Cherry Clough
Consultants

The Physical Basis of EMC

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Contents

1 Introduction

1.1 The aim of this article

My aim in writing this, is to try to impart an understanding of electromagnetic phenomena in a way that can be easily understood by practising engineers.

Armed with these few basic principles, electromagnetic compatibility (EMC) issues are easily *visualised*, and problems more easily solved, using only very simple mathematics and plain English.

Usually, articles like this start off by expounding Maxwell's famous equations, sometimes even deriving them from first principles. These are then used to solve a few simple constructions, such as an infinitely thin infinitely long perfectly-conducting wire suspended over an infinite plain of perfectly conducting material. Having to this far, they then usually abandon the reader, as if everything that could be said, had been said.

Unfortunately, no products are made using infinitely long wires or infinite planes. One reason is that they would be impossible to ship to customers – there would not be enough cardboard in the world to package them!

Another difficulty is that most modern products contain a great deal more than one conductor – typically they will have several million semiconductors (maybe hundreds of millions) and other linear and non-linear components interconnected by an equal number of wires (many of which will be metallised silicon).

So what use are Maxwell's equations to the product designer?

It turns out that they are of no *direct* use at all! (Do I hear thousands of academics howling in anguish? Sorry!)

However, they *are* used to create computer simulations that *are* of practical use; plus they underpin the basic physical principles that this article aims to communicate.

These principles are of great practical utility – I have been successfully using them for over 20 years in a huge range of applications, from µW to MW, and many other EMC specialists have been using them for even longer. But you won't often find them clearly described in textbooks!

I learnt Maxwell's equations from renowned electromagnetic experts in my final year at University in London, UK, and passed the exam well enough. But when I met my first interference problem as a designer, I eagerly pulled out my lecture notes only to find that there was not even one word in them that appeared to bear any kind of relationship at all to the practical problem I had to solve.

I never opened those notes again, and after the mid-1970s I couldn't even if I wanted to – because they had been lost in Beirut, Lebanon, during the civil war it was experiencing at that time.

Low-frequency EMC issues (say, below 100kHz) are relatively easy to understand using normal circuit design techniques, so this article will focus mostly on the physical basics of EMC at radio frequencies (RF), which most electronic circuit designers find difficult because of the apparently weird, and sometimes counter-intuitive things that can happen at frequencies above a few MHz.

1.2 Who will benefit from understanding these principles?

This article is not about the good EMC design techniques I normally write about, but instead about *the reasons why these techniques work so well*.

These same reasons underpin all EM phenomena – which means all signals and power too – so they are important for achieving signal integrity, power integrity, and for determining good EMC engineering techniques when designing, constructing and installing electrical and electronic devices, products, equipment, systems and installations of all types, and of any size from vanishingly small to continent-spanning.

1.3 A brief description of some basic EM principles

I make no apologies for mangling the academic approach to EMC in what follows. I am trying to use common English terms familiar to practising engineers who – if they ever learnt Maxwell's equations – have forgotten everything about them.

Understanding the following eight basic principals make it easy to *visualise* EM phenomena, without using any maths, in any situation. There are many other principles that will be described later on, but these are the ones that really help visualise EMC problems and their solutions.

i) Everything that we think of as being an AC voltage or current is really EM energy (Watts, Joules) propagating as a wave.

It makes no matter how small or large are the voltages, currents or powers, or what the application is.

If we have 1mW of power in a signal, or 10kW of power in a mains-powered load, then there is 1mW and 10kW of EM energy, respectively, in the EM fields associated with their send and return conductors.

Waves propagate in the three dimensions of space, and the one of time, and have different amplitudes at different places and different times.

The distribution of EM wave energy in space is called an EM field. It is rather like dropping a stone into calm pool of water causes waves to spread out over the water, leaving behind a 'field pattern' of ripples.

The way most of us were taught about electric current, as if it was caused by electrons rushing backwards and forwards along a wire, is no help at all at RF frequencies, and leads to mistakes in design – such as imagining that single-point earthing/grounding could possibly control where return currents flow at RF.

Of course, electrical currents do involve movement of electrons, but their velocity is about walking speed, about 3 miles per hour – rather less than the velocity of electromagnetic wave propagation at the speed of light!

ii) EM waves and fields consist of both Electric (E) and Magnetic (H) waves and fields.

This is exactly why they are called "electro-magnetic"!

E waves and fields are measured in Volts per metre, H measured in Amps per metre.

iii) When a conductor is exposed to an EM wave, its free electrons move around in response to the wave – generating what we call a current.

When a circuit's voltages and currents move the electrons in its conductor around, they cause the same EM waves to arise as would have made the electrons move in that same way.

This is called the ***Principle of Reciprocity***, and it means that a conductive structure (e.g. a cable, printed circuit board (PCB) trace, etc.) that has a certain antenna behaviour when it 'picks up' an EM wave and gives rise to currents and voltages in a circuit, has exactly the same antenna behaviour as an emitter of EM waves – when currents and voltages are made to occur in it by some circuit.

This understanding also shows that "conducted" and "radiated" EM phenomena are just 'two sides of the same coin' – they are not, in fact, different types of phenomena.

We classify EM phenomena as either conducted or radiated purely on the basis of whether we use conducted or radiated tests to measure them. All that a propagating EM wave knows is the impedance and velocity of the medium it is travelling through, both of which are affected by the presence of conductors.

Note that it is impossible to have electrons moving without there being an associated wave propagating in the nearby space. This wave is not confined to the locality of the conductors carrying the electrons, and how it spreads through space is defined by the relationship between its send and return current paths – which brings us to point iv).

iv) EM wave propagation, and its associated fields, are shaped by the "accidental antenna" structures that we think of as send and return current paths and the dielectric materials (insulators) that surround them.

And, as we learned in iii) above, by the principle of reciprocity, the voltages and currents that are picked-up by conductors from their ambient fields depend upon the shape of the accidental antenna created by that circuit's send and return current paths and their insulators.

To reduce emissions and increase immunity, we design using conductive structures that behave as poor accidental antennas.

Many low-cost (often free) EMC and signal integrity design techniques are based on this approach.

v) Induced and radiated coupling

If we consider the EM wave associated with voltages and currents in a given arrangement of send/return conductors, for example an individual part of an

electronic circuit, we see that any other conductors that this wave meets as it spreads through space will also experience currents and voltages as a result.

This fact of wave (field) coupling is used to design intentional EM couplers, for example transformers, and radio systems (i.e. transmitters and receivers).

It also explains some of the ways in which crosstalk, differential-mode and common-mode interference arise. A proportion of the EM energy associated with the currents and voltages in one circuit (i.e. arrangement of conductors) has coupled through the air (or other insulators) into a quite separate set of conductors.

The energy lost by the original circuit, the "source", has an effect on its voltages and currents. The energy picked-up by the other circuit, the "victim", is a noise – an unwanted signal.

We are all familiar with the idea that electrical energy (Watts, Joules) