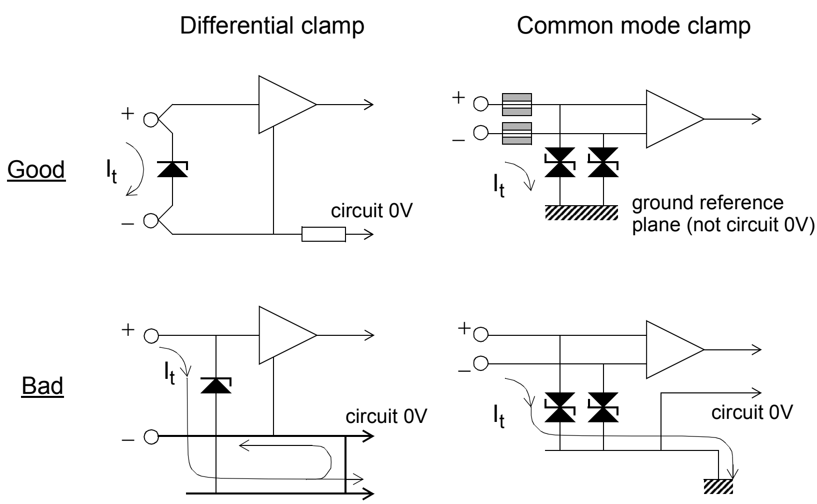


Figure 13.34 Layout and configuration of I/O transient suppressors

13.2.6 Contact suppression

An opening contact which interrupts a flow of current – typically a switch or relay – will initiate an arc across the contact gap. The arc will continue until the available current is not enough to sustain a voltage across the gap (Figure 13.35). The stray capacitance and inductance associated with the contacts and their circuit will in practice cause a repetitive discharge until their energy is exhausted, and this is responsible for considerable broadband interference [11][83]. A closing contact can also cause interference because of contact bounce.

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

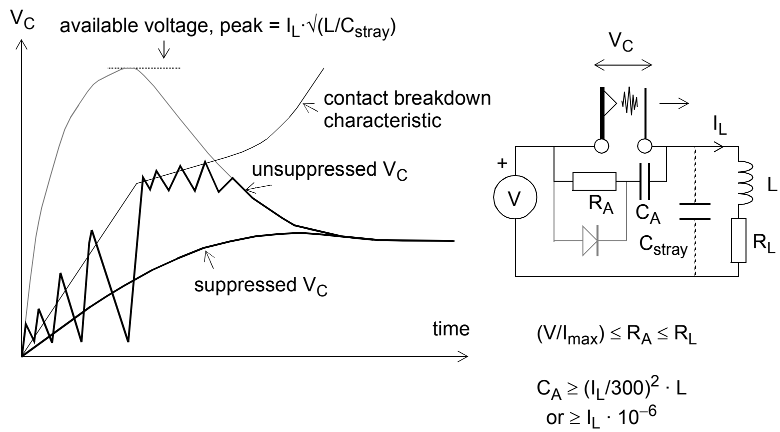


Figure 13.35 Contact noise generation and suppression

13.2.6.1 Suppression of inductive loads

When current through an inductance is interrupted a large transient voltage is generated, governed by $V = -L \cdot di/dt$. Theoretically if di/dt is infinite then the voltage is infinite too; in practice it is limited by stray capacitance if no other measures are taken, and the voltage waveform is a damped sinusoid (if no breakdown occurs) whose frequency is determined by the values of inductance and stray capacitance. Typical examples of switched inductive loads are motors, relay coils and transformers, but even a long cable can have enough distributed inductance to generate a significant transient amplitude. Switching can either be via an electromechanical contact as in the previous section, or a semiconductor, and the latter can easily suffer avalanche breakdown due to the overvoltage if the transient is unsuppressed. RF interference is generated in both cases at frequencies determined by stray circuit resonances, and is usually radiated from the wiring between switch and load.

The RC snubber circuit can be used in some cases to damp an inductive transient. Other circuits use diode, Zener or varistor clamps as shown in Figure 13.36. In all cases the suppression components must be mounted immediately next to the load terminals, otherwise a radiating current loop is formed by the intervening wiring. Protection of a driver transistor mounted remotely must be considered as a separate function from RF suppression.

13.2.6.2 Motor suppression

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

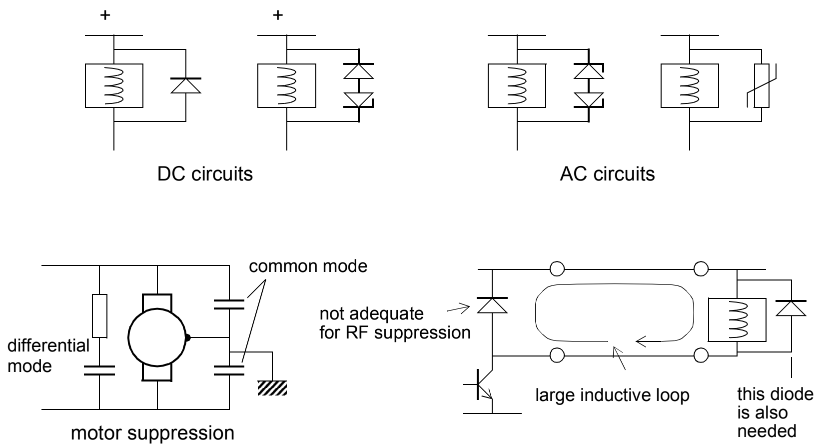


Figure 13.36 Inductive load suppression

mode noise across the terminals and common mode with respect to the housing, coupled through stray capacitance.

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

Chapter 14

Shielding

This chapter looks at that classic of EMC design, the shielded enclosure. It is only too easy to say “the product’s in a metal box, it must be shielded, that’ll do”. In fact there is much more to designing a shielded product than simply putting it in a metal box. This section explores the theory of ideal shields and the limits that this theory encounters when you look at the performance of real shields, and then the practical design measures that can be taken to deal with these limitations.

14.1 Shielding theory

Shielding and filtering are complementary practices. There is little point in applying good filtering and circuit design practice to guard against conducted coupling if there is no return path for the filtered currents to take. The shield provides such a return, and also guards against direct field coupling with the internal circuits and conductors. Shielding involves placing a conductive surface around the critical parts of the circuit so that the electromagnetic field which couples to it is attenuated by a combination of reflection and absorption. The shield can be an all-metal enclosure if protection down to low frequencies is needed, but if only high frequency (> 30MHz) protection will be enough then a thin conductive coating deposited on plastic is adequate.

Will a shield be necessary?

Shielding is often an expensive and difficult-to-implement design decision, because many other factors – aesthetic, tooling, accessibility – work against it. A decision on whether or not to shield should be taken as early as possible in the project. Chapter 10, sections 10.2 and 10.3 showed that interference coupling is via interface cables and direct induction to/from the PCB. You should be able to calculate to a rough order of magnitude the fields generated by PCB tracks and compare these to the desired emission limit (see section 12.1.2). If the limit is exceeded at this point and the PCB layout cannot be improved, then shielding is essential. Shielding does not of itself affect common mode cable coupling and so if this is expected to be the dominant coupling path – generally only experience of similar products will tell this – a full shield may not be necessary. It does establish a “clean” reference for decoupling common mode currents to, but it is also possible to do this with a large area ground plate if the layout is planned carefully.

A description of shielding issues is best split into two parts:

- the theory of electromagnetic attenuation through a conducting barrier of infinite extent, and
- the degradation of theoretically achievable shielding effectiveness by practical forms of shield construction.

To shield or not to shield

- if predicted differential mode fields will exceed limits, shielding is essential
- if layout requires dispersed interfaces, shielding will probably be essential
- if layout allows concentrated interfaces, a ground plate may be adequate
- consider shielding only critical circuitry

14.1.1 Shielding theory for an infinite barrier

14.1.1.1 Reflection and absorption

An AC electric field E_0 impinging on a conductive wall of finite thickness but infinite extent will induce a current flow J_i in that surface of the wall, which in turn will generate a reflected wave E_R of the opposite sense. This is necessary in order to satisfy the boundary conditions along the wall, where the electric field must approach zero – although not reach it, since there will be a voltage along the length of the wall which is determined by the current density times the impedance of the wall, which is not zero. The difference between the impinging and reflected wave amplitudes determines the reflection loss of the wall. Because shielding walls have finite conductivity, part of this current flow penetrates into the wall and a fraction of it J_t will appear on the opposite side of the wall, where it will generate its own field E_t (Figure 14.1). The ratio of the impinging to the transmitted fields E_0/E_t is one measure of the shielding effectiveness of the wall. As with an optical mirror, the reflection loss is a surface effect only and is not affected by the thickness of the barrier.

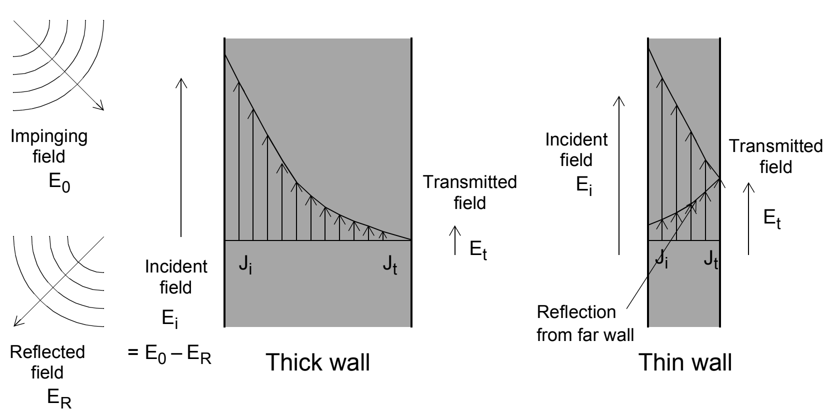


Figure 14.1 Reflection and absorption in an infinite barrier

The thicker the wall, the greater the attenuation of the current through it. This absorption loss depends on the number of "skin depths" through the wall. The skin depth (defined in Appendix D, section D.4.1) is an expression of the electromagnetic property which tends to confine AC current flow to the surface of a conductor, becoming less as frequency, conductivity or permeability increases. Fields are attenuated by 8.6dB (1/e) for each skin depth of penetration. For example, skin depth in aluminium at 30MHz is 0.015mm. This explains why thin conductive coatings are effective at high frequencies – the current only flows in a thin layer on the surface, and the bulk of the material does not affect the shielding properties.

Reflection loss

The reflection loss R depends on the ratio of wave impedance to barrier impedance. The concept of wave impedance has been described in section 10.1.4.2. The impedance of the barrier is a function of its conductivity and permeability, and of frequency. Materials of high conductivity such as copper and aluminium have a higher E-field reflection loss than do lower conductivity materials such as steel. Reflection losses decrease with increasing frequency for the E-field (electric) and increase for the H-field (magnetic). In the near field, closer than $\lambda/2\pi$, the distance between source and barrier also affects the reflection loss. Near to an E-field source, the electric field impedance is high and the reflection loss is also correspondingly high. Vice versa, near to an H-field source the magnetic field impedance is low and the reflection loss is low. When the barrier is far enough away to be in the far field, the impinging wave is a plane wave, the wave impedance is constant and the distance is immaterial. Refer back to Figure 10.9 on page 231 for the distinction in impedances between near and far field.

The re-reflection loss B is insignificant in most cases where absorption loss A is greater than 10dB, but becomes important for thin barriers at low frequencies.

Absorption loss

Absorption loss depends on the barrier thickness and its skin depth and is the same whether the field is electric, magnetic or plane wave: that is, it doesn't depend on the wave impedance, in contrast to reflection loss. The skin depth in turn depends on the barrier material's properties; steel, for instance, offers higher absorption than copper of the same thickness. At high frequencies, as Figure 14.2 shows, absorption becomes the dominant term, increasing exponentially with the square root of the frequency. Appendix D (section D.4) gives the formulae for the values of A, R and B for given material parameters.

14.1.1.2 Shielding effectiveness

Shielding effectiveness (SE) of a solid conductive barrier describes the ratio between the field strength without the barrier in place, to that when it is present. It can be expressed as the sum of reflection, absorption, and re-reflection losses, as shown in Figure 14.2 and given by equation (14.1):

$$\mathrm{SE(dB)} = \mathrm{R(dB)} + \mathrm{A(dB)} + \mathrm{B(dB)} \tag{14.1}$$

This is known as the "transmission line model" for shielding effectiveness, and among other things it makes the gigantic assumption that the coupling between the shield currents and the source of the incident field is negligible. In most shielding applications this is an unwarranted assumption, but it does at least simplify the model to the point at which it is comprehensible. The real problems are expanded in section 14.1.3.

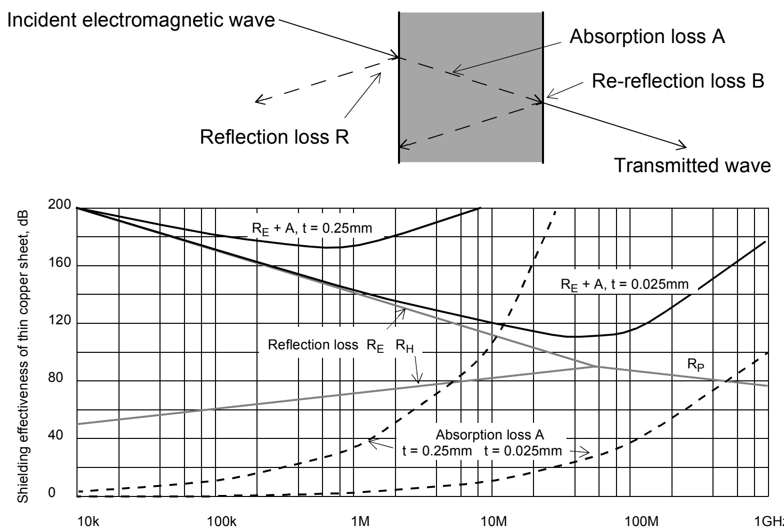


Figure 14.2 Shielding effectiveness versus frequency for a copper sheet of infinite extent

14.1.2 LF magnetic fields

Shielding against magnetic fields at low frequencies is to all intents and purposes impossible with purely conductive materials. This is because the reflection loss to an impinging magnetic field (R_H) depends on the mismatch of the field impedance to the barrier impedance. The low field impedance is well matched to the low barrier impedance, so there is little reflection, and the field is then transmitted through the barrier with only a few dB attenuation or absorption because of its low frequency. Fortunately, from the point of view of compliance the requirements of the EMC Directive and similar regimes generally don't extend to magnetic shielding at low frequencies, with the possible exception of some types of apparatus that may be susceptible to power frequency fields. But there are some applications where good magnetic field shielding is necessary for functional reasons: one example is CRT displays in high-current environments such as railway trackside locations, another is magnetically sensitive transducers such as electron beam microscopes.

A high-permeability material such as mu-metal or its derivatives can give LF magnetic shielding by concentrating the field within the bulk of the material, but this is a different mechanism to that discussed above, and it is normally only viable for sensitive individual components such as CRTs or transformers. For an infinitely long cylinder in a DC field the shielding factor is:

$$S_M = (\mu_r/2) \times t/d \qquad (14.2)$$

where μ_r is the material's relative permeability and t/d is the ratio of material thickness to cylinder diameter

Practically, there is a fall-off of shielding performance towards the ends, but for distances inside the cylinder greater than the diameter the shield can be regarded as infinitely long; alternatively a high-permeability end cap can be used. Both welded seams and high-intensity fields have the effect of reducing the material permeability μ_r. High flux densities will tend to saturate the material; if the expected flux density is greater than the material's saturation flux density then increase the material thickness or use a double shield with a "nested" construction, with a higher saturation flux density material facing the impinging field. For prototyping, you can use foil magnetic shielding material which can easily be worked by hand. Production shields should be properly fabricated and this needs to be done by a specialist.

You should be aware that this form of shielding is specifically for low frequencies. Permeability of all high-mu materials falls off at frequencies above a few tens of kHz. But as the frequency increases, so does the wave impedance at the barrier and therefore simple metal barriers become increasingly effective. If for some particular application you need extremely wideband shielding effectiveness, then you will use a combination of high-mu metals for the low frequency magnetic effects and conductive but non-magnetic metals for everything else.

14.1.3 The effect of apertures

The curves of shielding effectiveness in Figure 14.2 suggest that upwards of 200dB attenuation is easily achievable using reasonable thicknesses of common materials. In fact, the practical shielding effectiveness is not determined by material characteristics but by necessary apertures and discontinuities in the shielding. You will need apertures for ventilation, for control and interface access, and for viewing indicators; seams at the joints between individual conductive members act as apertures too. Also, shielding is almost invariably applied in the near field of the circuits inside an enclosure. The theoretical material- and field impedance-related attenuation is merely an upper bound on what is achievable, and much lower values are found in practice.

There are different theories for determining SE degradation due to apertures. The simplest assumes that SE is directly proportional to the ratio of longest aperture dimension L and frequency, with zero SE when $L = \lambda/2$: SE = $20\log(\lambda/2L)$. Thus the SE increases linearly with decreasing frequency up to the maximum determined by the barrier material, with a greater degradation for larger apertures. A correction factor can be applied for the aspect ratio of slot-shaped apertures. This simple approximation finds favour in evaluating design options, since it doesn't require any knowledge of the actual mechanical structure or field parameters, but neither does it correctly predict the actual value of attenuation that is obtained.

14.1.3.1 Transmission line theory of a rectangular box with a single aperture

Work done in the later 1990s by the universities of York and Nottingham [120][50] has refined a theory which does show good correlation with actual results, under certain well-controlled conditions. This theory treats the case of a rectangular shielding box with a single slot in one of its faces (Figure 14.3(a)), as if it were a length of shorted waveguide. The slot is modelled as a length of transmission line shorted at either end, and the incident field is represented as a voltage source with an impedance equivalent to that of free space (Figure 14.3(b)). To determine the shielding effectiveness, the incident voltage is compared to the voltage at the desired location along the inside of the waveguide. By including some loss in the waveguide propagation the effect of the enclosure contents can also be represented: loading the box with PCBs, for instance, which needs an accurate model for these loss effects.

Figure 14.3(c) shows the use of this model to calculate the shielding effectiveness

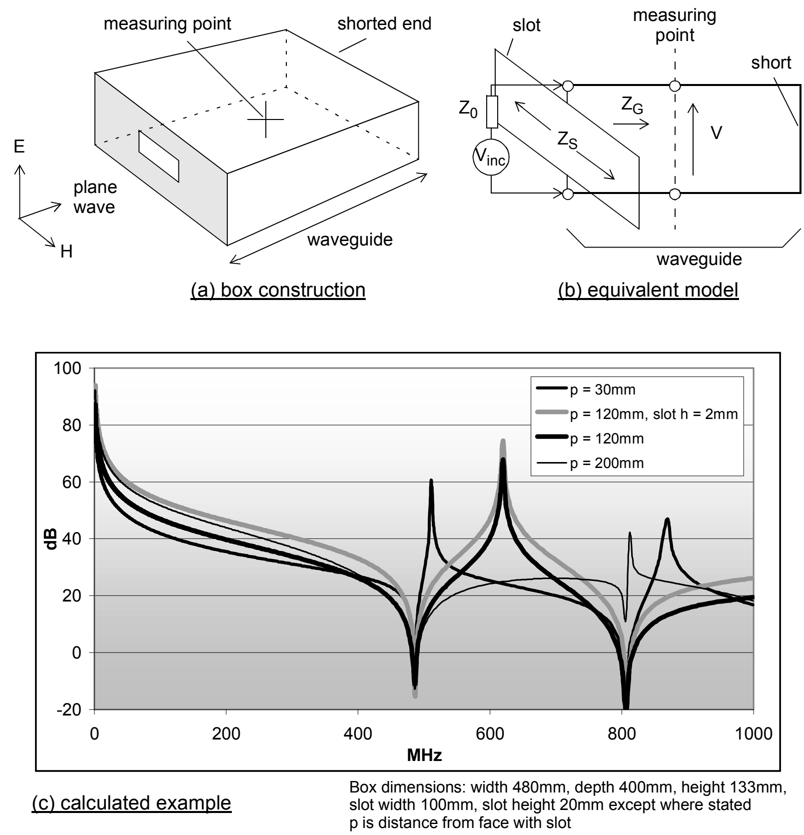


Figure 14.3 Modelling a rectangular box with a slot

of a box of a size typical of rack-mounted enclosures, with a display window in the front. This shows that shielding effect differs, as one might expect, depending on how far into the box the measuring point is located, and it also shows that the slot height does not have a large effect. Most noticeable, though, is that the shielding effectiveness becomes *negative* – that is, the field inside the box is higher than the incident field – at the box resonances. The ability of the enclosure to reduce the coupling of its contents with the environment is seriously compromised at these frequencies. How do they arise?

Enclosure resonance

As mentioned in section 10.3.1.3, a shielded enclosure can form a resonant cavity; standing waves in the field form between opposite sides when the dimension between the sides is a multiple of a half-wavelength. The electric field is enhanced in the middle of this cavity while the magnetic field is enhanced at the sides (Figure 14.4). For an empty cavity, resonances occur at:

$$F = 150 \cdot \sqrt{(k/l)^2 + (m/h)^2 + (n/w)^2} \text{ MHz} \qquad (14.3)$$

where l, h and w are the enclosure dimensions in metres
k, m and n are positive integers, but no more than one at a time can be zero

For approximately equal enclosure dimensions the lowest possible resonant frequency will be given by equation (14.4):

$$F \approx 212/l \approx 212/h \approx 212/w \text{ MHz} \tag{14.4}$$

The effect of resonances is to worsen the shielding effectiveness at the resonant frequencies. At these frequencies the field distribution within the cavity peaks, maximum current flows within the walls and hence maximum coupling occurs through the apertures of an imperfect enclosure, as can be seen in Figure 14.3(c). The resonances become more and more closely spaced as the frequency increases and higher order modes (k, m and n) are supported, so that actual shielding effectiveness tends to become extremely variable.

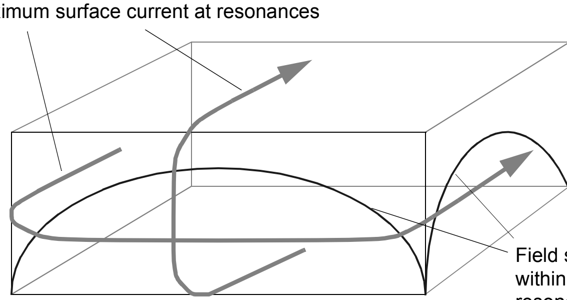


Figure 14.4 Resonances degrade shielding effectiveness

The above equations are accurate for a totally empty enclosure, but loading the enclosure with components, conducting structures and PCBs will detune the resonances and reduce their amplitude, often quite significantly [133]. This can have helpful consequences: for instance, if you have a particular emission or immunity problem at a box resonance, changing the dimension of the resonant cavity inside the box by adding, say, an internal screen or tie-bar may shift the resonance far enough away from the problem frequency to cure the effect. Conversely, if making such a modification doesn't bring relief, this is also a diagnostic hint. It means that you haven't addressed the particular mode that is causing the resonance, or that the problem is not related to resonant modes in the first place.

Exploring the effect of the slot

The worst shape for shielding is a long thin slot, especially if it is perpendicular to the E-field vector, which produces the maximum disturbance in the surface currents. The best shape is a round hole, but squares and hexagons are almost as good. Dividing a long slot into two shorter ones improves both the magnetic and electric shielding by about 6dB. The distance across the slot has only a second-order effect on the result. The slot's length also shows resonances, theoretically at every odd multiple of a half-wavelength established along its length. In practice, coupling between the slot resonances, the box resonances, and the content of the box detunes all of the resonances, so that the simple theory doesn't accurately predict what really happens. If you need to be able to evaluate the full impact of different enclosure design options before committing to tooling, the only way to do it with any confidence is to use electromagnetic modelling software [62].

14.1.3.2 The effect of seams

An electromagnetic shield is normally made from several panels joined together at seams. Unfortunately, when you join two sheets the electrical conductivity across the joint is imperfect. This may be because of distortion, so that surfaces do not mate perfectly, or because of painting, anodizing or corrosion, so that an insulating layer is present on one or both metal surfaces. Even if the surfaces appear conductive, when they are placed together contact is only created at points of pressure between one and the other.

Consequently, the shielding effectiveness is reduced by seams almost as much as it is by apertures (Figure 14.5). The fastener spacing d is critical in determining how much the shielding effectiveness is degraded, just as if the separation between fasteners was an actual aperture. The problem is especially serious for hinged front panels, doors and removable hatches that form part of a shielded enclosure. It is mitigated to some extent if the conductive sheets overlap, since this forms a capacitor which provides a partial current path at high frequencies. Figure 14.6 shows preferred ways to improve joint conductivity. Conductive gaskets are discussed in section 14.2.1.

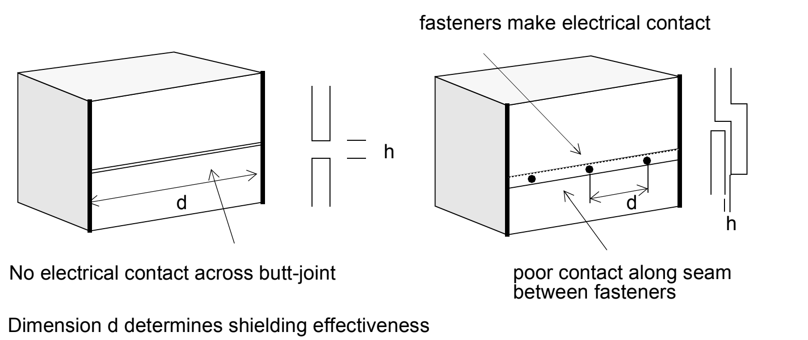


Figure 14.5 Seams between enclosure panels

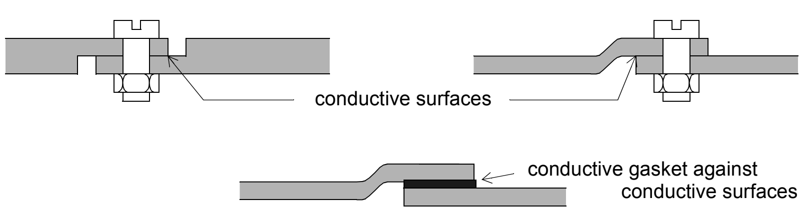


Figure 14.6 Cross-sections of joints for good conductivity

14.1.3.3 Seam and aperture orientation

The effect of a joint which creates a discontinuity is to force shield current to flow around the discontinuity. If the current flowing in the shield were undisturbed then the field within the shielded area would be minimized, but as the current is diverted so a localized discontinuity occurs, and this creates a field coupling path through the shield as described in section 14.1.3.1. The shielding effectiveness calculation assumes a worst case orientation of current flow. A long aperture or narrow seam will have a greater effect on current flowing at right angles to it than on parallel current flow†. This

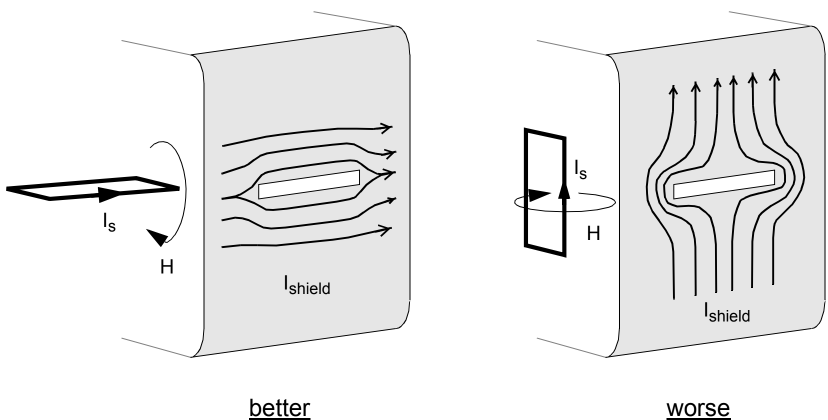


Figure 14.7 Current loop versus aperture orientation

effect can be exploited if you can control the orientation of susceptible or emissive conductors within the shielded environment (Figure 14.7).

The practical implication of this is that if all critical internal conductors are within the same plane, such as on a PCB, then long apertures and seams in the shield should be aligned parallel to this plane rather than perpendicular to it. Generally, you are likely to obtain an advantage of no more than 10dB by this trick, since the geometry of internal conductors is never exactly planar. Cables or wires, if they must be routed near to the shield, should be run along or parallel to apertures rather than across them. But because the leakage field coupling due to joints is large near the discontinuity, internal cables should preferably not be routed near to apertures or seams.

14.1.4 The shield as ground reference

A major advantage of a shielded enclosure is that the shield metalwork (or metallization) offers a low-inductance RF ground reference. This allows interfaces to be dispersed without suffering the penalty of high ground noise voltages developed by common mode noise currents flowing from one port to another. To achieve this, the interface cable screens must be bonded with low inductance directly to the shield as must any common mode decoupling capacitors. But, in using enclosure metalwork like this you can see that it is not absolutely essential that the shield be continuous in all

† Antenna designers will recognize that this describes a slot antenna, the reciprocal of a dipole.

dimensions around the circuit. The shield should be as complete (i.e. without apertures or seams) as possible *between the interface ports*, as any breaks will have the effect of diverting noise currents and thereby increasing the effective inductance of the ground reference. But this can quite easily be achieved with a U-shaped chassis (see Figure 14.8 for example, where a power supply comes in on one side of a circuit and the signal lines are connected on the other) without having to have a cover over the chassis.

You can choose to tie the circuit 0V to the shield in two ways. For good immunity of wideband analogue circuits it should be connected at the most sensitive part of the circuit to minimize noise coupling by the stray circuit-shield capacitances. On digital circuits on the other hand, multi-point grounding of the circuit 0V plane via several bonding points to the chassis will ensure the minimum common mode potential difference at any point in the circuit, and hence the best immunity to transient interference.

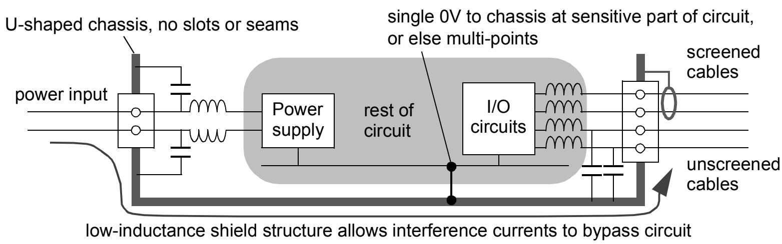


Figure 14.8 Shield metalwork as ground reference

14.1.5 The image plane

If you have a PCB which is poorly laid out and you are not in a position to revise the layout, a useful fix may be the image plane [72]. At first sight the image plane is a partial shield but in fact its mode of operation is distinct from that of a shield. The image plane is simply a flat conducting plane, typically made from a layer of foil laminated in plastic for insulation, placed as close as possible to the underside of the PCB (Figure 14.9). Alternatively, it could be a layer of metallization on the inner surface of the plastic enclosure on which the PCB is mounted. This plane should be at least the same size as the PCB and preferably larger so as to overlap the edges, especially those edges where (due to an oversight) there may be placed critical tracks. The criterion for overlap is that it should be greater than $10 \cdot h$, exactly as with PCB planes (section 11.2.2.7) and for the same reason: magnetic field coupling at the edges of the plane.

14.1.5.1 PCB with no external wires

If the PCB has no external connections, then the image plane can be electrically floating: there is no need to connect the PCB and the plane together. In this case the emissions radiated from the board on its own are principally due to common mode currents flowing on the various tracks. Adding an image plane creates a phantom reflection of these currents, effectively located at the "image" position on the opposite side of the plane (Figure 14.9(b)) – hence the name. These image currents are in the opposite polarity to the source currents. The effect of the ensemble is that the radiated emissions from the source currents are partially cancelled by the near-equal but opposite emissions from the image currents. The closer the plane is to the PCB (the

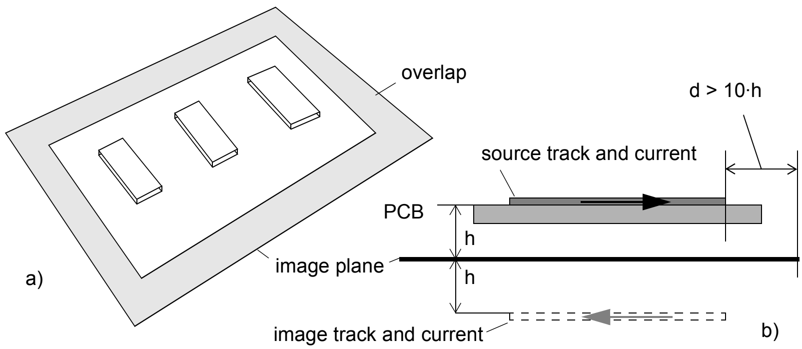


Figure 14.9 Image plane under a PCB

distance h in Figure 14.9) then the more effective is this cancellation, since the separation of source and image is $2 \cdot h$.

The plane is most effective at reducing emissions if the original PCB suffered from bad layout, i.e. if it includes large loop areas (see section 11.2.1). Here the plane can often be brought much closer to the board than the distance that may separate signal and return paths on the PCB. There is therefore a worthwhile reduction in the effective radiating area, which is now determined by the separation between source and image locations rather than the original signal and return. On the other hand, with a well laid out PCB, especially one with a ground plane already incorporated, adding an image plane will have little effect unless for some reason there are large common mode currents flowing across the board.

14.1.5.2 PCB with connected cables

Another effect of the image plane is to reduce the inductance of each track, including the ground track(s), because of the mutual coupling of each track with its image. This leads to a reduction in the ground noise voltages developed along the tracks, and this in turn will reduce emissions due to common mode currents injected into connected cables (section 10.2.1.3). As before, the reduction is most worthwhile when the board has been badly laid out so that its initial ground track inductance is high; but even here the reduction is only a few tens of per cent, rarely more than 6dB.

If though, instead of floating, the image plane is connected to the ground return track on the PCB *at the same point as the connected cable*, then there is now a return path for the generated common mode current and the net current being fed into the cable is near zero. This will allow a significant drop in emissions. In practice, all wires in the cable must be decoupled to this point at the emission frequencies (unless it is a shielded cable) and therefore a capacitor between each signal line and ground at the interface is mandatory. In this application, the image plane acts in much the same way as a chassis and can be used as a surrogate for a chassis if one doesn't exist; or you can regard the chassis, or metallization on a plastic enclosure, as acting like an image plane. In the limit, you can implement the image plane as an extra plane layer on the bottom of a multilayer PCB, as long as there are no components on this bottom side, but the quality of the image plane will be compromised to some extent by its penetration by via holes.

14.2 Shielding practice

14.2.1 Shielding hardware

Many manufacturers offer various materials for improving the conductivity of joints in conductive panels. From the advertising hype, you might think that these are all you need to rid yourself of EMC problems for good. In fact such materials can be useful if properly applied, but they must be used with an awareness of the principles discussed above, and their expense will often rule them out for cost-sensitive applications except as a last resort. A further difficulty which is now relevant is the EU's twin constraints on the materials used in products, the Restriction on Hazardous Substances (RoHS, 2002/95/EC) and Waste Electrical & Electronic Equipment (WEEE, 2002/96/EC) Directives. The second of these mandates a high degree of recyclability of products at their end of life, and the first limits the use of certain materials in any product's construction. The WEEE Directive may impact on techniques such as conductive coatings on plastic, or form-in-place gaskets. In cases where different materials are combined into one component, the recyclability of such components becomes very difficult.

The RoHS Directive places restrictions on the content of the following materials:

- lead;
- mercury;
- cadmium;
- hexavalent chromium;
- polybrominated biphenyls (PBB);
- polybrominated diphenyl ethers (PBDE).

Of these, really only hexavalent chromium has implications for EMC, since it is used in chromate conversions of aluminium; as long as you specify trivalent chromate conversion then you will stay clear of this restriction.

14.2.1.1 Gaskets and finger strip

Shielding effectiveness can be improved by reducing the spacing of fasteners between different panels. If you need effectiveness up to 1GHz and beyond, then the necessary spacing becomes unrealistically small when you consider maintenance and accessibility. In these cases the conductive path between two panels or flanges can be improved by using any of the several brands of conductive gasket, knitted wire mesh or finger strip that are available. The purpose of these components is to be sandwiched in between the mating surfaces and conform with the irregularities of each surface, to ensure continuous contact across the joint, so that shield current is not diverted (Figure 14.10). Their effectiveness depends entirely on how well they can match the impedance of the joint to that of the bulk shield material.

You must bear in mind a number of factors when selecting a conductive gasket or finger material:

- its conductivity, which should be of the same order of magnitude as the panel material;
- compressibility and compression set, which should be matched to the mechanical deflection expected of the mating surfaces or mounting channel;
- ease of mounting: gaskets should normally be mounted in channels

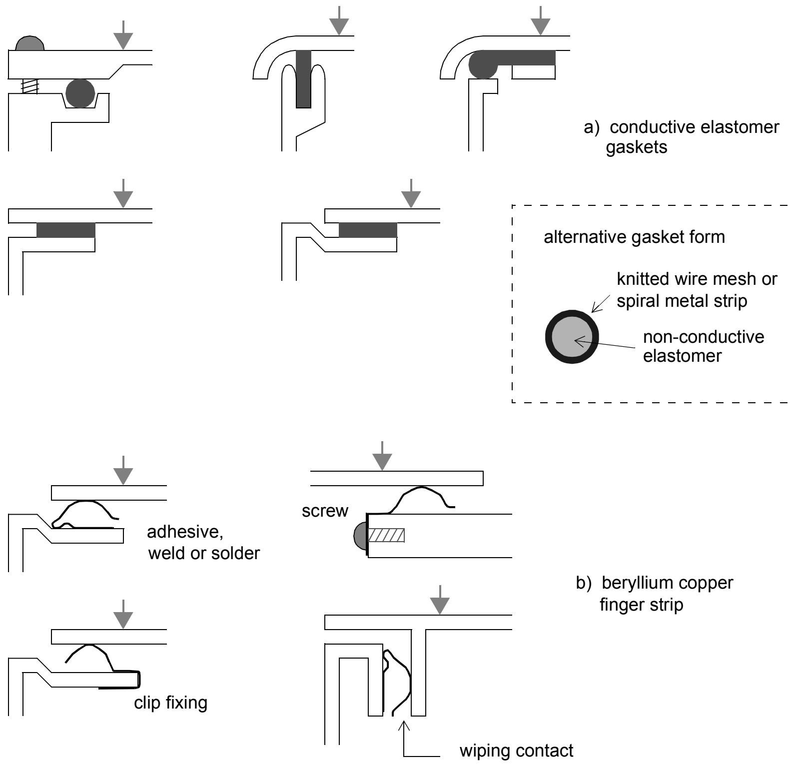


Figure 14.10 Usage of gaskets and finger strip

machined or cast in the housing, and the correct dimensioning of these channels is important to maintain an adequate contact pressure without over-tightening. Finger strip can be mounted by adhesive tape, welding, riveting, soldering, fasteners or by just clipping in place. The right method depends on the direction of contact pressure;

- galvanic compatibility with its host: to reduce corrosion, the gasket metal and its housing should be close together and preferably of the same group within the electrochemical series (Table 14.2, and see next section);
- environmental performance: conductive elastomers can offer combined electrical and environmental protection, but may be affected by moisture, fungus, weathering or heat. If you choose to use separate environmental and conductive gaskets, the conductive gasket should be placed *inside* the environmental seal, and also inside screw mounting holes.

There are a number of varieties of conductive gasket, and Table 14.1 describes their individual properties.

Table 14.1 Conductive gaskets

Material	Features
Form in Place (FiP)	A conductive elastomer paste is dispensed by numerically controlled machine as a continous strip or semi-circular profile bead onto a metal or metallized part and then cured, creating a single piece part before assembly and eliminating waste; generally a high volume process, conductivity depends on filler
Mould in Place (MiP)	Conductive elastomer ribs are moulded directly onto a substrate instead of being mechanically attached; a rib pattern can be moulded onto any metal or plastic substrate, creating a multi-compartment shielded enclosure, usually used in conjunction with PCBs
Fabric over foam	Metallized textile formed over urethane foam core, various profiles, flame retardant; good for applications requiring low mating pressure and large deflection, high conductivity
Conductive elastomer	Can be extruded in various profiles including hollow for increased deflection or die cut from flat sheet, variety of elastomer compounds and metallic fillers, can provide electrical and environmental seal in one part if carefully applied, mostly used in bolt-down applications; conductivity depends on filler, can be quite expensive
Spring finger stock	Historically the original conductive gasket, wide range of configurations available from many suppliers, high conductivity, good for repeated cycle applications such as doors with wiping or compression action, wide temperature range but relatively susceptible to damage; available in beryllium copper, stainless steel and phosphor bronze
Spiral wrap	Similar to finger stock in application, only mounted in a groove, usual material stainless steel, may be combined with elastomer
Oriented wire	Monel or aluminium wire bonded into flat silicone sheet provides a multiple spring effect with individual contact points cutting into metal surfaces; can be die cut, provides environmental seal
Knitted wire mesh	Monel, steel or aluminium wire knitted into various cross-section profiles, high conductivity, economical and resilient but only suited for occasional cycling, can be sheathed over elastomer core

14.2.1.2 Surface treatment

Apart from actually ensuring physical contact, the difficulty with the conductivity of metal surfaces is twofold:

- they will oxidize in contact with air, and some oxides are non-conductive;
- in contact with other metal surfaces, they may suffer electrochemical corrosion.

Any mating surfaces should be conductively finished to prevent insulating oxides from forming – alochrome or alodine for aluminium, nickel or tin plate for steel. (It should be blindingly obvious that mating surfaces should never be painted! Where you have a painted metal enclosure, the surfaces which are to be in contact should be masked from the paint and conductively finished.) For surfaces which will be subject to repeated wear, the hardness of the plating which is used is important. Bright tin and nickel offer an adequate combination of conductivity and hardness for general applications. For best conductivity combined with low wear, use gold or rhodium; silver has the highest conductivity, but the poorest wear resistance.

Corrosion is a significant factor in contact resistance. The by-products of corrosion are usually effective insulators. Corrosion occurs continually at any surface, but you can slow down its rate enough to avoid problems within the lifecycle of the unit that is being shielded. To control it, you need to select mating metals to have similar electrochemical potentials, within adjacent or preferably the same groups in the electrochemical series (Table 14.2); and you need to create a barrier between the mating faces and the atmosphere, to starve the chemical reaction of oxygen and electrolytes such as atmospheric salts. This can mean painting the assembled piece parts, which precludes separating them again, or ensuring a gas-tight joint, which means tightening fasteners adequately and providing an environmental seal.

Table 14.2 The electrochemical series

Anodic – most easily corroded ⟶				
Group I	Group II	Group III	Group IV	Group V
Magnesium	Aluminium + alloys			
	Zinc	Carbon steel		
	Chromium	Iron	Nickel	
	Galvanized iron	Cadmium	Tin, solder	Copper + alloys
			Lead	Silver
			Brass	Palladium
			Stainless steel	Platinum
				Gold
		⟶	Cathodic – least easily corroded	
Corrosion occurs when ions move from the more anodic metal to the more cathodic, facilitated by an electrolytic transport medium such as moisture or salts				

14.2.2 Conductive coatings

Many electronic products are enclosed in plastic cases for aesthetic or cost reasons. These can be made to provide a degree of electromagnetic shielding by covering one or both sides with a conductive coating [36][116]. Normally, this involves both a moulding supplier and a coating supplier. You can also use conductively filled plastic composites to obtain a marginal degree of shielding (around 20dB); it is debatable whether the extra material cost justifies such an approach, considering that better shielding performance can be offered by conductive coating at lower overall cost [43]. Conductive fillers affect the mechanical and aesthetic properties of the plastic, but their major advantage is that no further treatment of the moulded part is needed. Another problem is that the moulding process may leave a "resin rich" surface which is not conductive, so that the conductivity across seams and joints is not assured. As a further alternative, metallized fabrics are available which can be incorporated into some designs of compression moulding.

14.2.2.1 Shielding performance

The same dimensional considerations apply to apertures and seams as for metal shields;

an added factor is that any scratch or crack which breaks through the coating acts as an aperture and degrades the shielding effectiveness. Thin coatings will be almost as effective against electric fields at high frequencies as solid metal cases but are ineffective against magnetic fields. The major shielding mechanism is E-field reflection loss (Figure 14.2, R_E) since absorption is negligible except at very high frequencies, and re-reflection (B) will tend to reduce the overall reflection losses. The higher the resistivity of the coating, the less its efficiency. For this reason nickel paints, which have a resistivity of around 1Ω/square, make poorer shields than silver or copper paints or the various types of metallization (see Table 14.3) which offer resistivities below 0.1Ω/square.

14.2.2.2 Enclosure design

Resistivity will depend on the thickness of the coating, which in turn is affected by factors such as the shape and sharpness of the moulding – coatings will adhere less to, and abrade more easily from, sharp edges and corners than rounded ones. Ribs, dividing walls and other mould features that exist inside most enclosures make application of sprayed-on coatings (such as conductive paint or zinc arc spray) harder and favour the electroless plating methods. Where coatings must cover such features, your moulding design should include generous radii, no sharp corners, adequate space between ribs and no deep or narrow crevices (Figure 14.11).

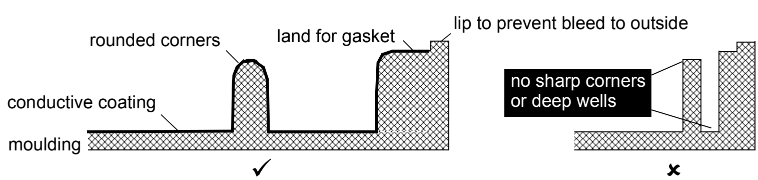


Figure 14.11 Moulding design for conductive coatings

14.2.2.3 Coating properties

Environmental factors, particularly abrasion resistance and adhesion, are critical in the selection of the correct coating. Major quality considerations are:

- will the coating peel or flake off into the electrical circuitry?
- will the shielding effectiveness be consistent from part to part?
- will the coating maintain its shielding effectiveness over the life of the product?

Adhesion is a function of thermal or mechanical stresses, and is checked by a non-destructive tape or destructive cross-hatch test. Typically, the removal of any coating as an immediate result of tape application constitutes a test failure. During and at completion of thermal cycling or humidity testing, small flakes at the lattice edges after a cross-hatch test should not exceed 15% of the total coating removal.

Electrical properties should not change after repeated temperature/humidity cycling within the parameters agreed with the moulding and coating suppliers. Resistance measurements should be taken from the farthest distances of the test piece and also on surfaces critical to the shielding performance, especially mating and grounding areas.

Table 14.3 compares the features of the more commonly available conductive coatings (others are possible but are more expensive and little used). These will give shielding effectiveness in the range of 30–70dB if properly applied. It is difficult to compare the shielding effectiveness figures from different manufacturers unless they specify very clearly the methods used to perform their shielding effectiveness tests; different methods do not give comparable results. Also, laboratory test methods may not correlate with the performance of a practical enclosure for a commercial product.

Table 14.3 Comparison of conductive coating techniques

	Cost £/m^2	E-field shielding	Thickness	Adhesion	Scratch resistance	Maskable	Comments
Conductive Paint (silver, copper)	5–15	Average	5–25μm	Poor	Poor	Yes	Suitable for lower quantities
Zinc Arc Spray (zinc)	5–10	Average/ good	0.1– 0.15mm	Depends on surface prep	Good	Yes	Rough surface, inconsistent
Electroless Plate (copper, nickel)	10–15	Average/ good	1–2μm	Good	Poor	No	Cheaper if entire part plated
Vacuum Metallization (aluminium)	10–15	Average	2–5μm	Depends on surface prep	Poor	Yes	Poor environ- mental qualities

Finally, be sure if you specify conductive coatings on an existing product that in doing so you are not compromising safety requirements. Particular issues are:

- the stability of adhesion of the coating – make sure as above that it cannot flake and bridge safety isolation paths or cause a fault;
- the effect of the coating on the flame retardant properties of the enclosure;
- the effect on creepage and clearance distances for safety insulation. A conductive coating applied without thought could reduce the isolation between safe parts and hazardous live parts by providing a new conducting path which is close to each of these.

14.2.3 Windows and ventilation slots

Viewing windows normally involve a large open area in the shield and you have to cover the window with a transparent conductive material, which must make good continuous contact to the surrounding shield, or accept the penalty of shielding at lower frequencies only. You can obtain shielded window components which are laminated with fine blackened copper mesh, or which are coated with an extremely thin film of gold or indium tin oxide (ITO). In either case, there is a trade-off in viewing quality versus a clear window, due to reduced light transmission (between 60 and 80%) and diffraction effects of the mesh. Shielding effectiveness of a transparent conductive coating is significantly less than a solid shield, since the coating will have a resistance of a few ohms per square† and attenuation will be entirely due to reflection loss. This

† The resistance across two sides of a square is independent of the size of the square, and hence the units of "ohms per square" are entirely reasonable and not a misprint. This and surface resistivity (ohms) are just two conventions for specifying the same thing, and are numerically equivalent.

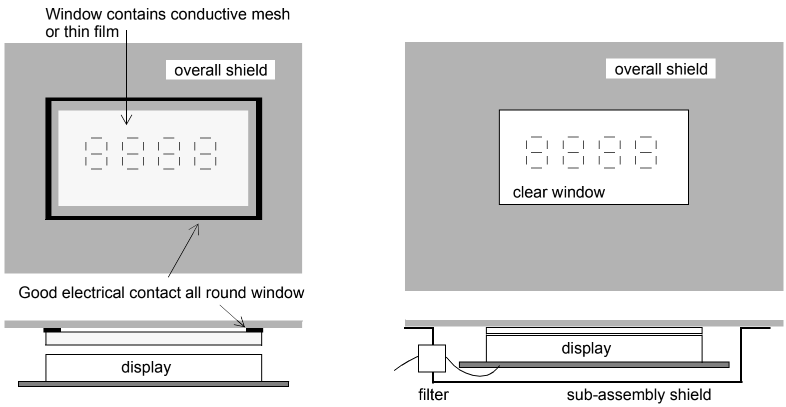


Figure 14.12 Alternative ways to shield a display window

is not the case with a mesh, but shielding effectiveness of better than 40–50dB may be irrelevant anyway because of the effect of other apertures. Shielded windows are certainly costly and rarely suited to consumer applications.

An ITO layer thickness of 1 micron will have a sheet resistance of about 10 ohms per square and a visible light transmission of around 80%. A simple and very approximate equation for the far field shielding attenuation versus frequency of such a layer is given by [63]

$$SE = 20 \log [(7 \cdot 10^{11}) / (f \times R)] \qquad (14.5)$$

Where f = frequency (Hz)
R = sheet resistance (Ohms/square)
SE = far field shielding effectiveness (dB)

14.2.3.1 Using a sub-enclosure

An alternative method which allows you to retain a clear window, is to shield behind the display with a sub-shield (Figure 14.12), which must make good all-round contact with the main panel. The electrical connections to the display must be filtered to keep the shield's integrity, and the display itself is unshielded and must therefore not be susceptible nor contain emitting sources. This alternative is often easier and cheaper than shielded windows. It can also be applied to other apertures besides windows: for example, a large number of connections may be made to the equipment via terminal blocks, and it is not effective to have these mounted within a shielded enclosure since the incoming cables breach the shielding. Instead, you can mount the terminal blocks in a compartment outside the primary shielding, and have the connections between the terminal blocks and the main circuit taken through the shielding at a filtered interface (Figure 14.13).

14.2.3.2 Mesh and honeycomb

Ventilation apertures are frequently a cause of difficulty for shielded enclosures. They can be covered with a perforated mesh screen, or the conductive panel may itself be perforated. If individual equally sized perforations are spaced close together (hole

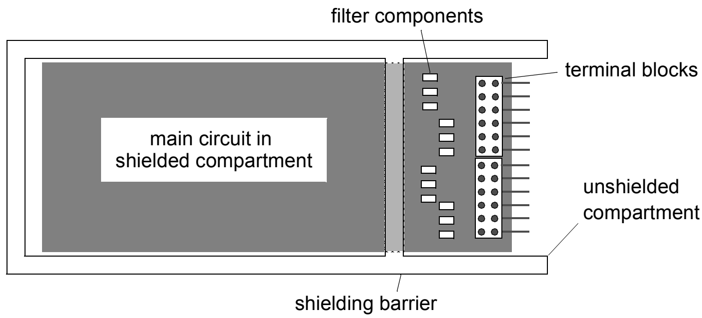


Figure 14.13 Terminals in an unshielded compartment

spacing << λ/2) then the reduction in shielding over a single hole is approximately proportional to the square root of the number of holes. Thus a mesh of 100 4mm holes would have a shielding effectiveness 20dB worse than a single 4mm hole. On the other hand, *for a fixed open area*, the shielding effectiveness *improves* proportionally to the square root of the number of holes: in other words, ventilation is always better provided by a mesh of many small holes rather than a few large ones. Two similar apertures spaced greater than a half-wavelength apart do not create any significant extra shielding reduction over a single aperture.

Wire mesh rather than perforated holes can be similarly treated, provided that the wires make good electrical contact at each crossover or intersection, and provided that the mesh is bonded to the main panel all around its edge. If the wires are corroded or loose at each intersection, the mesh will not provide the expected shielding results.

Honeycomb

You can if necessary gain improved shielding of vents, at the expense of thickness and

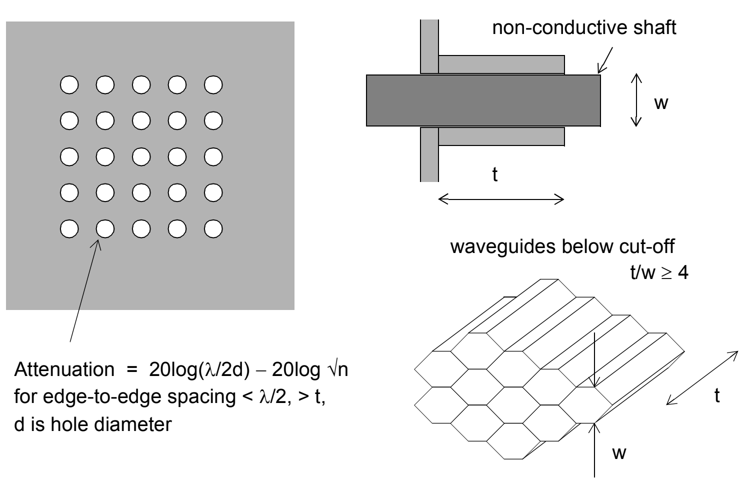


Figure 14.14 Mesh panels and the waveguide below cut-off

weight, by using “honeycomb” panels in which the honeycomb pattern functions as a waveguide below cut-off (Figure 14.14). In this technique the shield thickness is several times that of the width of each individual aperture, and the honeycomb is a cluster of waveguides. Each aperture is less than a half-wavelength in dimension w at the highest frequency which must be blocked. Thus for instance below 1GHz any tube of diameter less than 15cm will function as such a cut-off waveguide.

Under these circumstances and for frequencies several times less than the cut-off, field propagation down each aperture is attenuated at a rate of about 27dB (for rectangular cross-section, 32dB for circular) over each distance w in the direction t, through the thickness of the panel. A common t/w ratio is 4:1 which offers an intrinsic shielding effectiveness of over 100dB. This method can also be used to conduct insulated control spindles (*not* conductive ones, or cables!) through a panel.

In the construction of honeycomb vent panels, the foil strips making up the cells are treated so that the individual cells make good electrical contact with each other and with the mounting frame in all directions. Shielding effectiveness of honeycomb material is significantly improved when the complete panel is plated with a conductive material, such as tin. For these panels, with many waveguides in parallel, the loss down a single waveguide is reduced by a factor $10 \cdot \log_{10}(n)$ related to the total number of openings n in the panel. Thus a panel with 100 individual cells will perform 20dB worse than a single individual cell.

14.2.4 Shields on the PCB

Carrying on from the idea of shielding only part of an enclosure, it is logical to go one step further and apply a sub-shield as a single component on a PCB. This allows you to shield only those components or sections that need it, and if necessary to create multiple shielded sections on one PCB with a single formed part. This practice is not just for external EMC – maybe there is a wireless interface on the board which needs its own shielding regime; or it may be necessary for purely functional reasons to protect a low-level radio receiver from local microprocessor noise.

In virtually all cases the shielding component will work in conjunction with the circuit’s 0V plane. It is very rare, though not impossible, to apply shields to a board without a 0V plane. The plane is normally expected to provide the bottom face of the shield, which then itself prevents E-field coupling to or from the components and upper surface tracking. For this to work into the GHz range, the connection between the shield and the plane must have a very low inductance. This, as ever, means multiple connections all around the sides of the shield. The equivalent circuit for such an arrangement, assuming that various ICs are the source of E-field noise, can be viewed as shown in Figure 14.15.

What we are normally trying to prevent is coupling of V_N to external structures, so that unwanted disturbance currents don’t flow in these structures. The shield will couple it instead back to 0V, but only provided that the mounting inductance L is negligible. Purely for illustration, consider two sets of figures (Figure 14.16): with coupling capacitance of 10pF on both sides of the shield, the mounting inductance may be 5nH (poor) or 0.5nH (relatively good). The result is a peak in coupling (zero shielding effect) well below 1GHz for the high inductance, but reasonably above it for the low.

The best form of mounting is to have a surface land on the PCB that mates with a lip on the wall of the shield, all around its perimeter, with many vias through to the 0V plane, so that the shield wall is soldered all along its length. Mounting with individual

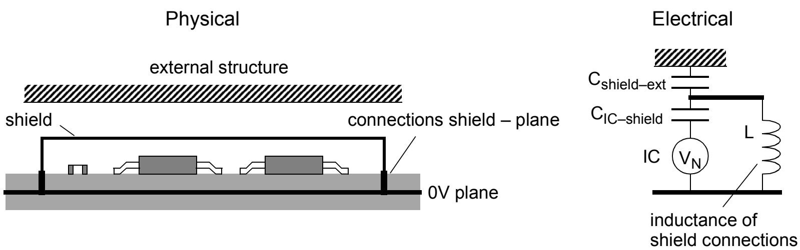


Figure 14.15 Equivalent circuit for E-field shield on a PCB

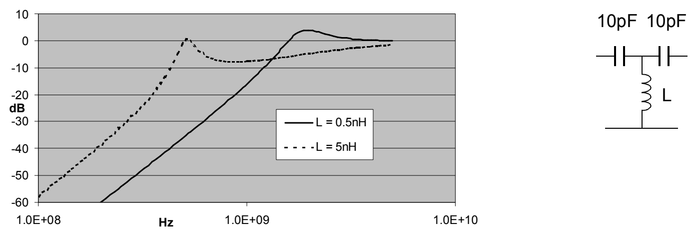


Figure 14.16 Attenuation through a shield with mounting inductance L

pins through the board is a second best; if you're going to use it, as a through-hole component, there should be as many pins as possible around the perimeter.

With this fundamental requirement addressed, actually designing the shield is usually simple. If you will need access to the top of the PCB later, for testing or rework, then a two-part shield is necessary: a wall or fence which is soldered to the PCB, and a lid which is subsequently clipped to the wall and is removable. Alternatively a five-sided box permanently soldered to the PCB can be used if no access is needed. Usually, tin plated steel is more than adequate for the purpose, both electrically and mechanically. Depending on volume, the part can be photochemically machined or stamped, and supplied ready for surface mounting in pick-and-place waffle packs. As is to be expected, the cost of the part will depend on its size and the complexity of forming.

14.2.5 Standardization of enclosure SE

A recent departure in the IEC 61000 series of standards has been to publish a method for the description of shielding effectiveness of commercial enclosures, similar to the IP rating scheme for environmental protection. IEC 61000-5-7 [157] specifies the "EM" rating, in which the designator EM*ABCDEF* specifies the shielding performance in each of six frequency ranges. Table 14.4 gives the meaning of each of these designators. The standard also specifies the SE test methods to be employed for each frequency range. It will be interesting to see how popular this designation becomes in the future.

Table 14.4 The IEC 61000-5-7 EM shielding code

Frequency range	Shielding designator	Shielding performance (dB)	Designator value
10kHz–100kHz	*A*	Untested	x
100kHz–1MHz	*B*	<10	0
1MHz–30MHz	*C*	≥10	1
30MHz–1GHz	*D*	≥20	2
1GHz–10GHz	*E*	≥30	3
10GHz–40GHz	*F*	≥40	4
		≥50	5
		≥60	6
		≥70	7
		≥80	8
		≥100	9

Example: EM66644x provides ≥60dB SE from 10kHz to 30MHz, and ≥40dB from 30MHz to 10GHz

Chapter 15

Systems EMC

15.1 System versus product EMC

The title of this book is *EMC for Product Designers* and that is its primary focus. But many individual products, that is electronic equipment and apparatus, are actually put together in systems which are required to function as a whole. This then changes the EMC context. There are now two aspects to consider: the compatibility with each other of the various items of equipment within the system, which we can call *intra*-system EMC; and the compatibility of the whole system with its environment and with other systems and apparatus in that environment, which we can call *inter*-system EMC. A complete view must take into account both aspects.

Designers of products which are going to be used in this way need to be aware of the two aspects, as do the designers and installers of the systems themselves. A companion book to this one, called *EMC for Systems and Installations* [22], treats the subjects raised by EMC at this level in much greater detail, and the systems designer is recommended to that book for further reading. Meanwhile, this chapter will serve as an introduction to the issues that face systems designers, from the point of view of a product designer whose projects must be capable of installation and interfacing at the systems level. Some of these issues have to do with the EMC compliance of the whole system, but more of them are related to interactions between equipment that may be installed either in close proximity or in different locations, and which therefore require EMC precautions that go beyond simple compliance.

15.1.1 Compliance requirements

As we have seen in Chapter 2, the 2nd edition EMC Directive has noticeably extended the requirements that apply to fixed installations. The substantive requirement is:

> A fixed installation shall be installed applying good engineering practices and respecting the information on the intended use of its components, with a view to meeting the protection requirements. These good engineering practices shall be documented and the documentation shall be held by the responsible person(s) at the disposal of the relevant national authorities for inspection purposes as long as the fixed installation is in operation.
>
> – [183] Annex 1(2)

To help to meet this requirement, whenever practicable the project manager, system designer or installation engineer should use recognized practices and procedures for EMC as given in international, European or national standards, technical reports, specifications or codes of practice. There aren't that many of them, but this chapter makes reference to the few standards that exist and discusses some of the methods recommended by them.

15.1.2 Functional requirements

Irrespective of the need for compliance with external compatibility legislation, a system or installation must actually work properly. Part of this is that its sub-systems should not interfere with each other. Issues arise when equipment either is located in close proximity to disturbance sources or victims, or is subject to disturbances which are developed in other parts of the installation and unintentionally coupled in by structural components. Under these circumstances, the construction of the interconnections and layout within a system becomes important [39][67]; earthing, bonding and cable layout must be designed rather than allowed to go by default, and they should be designed to separate the unavoidable interference currents from the required functional interfaces.

The installation techniques described in this chapter can be regarded as best practice. Nevertheless, it is still true that the best equipment design will be one which puts no restrictions on earthing, cable routing and segregation – i.e. one where the major EMC design measures are taken internally. There are many application circumstances when the installation is carried out by unskilled and untrained technicians who ignore your carefully specified guidelines, and the best product is one which works even under these adverse circumstances.

15.2 Earthing and bonding

15.2.1 The purpose of the earth

In a system context we can identify four purposes for earthing.

15.2.1.1 Safety earth

The purpose of the safety earth is to guarantee personnel safety under fault conditions. The IEE Wiring Regulations (BS 7671, [175]) define "earthing" as:

> Connection of the exposed conductive parts of an installation to the main earthing terminal of that installation.

Earthing provides a low-impedance path in which current may flow under fault conditions. Exposed conductive parts are those conductive parts of equipment which may be touched and which may become live in the case of a fault. The earthing connection prevents such live parts from reaching a hazardous voltage. The protective conductor (typically colour coded green-and-yellow) provides an electrical connection which maintains various exposed and extraneous conductive parts at substantially the same potential under both operational and fault conditions, and also connects the conductive parts to the installation's main earthing terminal. The prospective touch voltage within the installation is then the product of the impedance of the protective conductor and the earth fault current.

This creates a zone within which exposed and extraneous conductive parts are maintained at "substantially" the same potential. Although the voltages within such a zone may be safe, they are not necessarily, and not even usually, zero. Continuous currents from various sources, including equipment earth leakage, are likely to be flowing, even in a "healthy" circuit. (Allowable earth leakage levels from individual items of equipment are covered in section 13.2.3.3.) Such an "equipotential" zone may protect people but is not guaranteed to protect equipment or wiring.

Protection against electric shock is typically provided by earthing in conjunction with automatic disconnection of the supply. For this purpose, the protective device must be co-ordinated with the installation's earth fault impedance, to disconnect quickly enough to prevent the touch voltage from reaching a hazardous level. The sizing and hence resistance of the earth protective conductor will therefore be determined largely by the prospective fault current available from the rest of the system. Since the concern is low frequencies, it is resistance rather than inductance which determines the conductor impedance; this is not the case for high frequency earths.

15.2.1.2 Functional earth

For an electrical circuit to interface correctly with other equipment, there must be a means both of relating voltages in one equipment to those in another, and of preventing adjacent but galvanically separate circuits from floating.

This is the purpose of the functional earth, and it must be distinguished from the safety protective earth. Because of the threat of circulating currents and potential differences between earthing zones, there may be practical constraints on the widespread use of functional earthing on large systems, especially since there is normaly no explicit requirement for conductor cross-section to maintain low impedance. Signal circuits of equipment should normally be specified for a maximum common mode voltage, which will be the voltage that appears between different parts of a functionally-earthed system. If this is impractical or inadequate, isolated circuit interfaces are the normal solution.

15.2.1.3 Lightning protection earth

In building installations, a further important safety-related earthing function is to provide a return connection for currents induced by a lightning strike. In many respects this is the *only* correct use of the term "earth", since this function is normally provided by ensuring a low-impedance connection throughout the building fabric to the literal earth on which the building sits. Since lightning potentials are built up between the cloud structure of a thunderstorm in the atmosphere and the surface of the Earth, connection to earth is the correct way to complete the circuit in the shortest manner.

Several standards for lightning protection have been published by the various standards bodies (e.g., BS 6651 [172]) and you should look to these for detailed advice (Table 15.1). Section 15.5 reviews the main principles.

Table 15.1 Standards for lightning protection

	Protection of structure	Protection of contents	Risk assessment
IEC/ CENELEC	IEC 62305-3: 2006 EN 62305-3: 2006	IEC 62305-4: 2006 EN 62305-4: 2006	IEC 62305-2: 2006 EN 62305-2: 2006
BSI	BS 6651:1999	BS 6651:1999 App C	BS 6651:1999

15.2.1.4 EMC earth

The EMC earth has the sole purpose of ensuring that interfering voltages are low enough compared to the desired signal that incorrect operation or excessive emission does not occur [135]. It has no explicit safety or operational function. Because of this, and because of the wide frequency range over which it must work, earthing for EMC usually takes advantage of distributed structural components that are part of the whole system – typically, chassis members, enclosure panels and so on. The value of an EMC earth is directly related to its physical geometry. This means that design and implementation of an EMC earthing system is not restricted to the electrical engineering discipline alone – it must also involve constructional aspects, that is, the mechanical designers and installers.

15.2.2 Installation techniques for multi-purpose earthing

15.2.2.1 Three-dimensional meshed equipotential earth-bonding

The preferred earth-bonding method in buildings is the Common Bonded Network, or CBN, shown by Figure 15.1. Complex installations require an equipotential system

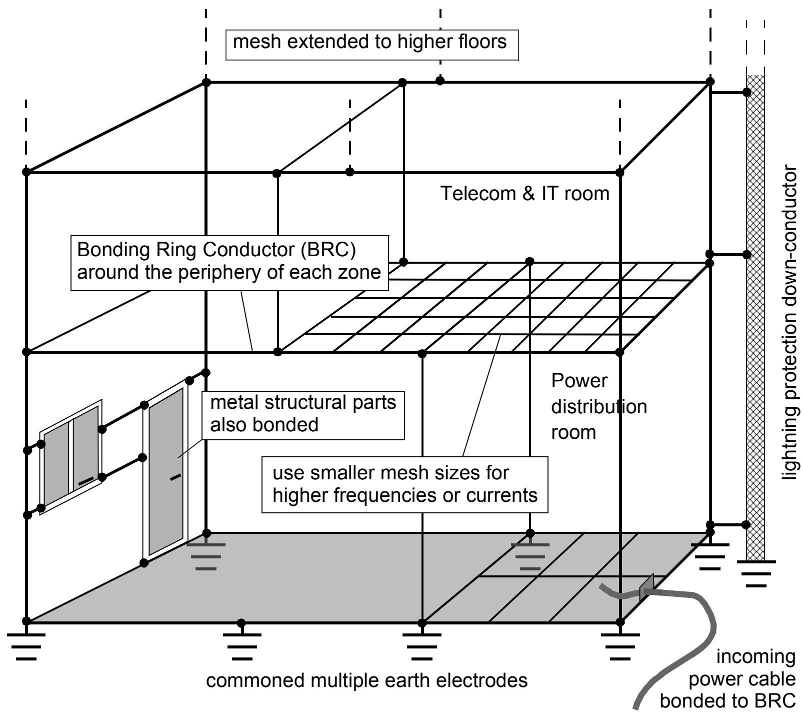


Figure 15.1 Common bonded network

meshed in three dimensions, often referred to as a MESH-BN (for Mesh Bonding Network). This bonds every piece of structural and non-structural metalwork together to make a very highly interconnected system, which is then connected to the lightning protection system (LPS) at ground level and possibly other floors. Whilst Figure 15.1 shows the application within buildings, the principle applies to an installation in any circumstances: aircraft, ships and vehicles should all follow the same practice.

This highly-meshed three-dimensional system is then interconnected to the screens and armouring of all electrical cables, and the frames or chassis of every piece of electronic equipment. Where existing metalwork or conductors do not already exist, heavy gauge conductors are added to complete the mesh either vertically or horizontally so that nowhere is the mesh size greater than about 3 or 4m. The main earthing terminal for the incoming power supply to the building needs a number of bonds to the MESH-BN.

The MESH-BN gives safety, functional, and EMC earthing, all at the same time – an integrated earth-bonding system. With it you can achieve the various aims of safety, signal integrity, equipment reliability, and EMC, which are often seen as being in conflict, at reasonable cost in a reasonable time, without compromises, and without restricting future modifications.

Although each of the MESH-BN elements will resonate and disturb the earth bond at the resonant frequencies – refer to section 11.1.3.4 which discusses this phenomenon – its highly interconnected nature will ensure that there are alternative current paths that are not resonating, and so provide a high degree of equipotentiality over a wide frequency range. One consequence of this is that very regular bonding structures should be avoided, since all their elements would exhibit similar resonances at the same frequencies.

To limit voltage differences in the earth structure at higher frequencies for the same level of power, or at higher powers for the same frequency, the mesh size of the MESH-BN needs to be smaller. With the different types of apparatus having been segregated according to whether they are "noisy" or "sensitive", the building should then be partitioned into areas with different earth mesh sizes, depending on the earthing needs of each [156], as shown by Figure 15.1.

Note that each segregated apparatus area with its individual meshing or bonding is surrounded by a complete conductor. These are known as bonding ring conductors (BRCs) and protect the apparatus in an area against lightning transients, earth faults, and other low frequency surges originating outside of their area.

15.2.2.2 The bogey of ground loops

A common objection to the meshed earthing system is that it creates "ground loops" (equally known as "earth loops"). Historically, currents flowing in ground loops, and their associated driving potential differences across different parts of the earth network, have been found to be particularly serious contributors to interference problems, and therefore a practice has developed of trying to eliminate all such loops. This practice, although often superficially successful, is unfortunately misguided.

In a situation where high earth potential differences exist, closing a loop between two such earth points will allow a high current to flow in the structure. If the conductors in that loop include a segment which either forms part of, or is closely coupled to, a signal or low-level power cable, then substantial interference can be induced in the circuits of that cable. If the loop is opened, the current no longer flows, and the interference disappears – although the high potential differences remain, ready to create problems again when another loop is closed somewhere else. This is the principle which is formalized in the star or single point earth regime: remove all ground loops and live with the resultant high voltages between different parts of the earthing system.

Such an approach is fairly easy to implement and quite successful in simple low-frequency systems, but it represents a retreat from best practice. With interference frequencies measured in MHz rather than Hz, it is untenable. This is because the star earthing conductors present a high impedance to these frequencies and therefore *de*couple a system from earth, rather than couple to it. Also unfortunately, larger star systems tend to degenerate into accidentally ground-looped systems as time passes and systems and buildings are modified and added to, requiring a heavy management and control burden to maintain their efficiency and ensure safety and equipment reliability.

In these circumstances the only reliable earthing system is a mesh. The mesh does indeed provide a multiplicity of ground loops, but they are small and controlled: voltage differences between parts of the structure are minimized, resulting currents are low and the interference consequences, if any, are negligible.

15.2.3 Earth conductors

15.2.3.1 Short fat straps

Elsewhere in this book we have discussed the impedance of wires and structures (section 11.1.3.4). In the context of systems earthing, a number of points are relevant:

- any length of wire becomes predominantly inductive above a few kHz; short fat wires have a higher transition frequency than long thin ones;
- the inductive impedance of typical lengths of wire reaches ohms at around 1MHz, and tens of ohms in the tens of MHz range;
- the impedance of a length of wire connected at one end to an earth reference plane reaches a resonant maximum when its length is a multiple of a quarter wavelength, and falls to a resonant minimum at multiples of a half wavelength;
- the exact frequencies at which these resonant peaks and nulls occur are strongly affected by layout; if any of them coincides with a susceptible or emissive frequency of the equipment, surprising and unpredictable variations in equipment performance will be brought about simply by moving such a wire by a few centimetres.

The general rule with earth wires is: short fat straps have the lowest impedance, as suggested in Figure 15.2. But even short straps are not perfect. A tinned copper braid 10cm long by 9mm wide by 2mm thick, for instance, still has substantial impedance in the hundreds of MHz. Its merit is that those resonances which still exist are pushed much higher in frequency and exhibit a much lower Q, thus reducing their impact usually to negligible proportions.

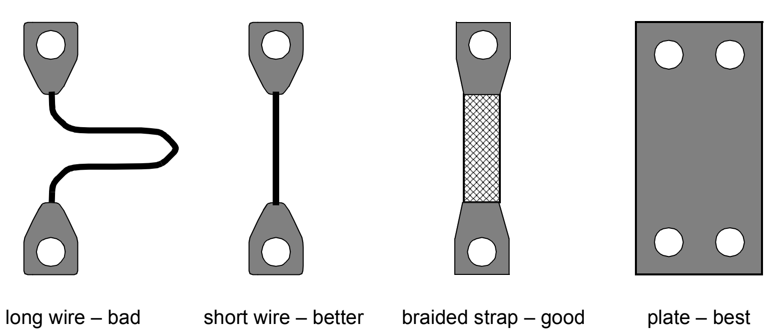


Figure 15.2 Hierarchy of earth conductors

15.2.4 Bonding techniques

The safety bond – a length of green-and-yellow wire interconnecting panel and frame, or different structural parts – is familiar to system builders. It is vital to realize that *this* *is not adequate* for EMC bonding. The purpose of a wired safety bond is to prevent different parts of the structure from assuming different potentials and hence presenting an electric shock hazard at power frequencies: it has no other purpose. It cannot give a low-impedance connection at RF.

This is not to say that the safety bond is forbidden for EMC purposes: it can coexist quite happily with a proper EMC bond. But the one is no substitute for the other. If you are intent on building an RF-adequate enclosure or structure, then full metal-to-metal connection at all joints is required:

- bonds are best made by surface-to-surface conductive contact at frequent intervals, or preferably continuously along a seam; "bonding" straps, although necessary in many circumstances and preferable to wire connections, are a second best option;
- bonding between parts requires removal of insulating layers, for instance paint or anodizing, and often the treatment of mating surfaces to ensure conductivity, for instance zinc plating or chromate conversion;
- positive pressure is required to make a bond; fasteners will provide this but usually the gap between two fasteners does not allow pressure to be maintained, hence the use of conductive gaskets;
- once a bond is made between two surfaces, it should be protected from corrosion by being made gas-tight or by applying some type of overall coating.

15.2.4.1 Bonding of equipotential mesh structures

As described in section 15.2.2.1, a three-dimensional earth structure is required to provide equipotentiality over a wide range of frequencies.

All structural metalwork and cable supports should be RF bonded across all their joints, and RF bonded between each other whenever they are close enough, to make a three-dimensional earth mesh. Plumbing, pipework, air ducts, chimneys, re-bars, I-beams, cable trays, conduits, walkways, ladders, ceiling supports, etc., should all be RF bonded, as shown in Figure 15.3. Building steel and reinforcing rods should have welded joints and a sufficient number of access points to them for frequent bonds to the earthing network to create the appropriate mesh size for the MESH-BN.

The length of the connection between a structural item and the common bonded network should not be more than 0.5m, and an additional connection should be added in parallel some distance away. Connecting the earthing bus of the electrical switchboard of an equipment block, or the earth bonding bar of a local AC power distribution cabinet, to the bonding network, should use conductors of under 1m length and preferably under 0.5m. Achieving good signal integrity and EMC performance at frequencies of 100MHz and above requires direct metal-to-metal bonds at multiple points, preferably seam-welded, for each joint.

15.2.4.2 Bonding cable trays and ducts

Galvanized cable trays and rectangular conduits are best jointed by seam-welding, but it is often acceptable to use U-brackets with screw fixings every 100mm or less around the periphery of the U instead. Using lengths of wire will only control low frequencies (such as 50/60Hz). Shorter wires, or short fat braid straps, or multiples of each, all reduce impedance and so help increase the frequencies (or power levels) at which interference can be controlled.

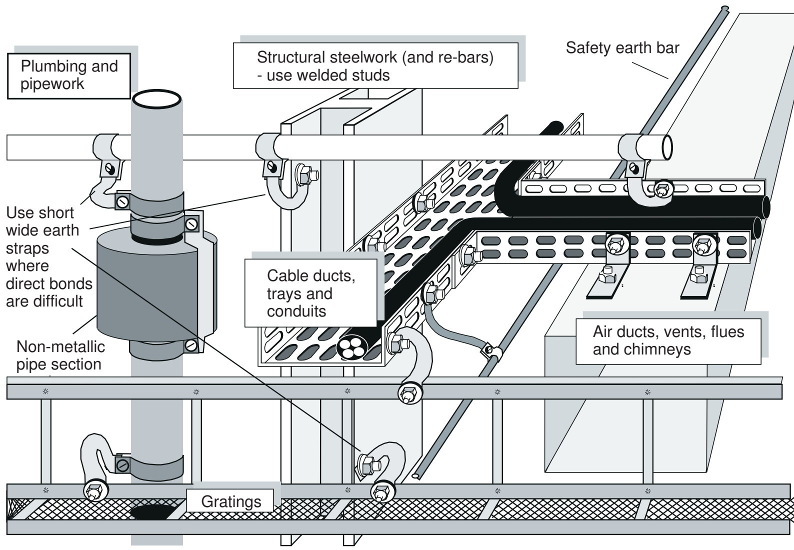


Figure 15.3 Bonding structural components of the equipotential earth mesh

Cable trays, ducts, and conduits will be required to act as Parallel Earth Conductors (PECs), as described in section 15.4.2. The bonding methods at their joints and end terminations should also be appropriate for the frequencies to be controlled. Where a rectangular cable tray or duct terminates at the wall of an equipment cabinet (or similar) two or more straps will give better control of higher frequencies (Figure 15.4). An alternative is to cut away a few inches from the sides of the tray, bend the remaining floor section over and bolt it to the cabinet wall in at least two places. A U-bracket may also be used.

Circular conduits are best jointed (either inline or at corners or junctions), using standard screwed couplings which make a 360° electrical bond. Similar 360° bonding glands should be used wherever a round conduit is terminated at cabinet walls, other types of cable ducts, or similar metal surfaces. These will generally employ some type of conductive gasket in their internal construction.

15.3 Cabinets, cubicles and chambers

In the context of systems EMC, a metal enclosure can have a number of purposes as well as its primary one of providing physical protection and mounting:

- to provide a local earth reference for the internal equipment;
- to provide and demarcate a zone of increased EM protection;
- to prevent radiated field coupling to and from the internal equipment.

You will notice that the conventionally understood function of a metal cabinet – to provide shielding – is placed last in the above list. This is deliberate. As we will see,

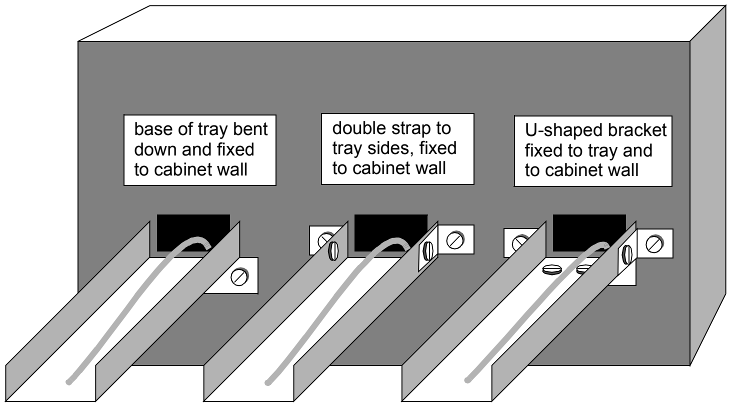


Figure 15.4 Methods of bonding cable trays and ducts to equipment cabinets

many examples of metal housings are likely to function very poorly as shields, because of their surfeit of inadequately treated apertures and seams. But this doesn't mean they have no EMC function. In fact, the first purpose – to provide a local earth reference – is nearly always the most important. This can be achieved by an enclosure which is effectively *un*shielded, provided that care is taken to use the bulk metal of the cabinet in the right manner. Such an approach is very cost-effective, since many of the assembly, installation and maintenance implications of a fully shielded enclosure can be dispensed with.

Simply constructing a metal cabinet or cubicle to house the equipment in the physical sense, is not adequate for electromagnetic purposes. The constructional and assembly methods of the individual structural parts, and the provisions for cable entry/exit, are at least as important as the mere fact of a conductive enclosure. Interference currents in the structure are expected and have to be controlled. So, electrical interconnections between the parts must be fully specified, and any discontinuities in the form of apertures, seams and cable penetrations must be avoided or controlled.

15.3.1 Transfer impedance of the earth reference

The function of providing a local earth reference is critical to the effectiveness of a metal enclosure, whether this enclosure is deliberately intended as a shield or not. The previous section already looked at the principles of earthing and the next will look at cable layout. In between is the area where the cables and equipment are terminated. This area must provide the lowest possible transfer impedance to the internal circuits and equipment, so that interference currents do not couple between the enclosure and the sensitive or noisy circuits within it.

15.3.1.1 Cabinet backplate earths

After a cylindrical or rectangular enclosure, a large flat plate over which the relevant circuits are mounted gives the best transfer impedance performance. Industrial control cabinets in general include a backplate for physical mounting, whilst smaller enclosures provide some sort of metal chassis. Telecomms and IT cabinets have an internal support structure. Mounting all electronic modules on, and terminating all cable screens and parallel earthing conductors to, a backplate or similar chassis provides the lowest achievable transfer impedance in practical terms. But since using the backplate in this way means that it must carry interference currents, the way in which contacts are made to it, and its conductivity, become significant. Zinc plating, and clamp-style cable screen connections, are both recommended.

At high frequencies only a metal area (mesh or plate) can give a reliable low-transfer-impedance earth, so you are best advised to use a solid backplate or chassis of an enclosure as the earth for all internal electronic equipment *instead of* using green/yellow wires to a star point. This calls for heavy zinc-plated metalwork, not painted; at the cost of some inconvenience, you can use painted metalwork as long as the following precautions are taken for all the earth connections:

- remove the paint;
- use star washers to bite into the metal;
- apply suitable corrosion protection after the joint is made.

Terminations of screened cables to the backplate or chassis should be carefully planned and implemented so that all common mode interference currents flow directly through it, and not into the circuits mounted on it.

15.3.2 Layout and placement within the enclosure

15.3.2.1 Cable runs

Coupling of external fields to internal cables, and of local electric and magnetic fields between cables, is greatly affected by the route a cable follows around a system. To minimize coupling of cables with external fields, run the cables close to a well-bonded metal structure which can act as a low-impedance earth reference – this is usually the backplate or chassis. Where a cable leaves the backplate/chassis, ensure that it follows a conductive structure which is electrically bonded to the backplate/chassis. Avoid running cables near to apertures in the structure or enclosure or near to breaks in the bond continuity (Figure 15.5), as the localized fields around these points are high. (This advice is really the same as saying that the internal construction of the enclosure acts as a continuation of the PEC for the cables, as described in section 15.4.2.)

To minimize coupling of cables with each other, segregate different classes of cable and run them with at least 150mm (see section 15.4.1) of separation. Do not allow long runs of closely spaced cable of different segregation classes.

15.3.2.2 Module placement

Carefully position the various items on the backplate/chassis to keep sensitive units such as PLCs, computers or analogue instrumentation away from electrical noise sources such as switches, relays or contactors, and to help achieve segregation of the different cable classes. The important principle here is to assess each item for its interference potential, and to specify the internal layout accordingly.

Figure 15.6 describes an industrial enclosure, showing the cable route to door-mounted equipment, with cables strapped along the short earthing braid between door and cabinet wall; and an example of backplate layout in a motor drive area. The purpose of the cable following the earth strap across the door opening is to minimize coupling of the cable with the door aperture. The earth strap provides continuity across the

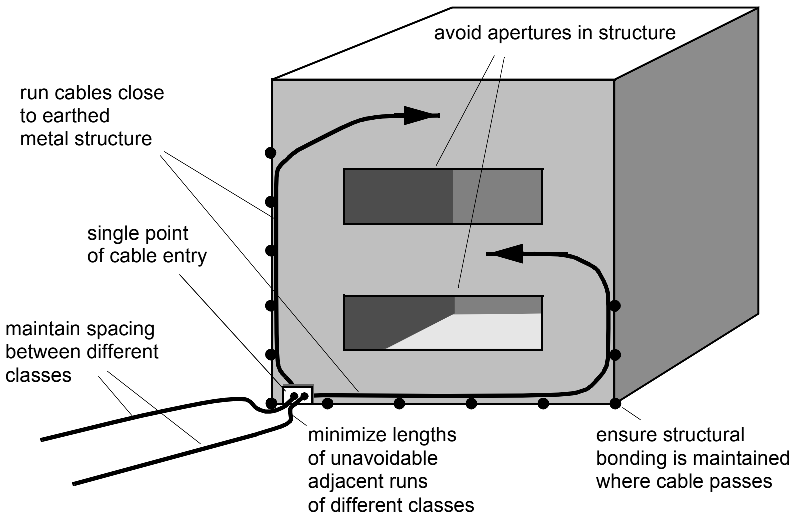


Figure 15.5 Cable runs within the enclosure

aperture, and hence keeps a low transfer impedance between the cable and the cabinet.

Because of their extremely aggressive emissions, inverter drive motor connections are always important [84]. A local return path to the filter has to be provided for the switching noise currents which are flowing back down the earthing conductor and/or the screen of the cable to the motor. These currents can easily pollute the rest of the cabinet and even other equipment in the immediate area of the drive-to-motor circuit, if proper high frequency earth bonding between the cable screen and the filter is not provided. Similar considerations apply for other "noisy" transducer drivers, such as RF-stabilized welding, spark erosion, and Class D audio amplifiers.

15.3.2.3 The clean/dirty box approach

A frequent and effective approach for industrial and other enclosures is to segregate the enclosure into a "clean" compartment and a "dirty" compartment (Figure 15.7). Either the cabinet can have a partition welded into it, or an additional "dirty" enclosure can be bolted or welded to the side of the main "clean" cabinet instead of a dividing plate.

The "clean" compartment is then used for all the electronics which must be shielded from the external environment. All apertures in this part of the enclosure are rigorously controlled. Connections through to the clean volume must be made via 360° screen bonds to the partition plate, or via effectively earthed filters – no untreated cables are allowed (compare this with Figure 14.13 on page 399). Through-bulkhead filters present no problem, but chassis-mounting filters must keep the leads passing through the partition plate as short as possible and preferably should be treated with ferrite sleeves to minimize HF propagation across the partition.

The great advantage of this approach is that the dirty volume can be used for many or all of the field-installed connections. The interface through the partition can be pre-wired and checked before the system is shipped from the factory. Then, all of the strictures about ensuring correct installation practices are taken out of the hands of the

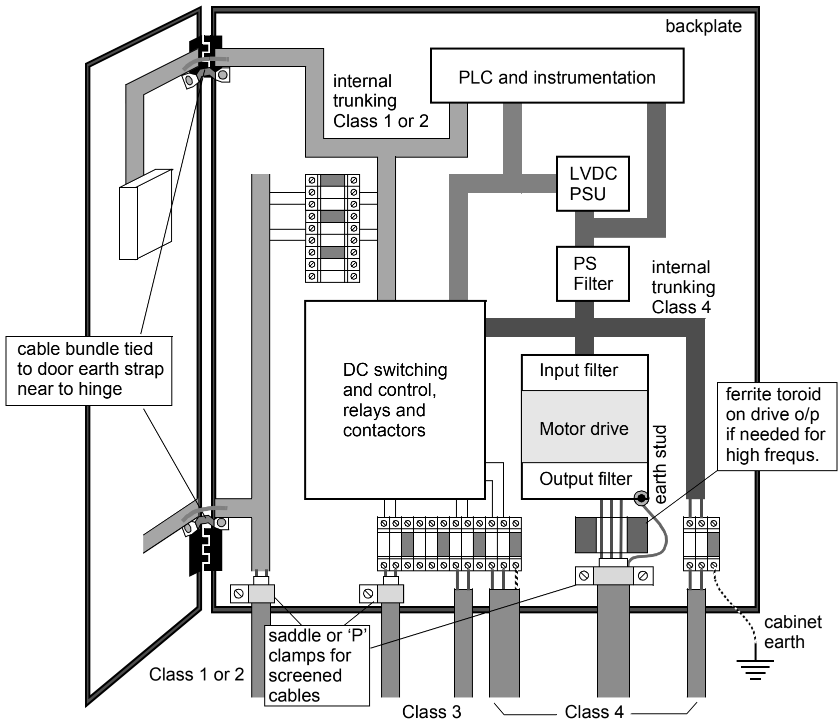


Figure 15.6 Layout and cable routing internally in cabinet

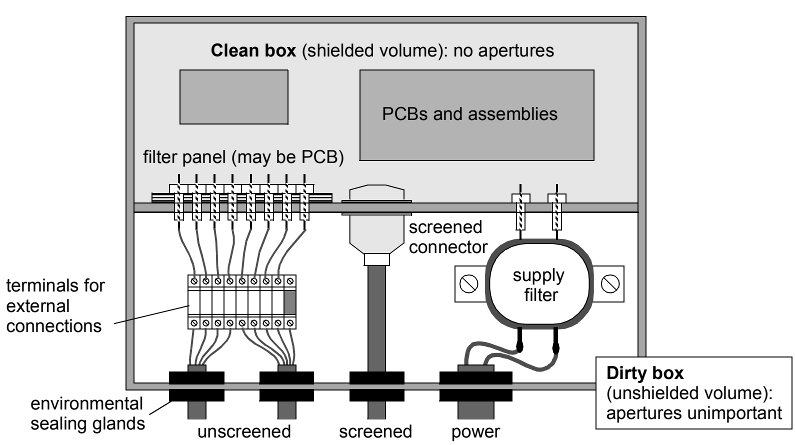


Figure 15.7 The "clean/dirty" segregated shielded cabinet

installation technician and given to the system assembler and designer. This is of particular benefit if the system supplier does not control the installation methods at all.

15.3.3 Conductive hardware

Whilst conductive gaskets are widely available for mating two surfaces, and have been covered in section 14.2.1, there is also a growing range of more specialized products for particular shielding applications.

15.3.3.1 Cable penetrations

All cables entering or leaving an enclosure should have their screens properly terminated to the enclosure wall. (*Un*screened cables should enter or leave via a suitable filter if the shielding effectiveness is to be maintained.) This means that full 360° contact should be maintained around the outer surface of the cable screen.

Mechanisms for ensuring this are similar to conventional cable glands for environmental sealing, except that the appropriate parts are fully conductive. Most of the traditional manufacturers of cable accessories are now aware of the importance of EMC aspects, and provide EMC-specific cable glands as part of their stock range. A typical construction using an iris-type spring compressed against the cable screen outer surface is shown in Figure 15.8. This is one of the most common methods of clamping to the screen, but others are possible, including collet or other clamping mechanisms, elastomeric compression modules, folding the screening braid back over a conductive tube or even passing the bare screen through a box of copper shavings.

Figure 15.8 Construction of a typical shielding cable gland assembly (KEC Ltd)

Aspects which you need to consider when specifying a screening cable gland system are

- mechanical compatibility: the cable screen outer diameter must match the gland's construction, often within quite tight tolerances;
- electrochemical compatibility: the materials used for screen, gland and enclosure panel should discourage corrosion;
- ease of assembly, especially if unskilled or poorly trained technicians are expected, or the working area is restricted;
- conductivity across the joints, which directly affects shielding effectiveness;

- least disturbance of the screen;
- whether or not an environmental seal is also required.

15.3.3.2 Ventilation panels

Fully shielded enclosures will often require ventilation. Pre-packaged units (Figure 15.9) known as "honeycomb panels", using waveguides below cut-off as discussed in section 14.2.3.2, can simply be fitted into the wall of an enclosure (observing the proper precautions regarding bonding all around the periphery of the assembly) to give any reasonable level of ventilation. These are much more effective at screening than a mesh of holes of the same open area in a thin panel, but of course are more costly and require some thickness in addition to the panel.

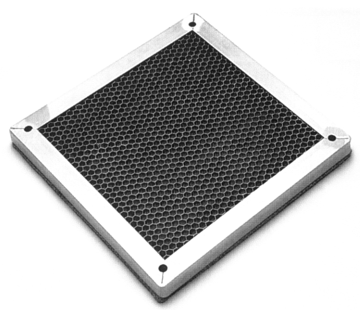


Figure 15.9 Honeycomb ventilation panel (TBA ECP Ltd)

15.3.3.3 Shielded windows

Viewing apertures can represent the largest size hole in the apparatus. If the display behind the window is a serious source or victim of disturbances, then the entire aperture needs to be shielded. Special conductively treated windows are available for this purpose; they need to be installed, as always, with great care to ensure that they are bonded to the surrounding panel all the way around the edge. The conductive treatment must be brought out to the edge of the window in such a way that good contact can be made to it with no breaks – a metal or metallized frame is often the best way of ensuring this.

Shielded window materials are discussed in section 14.2.3. Whatever the material, there is a trade-off to be made between electromagnetic protection and transparency or light transmission; generally, the more transparent a window, the less shielding effect it can give.

15.3.4 Installation and maintenance of screened enclosures

As apertures and seams in screened enclosures such as racks and cabinets can affect the screening performance drastically, it is very important to ensure that measures which are taken to control their effects at the system design stage are not degraded by installation and maintenance procedures (Figure 15.10).

Bonding integrity must be maintained continuously. Any surfaces which are intended to mate must not be allowed to corrode and must never be painted until after they have been assembled. Normally you will use a robust conductive finish, but if the environment is corrosive (such as in a naval installation) then more specialized measures, and more frequent maintenance, will be needed. Where fastenings provide a conductive path they must all be kept in place and at the correct torque. Replacement of short, wide bonding straps by loops of wire is unacceptable.

Doors, panels and hatches which make contact via gaskets or spring finger stock must be installed and treated with care so as not to damage or distort the contact surfaces, which should be regularly checked and cleaned if necessary. Filtered inlets and shield penetrations must make assured 360° contact to their host panels; a DC continuity check is rarely adequate to confirm that this is present.

It should be clear by now that *requiring* a cabinet or other enclosure to exhibit good shielding is not a simple or inexpensive option. The requirement affects all aspects of the installation throughout its intended life cycle. Also, maintenance, installation personnel and even users need to be trained in the principles and techniques involved, since they may otherwise unwittingly compromise shielding integrity just by following their own established practices, such as leaving a cabinet door open. For these reasons, shielding is best regarded as a means of last resort if other, lower-cost EMC options are unavailable or inadequate.

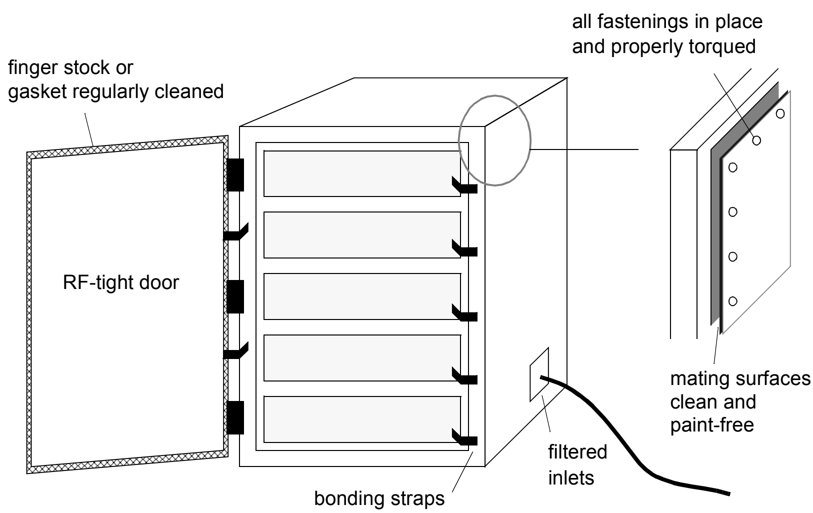


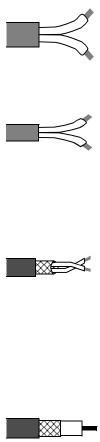
Figure 15.10 Screened enclosure maintenance

15.4 Cabling

Chapter 13 discusses cable issues in some detail and what it says will not be repeated here. Rather, we will consider the aspects of cable installation that are more specific to systems. These include cable classification, segregation and routing, and use of a parallel earth conductor.

15.4.1 Cable classification, segregation and routing

To minimize crosstalk effects within a cable, the signals carried by that cable should all be approximately equal (within, say, ±10dB) in current and voltage. This leads to the grouping of cable classifications shown in Figure 15.11. Cables carrying high frequency interfering currents should be kept away from other cables, even within shielded enclosures, as the interference can readily couple to others nearby. See Figure 10.4 on page 226 for the effect on mutual capacitance and inductance of the spacing between cables.



Class 4 Noisy: AC power and return, chassis ground, high-power RF and wideband signals; power inputs, outputs and DC links of adjustable speed motor drives, welding equipment, and similar electrically noisy equipment

Class 3 Slightly Noisy: DC power, suppressed switched loads, filtered AC; externally supplied low-voltage AC or DC power which does not also supply other noisy equipment, contactor and solenoid coil circuits

Class 2 Slightly Sensitive: low-power low frequency signals, low bit rate digital data; analogue instrumentation (e.g. 4–20 mA, 0–10V) and slow digital bus communications (e.g. RS232, RS422, RS485, Centronics); switched I/O such as limit switches, encoders, and the outputs of internal DC power supplies

Class 1 Sensitive: low-level analogue signals such as thermocouples, thermistors, RTDs, strain gauges, load cells, microphones; also wideband digital and analogue communications such as Ethernet, video, RF receiver inputs; and all other signals with full-scale range less than 1V or 1mA, or with a source impedance > 1kΩ, or signal frequency > 1MHz

Figure 15.11 Cable classification

15.4.1.1 Physically segregating cables by their classes

The purpose of determining which class a cable belongs to is to be able to choose the correct cable type and terminations, but it is also so that different classes of cables may be run segregated from each other to prevent them from interfering with each other. Figure 15.12 shows the minimum separation distances that should be maintained between the different cable classes. This assumes a continuous flat metal PEC (parallel earth conductor) under them all.

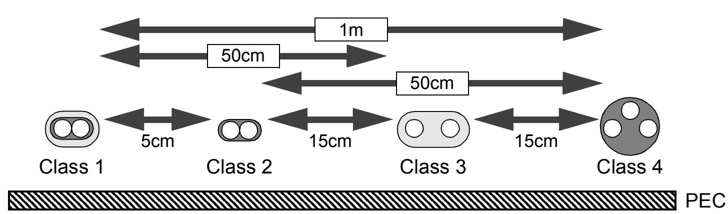


Figure 15.12 Minimum spacings between cable classes when run over a single PEC

Alternatively, IEC technical report IEC 61000-5-2 [156] on the installation of cables and earths in buildings simply recommends that cables should be segregated according to the type of signal they carry. Each loom should be 0.15m from the next if it is carried on a metallic conduit used as a PEC, and 10 times the diameter of the largest loom if it has no PEC.

15.4.1.2 Segregation within classes

This discussion of segregation has assumed that all the cables in a class may be bundled together, but this may not always be advisable, especially for the more extreme classes. Sensitive analogue Class 1 cables should not be bundled with high-rate digital signals in twisted pairs and neither of them should be bundled with high-rate digital in coaxial cables. These sub-classes should be bundled separately and not run next to each other (separation of at least 10mm between each pair), keeping each bundle as close as possible to the metal surface of the PEC at all times.

Different Class 4 cables may also require individual routings. The cables from adjustable speed inverter drives to their motors may be specified by the drive manufacturers to have 600mm or more spacing from any other parallel run of cable (drives are perhaps the noisiest devices on the planet!). It is difficult to make general rules for segregation within Class 4, because a cable may be very noisy in its own right, but still able to pick up sufficient interference from neighbouring Class 4 cables to affect the electronics to which it is connected.

15.4.1.3 Routing

All the cables between items of equipment should ideally follow a single route along a single PEC, whilst also maintaining their segregation by class. Figure 15.13 shows the two main principles of cable routing:

- the cables between two items of equipment must always follow the same route, and
- there should be a single interconnection panel for each item of equipment.

Where you need several routes and/or connection panels, each should have its own PEC, and higher PEC earth currents should be expected.

Stacking cable trays along a route

Because of the minimum segregation distances required between cable classes, it is generally impossible to run cables of all four classes along one cable tray (they are usually not wide enough). This is overcome by running a "stack" of cable trays. The

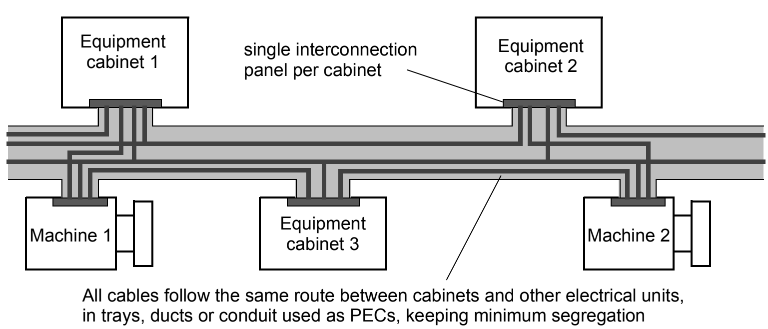


Figure 15.13 Installation cable routing

cable trays are stacked vertically and electrically bonded together at all of their support pillars. They all follow the same route between two items of equipment.

Connections to cabinets

There should only be a single connector panel for a cabinet. All external cables should enter a cabinet at only one side, rear, top, or bottom, and they should also enter the earthed backplate along one of its edges. This is so that, in conjunction with the other techniques described here, the high-level circulating currents flowing in the long cables in many industrial situations will flow from cable to cable via the connector panel or backplate edge via the screen-terminations or filters mounted in that area, and will not flow through the rest of the cabinet or backplate structure and hence affect the electronic units.

15.4.2 Parallel Earth Conductor (PEC) techniques

Modern best practices for EMC in installations (according to IEC 61000-5-2 [156] and EN 50174-2 [147]) require the use of cable trays, conduits, and even heavy-gauge earth conductors as Parallel Earth Conductors (PECs) to divert power currents away from cables and their screens. From the equipment designer's point of view, the cabinet and backplate should provide the means for the connection of the necessary PECs.

15.4.2.1 Constructing PECs

The first function of a PEC is to divert heavy earth loop currents from both screened and unscreened cables. Since earth currents are usually at 50/60Hz, and the surges from lightning events have most of their energy below 10kHz, it is enough for this purpose that the PEC has a very low resistance and a sufficient current-carrying capacity. Most cable support systems have enough metallic cross-sectional area to provide this low resistance and current capacity. Cables must be run very close to the metal of their PEC throughout the length of their run.

Any screen or earth conductor external to a cable should be treated as a PEC and bonded to earth at both ends. Cable armouring can be used as a PEC, but there must be no breaks in the electrical continuity of any armour used for this purpose. Cable installers traditionally regard armour merely as mechanical strengthening or protection, and may not be used to bonding it at joints and to the local earth at both ends.

PECs can also control higher frequencies. Figure 15.14 shows a variety of types of PECs, and ranks them by high frequency performance. (Compare this to Figure 11.5 on page 263.)

A cable tray is usually perforated with slots to make cable fixing easier, but these can detract from its high frequency performance. The problem is exactly the same as has been described in section 14.1.3 on shielding effectiveness: slots and gaps interrupt current flow and therefore increase the transfer impedance of the structure. Because of their open construction, ladder- and basket-type cable support systems are poor as PECs.

In extreme environments PECs may have to carry high continuous currents, and should do this without overheating or other damage, so they must have an adequate metallic cross-section. Conductively coated plastic conduit or trunking will obviously not be adequate, for this reason, and if used will require a heavy-gauge copper wire PEC inside to handle any heavy currents. On the other hand, in a building installed with a designed-in equipotential earth mesh any of the interconnected metalwork may be used as a PEC (I-beam girders, building steel, etc.).

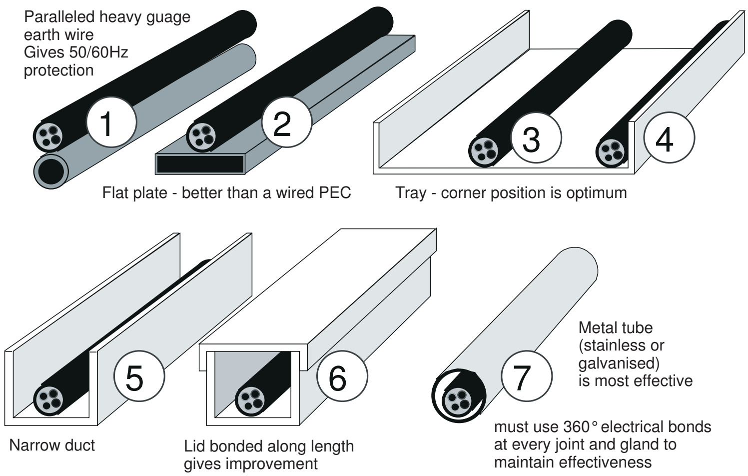


Figure 15.14 Some types of parallel earth conductors (in order of increasing HF effectiveness)

15.4.2.2 Bonding PECs

PECs must be electrically bonded to the local equipment earth at each end, and to all their support structures and any other earthed metalwork at every available opportunity. This helps to create a meshed earth structure, and it also helps the PEC to function effectively.

Joints and end-terminations in PECs must be bonded using appropriate methods. Cable trays and rectangular conduits will need to make electrical bonds directly to the cabinet wall (or floor, top, or rear) using U-brackets or similar with multiple fixings. Round conduit can bond to the cabinet wall with circular glands (see 15.3.3.1), remembering to remove the paint first (to ensure a 360° bond) and apply corrosion protection. For plain wire PECs, a cabinet will need appropriately-sized and positioned earth terminals.

15.5 Lightning protection

This short section does little more than review the basics of lightning protection for apparatus in buildings, which is a vast subject all to itself. Personnel safety issues are not covered.

15.5.1 How lightning phenomena can affect electronic apparatus

The issues raised by lightning protection for electronic apparatus are addressed by IEC 62305-4 [160] and by Appendix C of BS 6651 [172]. Lightning can cause damage to electronic equipment in a number of ways:

- Resistively induced voltage: the resistance of the soil and of earthing networks, when subjected to intense lightning discharge currents (considered to be between 2kA and 200kA, with 1% of strokes exceeding

200kA) creates potential differences between areas normally considered to be at the same "earth" potential, and this exposes electronics connected to these different areas to excessive surges. Long cables, and especially cables between buildings or structures, are particularly likely to cause damage due to this effect, which is sometimes known as "ground potential rise" or "ground lift".

- Magnetically induced voltage: excessive voltages may be induced into conductors and bonded earth structures due to the radiated magnetic fields from lightning discharge, for strikes at up to 100m distance, due to the rate of change of the discharge currents. A maximum rate of 200kA/µs is accepted for the arc channel itself, with lower values where the lightning discharge current is shared between a number of conductors. Even the pigtails traditionally used for bonding the screens of cables can present a serious risk to their equipment due to inductive voltage coupling (consider $V = -L \cdot di/dt$: a pigtail inductance of 20nH with a di/dt of 200.10^9A/s would give 4kV).
- Current injection from direct strike: there can be direct injection of the lightning main discharge current into any exposed external equipment and cables. The arcing flashovers associated with a direct strike to external equipment often results in damage to connected internal equipment, but may also cause damage to unrelated equipment by flashovers in shared cable routes or terminal cabinets. Here is another reason for cable segregation and good earth bonding at cabinet entries.
- Electric field coupling: the whole area around a lightning strike that is about to happen can be exposed to electric fields of up to 500kV/m (the breakdown voltage of air) over an area of up to 100m from the eventual strike point, with fluctuating fields of 500kV/m.µs occurring during a strike. These fields will induce voltages and currents into conductors and devices, but except for high-impedance circuits do not pose as much of a threat as the high-current effects.
- Lightning Electromagnetic Pulse (LEMP): this is a far field phenomenon, and may be caused by cloud-to-cloud lightning as well as by distant cloud-to-ground lightning. It is usually only a problem for exposed external conductors, and is effectively dealt with by the measures taken to protect equipment from other lightning threats.
- Thermal and mechanical effects from the intense energies associated with a lightning strike: these are more usually problems for the structure's fabric and the design of the lightning protection system itself.
- Multiplicity of the surges in a single "strike": a typical lightning event consists of many discharges (or "strokes"), of which the second one usually contains the most damaging energies. Multi-stroke flashes can exceed ten strokes and last for over a second, which is of great importance in the design of software for error-correction and for the recovery of systems.

15.5.1.1 Assessing the criticality of the apparatus

Lightning damage to electronic equipment can cause safety problems to personnel or damage to the structure, usually through electrocution or fires, but sometimes because the equipment has a safety-related function.

Safety concerns such as fire and electrocution must be addressed as part of the normal health and safety at work procedures. For EMC we are concerned with the response of each item of electrical and electronic apparatus to the effects of lightning. Each item of apparatus should be assessed against the following criteria:

(A) catastrophic failure requiring replacement of the apparatus is acceptable

(B) the apparatus is required merely to survive the lightning event undamaged, with no concern about its functionality during the event

(C) the apparatus must continue to operate during a lightning event, although reduced performance is acceptable (the degree of degradation needs to be specified for each function)

(D) the apparatus must continue to operate without any reduction in performance during a lightning event: safety- or mission-critical equipment.

The same equipment may have different criteria depending on where it is used in a structure, how it is installed (its exposure), and what it is used for (how critical is its function). Co-ordination is then required between three aspects:

a) the apparatus' functional criticality
b) the apparatus' ability to withstand lightning electromagnetic phenomena, which can be derived from surge immunity testing, as discussed earlier in section 7.2.3
c) the lightning electromagnetic phenomena that the installation exposes the apparatus to (especially voltage or current surges).

Apparatus must therefore be designed and tested to achieve the required degree of protection and reliable functional performance depending upon its exposure to various lightning phenomena when installed as specified. This may require the use of surge protection devices (SPDs) at exposed ports, particularly the power supply and any connections to external cables, to deal with both common-mode and differential-mode voltage surges. Meanwhile, the building in which it is installed should benefit from a properly designed lightning protection system.

15.5.2 Overview of design of a lightning protection system (LPS)

15.5.2.1 Basic design of an LPS

The design of a basic LPS for safety and protection of the structure typically requires:

- risk assessment based on lightning exposure and acceptability of consequential losses;
- design of the air termination network and down-conductors;
- design of the earth termination network and earth electrodes;
- either bonding of the metalwork within a structure (the "internal" LPS), and the metallic services entering a structure, to the external parts of the LPS, or separation from them.

The possible utilization of metal parts of the structure – so-called "natural" components – as parts of the LPS should be foreseen in the design of the structure itself, but only used with the agreement of the owner and the structural engineer. All metal parts so used (metal sheets, metal parts of roof construction, gutters, ornamentation, railings, pipes, tanks, etc.) must meet specified minimum requirements. Copper theft can be a serious concern and puts an external LPS at risk, and it is often difficult to persuade owners and their architects that an external LPS enhances the appearance of their building. For these reasons the use of natural components is preferred, although successful application requires consideration right from the start of a building design.

15.5.2.2 Documenting and maintaining an LPS

Both [172] and [160] specify that records are required to be kept throughout the design and construction process. Certain procedures are also specified for the regular inspection, maintenance, and upkeep of the LPS, and records must be kept of these too. These records are generally required to meet the requirements of safety laws and insurers, but are also recommended for aspects of the LPS that concern the protection of electronic equipment.

15.5.2.3 Construction of an LPS

A basic LPS consists of an air termination network, a down-conductor network, and an earth termination network (Figure 15.15). It is possible to construct an isolated LPS that protects a structure whilst being electrically separated from it, but the type of LPS described here is attached to the structure and bonded to its internal CBN.

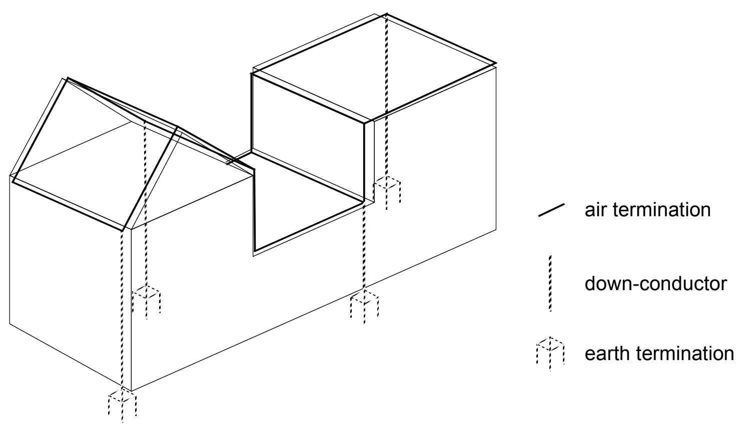


Figure 15.15 Components of building protection

Air termination network

The air termination network is intended to intercept the lightning strike and divert its currents via the down-conductors to the earth termination network, thereby protecting the structure from the strike. This can be a mesh arrangement of copper or aluminium conductors laid horizontally and vertically on the tops of roofs and the outsides of walls, with minimum spacings between conductors as specified by standards or codes of practice. “Natural” components such as gutters, railings, or metal-clad roofs may be usefully pressed into service and can even take the place of a separately installed air termination network.

Down-conductors

Down-conductors provide a low-impedance path for lightning currents from the air termination network to the earth electrode system, and in general there should be several, equally spaced around the structure to share lightning current amongst themselves. Metal structures such as radio masts or flagpoles may use their exposed metal structure as all or part of their air termination and down-conductor network simultaneously.

Down-conductors should be straight and vertical, fitted at least at the corners of a structure, equally spaced, and should provide the most direct route to the earth electrodes.

Earth termination network

The earth termination network is the system of earth electrodes which dissipates the lightning currents into the mass of the soil or rock beneath the structure to be protected. All soils and rocks have finite conductivity, which compromises their performance as an earth mass, so care must be paid to the design, construction and maintenance of earth electrodes. The earth termination network for a structure is generally required to provide an earth resistance of under 10Ω, although higher or lower resistances may be allowed or needed in special cases. The lightning standards and codes provide rules and formulae for designing different types of earth electrodes.

A typical earth electrode consists of a copper alloy rod electrode deep-driven vertically into the soil, sited at the foot of each individual down-conductor a metre or so from the boundary of the structure. The reinforcement in concrete foundations (a little while after construction) can achieve a very low earth resistance, especially concrete pilings. This is known as a foundation earth electrode, and it requires the reinforcing bars to be welded, or at least reliably bound together with tying wire, at their crossing points. Strip electrodes may also be used, especially to help reduce voltage gradients around a structure, when they are known as potential grading electrodes.

Bonding

For buildings of just a few storeys, say up to 15m high, with a properly designed LPS, it is usually enough to bond the LPS to the structure's internal common bonding network at ground level only. Structures higher than 20m should bond their non-LPS metalwork to their LPS at top and bottom, and at intervals of no more than 20m in between [172]. [160] recommends that bonds between LPS and CBN take place where there is already a horizontal ring conductor which bonds the LPS down-conductors.

Chapter 16

EMC management

16.1 Managing the EMC process

Implementing a strategy of EMC compliance within a company is much more than simply ensuring that the product designers make their designs meet the standards – although this is a first step. EMC awareness, like quality and safety awareness needs to extend vertically and horizontally throughout the company if it is to be effective. This means that each department must know what its role is in maintaining a product's EMC and why. This applies with particular force to the purchasing, production and installation groups.

It also means that business and management decisions affecting EMC should be made in a similar climate of awareness. Investment decisions on EMC test facilities are a clear case in point, as also are marketing decisions on the specification and life cycle for a particular product range, and the question of whether to launch a marginally compliant or non-compliant product. Appointing an "EMC specialist" to take responsibility for these issues is better than no co-ordinated policy at all, but should only be the first step towards an integrated understanding throughout the company.

16.1.1 Putting EMC in context

Most products are already subject to a range of environmental or safety specifications. These requirements are partly fixed by legislation and partly to ensure "fitness for use" of the product, and they include factors such as temperature and relative humidity ranges over which the equipment must operate, together with ergonomic aspects and control of hazards (such as heat or electric shock) associated with the unit. EMC is best regarded as simply another of these specifications.

Immunity to external EM disturbances is clearly an environmental issue. A product must function reliably in its intended environment, and if this environment includes EM disturbances then these should be incorporated in the environmental specification. One of the functions of EMC immunity standards is to provide some guidance in defining this environmental requirement. Control of emissions is more properly a regulatory function unless such emissions have an intra-system aspect – that is, if they affect the proper operation of the overall system or its installation.

16.1.1.1 Selling EMC within the company

Management throughout the company must be convinced of the worth of EMC before it will give its commitment to effort and resources. First and foremost, the impact of EMC legislation must be made absolutely clear. Beyond this, you can monitor pressure from customers, competitors and third party approvals bodies and use it to justify the need for effort. The link between good EMC and product quality should be continually stressed.

There are two particular difficulties in the way of a wide appreciation of EMC [37] – one is the sheer unfamiliarity of the topic in a world where most designers have avoided analogue subjects in favour of digital, the other is a perception of EMC as a cost rather than a benefit. These can be overcome by documenting and presenting details of EMC-specific actions that may already have been taken on a particular product that is familiar to all. Shared experience is a powerful educational force. Documenting the knowledge gained on a particular project will also avoid the risk of creating EMC "gurus", which in the long term benefits neither the company nor its engineers. Documenting solutions – rather than failures – results in a positive "get-it-right-this-time" culture in the design group.

16.1.1.2 Costs and timescales

Addressing EMC early in the product development cycle is easier and more cost-effective than any wait-and-see approach. The early stages of a design allow designers the most freedom to achieve EMC. The packaging and circuit layout, for example, are subject to the fewest constraints. As the product design evolves towards production, changes become increasingly difficult and expensive. For example, a radiated emission problem that might have been solved earlier by judicious PCB layout may require changes to the packaging and interfaces. The key to cost-effective EMC is to instil among circuit and packaging designers an appreciation for good EMC design practices.

A critical factor in product development is not just the overall development cost but also the timescale. In many product sectors the marketing window of opportunity is short, so lost time in bringing the product out translates to lost sales and poor profits. If ignored for long, EMC requirements will undoubtedly impact time to market. Every EMC test house has a long, sad list of companies which have optimistically brought their product along the day before launch only to find they were non-compliant.

The two ways to prevent this happening are to ensure that EMC is treated as part of the overall specification, so that its dictates are adhered to right from the start of the project; and to continually monitor the EMC profile throughout the development stage so that late surprises are eliminated. It is also wise to assume that one re-test *will* be needed – even when all the design principles are adhered to, initial compliance is not certain – and to factor this in to the development schedule (see section 16.3.1.3).

16.1.2 The EMC co-ordinator

A typical first step for a company implementing an EMC control policy is to appoint an EMC co-ordinator. By default this is often a design engineer who has more knowledge than anyone else of RF topics, although it may be viewed as a job for a quality engineer. The magnitude of the task facing this individual must be appreciated, and the powers that are delegated to him or her have to be sufficient to deal with all the issues that will arise. In particular, the distinction between engineering decisions† and business or legal decisions‡ must be strictly maintained; or else the co-ordinator's function must clearly include legal and business authority, and his or her training must reflect this responsibility [126].

Shepherding an individual project through EMC compliance can absorb (if it is done thoroughly) between 10 and 20% of the development cost. Thus for every 5–10 design engineers employed, there should be one EMC engineer. For companies with large development departments and a wide range of products, a separate EMC

† "It's 2dB under the limit; we need a 6dB margin to cover uncertainties; we can't ship it."

‡ "It's 2dB under the limit; we need the product on the market; of course we can ship it."

department (perhaps allied with the quality department) is justifiable. Conversely, for small companies the job may best be filled by a retained consultant. Some of the job functions that may be expected are to:

- oversee and advise the design process;
- interpret EMC standards requirements for the company's product range;
- devise test and control plans and schedules;
- control/sign-off EMC-related engineering changes;
- liaise with all departments on EMC matters;
- keep abreast of developments in the EMC field.

16.2 The design process

16.2.1 The product specification

A project begins, at least theoretically, with a specification. This will naturally concentrate on the functional requirements of the product or system. As soon as there is reasonably stability in these requirements, particularly once the interfaces have been settled, you should be looking at the EMC aspects.

The specification should at least be clear on what environment the product is intended for. Harmonised standards are linked to environments and the EMC Directive is explicit that compliance is expected when the apparatus is "used for its intended purpose", including a restriction on use in residential areas if necessary. Other environmental requirements, such as temperature range, vibration, humidity and hazardous area certification, also need to be set at the start and EMC should be simply another one of these.

If the project is negotiated to a contract then this may also specify EMC requirements, usually by reference to standards which are industry- or product-specific. This would be typical in the defence industry and in some other sectors, such as automotive and railway industries. It's not unusual for such requirements to be extremely vague in the early stages, before the product's outlines are clearly established, and because procurement engineers are often unfamiliar with the detail of EMC standards. This means that there will be room for detailed negotiation when specifics become available, but the existence of a clear management approach which takes EMC on board will make these negotiations easier.

Whether or not a contract is involved, the highest level of specification will generally concentrate on reference to standards, and your main aim is to understand the design implications of the tests that these standards set.

16.2.2 Design rules

As a matter of course, every design group should have some general rules to work against that address EMC performance. Appendix A and section 11.2.4 in this book give some suggestions, but it is impossible to give a single set of inviolate rules that will apply to all kinds of product. The variety of forms of construction (plastic case, metal case, rack, cabinet), PCB technology (single-sided through to 20 or more layers), circuit complexity (simple analogue through to multi-clocked dense processing) and of course EMC performance requirements, ensures this. Rules which may be essential for a high-performance military requirement can be inappropriate for a simple consumer device.

Rather, it is better to have a list of issues that should be addressed with an understanding of the principles involved. So, as a minimum, you should consider:

- circuit design: analogue bandwidth limitation, power segmentation and decoupling, choice of clock frequencies, clock distribution and buffering;
- PCB layout: 0V plane(s), power plane(s), layer stackup, decoupling placement, component placement, constant impedance layers, heatsink and mechanical aspects, general routing;
- interfaces: power supply filtering, low frequency unscreened cable port filtering, high frequency unscreened cable port balance and common mode chokes, connection of screen to chassis for screened cable ports;
- enclosure design: aperture size and location, bonding of structural parts, use of conductive gaskets, mating surface treatment and paint masking, PCB-to-enclosure connections, localized screening, moulding design for conductive coatings and choice of coating.

The implementation of these aspects for any given design is the responsibility of each designer, but it can be validated through the process of design review.

16.2.3 Design reviews

Reviews at every stage of a product's design can help to ensure proper EMC control. In these, the product's own designers are subjected to peer review by other designers not involved in that project, and consensus decisions are reached on the validity of each design aspect. The design reviews should be key milestones in the development process, and progress to the next development stage should be conditional on the successful outcome of each review. This self-imposed discipline offers a formal method for checking that EMC aspects (and others, such as safety) are being addressed throughout the process. There is a further hidden advantage – the minutes of design review meetings are convincing evidence, if challenged on its compliance statement, that a company has at least attempted to address EMC issues.

The practical implication of this is that the EMC co-ordinator should be present at each design review and should be included in the sign-off for the review. It is this person's responsibility to question each design decision that may have an implication for the product's EMC profile. The design team cannot (and indeed should not) follow EMC design rules absolutely rigorously because in many cases this would load the unit cost excessively. But there will be many occasions where design choices have an EMC dimension; and where problems have occurred, the designers must not only understand the cause but must also be able to follow the steps needed to eliminate the problem from a future design.

If you want to follow a specific format for the EMC part of design reviews, start with the lists given in Appendix A and section 11.2.4.

16.2.4 Identifying EMC-critical aspects

A vital part of the process, which should be emphasized in the design reviews, is the identification of those aspects of the design which are likely to create difficulties in the compliance tests. This will never be 100% successful, or it wouldn't be necessary to do the tests at all. Rather, the idea is to anticipate and deal with potential problems before they become show-stoppers. This process relies to a great extent on experience, on how previous products of a similar type fared in their tests; and this means that good records of test results need to be kept in order to nurture this experience. Just achieving a pass is not enough: you should know by what margin the product passes, either below the emissions limit at relevant frequencies, or above the minimum stress level for immunity. This will then guide you for the next design.

Aspects of the new design for which there is no experience automatically become critical. Instances might be a different interface (say, USB or Ethernet), or a new processor with a higher frequency clock, or a greater sensitivity A-D converter. Over and above the normal good practice in board layout and interface protection, such extensions need to be treated with extra care.

Any of the following should trigger more investigation in the early stages:

- wideband, high-speed signals entering or leaving via interfaces;
- clock and data bus signals distributed around a large board, or around a whole system via a backplane or inter-board cables;
- high-power, high frequency switching circuits;
- microvolt or millivolt level signal amplification.

16.3 Test management

Chapter 9 discusses the planning of EMC tests in detail. Managing these tests is itself a significant activity.

16.3.1 When to do the tests

One issue which faces project managers is, at what point in the product's development should EMC testing be carried out. The worst time to find out that a product has EMC problems is when the design has been finished, the tooling and PCB layout has all been signed off and the launch date is looming. But on the other hand, if you test a product too early, the test results won't reflect the performance of the final production build. So what timing is best?

16.3.1.1 Initial look-see tests

In fact, it is wise to plan for at least two sets of tests: an initial look-see, and a final compliance test. This gets around the dilemma, but at the apparent cost of doubling the testing budget. Actually, it needn't be as bad as that. The initial look-see doesn't have to follow the full compliance methods of testing as described in the standards. As long as you are familiar enough with the principles and limitations of the tests, what are known as "pre-compliance" methods can be adequate to flush out any major problems that will clearly cause failures in the final compliance stage. This means that they can be cheaper, allowing either a lower budget or more comprehensive coverage: different design options can be evaluated without the high per-hour cost of a fully compliant test lab.

If this two-stage approach is followed, you can afford to be flexible about the timing of the initial tests. In essence, you will want to test as early as possible while still using a representative build. One school of thought says that there is no point in testing anything until the final build is reached, because even minor changes in the run-up to production can have dramatic EMC effects. This is true, but it overlooks the value of knowing whether you are even in the right ballpark, and not being hit by nasty surprises when you can least afford them. The purpose of the early tests is to identify if there are any gross issues which have to be immediately addressed in the design, accepting that later design changes may affect these one way or the other. For this purpose it may even be appropriate to test just a part of the design about which you are unsure; for instance,

- a new high-speed interface, to choose between filtering and cable shielding options, and to identify particularly emissive or susceptible data rates and formats;
- a video circuit, to check the impact of different levels and frequencies of applied RF disturbance on picture quality;
- shielding effectiveness of different enclosure designs or conductive gaskets, before committing to tooling.

And, of course, not all tests may need to be done at this stage; usually the most relevant are the RF emissions and immunity tests, and possibly the fast transient and surge immunity. ESD immunity is highly dependent on layout, grounding and enclosure design and is only really valid once these aspects are settled and a representative model is available. The other tests are also affected by these factors but to a lesser extent, and it is worth doing them earlier, especially if you can identify and control likely critical coupling paths. For example, if the interface grounding regime can be made representative, it is often possible to do the conducted emissions and immunity tests without a proper enclosure, at least to indicate what aspects to address in the next build.

A further use for initial testing is to develop knowledge of the effects of operating mode and build state on the EMC profile. This will then allow you to justify the choice of particular configurations and modes as worst case in the final compliance test without so much time-consuming investigation at this late stage.

16.3.1.2 Compliance tests

The second question is, at what point should the compliance tests be done?

This question is at least as much determined by company policy as by engineering considerations. This is because, for the purpose of self certification to Directives, it is “the manufacturer’s responsibility” to declare compliance and therefore it must involve commercial and legal aspects as well as engineering ones. The same consideration applies to contractually-set obligations.

The problem with these tests is that they are normally going to be performed when the pressure to release the product is most intense: just before (or, worse, after) the published launch date. The criteria which you should apply to check whether the product is ready to test are:

- have all engineering changes except trivial ones (such as case colour, or the external moulding detail) been frozen: will the product being tested be essentially the same as that which goes to market?
- are all relevant software functions available, debugged and complete?
- is all necessary support equipment, including the correct cables, available and working properly?
- are enough samples of the product meeting the above criteria available to carry out the tests adequately?

It’s rarely possible to be able to tick all these boxes fully and still meet launch schedules. Any deviation from the best case should be evaluated for the risk it poses to the validity of the test outcome, particularly the questions of engineering and software changes. After the compliance test has been completed successfully, further changes should only be allowed if they can be shown by analysis to have no expected effect on the outcome if they were to be re-tested. Very rarely is the EMC engineer able to certify this completely: any such change will carry some risk with it. The project manager has to judge the moment of the compliance test to minimize this risk and still meet the schedule.

16.3.1.3 Planning for different outcomes

It is only natural that the project manager should be optimistic about the outcome of a set of EMC tests. But it is more realistic to expect that some failures will occur, and to identify the risk of this happening and plan to mitigate it. In the early days of EMC compliance, test houses were reporting that most products failed their tests at the first try. This depressing situation has eased somewhat with a greater understanding of the subject, but there would be no point in the tests if all products were guaranteed to pass.

There are various ways to anticipate difficulties at the compliance test stage. One is to allow sufficient schedule flexibility to cope with a board or enclosure re-spin between the initial compliance test and final sign-off, with an option on a further test date if this is needed. Another is to provide for several variations in the design itself, which can be selected on the fly until one is found that achieves the best pass, or which are implemented in different builds of the product that can be tested at the same time. If you have a good set of pre-compliance results, you will be in a better position to evaluate the risks of each design option. As with the timing of the tests, the project manager has to juggle the competing merits of design implementations of differing costs with the need to minimize the risk of long delays to the project caused by test failures.

16.3.2 In-house or external?

16.3.2.1 Options for testing

Small to medium-sized enterprises will not be able to afford their own full-scale test facilities and their choices are limited:

- join and help finance a consortium of similar companies which operates a test facility jointly for the benefit of its members;
- use an independent test house for all their EMC test requirements;
- establish a rudimentary EMC test capability in-house for confidence checking, and use an independent test house for compliance testing only.

The first option has not been established on a widespread basis in the UK, although there are precedents in the form of co-operative “research clubs” in other fields. The second option will be expensive and has the disadvantage that experience gained in testing your own products is not brought in-house to apply to future products. The expense could be diluted by using cheaper, non-accredited test houses for confidence checking and saving the accredited test houses for full compliance testing. It is though more preferable to develop a close relationship with one test house with which you feel comfortable than to change test houses at will. And unfortunately the nature of EMC testing is that there are large measurement uncertainties to contend with, and there is no guarantee that a test at one facility will produce the same results as an apparently identical test at another. (This has given rise to the rather cynical strategy of hawking a marginal product around several test houses until a “pass” is achieved, on the basis that this is cheaper than dealing with the product design!)

16.3.2.2 In-house testing

The problem of measurement uncertainty also applies to the third option, with possibly greater force because the confidence checks are done in a largely uncontrolled environment. Even for confidence checks, the equipment budget needed to carry them out is by no means negligible. It can be reduced by hiring expensive equipment at the appropriate time if the work load is light. A further but less obvious disadvantage is that not only must you invest in test equipment and facilities, but also in training staff to use them and in keeping up to date with the highly fluid world of EMC regulations and test methods. An external test house will have (reasonably) up-to-date equipment, facilities and expertise.

The advantage of the in-house approach is that you can carry out testing at any stage of the product design and production cycle, and the process of EMC confidence testing helps to instil in the design team an awareness not only of the test techniques, but also of the effectiveness of the various design measures that are taken to improve EMC. The benefit of this will be gained in future designs. Also, designers will be under much less stress if they have the ability to test and re-test modifications made at the bench without too much concern for the money that is being spent in the process.

If the product will be self certified to harmonised standards then there is no need to use an external test house at all, provided that you are confident in the capability and accuracy of your own tests. Nevertheless many firms, and especially their empowered signatory who signs the declaration of conformity, are happier having independent confirmation of compliance from an organization whose competence in the field is recognized – and this is sometimes a commercial requirement anyway. It would be perfectly in order to choose some tests, perhaps those involving RF emissions or immunity, to be performed outside while others such as transient, ESD and mains disturbance immunity are done in-house.

16.3.2.3 Accreditation

The requirement for confirmation of integrity of a test facility is met by accreditation, which is based on the ISO 17025 standard. This covers organization and management, calibration and maintenance of test equipment, measurement traceability and procedures, records and reports, the quality system, and staff competence. In the UK the body which handles accreditation is UKAS, the UK Accreditation Service. Mutual recognition of test house accreditation throughout Europe has yet to be achieved, and this is a major aim of the European Organization for Testing and Certification (EOTC). The European groups responsible for accreditation of test facilities (members of EA, the European co-operation on Accreditation) are given in Table 16.1.

16.3.3 The various types of report

The report can range from a test house certificate which merely states whether the EUT did or did not meet its specification, to a detailed test report which includes all results and test procedures. Most UK test houses would be able to deliver a report to UKAS standards, which contains essential results and information without detailing lab procedures. For the purpose of complying with the EMC Directive, you need to know whether you want a certificate of compliance with an EN standard or a report for insertion into a technical file. For self declaration to harmonised standards, the form of the report is set by the level of detail you need for your own purposes.

A simple certificate stating “pass” (preferably not “fail”!) is cheap and quick; more detailed reports will be more costly and can take several weeks to be produced by the

Table 16.1 European organizations responsible for test accreditation

Austria	BMWA
Belgium	BELAC
Denmark	DANAK
Finland	FINAS
France	COFRAC
Germany	DAR
Greece	ESYD
Iceland	ISAC
Ireland	INAB
Italy	SINAL
The Netherlands	RvA
Norway	NA (Justervesenet)
Portugal	IPAC
Spain	ENAC
Sweden	SWEDAC
Switzerland	SAS
UK	UKAS

lab. But there are advantages to having more detail, if the report is going to be scrutinized in future, either because of a legal challenge to your declaration of conformity or more likely because a customer wants to satisfy themselves of the compliance of the product. If the report clearly states what individual tests were done, to what ports, under what conditions of operation, with what monitoring equipment and with which performance criteria, then it can be reviewed by a third party for completeness and relevance and can if necessary allow the tests to be repeated at another lab. If it lacks this detail, then its worth is questionable, and in the worst case a customer or an enforcement authority may not accept its validity.

Another debate is whether the report should contain all the test results, quantified: for instance, the actual levels of emissions. EN 55022 requires the six highest to be reported and this is often taken as a guide for any emissions report. It allows an external reader to judge how much margin the product has and at what frequencies, and this is valuable if the product will be used as part of a system which must itself be certified. An emissions plot is also sometimes provided. By itself this is of little use for knowing the actual emissions levels, but can be helpful in showing the general profile of the product. In fact, providing these details is more beneficial to the product's designer than to most other users of the report.

Chapter 9 discusses test planning in depth, and a good test plan can be used in conjunction with the test report to expand on and reinforce its content.

16.4 Compliance during production and beyond

EMC control does not stop when the product has moved from design into production. Both the EMC and R&TTE Directives require the manufacturer to take "all measures necessary" to maintain the compliance of each individual item (see section 2.2.4), but do not specify what these measures might be.

16.4.1 Degrees of quality assurance

The R&TTE Directive differs from the EMC in that it offers a number of different options for compliance of different classes of radio product, from straight self certification including technical documentation, through additional tests agreed by a Notified Body, to full quality assurance with assessment and surveillance by a Notified Body. This last is equivalent to Module H as defined in the framework Directive 93/465/EEC on conformity assessment procedures. The requirements of the quality system that is mandated by this Module are given in the R&TTE Directive [187] and are summarized as follows:

> The quality system must ensure compliance of the products with the requirements of the Directive that apply to them. All the elements, requirements and provisions adopted by the manufacturer must be documented in a systematic and orderly manner in the form of written policies, procedures and instructions. This quality system documentation must ensure a common understanding of the quality policies and procedures such as quality programmes, plans, manuals and records.
>
> It must contain in particular an adequate description of:
>
> – the quality objectives and the organizational structure, responsibilities and powers of the management with regard to design and product quality,
>
> – the technical specifications, including the harmonised standards and technical regulations as well as relevant test specifications that will be applied and, where the standards will not be applied in full, the means that will be used to ensure that the essential requirements of the Directive that apply to the products will be met,
>
> – the design control and design verification techniques, processes and systematic actions that will be used when designing the products pertaining to the product category covered,
>
> – the corresponding manufacturing, quality control and quality assurance techniques, processes and systematic actions that will be used,
>
> – the examinations and tests that will be carried out before, during and after manufacture, and the frequency with which they will be carried out, as well as the results of the tests carried out before manufacture where appropriate,
>
> – the means by which it is ensured that the test and examination facilities respect the appropriate requirements for the performance of the necessary test,
>
> – the quality records, such as inspection reports and test data, calibration data, qualification reports of the personnel concerned, etc.,
>
> – the means to monitor the achievement of the required design and product quality and the effective operation of the quality system.

This is as good a description as any of the requirements of a quality system that would be necessary fully to control the EMC compliance process. Naturally it aligns closely with global standards such as the ubiquitous ISO-9000, but extends it to deal with issues specific to EMC.

16.4.2 Production QA testing

There are two reasons why the EMC performance of a product in series production might differ from that of the product when it was tested. One is that the formal build state may be different, as a result of engineering design changes. The other is that production tolerances, both in electronic components and mechanical construction, will vary the emission level from noise sources or the susceptibility of victim circuits.

Strictly speaking, to be absolutely sure that these variations have not taken the product outside compliance you would need to EMC test each unit as it comes out of production. In most cases this is not realistic. A more tenable option is to operate a sample testing system. The tests to be applied to each sample could be a full suite of standard compliance tests, or could be simpler comparative tests of key parameters – such as certain emission or susceptible frequencies measured on critical cable ports, to avoid radiated tests – compared against a transfer standard or "golden product" [140] which is known to comply with the full test suite.

16.4.2.1 Sample testing regime

The question then arises, how often should samples be tested? One suggested solution is to base the sampling only on known changes, assuming that a margin has been built into the design for production tolerances, and to institute a points system for engineering changes. Each change is awarded a given number of EMC points, depending on its likely seriousness in affecting the EMC profile. When, as a result of changes, a certain number of points have accumulated, the product is submitted to a re-test. Of course, the key to this system is knowing how to allocate points to changes and deciding what the re-test trigger level should be. This is likely to be based largely on guesswork at first, although experience and knowledge of the critical areas gained during development will help refine the process.

Another approach is to base the sample test period on the measured margin exhibited during the initial compliance test [68]. The closer to the emissions limits was the initial measurement, then the more frequent should be testing of production samples. Continual sampling in this way will help build up a database of the statistical spread of levels, enabling a good degree of confidence that the product complies with the 80/80 rule (see section 2.2.4.1 on page 35). This is also one of the hidden functions of the signature on the declaration of compliance (see section 2.2.3.4): the signatory must be of sufficient seniority to be able to insist on a commitment to re-verification (since the signature commits the company to compliance of every product that is sold)[121]. The EMC co-ordinator, or test authority, should produce a document stipulating the re-verification testing that will be required during the product's life, related to the quality of pass during the initial compliance test. This forms part of the EMC-specific documentation for that product, is signed by the product manufacturing manager, and is maintained along with the compliance declaration.

16.4.2.2 ISO-9000

Within a company that has accreditation to ISO-9000, procedures can be set up and documented to ensure that EMC is maintained during production. These procedures will depend on a thorough knowledge of what aspects of the production process are important to EMC. Therefore it will be necessary to revise the company's operations in three ways [26]:

- determination by the design authority of all EMC-critical parts, assembly and installation methods, and processes;
- clear marking of these parts and work instructions to make plain to all concerned that they *are* critical, and a procedure to invoke the EMC authority when changes are needed;
- optimizing the level of in-house testing for early detection of variations.

Each of these issues can be addressed through the procedures listed in the company's quality manual.

16.4.2.3 Other parts of the company

EMC responsibility extends further than just the design and manufacture of a product. Larger equipment and distributed systems are built on- or off-site by production fitters and installers. Equipment in the field is repaired by service technicians. Each of these groups will either need to have a knowledge of EMC approaching that of the designers, or they will need to follow tightly-controlled procedures.

Similarly, the purchasing group can influence the EMC performance of a product. Components and sub-assemblies must be procured to EMC-related specifications if they are expected to have an effect on the final system's EMC profile: for instance, alternative sources of digital parts must not include devices which have faster switching times. The purchasing managers should be made aware of their responsibility through training, and procurement specifications must include a section devoted to EMC requirements.

16.4.3 Engineering change control

Section 16.4.2.1 above referred to a re-test regime that was based on an accumulation of changes during production. This implies that a system is in place to record and evaluate these changes. In most companies, certainly those that operate a quality system, engineering change control is a formal process, in which every design change after a product has gone into production is reviewed and documented before it is actioned. For EMC purposes, this process must cover any change that has implications for the compliance status.

Whatever change control is in place, the change document template should include a decision box to indicate whether it is an EMC-related change or not. Who actually makes this decision is a moot point; it depends on the degree of EMC expertise in the company and in whom it resides. Typically, it will require review by the EMC co-ordinator, but it may be the responsibility of the appropriate design engineer. Changes which are EMC-related should then be evaluated to decide whether they do not require a re-test at all, whether they should contribute to an aggregated re-test requirement or whether they justify a re-test in themselves. To aid traceability throughout the company and to support a compliance declaration, the decisions and their rationale should be documented within the system.

16.5 The control plan and documentation for Directives

All of the management activity discussed in this chapter implies that some level of EMC-related documentation will be necessary for each project. This is best described as an "EMC Control Plan".

16.5.1 The purpose of the control plan

The EMC control plan is a document, part of the specification of a new product, which lays down a schedule and a method to define how work towards the product's EMC compliance will be undertaken. Some projects within the military and aerospace sectors require this plan to be submitted as part of the tender documentation. Even if it is not a contractual requirement, the use within the design team of a detailed plan, showing which actions need to be taken and when, is a discipline that will reap benefits at the tail end of the design process when EMC performance comes to be evaluated. Sadly, if this discipline has not been applied the EMC performance often turns out to be the

Achilles heel of the product. Within a properly structured design environment the incorporation of an EMC control plan is not a major overhead.

Responsibility for EMC can be vested in the EMC co-ordinator or in an individual who is a member of the product design team. Their task is to develop the EMC control and test plans and within these to define the test set-up, operating modes and test failure criteria for the product. Familiarity with the appropriate standards and test methods as they apply to the product is essential. Structuring the EMC control in this way results in a strong "sense of ownership" of the EMC aspects of the product, but may not be an efficient way to establish and maintain the expertise in EMC that will be needed throughout the company. Alternatively, the company may have established a separate EMC control group, which oversees the EMC aspects of all product developments on an internal consultancy basis. Whilst this approach allows a company to build up a strong core of EMC-specific expertise, it may result in friction between the product development team and the EMC control engineers which may not eventually give the optimum product-specific solution.

DEF STAN 59-41

For military projects where the UK MoD procures to DEF STAN 59-41, part 2 of this standard [202] has explicit requirements for the formulation of both an EMC Control Plan and an EMC Test Plan.

The Control Plan essentially states how EMC will be achieved, particularly:

> The management and organisational procedures by which the EMC control programme will be implemented and liaison effected … The electrical, mechanical and installation design for EMC, including production and maintenance implications. The screening of the equipment cases, cable form screening policy, filtering and system grounding together with bonding policies … A Test and Qualification Programme for development models and all qualification testing.

Its purpose is to ensure that EMC requirements as defined by the Project Specification and DEF STAN are adequately addressed, to ensure the development of a cost-effective EMC control programme and to identify responsibilities.

16.5.2 The EMC Assessment

Under the old EMC Directive, a properly structured and documented control plan formed a valuable and major part of the Technical Construction File if this route to compliance was being followed. Under the new EMCD, it is pretty much a necessity in order to comply with the requirement for an EMC Assessment.

The requirements of the new EMCD were discussed in depth in Chapter 2. To briefly repeat its requirement here:

> 1. The manufacturer shall perform an electromagnetic compatibility assessment of the apparatus, on the basis of the relevant phenomena, with a view to meeting the protection requirements set out in Annex I, point 1. The correct application of all the relevant harmonised standards whose references have been published in the Official Journal of the European Union shall be equivalent to the carrying out of the electromagnetic compatibility assessment.
>
> 2. The electromagnetic compatibility assessment shall take into account all normal intended operating conditions. Where the apparatus is capable of taking different configurations, the electromagnetic compatibility assessment shall confirm whether the apparatus meets the protection requirements set out in Annex I, point 1, in all the possible configurations identified by the manufacturer as representative of its intended use.
>
> 3. In accordance with the provisions set out in Annex IV, the manufacturer shall draw up technical documentation providing evidence of the conformity of the apparatus with the essential

requirements of this Directive.

– Annex II

1. Technical documentation

The technical documentation must enable the conformity of the apparatus with the essential requirements to be assessed. It must cover the design and manufacture of the apparatus, in particular:

— a general description of the apparatus;

— evidence of compliance with the harmonised standards, if any, applied in full or in part;

— where the manufacturer has not applied harmonised standards, or has applied them only in part, a description and explanation of the steps taken to meet the essential requirements of the Directive, including a description of the electromagnetic compatibility assessment set out in Annex II, point 1, results of design calculations made, examinations carried out, test reports, etc.;

— a statement from the notified body, when the procedure referred to in Annex III has been followed.

– Annex IV

The control plan is clearly able to form the overall framework for this requirement for technical documentation. Much of the content that is required by Annex IV above will be delegated to the test plan and the ensuing test reports, if testing to harmonised standards forms the main part of the EMC compliance approach. The explicit requirement in Annex II for a confirmation that all possible configurations meet the essential requirements will often mean that a more extensive analysis to support testing of limited configurations is needed, and this should be documented in the control plan, as should any rationale for deviations from or limitation of the full tests required by the harmonised standards.

16.5.3 Contents

The control plan can be divided into two major sections, one defining the EMC risks, analyses and best design practice which is to be followed in developing the product, the other defining responsibilities, project reviews and control stages.

Part I

1. Definition of EMC phenomena to be addressed; catalogue of probable sources and victims; site- and installation-specific aspects; rationale for choice of applicable harmonised standards, and listing of these standards
2. Reference to test plan, and if necessary, analysis matrix for products that may take multiple configurations to support limited testing
3. Design practice

   This section can draw on the various EMC control methods as discussed in Chapters 11 to 14 of this book. Coverage should be as detailed as possible given what is already known about the design of the product.

   - Grounding regime, including a ground map
   - Control and layout of interfaces
   - Use of screened connectors and cable
   - PCB layout

- Circuit techniques
- Choice of power supply
- Filtering of power ports
- Filtering and isolation of signal ports
- Packaging design including screening

Part II

4 Project management

Mandatory EMC design reviews, control stages and checkpoints, and who is responsible for overseeing them; these can normally be incorporated into the overall project management scheme but there may be advantages in separating out the EMC functions.

- Responsibility for progressing EMC aspects
- Preliminary design review
- Design testing
- Detailed design review
- Pre-compliance confidence testing
- Final design review
- Compliance test or certification
- Responsibility for Declaration of Conformity
- Production quality assurance

Appendix A

Design checklist

Many factors must be considered when looking at the EMC aspects of a design, and it is easy to overlook an important point. This generic checklist is provided for you to assess your design against as it proceeds. For particular classes of design, you will want to expand it with your own experience.

- Design for EMC from the beginning; know what performance you require
- Partition the system into critical and non-critical sections:
  - determine which circuits will be noisy or susceptible and which will not
  - lay them out in separate areas as far as possible
  - select internal and external interface locations to allow optimum common mode current control
- Select components and circuits with EMC in mind:
  - use slow and/or high-immunity logic; apply slew rate limiting to data transmission interfaces
  - use series R buffering on all high-speed clock and data lines
  - use good power decoupling techniques: small, low-inductance capacitors adjacent to the ICs they are decoupling
  - use series ferrite chips in the supplies to create power segments
  - reduce fan-out on clock circuits by liberal use of buffers
  - minimize analogue signal bandwidths
  - maximize dynamic range of analogue signal paths
  - check stability in wideband amplifiers
  - don't leave unused IC input pins floating: tie them to 0V or V_{CC}
  - include resistive, ferrite or capacitive filtering at all sensitive analogue inputs
  - incorporate a watchdog circuit on every microprocessor
  - avoid edge triggered digital inputs if possible, protect them if unavoidable
- PCB layout:
  - refer to separate checklist in section 11.2.4
- Cables:
  - segregate, and avoid parallel runs of, signal and power cables

- choose RF-screened cables if the wanted signal cannot be properly filtered
- avoid screened cable with the screen connected only at one end; if unavoidable, treat the cable as unscreened at RF
- use twisted pair both within and outside an enclosure, for balanced or high di/dt lines
- use properly designed looms, ribon or flexi for internal wiring – avoid loose wires or bundles
- run cables away from apertures in the shielding, tied close to conductive grounded structures
- apply ferrite suppressors to damp resonances and control common mode currents
- ensure that cable screens are properly terminated to the connector backshell; avoid pigtails
- terminate lines carrying high frequency signals with the correct transmission line impedance

- Grounding:
  - design and enforce the ground system at the product definition stage
  - consider the ground system as a return current path, not just as 0V reference
  - provide for parallel earth conductors at the system level
  - ensure metal-to-metal bonding of screens, connectors, filters, and enclosure panels
  - ensure that bonding methods will not deteriorate in adverse environments
  - mask paint from, and apply a conductive finish to, any intended contact surfaces
  - keep earth straps short and define their geometry
  - avoid common ground impedances for different circuits
  - provide an interface ground area for decoupling and filtering
- Filters:
  - assume that a supply filter is needed: design the filter for the application
  - filter all I/O lines, using either or both of three-terminal capacitors to interface ground, and common mode chokes
  - apply π filters at the DC power input to each board, in multi-board designs
  - ensure a defined ground return for each filter
  - apply filtering to interference sources, such as switches or motors, directly at their terminals
  - locate all filter components and associated wiring or tracks adjacent to the interface being filtered

- Shielding:
  - design all metallic structures as if they were electrical components: account for their stray capacitance and inductance
  - consider segregated enclosures: enclose particularly sensitive or noisy areas with extra internal shielding
  - avoid large or resonant apertures in a shield, or take measures to mitigate them
  - avoid dipole-like structures in a metallic enclosure
  - ensure that separate panels are well bonded along their seams using conductive gaskets: apply good bonding practice as in “grounding” above
  - design plastic enclosures to allow internal conductive coating if necessary
  - decide on and implement DC or RF tie points between circuit 0V and the shield
  - use multiple internal tie-points to minimize box resonances
- Test and evaluate for EMC continuously as the design progresses

Appendix B

CAD for EMC

B.1 Overview

It may seem strange to devote no more than a few pages at the back of a book on product design for EMC to the important subject of computer-aided design (CAD). The prospect of non-compliance with EMC requirements is sufficiently threatening that manufacturers are having to give these requirements serious consideration at the design stage, and the second part of this book discusses the major design principles that this involves. Many other aspects of the circuit design process are now automated and simulated to the extent that a breadboard stage, to check the correctness of the basic design concepts, is often no longer necessary. A very attractive option to the product designer would be a CAD tool that predicted RF emissions and susceptibility, with enough accuracy for initial evaluation purposes, from the design data of the product. This would allow alternative EMC techniques to be tried out before the costly stage of committing to tooling and pre-production had begun.

The reason why this subject is relegated to an appendix is because no such tool yet exists. Many problems of a specific nature can be solved by electromagnetics computation packages that have been available for many years, but these generally address the needs of EMC experts rather than product designers. Several groups are working on the production of software packages that need less expertise for successful use, and the next few years may well see the successful introduction of such tools, but history does not give much cause for optimism.

The difficulties facing these researchers are many:

- modelling a small collection of electromagnetic emitting elements (usually a current segment) is easy. As the number of elements n increases, the computational memory increases as n^2 and the processing time as n^3, so that computer speed and storage limitations, despite having vastly increased over the years, still restrict the size of the problem being handled.
- as the frequency increases, so the element size must be reduced to maintain accuracy, further increasing the required number of elements. Many computational methods work in the frequency domain and so the calculations must be repeated for each frequency of interest.
- PCBs with ground plane layers can effectively be approximated as conducting sheets, which simplifies the structure model, whereas those without ground planes cannot be so easily modelled.
- connecting wires and cables have a large effect on the coupling and must be modelled explicitly. This may invalidate the results if the layout is not properly defined in the final product. Some early attempts modelled the performance of the PCB only and their results bear little relevance to reality.

- the driving currents can theoretically be derived from device models and calculated track and circuit impedances, but these may differ substantially from the real circuit with real tolerances. For example the amplitudes of higher order clock harmonics depend heavily on risetime, which is a poorly specified parameter and is affected by circuit capacitances and device spread.

These difficulties relate only to emission predictions. A further set of variables come into play when you try to simulate immunity, and apart from some research into immunity at the IC level these have hardly been addressed yet. There are other obstacles to implementing practical tools, which are not related to the technical problems of electromagnetic modelling [95]:

- EMC design aspects do not respect the demarcation common in design labs of circuit, layout and mechanical disciplines. Considerable interaction is needed between these which presupposes a common body of EMC knowledge among the different fields, which is unlikely to exist.
- To overcome this demarcation, the CAD package must integrate these aspects and must therefore accept input on all fronts: circuit schematics, PCB layout designs and mechanical and wiring drawings. Although integrated CAD environments are installed which could provide such input automatically, interaction between the different parts is often poor and manual input is not realistic given the time constraints facing a typical design department.
- The output of the package must also be in a form which is useable and comprehensible by these different designers. It must be structured to be of maximum assistance at the most appropriate phase(s) of the design process. Some re-training of the users may be required so that they can actually use the output.

B.2 Modelling packages

Those software packages which are currently available for electromagnetic modelling purposes have been developed for solving EMI coupling problems in certain well-defined applications. Every EMC problem can be described in terms of a source of interference, a coupling path and a receptor (or victim) of the interference. The structure of the coupling path may include either or both radiative or conductive mechanisms, and these are often amenable to analysis so that, for instance, the voltage or current present at an interface with an item of equipment can be derived as a result of coupling of the structure containing that equipment with an external field.

A typical application may be to model the surface currents on an aircraft fuselage which is illuminated by a plane-wave field, and deduce the currents flowing in the cable bundles within the fuselage. Another might be to determine the RF energy transport by penetrations through a shield such as electrical conductors or pipework. The codes which perform the electromagnetic modelling of these situations use finite difference, method of moments, transmission line modelling or finite element methods to solve Maxwell's equations directly. Each has its advantage in a particular situation, for instance the method of moments code NEC deals efficiently with wire coupling problems such as occur in antenna design, whereas the finite element code EMAS or transmission line modelling (TLM) are more suited to problems with inhomogeneous regions and complex geometries, such as surface currents on enclosures.

Many approximations need to be made even for well-defined problems to allow computation with a reasonable amount of effort. For instance near field coupling is much more complicated than far field, since the nature of the source strongly affects the incident wave. Shield aperture size is critical since for electrically large apertures (size comparable to a wavelength) the internal and external regions must be treated consistently, whereas the modelling of fields penetrating through small apertures can be simplified.

The major difficulty with applying these packages to commercial product design is that they have been developed for a rather different set of purposes. Mostly they have been derived from research on military EMC problems. This has two consequences:

- technically, they are appropriate only for situations which are clearly defined: cable routes, connector terminations, mechanical structures and shielded enclosures are all carefully designed and controlled through to the final product and its installation. This contrasts with the commercial environment where the costs of doing this, and indeed the impossibility in many cases, puts such an approach out of court. Thus the parameters and approximations that would have to be made in the simulation are not valid in reality.
- operationally, there are difficulties in performing the simulation. Only a few of the packages (which started life in university research departments) have been developed to the point of being user-friendly. A great deal of input data, usually in the form of structure co-ordinates, must be loaded and validated before the program can run, and its output must then be interpreted. A typical product design engineer will have neither the time nor the training to do this – only those companies with the resources to run a specialist group devoted to the task can handle it successfully.

Nevertheless, these packages can find a use in areas of EMC not directly related to product design, such as those to do with radiated field testing: predicting reflections in screened rooms and damping them, predicting proximity effects between the antenna and the EUT, and predicting calibration errors when antennas are used in screened rooms [49]. The University of Missouri-Rolla EMC lab website (http://www.emclab.umr.edu) lists 55 commercial suppliers of electromagnetic modelling codes, and 21 different free sources. The vast majority of these use specific classes of model and are appropriate only to a given type of problem, and most are for electromagnetic applications other than EMC.

One hurdle which still needs to be overcome is the simple one of gaining confidence in the output of a particular simulator. One worker in the field has said [61]

> Uncertainty arises when the predicted results using one type of CEM code do not agree favourably or consistently with the results of other codes of comparable type as well as against measured data on benchmark models. Many examples can be cited where fairly significant deviations among analytical or computational techniques or between empirical based methods have been observed. Differences are not unexpected, but the degree of disparity in certain cases cannot be readily explained nor easily discounted, which leads to the fundamental question, "... which result is correct?"

B.3 Circuit CAD

Some circuit design and PCB layout CAD packages already offer transmission line analysis for high-speed logic design, taking into account track parameters calculated from their geometries and the board dielectric. Extending these capabilities into the domain of RF properties of the board interacting with its environment – as will be needed for EMC prediction – will be difficult. A possible approach is to segment the overall problem into e.g. the PCB, the internal cables, the case and the external cables, and then apply appropriate numerical methods sequentially and assemble the outputs into an overall result. This requires the analysis to be done by someone with expertise both in the appropriate segmentations and in defining acceptable approximations.

Design advisers

Commercial offerings of various types are now available, though. One of these types is a rule-based design adviser. The design adviser takes as its input a set of rules, such as allowable loop area or decoupling capacitor placement, which are supplied with the package but can be enhanced with in-house rules as required. A particular advantage of this approach is that it is not limited to EMC: other design rules, for instance to do with thermal management or manufacturability, can also be incorporated and trade-offs between different constraints can be evaluated quickly and automatically.

The circuit designer must specify at the schematic capture stage those nodes that are to be checked; the rules can be individually selected and weighted. The PCB layout draughtsman can then initiate a rule check on the layout as it develops and rule violations can be highlighted. The advantage of working in this way is that it limits the EMC expertise needed by the layout draughtsman, since the rules act as a form of "expert system", but it is important that the rules are soundly based and do not become so codified that they constrain the development too much to produce cost-effective products. The technique does not claim to perform any EMC prediction or calculations at all.

Emissions prediction

Some packages are available which do offer emissions prediction based on calculations. It is possible, for instance, to analyse transmission line structures on multilayer ground plane boards, as well as those with a discrete return path, for near field radiation. Interfaces to various layout packages for layout data extraction, and macromodels for some circuit components, are available.

Another approach is to integrate EMC aspects with other parts of the mechanical design which can be automated, particularly thermal issues. Thus for instance design of heatsink size and placement, and ventilation apertures in a metal case, have both EMC and thermal consequences. Flomerics, who have established themselves firmly in the thermal design and analysis sector, have extended their capability with an EMC simulation package using TLM which looks at the effect on radiated emissions of enclosure resonances and apertures.

It is still necessary, though, to be sceptical of the capabilities of these tools when faced with the demand to predict properly the EMC of a whole product. At the time of writing, few vendors are offering tools which have been extensively validated against results from real designs. Some years ago, one commentator made the wry but accurate observation [132] that

> Once again, we are in danger of being seduced by our desire for a comprehensive solution into the belief that, perhaps, this time the marketing is based on reality. Alas ... this set of EDA technologies [based on analytical tools] ignores many aspects of EMI.

Not much has changed.

Appendix C

Case studies

This section discusses a few real-life case studies of electronic product design, each of which illustrates some aspects of the EMC design principles discussed earlier in this book. While based on real products, the details have, of course, been dis-identified.

C.1 Cockpit display

A common EMC problem arises with large (say, 10” or 12”) LCD displays driven with standard VGA or high resolution video. The product featured here was cost limited and was required to be daylight visible, but also had to meet stringent radiated RF emission limits, not just the standard Class B ITE limits of EN 55022. Because of the first two requirements, the third couldn’t be met by a shielded window – it would have been too expensive and would have attenuated the light transmission too much. So although a shielded enclosure was acceptable and necessary for the electronics, which included a 100MHz processor, the display had to be outside the shield. The challenge was to find a way of doing this without compromising the emissions limits.

Emissions from these displays tend to be dominated by the pixel clock and its harmonics, and come via two coupling routes: direct radiation from the glass face, and common mode radiation from the whole assembly. The face radiation is entirely controlled by the LCD manufacturer – it depends on the transition speed of the edges created by the transistor matrix drivers, and the physical configuration of the conductor pattern on the glass – so the first task was to evaluate the emissions from several manufacturers’ samples in a standardized jig. From this exercise a couple of front-runners emerged, the final choice being made on other features.

The emissions limits in this application were not uniform across the frequency range – they were tighter in those frequency bands that included receivers in close proximity to the display unit. The second task then was to identify clock frequencies whose fundamental and harmonics would fall within these bands and try to eliminate them from the design. This eventually meant outlawing certain display modes, and subtly tweaking others to shift their pixel clocks slightly up or down. Choosing a display and a graphics controller with maximum flexibility in acceptable clock frequencies aids this process; it’s not too difficult if the only display is the one on-board, but harder if you have to feed the video out in parallel to an external monitor so that it must meet the specified VESA standards. As well as the display frequencies, the processor clock frequencies had to be vetted and tweaked in the same way.

The major design work involved the creation of a structure that would minimize the common mode radiation. This radiation is due to noise voltages appearing on the metal case of the LCD relative to the shield, which then emits in much the same way as a patch, or transmission line, antenna. These voltages are amenable to control by careful design. This took a number of forms, referring to Figure C.1:

- the shielded enclosure for the rest of the electronics was maintained along the back of the module, behind the front face of the unit but enclosing the electronics;
- the separation distance between the back of the LCD module and the outside of the shield was kept to a minimum: radiating efficiency is directly proportional to this distance;
- the back of the LCD module was grounded at multiple points to the shield, using conductive gasket material (in some applications the LCD is at circuit 0V, and this has to be separate at DC from the shield. This means that capacitance has to be inserted in series with the grounding points, which is tricky but possible);
- most importantly, the signal cable to the LCD should not introduce common mode noise along its length, despite carrying the pixel clock and the video data, and despite the LCD connector being poorly specified for screening. Wire bundles are ruled out. The best solution is a flexi connection between the driver on the electronics PCB and an adapter to plug into the LCD module (see section 13.1.8.3). The double-sided flexi has one side devoted to a ground plane that is taken through multiple pins to the driver PCB 0V, which is locally grounded to the shield to minimize the noise voltage difference between the ground plane and the shield. The adapter converts the flexi to the LCD connector, and takes a separate low-inductance connection to the LCD case.

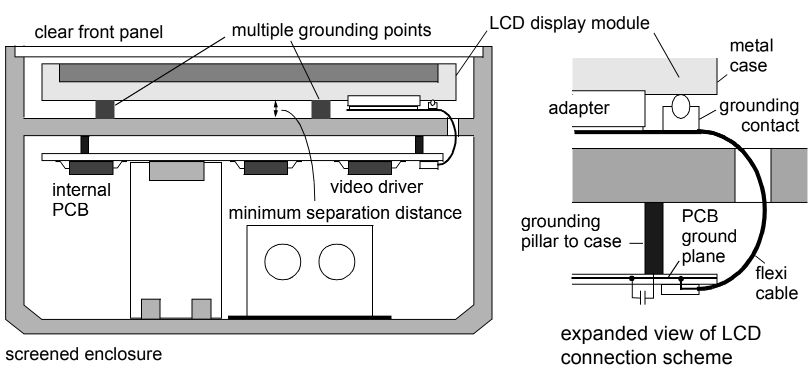


Figure C.1 General assembly of cockpit display

C.2 Liquid tank sensor

A manufacturer of instrumentation, which included temperature and liquid level sensing transmitters, discovered a market in the marine sector; but this entailed hardening the electronics against RF to a level of 20V/m whereas before it only needed 3V/m. The least possible and lowest cost change to the design was desired. The original unit had a remote thermistor and level sensing potentiometer, connected by unscreened wires to terminal blocks on the two-layer electronics PCB in a plastic box (Figure C.2).

The unit was connected to the system control by a long distance 4–20mA link, again over unscreened wires.

To improve the hardening, a classical interface filtering plus shielding approach was needed. The PCB design was expanded to four layers and a 0V plane was implemented. The terminal blocks remained on the upper side of the board and the circuit was enhanced with filter capacitors to the ground plane on all terminal pins, and capacitors applied for bandwidth limitation at selected points in the analogue circuit – generally at the inputs to the operational amplifiers. Finally, the interior of the box was specified to be sprayed with low-cost conductive paint (at the time, nickel), and the edge of the board was fitted with a series of surface mount spring fingers to make contact between the 0V plane and the conductive coating. After all this, there was no susceptibility discernible up to at least 30V/m. The manufacturer later found that he could omit the nickel shielding on the plastic box, and the unit was still adequately immune to 20V/m due solely to the PCB-level changes.

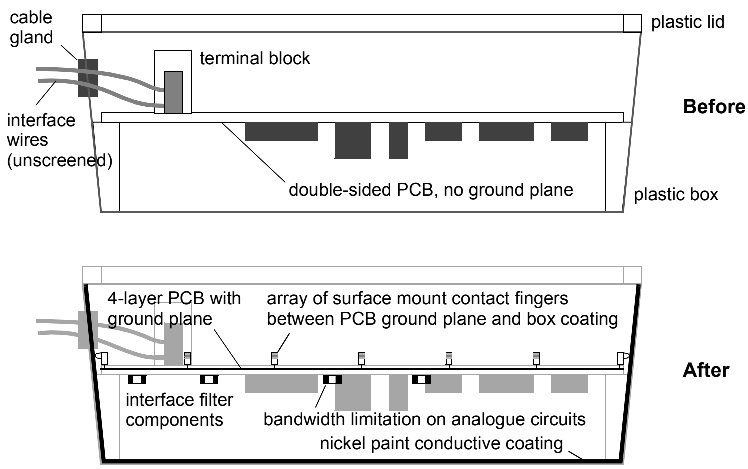


Figure C.2 Changes to the level/temperature transmitter

C.3 The problem with wall-warts

Because many small products – particularly IT and consumer products – use low levels of power at DC (typically 12V) but need to be supplied at mains voltage, a large market has grown up for external DC switchmode power supplies which plug directly into the mains outlet (hence, “wall-wart”) and which supply power to the application unit via a short two-wire low-voltage cable. The product manufacturer normally buys these in from an external supplier, often in the Far East, and has no control over the design of the supply itself. The wall-wart supplier will typically offer a product which does in fact meet at least the conducted emissions requirements of the IT standard EN 55022, and this might appear to be enough for the product manufacturer who will package it with their own unit. But the supplier of such a “system” must ensure that the whole system complies when it is used in a typical configuration, and the connection of a DC load to the output of the wall-wart may change its emissions profile, as well as adding its own emissions coupled back down the DC power lead.

This isn't the only problem. EN 55022 Edition 3 includes requirements for conducted emissions tests on telecom ports. In these tests, you again bump into the problem of conducted emissions generated by the wall-wart power supply. To understand the mechanism, we must delve more deeply into the equivalent circuit.

C.3.1 The conducted emissions equivalent circuit

The equivalent circuit of the conducted emissions test, both for mains port and telecom port emissions, can be drawn as shown in Figure C.3. For the sake of this discussion, emissions generated within the unit – for illustration, call it a LAN switch – itself are ignored. The wall-wart is Safety Class 2, i.e. it has no connection to its safety earth pin.

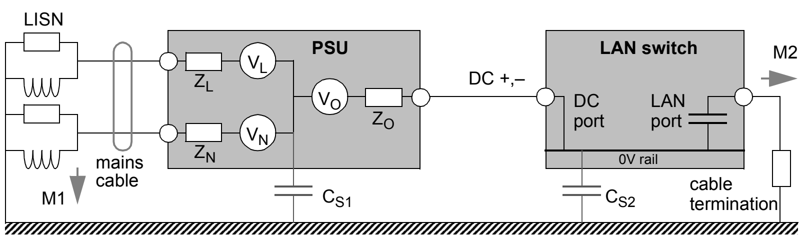


Figure C.3 Equivalent circuit for conducted emissions

The interference is generated within the wall-wart PSU due to its switching power supply and is represented by V_N and V_L, feeding the live and neutral mains connections, and V_O, feeding the DC output connection. Each has a source impedance Z_L, Z_N or Z_O, and each is referred to the enclosure and thence by stray coupling to the ground reference plane of the test. These impedances are likely to be predominantly capacitive and affected by layout aspects and transformer construction of the power supply. (If the enclosure is entirely plastic then the capacitive interconnections become more complicated, but do not vanish.)

It is vital to appreciate that the components at the output of the wall-wart form the *common mode* circuit; that is, neither the DC link voltage nor the LAN output signals are of any interest to us and are not shown, only the entire cable, considered as a whole, is shown. It doesn't actually matter whether or not the LAN switch itself is powered up and operating for this circuit and its effects to be valid.

V_O can only be referred to the case of the unit, to which V_N and V_L are also referred. This means that V_O will be added to the measurement M1 of both V_L and V_N, but its level will depend on the ratio of the various impedances in the power supply circuit (Z_O, Z_L, Z_N and the LISN impedance) as well as the impedance of whatever the output is connected to. In the circuit this is determined by the connection of the LAN switch's signal port. Clearly then, mains emissions in this case will be affected by the test set-up at the output of the wall-wart.

If there were no output connection, V_O would be irrelevant, and the levels measured at live and neutral would be entirely determined by V_L, V_N, their source impedances and the LISN impedance. But with the LAN switch connected, V_O is added to each of V_N and V_L from a source impedance determined at low frequency by the series combination of Z_O, the LAN port output impedance and the cable termination impedance. From this you can see that if the LAN cable termination is a high impedance (say, a short unscreened cable to a laptop) the effect of adding V_O will be negligible; but if it is low, such as a desktop PC fed via a screened cable, V_O will be significant. So it would be quite reasonable for two different but equally valid test set-ups to produce markedly different results.

The situation is complicated at high frequencies by the presence of stray capacitances C_{S1} and C_{S2}, and also by the inductive impedance of the DC power cable and the signal cable. These can often combine with other strays into a series resonant circuit that gives a characteristic hump in switchmode emissions around 10–25MHz (see Figure 12.18 on page 311), but usually they are less significant below, say, 1MHz.

C.3.2 Telecom port emissions

The previous discussion has concerned mains conducted emissions, but Figure C.3 shows that similar issues apply to telecom port emissions, i.e. those measured in common mode on the LAN cable (M2). Although some such emissions will be generated within the LAN switch, particularly the common mode component of the LAN data, there will also be a significant component that is passed directly through the LAN switch from the wall-wart. Here, the worst case will be if the reference for V_O is in fact directly connected to earth (Safety Class 1 supplies). This allows the maximum current developed by V_O to flow through Z_O and out through the terminated LAN cable. Even if V_O is negligible, the mains sources V_L and V_N will also contribute to levels at the telecom port if the earth is removed.

C.3.3 Mitigation

Because all these emissions are propagated from the wall-wart to the LAN switch in common mode, any differential mode fixes such as capacitive filtering at the DC supply input of the LAN switch will be utterly useless. Indeed, this is a simple diagnostic technique: put a big capacitor, say 1μF, across + and – at the input to the LAN switch and check to see if it makes any difference. If it doesn't, and if you have confirmed that the emissions frequency spectrum implicates the wall-wart, this is conclusive evidence of the coupling path.

In this case, shielding of the LAN switch or providing a metal chassis or improving its PCB layout will also all be useless, since the interference is being passed directly from DC input to LAN output.

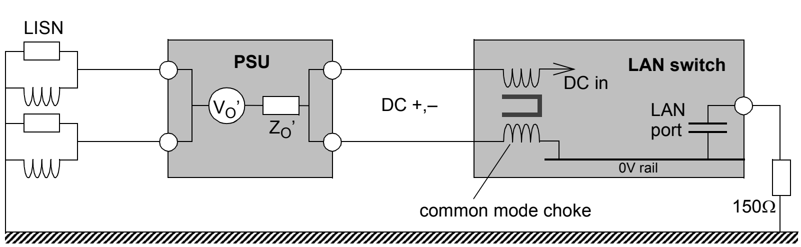


Figure C.4 Inserting a common mode choke

The first solution that will work is to put a common mode choke right at the DC input to the LAN switch (Figure C.4). The choke increases the impedance in series with Z_O' (the total common mode source impedance). This will reduce the disturbance currents flowing through the LAN cable in direct proportion to the difference in impedances with and without the choke. Thus if the dominant impedances without the choke are (Z_O' + 150) ohms, and you need a 6dB improvement in emissions, the CM choke must also have an impedance of (Z_O' + 150) ohms at the appropriate frequency. This works well if Z_O' is relatively low and the frequencies are high; there are many small surface mount CM chokes for DC inputs that can give well over 1kΩ impedance. It also has the benefit of reducing the effect of any common mode emission sources within the LAN switch. If the frequency is down towards 150kHz though, and Z_O' is already high (implying that the wall-wart is acting more like a common mode current source), then you will need a potentially very large CM choke to have enough effect.

The next "solution" that will work is to reduce the impedance of the LAN switch itself to the measurement ground plane, by making a direct or indirect connection to it. If the unit were to have an earth point, then this would be directly connected to the ground plane and the majority of the power supply emissions would pass into the plane rather than through the LAN port. But how many such devices have an intentional earth connection?

You may, though, have the option of making additional connections to other ports. If these have a low-impedance connection to the LAN switch's 0V plane, and also a low impedance termination to the ground plane (or at least stabilized to 150 ohms, as with the LAN port) then they will also pass a proportion of the power supply emission currents to the plane (Figure C.5). The proportion will be directly related to the ratio of impedances at the various ports; if the impedance of one additional connected port is the same as that of the measured LAN port, then a reduction in the LAN port emissions of 6dB can be expected.

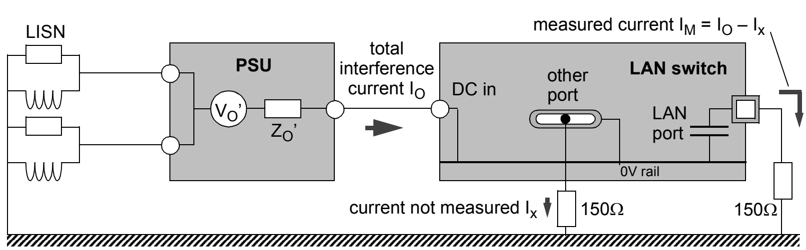


Figure C.5 Connections to other ports

Lastly, the best way of dealing with the whole problem is to specify the emissions of the wall-wart supply properly. Instead of leaving it to the wholly unsatisfactory and uncontrolled requirement that the wall-wart be CE Marked, place explicit specifications on both the input and output emissions levels. It will generally be sufficient to quote EN 55022 Class B levels for the mains conducted emissions, with the output terminated in a DC load equivalent to its maximum rated power and at the same time a common mode impedance of 150 ohms to the ground plane. The crucial extra requirement is that the DC output common mode voltage and/or current is also specified to EN 55022 Class B telecom port emission levels, under the same conditions. If the supplier can be contractually persuaded to meet these limits, then you have good confidence that any emission failures are due to your own product, not the wall-wart.

C.4 The dipole problem: a box in two halves

The example used here is a fingerprint analyser, but the principles apply to a wide range of products which are mechanically designed in the same way. The form of construction (Figure C.6) neatly illustrates a number of points.

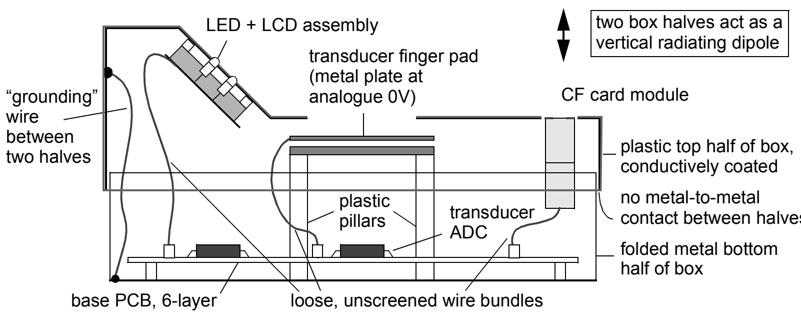


Figure C.6 Two half-boxes acting as a vertical dipole

This design had serious ESD susceptibilities as well as over-limit radiated emissions. It might also have suffered radiated RF susceptibilities, but nobody found this out because the test programme was stopped once the early problems were discovered.

The base (processor) board design was fundamentally sound, with both ground and power planes on a six-layer PCB. But the various peripherals, all mounted on the upper half of the box, fed down to several connectors on the base board through loose wire looms. Anticipating possible EMC problems, the manufacturer had elected to "shield" the box by conductively coating the inside of the top part, and had "grounded" the two halves together by a 15cm length of (admittedly green-and-yellow) wire.

The ESD susceptibilities manifested in three ways:

- air discharges to the LEDs and LCD bezel, both of which were accessible through the front panel, would immediately corrupt the LCD display and, after an indeterminate number, crash the processor;
- contact discharges to the finger pad would destroy the analogue-to-digital converter (ADC), sometimes immediately but always within a few discharges;
- air discharge to the edge of the Compact Flash (CF) card would corrupt the CF memory and crash the processor.

In addition, although the unit's radiated emissions were satisfactory under quiescent conditions, when the CF card was addressed (which happened frequently) they shot up above the limit. This problem was clearly due to the signals present on the card interface wires as it could be traced to the relevant clock frequencies, creating noise on the CF module which then coupled capacitively to the metallization on the top half of the box. Since the top and the bottom were only connected via a length of wire, the pair formed an effective radiating dipole antenna in the vertical plane. The wire's inductance merely re-tuned the antenna's worst-case resonance, it didn't reduce the radiating efficiency.

C.4.1 Mitigation

The various changes that were made to this design are shown in Figure C.7. To solve the radiated emissions problem, the most effective method was in fact to *eliminate* the top half of the shield, while "shielding" the cables to the CF module. The conductive coating was removed from the top plastic half of the case (which pleased the designer, since it was a noticeable cost reduction). The loose wire cable was enclosed in a metallized sheath bonded at one end to the 0V plane on the base board, and at the other to the case of the CF module; this reduced the common mode noise carried up to the CF module, and the lack of shielding on the top half meant that the radiating dipole effect was limited only to the module itself, not the whole case. It would have been possible to have left the top half shielded, with all the other measures to be described in place, and improved the bond between the top and bottom halves by using appropriate conductive fasteners and gaskets so as to reduce the dipole effect; but this would have been a lot more expensive and would have probably created further ESD problems.

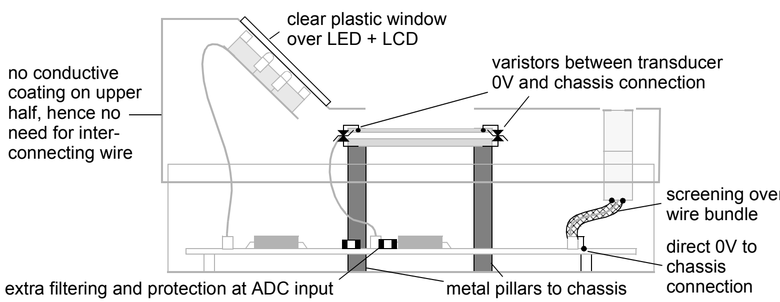


Figure C.7 Modifications to the box in two halves

A preferable way to have reduced the CF module noise would have been to redesign the cable connection to it for minimum transfer impedance, by using either multi-ground connections in a ribbon or a ground plane flexi. But this wasn't an option; shielding the cable assembly was. In other circumstances, the better options would be possible.

The effect of making this change to the cable was also to increase the ESD immunity of the CF module. It was important that the cable screen at the base board was taken to the 0V plane very close to a grounding connection to the chassis. Otherwise, the discharge current would have had to flow in a large area of the 0V plane, with greater stress to the processor operation.

The ESD susceptibility of the LED/LCD assembly was reduced to some extent by a ferrite sleeve on the wires coupling to the base board, but this was not enough. Rather than spend a lot of effort on improving the immunity of this connection, the easiest way around that problem was to prevent a discharge occurring in the first place: both the LEDs and the LCD were protected by a clear plastic window placed over the front of the enclosure, sufficiently large so that no short creepage path existed around its edge to any conductive parts of the components. This is invariably the preferred method of dealing with discharges to any similar parts.

The ESD susceptibility of the transducer ADC was dealt with in the opposite way. There was no way around the fact of a discharge occurring to the metal finger pad. But the discharge current could be diverted to the base of the unit, simply by using metal instead of plastic pillars, and putting varistors between the analogue 0V of the contact pad and its now chassis-connected metal backing plate. At the same time, the protection of the ADC input was beefed up by series resistors and parallel steering diode pairs. Then, the discharge current was encouraged to flow through the pillars to chassis rather than through the high impedance of the ADC input, which in turn could cope with the remanent spike that did make it onto the base board.

Appendix D

Useful tables and formulae

D.1 The deciBel

The deciBel (dB) represents a logarithmic ratio between two quantities. Of itself it is unitless. If the ratio is referred to a specific quantity (P_2, V_2 or I_2 below) this is indicated by a suffix, e.g. dBμV is referred to 1μV, dBm is referred to 1mW.

Common suffixes

suffix	refers to	suffix	refers to
dBV	1 volt	dBA	1 amp
dBmV	1 millivolt	dBμA	1 microamp
dBμV	1 microvolt	dBμA/m	1 microamp per metre
dBV/m	1 volt per metre	dBW	1 watt
dBμV/m	1 microvolt per metre	dBm	1 milliwatt
		dBμW	1 microwatt

Originally the dB was conceived as a power ratio, hence it is given by:

$$\text{dB} = 10 \log_{10} (P_1/P_2)$$

Power is proportional to voltage squared, hence the ratio of voltages or currents across a constant impedance is given by:

$$\text{dB} = 20 \log_{10} (V_1/V_2) \quad \text{or} \quad 20 \log_{10} (I_1/I_2)$$

Conversion between voltage in dBμV and power in dBm for a given impedance Z ohms is:

$$V(\text{dB}\mu\text{V}) = 90 + 10 \log_{10} (Z) + P(\text{dBm})$$

dBμV versus dBm for Z = 50Ω

dBμV	μV	dBm	pW	dBμV	mV	dBm	nW
–20	0.1	–127	0.0002	30	0.03162	–77	0.02
–10	0.316	–117	0.002	40	0.10	–67	0.2
				50	0.3162	–57	2.0
0	**1.0**	**–107**	**0.02**	60	1.0	–47	20.0
							μW
5	1.778	–102	0.063	70	3.162	–37	0.2
7	2.239	–100	0.1	80	10.0	–27	2.0
10	3.162	–97	0.2	90	31.62	–17	20.0
15	5.623	–92	0.632	100	100.0	–7	200.0
20	10.0	–87	2.0	**120**	**1.0V**	**+13**	**20mW**

Actual voltage, current or power can be derived from the antilog of the dB value:

$V = \log^{-1}(\text{dBV}/20)$ volts
$I = \log^{-1}(\text{dBA}/20)$ amps
$P = \log^{-1}(\text{dBW}/10)$ watts

Table of ratios

dB	Voltage or current ratio	Power ratio	dB	Voltage or current ratio	Power ratio
–30	0.0316	0.001	12	3.981	15.849
–20	0.1	0.01	14	5.012	25.120
–10	0.3162	0.1	16	6.310	39.811
–6	0.501	0.251	18	7.943	63.096
–3	0.708	0.501	20	10.000	100.00
0	**1.000**	**1.000**	25	17.783	316.2
1	1.122	1.259	30	31.62	1000
2	1.259	1.585	35	56.23	3162
3	1.413	1.995	40	100.0	10,000
4	1.585	2.512	45	177.8	31,623
5	1.778	3.162	50	316.2	10^5
6	1.995	3.981	60	1000	10^6
7	2.239	5.012	70	3162	10^7
8	2.512	6.310	80	10,000	10^8
9	2.818	7.943	90	31,623	10^9
10	3.162	10.000	100	10^5	10^{10}
			120	10^6	10^{12}

D.2 Antennas

D.2.1 Antenna factor

$AF = E - V$
where AF = antenna factor, dB/m
E = field strength at the antenna, dBμV/m
V = voltage at antenna terminals, dBμV

D.2.2 Gain versus antenna factor

$G = 20 \log F - 29.79 - AF$
where G = gain over isotropic antenna, dBi
F = frequency, MHz
AF = antenna factor, dB/m

D.2.3 Dipoles

Gain of a $\lambda/2$ dipole over an isotropic radiator:

$G = 1.64$ or 2.15dB

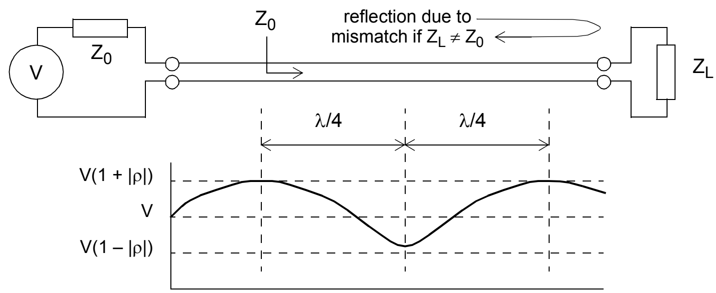
Input resistance of short dipoles of length L [20]:

$0 < L < \lambda/4$: $R_{in} = 20 \cdot \pi^2 \cdot (L/\lambda)^2$ ohms

$\lambda/4 < L < \lambda/2$: $R_{in} = 24.7 \cdot (\pi \cdot L/\lambda)^{2.4}$ ohms

D.2.4 VSWR

The term Voltage Standing Wave Ratio (VSWR) describes the degree of mismatch between a transmission line and its source or load. It also describes the amplitude of the standing wave that exists along the line as a result of the mismatch.



VSWR K = $(1 + |\rho|)/(1 - |\rho|)$ = (Z_0/Z_L) or (Z_L/Z_0)

Reflection coefficient $|\rho|$ = $(K - 1)/(K + 1)$

D.3 Fields

D.3.1 The wave impedance

In free space:

$Z_0 = (\mu_0/\varepsilon_0)^{0.5} = E/H = 377\Omega$ or 120π

μ_0 = permeability of free space = $4\pi \cdot 10^{-7}$ Henries per metre
ε_0 = permittivity of free space = $8.85 \cdot 10^{-12}$ Farads per metre

D.3.2 Near field / far field

$d < \lambda/2\pi$: near field; $d > \lambda/2\pi$: far field

(see Figure 10.9 on page 231)

In the near field, the impedance is either higher or lower than Z_0 depending on its source. For a high-impedance field of F Hz at distance d metres due to an electric dipole:

$|Z| = 1/(2\pi F \cdot \varepsilon \cdot d)$

For a low-impedance field due to a current loop:

$|Z| = 2\pi F \cdot \mu \cdot d$

D.3.3 Power density

Conversion from field strength to power density in the far field:

$P = E^2/(120 \cdot \pi)$
where P = power density, mW/cm^2
E is field strength, volts/metre

or for an isotropic antenna:

$$P = P_T/4\pi \cdot R^2$$

where R is distance in metres from source of power P_T watts

D.3.4 Field strength

For an equivalent radiated power of P_T, the field strength in free space at R metres from the transmitter is:

$$E = (30 \cdot P_T)^{0.5} / R$$

or $$E\ (\text{mV/m}) = 173 \cdot (P_T \text{ in kW})^{0.5} / (R \text{ in km})$$

Propagation near the ground is attenuated at a greater rate than 1/R. For the frequency range between 30 and 300MHz and distances greater than 30 metres, the median field strength varies as $1/R^n$ where n ranges from about 1.3 for open country to 2.8 for heavily built-up urban areas [161].

D.3.5 Field strength from a small loop or monopole [11]

"Small" in this context means substantially shorter than $\lambda/4$. For a loop in free space of area A m^2 carrying current I amps at a frequency f Hz, electric field at distance R metres and an elevation angle θ is:

$$E = 131.6 \cdot 10^{-16} (f^2 \cdot A \cdot I)/R \cdot \sin\theta \quad \text{volts/metre}$$

Correcting for ground reflection (x2) with a measuring distance of 10m at maximum orientation:

$$E = 26.3 \cdot 10^{-16} (f^2 \cdot A \cdot I) \quad \text{volts/metre}$$

Short monopole of length L (<< $\lambda/2$) over ground plane at distance R driven by common mode current I:

$$E = 4\pi \cdot 10^{-7} \cdot (f \cdot I \cdot L)/R \cdot \sin\theta \quad \text{volts/metre}$$

Maximum orientation at 10m:

$$E = 1.26 \cdot 10^{-7} \cdot (f \cdot I \cdot L) \quad \text{volts/metre}$$

D.3.6 Field strength from a resonant cable [17]

When cable length approaches or exceeds a wavelength, the resonances drastically change the emission patterns resulting in multiple lobes depending on the L/λ ratio. Maximum field intensity occurs when the radiator length is $\lambda/2$. Now the field strength (uncorrected for ground reflections, since these are unpredictable) is:

$$E_\theta = \{(60 \cdot I)/R\} \cdot \{\cos(\beta \cdot L \cdot \cos\theta/2) - \cos(\beta \cdot L/2)\}/\sin\theta$$

where β is the phase constant $2\pi/\lambda$.

D.3.7 Electric versus magnetic field strength

In the far field the electric and magnetic field strengths are related by the impedance of free space, Z_o (377Ω):

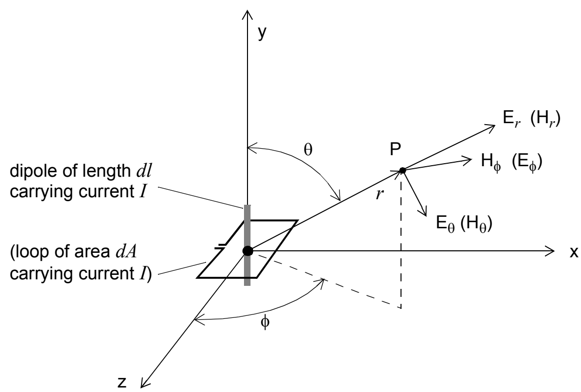
$$E(\text{dB}\mu\text{V/m}) = H(\text{dB}\mu\text{A/m}) + 51.5$$

H can be expressed in Amps per metre, Tesla or Gauss:

1 Gauss	=	100μT	=	79.5 A/m
1 A/m	=	$4\pi \cdot 10^{-7}$ T		

D.3.8 The field equations [8]

The following equations characterize the E and H fields at a point P due to an elementary electric dipole (current filament) and an elementary magnetic dipole (current loop). They use the spherical co-ordinate system shown below.



For the electric dipole:

$$E_r = Idl\cos\theta\left(\frac{\beta^3}{2\pi\omega\varepsilon_0}\right)\left(\frac{1}{(\beta r)^2} - \frac{j}{(\beta r)^3}\right)e^{-j\beta r}$$

$$E_\theta = Idl\sin\theta\left(\frac{\beta^3}{4\pi\omega\varepsilon_0}\right)\left(\frac{j}{(\beta r)} + \frac{1}{(\beta r)^2} - \frac{j}{(\beta r)^3}\right)e^{-j\beta r}$$

$$H_\phi = Idl\sin\theta\left(\frac{\beta^2}{4\pi\omega}\right)\left(\frac{j}{(\beta r)} + \frac{1}{(\beta r)^2}\right)e^{-j\beta r}$$

For the magnetic dipole:

$$H_\theta = IdA\sin\theta\left(\frac{\beta^3}{4\pi}\right)\left(\frac{-1}{(\beta r)} + \frac{j}{(\beta r)^2} + \frac{1}{(\beta r)^3}\right)e^{-j\beta r}$$

$$H_r = IdA\cos\theta\left(\frac{\beta^3}{2\pi}\right)\left(\frac{j}{(\beta r)^2} + \frac{1}{(\beta r)^3}\right)e^{-j\beta r}$$

$$E_\phi = IdA\sin\theta\left(\frac{\beta^4}{4\pi\omega\varepsilon_0}\right)\left(\frac{1}{(\beta r)} - \frac{j}{(\beta r)^2}\right)e^{-j\beta r}$$

In all the above,

β = the phase constant $2\pi/\lambda$
ω = the angular frequency of I in rad/s
ε_0 = the permittivity of free space (see D.3.1)
r, θ describe the co-ordinates of point P
E_r, E_θ, E_ϕ are the electric field vectors in V/m
H_r, H_θ, H_ϕ are the magnetic field vectors in A/m

These equations show that:

a) for $\beta r << 1$ (the near field) the higher order terms dominate with E varying as $1/r^3$ and H as $1/r^2$ for the electric dipole, and vice versa for the magnetic. The $1/r^2$ terms are known as the induction field.

b) for $\beta r >> 1$ (the far field) the radial term (E_r or H_r) becomes insignificant and the transverse terms (θ and ϕ) propagate as a plane wave, varying as $1/r$.

D.4 Shielding

D.4.1 Skin depth

$\delta = (\pi \cdot F \cdot \mu \cdot \sigma)^{-0.5}$ metres

For a conductor with permeability μ_r and conductivity σ_r, F in Hz:

$\delta = 0.0661 \cdot (F \cdot \mu_r \cdot \sigma_r)^{-0.5}$ metres
or $2.602 \cdot (F \cdot \mu_r \cdot \sigma_r)^{-0.5}$ inches

Typical skin depth for copper ($\mu_r = \sigma_r = 1$) is 66µm ($6.6.10^{-5}$ m) at 1MHz, 6.6µm at 100MHz

D.4.2 Reflection loss (R)

The magnitude of reflection loss depends on the ratio of barrier impedance to wave impedance, which in turn depends on its distance from the source and whether the field is electric or magnetic (in the near field) or whether it is a plane wave (in the far field). The following expressions are for F in Hz, r in metres and μ_r and σ_r as shown above.

$R = 168 - 10 \cdot \log_{10}((\mu_r/\sigma_r) \cdot F)$ dB   Plane wave

$R_E = 322 - 10 \cdot \log_{10}((\mu_r/\sigma_r) \cdot F^3 \cdot r^2)$ dB   Electric field

$R_H = 14.6 - 10 \cdot \log_{10}((\mu_r/\sigma_r)/F \cdot r^2)$ dB   Magnetic field

D.4.3 Absorption loss (A)

$A = 8.69 \cdot (t/\delta)$ dB
where t is barrier thickness, δ is skin depth

D.4.4 Re-reflection loss (B)

$B = 20 \cdot \log_{10}(1 - e^{-2\sqrt{2}(t/\delta)})$ dB

B is negligible unless material thickness t is less than the skin depth δ;
e.g. if $t = \delta$, $B = -0.53$dB; if $t = 2\delta$, $B = -0.03$dB
B is always a negative value, since multiple reflections degrade shielding effectiveness

D.4.5 Shielding effectiveness (see section 14.1.1)

Intrinsic shielding effectiveness of a homogeneous conducting barrier of infinite extent:

$$SE_{dB} = R_{dB} + A_{dB} + B_{dB}$$

Properties of typical conductors [14]

Material	Relative conductivity σ_r (copper = 1)†	Relative permeability @ 1kHz * μ_r
Silver	1.08	1
Copper	1.00	1
Gold	0.70	1
Chromium	0.66	1
Aluminium	0.61	1
Zinc	0.30	1
Tin	0.15	1
Nickel	0.22	50–60
Mild steel	0.10	300–600
Mu-metal	0.03	20,000

*: relative permeability approaches 1 above 1MHz for most materials
†: absolute conductivity of copper is $5.8 \cdot 10^7$ mhos

D.5 Capacitance, inductance and PCB layout

D.5.1 Capacitance

Capacitance between two plates of area A cm^2 spaced d cm apart in free space:

$$C = 0.0885 \cdot A/d \text{ pF}$$

The self-capacitance of a sphere of radius r cm:

$$C = 4\pi \cdot 0.0885 \cdot r = 1.1 \cdot r \text{ pF}$$

The capacitance per unit length between concentric circular cylinders of inner radius r_1, outer radius r_2 in free space:

$$C = 2\pi \cdot 0.0885 / \ln(r_2/r_1) \text{ pF/cm}$$

The capacitance per unit length between two conductors of diameter d spaced D apart in free space:

$$C = \pi \cdot 0.0885 / \cosh^{-1}(D/d) \text{ pF/cm}$$

The factor 0.0885 in each of the above equations is due to the permittivity of free space ε_0 (see D.3.1); multiply by the dielectric constant or relative permittivity ε_r for other materials.

Relative permittivities ε_r of some dielectrics

Air	1.0
PTFE (Teflon)	2.1
Polyethylene	2.3
Polystyrene	2.5
PVC, Polycarbonate	3.2
Polyimide	3.4
Epoxy glass	4.2–4.7
Glass (borosilicate)	5.0
Porcelain	5.5
Phenolic resin fabric	5.5
Alumina (pure)	8.5
Methanol @ 900MHz	31
De-ionised water	80

D.5.2 Inductance [6]

The inductance of a straight length of wire of length l and diameter d:

$$L = 0.0051 \cdot l \cdot (\ln(4l/d) - 0.75) \ \mu H \text{ for } l, \text{ d in inches, or}$$
$$0.002 \cdot l \cdot (\ln(4l/d) - 0.75) \ \mu H \text{ for } l, \text{ d in cm}$$

A useful rule of thumb is 20nH/inch.

The inductance of a return circuit of parallel round conductors of length l cm, diameter d and distance apart D, for $D/l << 1$:

$$L = 0.004 \cdot l \cdot (\ln(2D/d) - D/l + 0.25) \ \mu H$$

The mutual inductance between two parallel straight wires of length l cm and distance apart D, for $D/l << 1$:

$$M = 0.002 \cdot l \cdot (\ln(2l/D) - 1 + D/l) \ \mu H$$

The mutual inductance between two conductors spaced D apart at height h over a ground plane carrying its return current [98]:

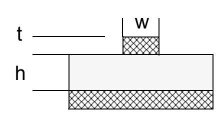
$$M = 0.001 \cdot \ln(1 + (2h/D)^2) \ \mu H/cm$$

The inductance of a single wire of diameter d at height h over a ground plane carrying its return current [98]

$$L = 0.002 \cdot \ln(4h/d) \ \mu H/cm$$

D.5.3 PCB track propagation delay and characteristic impedance [100]

<u>Surface microstrip</u>



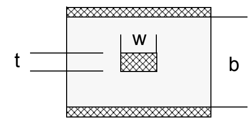
$$T_{pd} = 1.017 \cdot \sqrt{(0.475 \cdot \varepsilon_r + 0.67)} \ \text{ns/ft}$$

$$Z_0 = (87/\sqrt{(\varepsilon_r + 1.41)}) \cdot \ln[5.98h/(0.8w + t)] \ \Omega$$

e.g. for h = 1.6mm, w = 0.3mm, ε_r = 4.2 and t << w in surface microstrip, $Z_0 = 130\Omega$

NB [42] shows that this equation (which is widely referenced) is inaccurate compared to numerical model results for lower values of Z_0, i.e. large values of w/h, and gives a more accurate but more complex equation derived from Wadell [137]

Embedded stripline



T_{pd} = $1.017 \cdot \sqrt{\varepsilon_r}$ ns/ft

Z_0 = $(60/\sqrt{\varepsilon_r}) \cdot \ln[4b/(0.67\pi \cdot (0.8w + t))]\ \Omega$

For FR4 epoxy fibreglass PCB material at high frequencies ε_r is typically 4.2 which gives a propagation delay T_{pd} of 1.7ns/ft (56ps/cm) for surface microstrip and 2.1ns/ft (69ps/cm) for embedded stripline.

e.g. for b = 1.6mm, w = 0.3mm, $\varepsilon_r = 4.2$ and t << w in embedded stripline, Z_0 is 74Ω.

When a track is loaded with devices, their capacitances modify the track's propagation delay and Z_0 as follows:

T_{pd}' = $T_{pd} \cdot \sqrt{(1 + C_D/C_o)}$

Z_0' = $Z_0/\sqrt{(1 + C_D/C_o)}$

where C_D is the distributed device capacitance per unit length, i.e. the total load capacitance divided by the track length, and C_o is the intrinsic capacitance of the track calculated from:

C_o = $1000 \cdot (T_{pd}/Z_0)$ pF/length

D.5.4 Distributed coupling [19]

For a two-wire transmission line coupled to a plane wave electromagnetic field:

s = line length in m, b = vertical wire separation in m, a = wire diameter in m

Z_1 = source end impedance in ohms, Z_2 = load end impedance in ohms, Z_0 is line characteristic impedance, derived from geometry: $Z_0 = 276 \cdot \log(2b/a)$

β is phase constant = 2π/λ

D is "denominator function":

$$D = (Z_0 \cdot Z_1 + Z_0 \cdot Z_2) \cdot \cosh(j \cdot \beta \cdot s) + (Z_0^2 + Z_1 \cdot Z_2) \cdot \sinh(j \cdot \beta \cdot s)$$

There can be three equations for different wave conditions (variable X is the coupling factor, V/E, for the load end):

(a) E vertical, travelling towards line

$$X = Z_2 \cdot b/D \cdot (Z_0 \cdot (1 - \cos\beta s) + Z_1 \cdot j \cdot \sin\beta s)$$

(b) E vertical, travelling along line

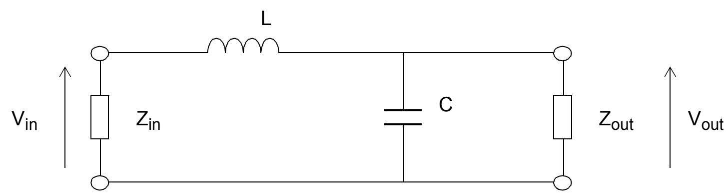
$$X = Z_2 \cdot \frac{b \cdot (Z_0 - Z_1)}{2D} \cdot ((1 - \cos 2\beta s) + j \cdot \sin 2\beta s)$$

(c) E horizontal, travelling towards line

$$X = Z_2 \cdot \left(\frac{-2j \cdot \sin\beta\frac{b}{2}}{D \cdot \beta}\right) \cdot (Z_0 \cdot \sin\beta s + Z_1 \cdot j \cdot (1 - \cos\beta s))$$

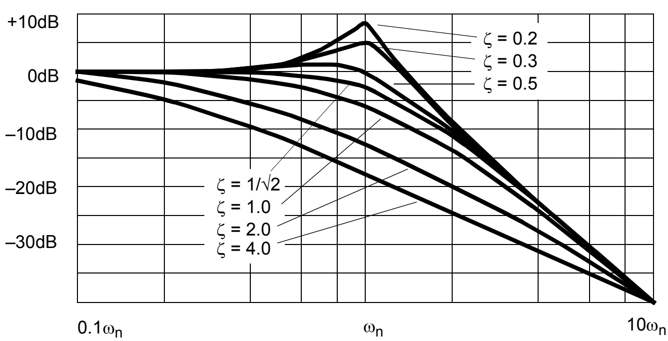
D.6 Filters

D.6.1 Second order low-pass filter [124]



$$\frac{V_{out}}{V_{in}} = \frac{1}{1 + 2j\zeta(\omega/\omega_n) - (\omega/\omega_n)^2} \qquad \omega_n = 1/(\sqrt{LC})$$

$$\zeta = \frac{L}{2Z_{out}\sqrt{LC}}$$



The damping factor ζ describes both the insertion loss at the corner frequency and the frequency response of the filter. ζ is affected by the load impedance and low values may cause insertion gain around the corner frequency.

The following design procedure may be applied to any low-pass LC filter and to the typical mains filter configuration (see Figure 13.27 on page 369) to design both the differential components and the common mode components, remembering that the latter are symmetrical about earth and can therefore be treated as two separate circuits.

1. Identify the required cut-off (corner) frequency ω_n: the second order filter rolls off at 40dB/decade, so the desired attenuation A_{dB} at some higher frequency F will put the corner frequency at:

   $\omega_n = 2\pi F / \log^{-1} (A/40)$

2. Identify the load resistance Z_{out} and desired damping factor ζ. A value for ζ between 0.7 and 1 will normally be adequate if Z_{out} is reasonably well specified. Values much larger than 1 will cause excessive low frequency attenuation while much less than 0.7 will cause ringing and insertion gain.

3. From these calculate the required component values:

$L = 2 \cdot Z_{out} \cdot \zeta / \omega_n$

$C = 1/(L \cdot \omega_n^2)$

4. Iterate as required to obtain useable standard component values.

D.6.2 Filter insertion loss vs. impedance

Standard filters are nearly always characterized between 50Ω resistive impedances. This is unlikely to match the actual circuit impedance. However, if you know the actual circuit impedances and they are also resistive, you can calculate the expected insertion loss from the published 50Ω value. First, derive the transfer impedance Z_T of the filter:

$Z_T = 25/\{\text{antilog}(IL_{dB}/20) - 1\}$ where IL_{dB} is the 50Ω published insertion loss

Now the insertion loss between other resistive impedances Z_S (source) and Z_L (load) is:

$IL_{dB} = 20 \log \{1 + (Z_S \cdot Z_L)/(Z_T \cdot (Z_S + Z_L))\}$

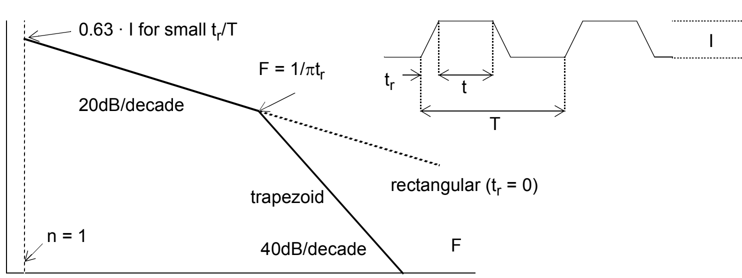
Reactive source or load impedances modify the performance and this equation is no longer applicable.

D.7 Fourier series

For a symmetrical trapezoidal wave of rise time t_r, period T and peak-to-peak amplitude I, the harmonic current at harmonic number n is:

$$I(n) = 2I((t+t_r)/T)\left(\frac{\sin n\pi((t+t_r)/T)}{n\pi((t+t_r)/T)}\right)\left(\frac{\sin n\pi(t_r/T)}{n\pi(t_r/T)}\right)$$

This gives the envelope shown below.



Straight line envelope of harmonic amplitudes

The general form [2] of the Fourier Series is:

$$f(t) = 0.5A_0 + \sum_{n=1}^{\infty} (A_n \cos\omega_n t + B_n \sin\omega_n t)$$

where the coefficients A_n and B_n are:

$$A_n = \frac{2}{T}\int_{-T/2}^{T/2} f(t)\cos\omega_n t dt$$

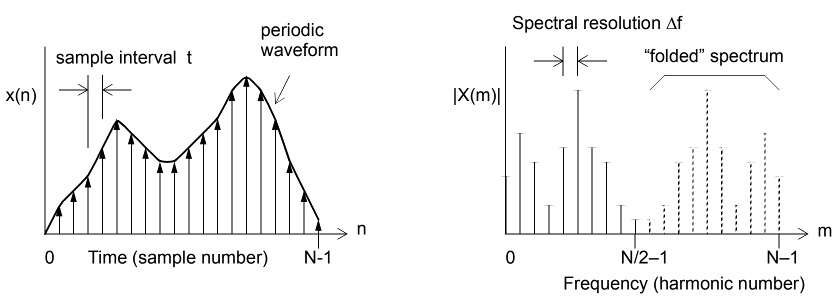
$$B_n = \frac{2}{T}\int_{-T/2}^{T/2} f(t)\sin\omega_n t dt$$

Any arbitrary waveform can be analysed by sampling it at discrete time intervals and taking the discrete Fourier transform (DFT) [4]. This is achieved by replacing the integrals above by a finite weighted summation:

$$A(m) = \frac{1}{N}\sum_{n=0}^{N-1} x(n)\cos 2\pi m\left(\frac{n}{N}\right)$$

$$B(m) = \frac{1}{N}\sum_{n=0}^{N-1} x(n)\sin 2\pi m\left(\frac{n}{N}\right)$$

Here the time axis is given by (n/N) where N is the total number of samples and x(n) is the sample value at the nth sample. m represents the frequency axis and the DFT calculates A(m) and B(m) for each discrete frequency from m = 0 (DC) through to m = (N/2) – 1.



The spectral resolution Δf in the frequency domain is the reciprocal of the total sample time, 1/Nt, which is equivalent to the period of the time domain waveform. m = 1 represents the fundamental frequency, m = 2 the second harmonic, and so forth. The spectrum image is folded about m = N/2 and therefore the maximum harmonic frequency that can be analysed is half the number of samples times the fundamental, or 1/2t.

For EMC work A(0) and B(0) are normally neglected (they represent the DC component) as also is phase information, represented by the phase angle between the real and imaginary components arg(A(m) + jB(m)). The mean square amplitude:

$$|X(m)|^2 = |A(m)|^2 + |B(m)|^2$$

represents the power in the mth harmonic. A simple program to calculate $X(m)$ for an array of samples $x(n)$ in the time domain is easily written.

Appendix E

The EU and EEA countries

The European Union

Established as the 'European Communities' by the Treaty of Rome, 1957.
* = founder members

Country	Date Joined
Austria	1995
Belgium	*
Cyprus	2004
Czech Republic	2004
Denmark	1973
Estonia	2004
Finland	1995
France	*
Germany	*
Greece	1981
Hungary	2004
Ireland	1973
Italy	*
Latvia	2004
Lithuania	2004
Luxembourg	*
Malta	2004
The Netherlands	*
Poland	2004
Portugal	1986
Slovenia	2004
Slovakia	2004
Spain	1986
Sweden	1995
United Kingdom	1973

The European Economic Area (EEA) includes the EU Member States as above, plus Norway, Iceland and Liechtenstein.

CENELEC includes all the above, plus Switzerland.

Two more countries – Bulgaria and Romania – are expected to join the EU in 2007. Two others, Croatia and Turkey, are also beginning negotiations for EU membership.

Glossary

ACB	Association of Competent Bodies (Europe)
ADSL	Asymmetrical Digital Subscriber Line
AE	Associated (ancillary) equipment
ALSE	Absorber Lined Shielded Enclosure
AM	Amplitude modulation
AMN	Artificial Mains Network
ANSI	American National Standards Institute
ASIC	Application Specific Integrated Circuit
BBC	British Broadcasting Corporation
BCI	Bulk current injection
CAD	Computer-aided design
CB	Citizen's band
CBN	Common bonded network
CCTV	Closed circuit television
CDN	Coupling/decoupling network
CEN	European Committee for Standardization
CENELEC	European Committee for Electrotechnical Standardization
CISPR	International Special Committee for Radio Interference
CMAD	Common mode absorbing device
CMR	Common mode rejection
CW	Carrier wave (unmodulated continuous RF)
DIS	Draft International Standard
DTI	Department of Trade and Industry
E field	Electric field
EC	European Commission
EEA	European Economic Area
EED	Electro-explosive device
EEPROM	Electrically erasable programmable read-only memory
EFT/B	Electrical Fast Transient/Burst
EMC	Electromagnetic compatibility
EMCTLA	EMC Test Laboratories Association (UK)
EMF	Electromagnetic Fields
EMI	Electromagnetic interference
EMP	Electromagnetic pulse
EMR	Electromagnetic radiation
EOTC	European Organization for Testing and Certification
ESA	Electronic sub assembly
ESD	Electrostatic discharge

ETSI	European Telecommunications Standards Institute
EU	European Union
EUT	Equipment under test
FAR	Fully Anechoic Room
FCC	Federal Communications Commission (US)
FEXT	Far-end crosstalk
GSM	Groupe Special Mobile
GTEM	Gigahertz TEM cell
H field	Magnetic field
HF	High frequency
HV	High voltage
I/O	Input/output
ICNIRP	International Commission on Non-Ionizing Radiation Protection
IEC	International Electrotechnical Commission
IEEE-488	IEEE standard for data communications between test instruments
IF	Intermediate frequency
ISDN	Integrated Services Digital Network
ISM	Industrial, scientific and medical
ISN	Impedance stabilization network
ISO	International Standards Organization
ITE	Information technology equipment
ITO	Indium Tin Oxide
ITU	International Telecommunications Union
LAN	Local Area Network
LCL	Longitudinal Conversion Loss
LEMP	Lightning Electromagnetic Pulse
LF	Low frequency
LISN	Line impedance stabilizing network
LLA	Large Loop Antenna
LP	Log periodic
LPS	Lightning protection system
LRU	Line replaceable unit
LV	Low voltage
MESH-BN	Mesh bonded network
MF	Medium frequency
MS	Mains signalling
MV	Medium voltage
NEMP	Nuclear electromagnetic pulse
NEXT	Near-end crosstalk
NPL	National Physical Laboratory
NSA	Normalized Site Attenuation (also NSIL, Normalized Site Insertion Loss)
NVRAM	Non Volatile Random Access Memory
OATS	Open area test site
OJEC/OJEU	Official Journal of the European Communities/Union
PC	Personal computer

PCB	Printed circuit board
PEC	Parallel earth conductor
PFC	Power factor correction
PLC	Programmable logic controller
PRF	Pulse repetition frequency
PSU	Power supply unit
QP	Quasi-peak
RAM	Random access memory (also RF absorbent material)
RF	Radio frequency
RFI	Radio frequency interference
RMS	Root mean square
ROM	Read only memory
RTCA	Radio Technical Commission of America
SAE	Society of Automotive Engineers
SI	Statutory Instrument (also Système Internationale)
SMPS	Switched mode power supply
SMT	Surface mount technology
SPD	Surge protection device
STI	Surface transfer impedance
TCB	Telecommunications Certification Body (US)
TEM	Transverse electromagnetic mode
TETRA	Trans European Trunked Radio Access
THD	Total harmonic distortion
TTE	Telecommunications terminal equipment
UART	Universal asynchronous receiver/transmitter
UKAS	United Kingdom Accreditation Service
VDE	Verband Deutscher Elektrotechniker (Association of German Electrical Engineers)
VDR	Voltage dependent resistor
VDSL	Very high speed digital subscriber line
VDU	Visual display unit
VLSI	Very large scale integration
VSWR	Voltage standing wave ratio
ZnO	Zinc Oxide (varistor)

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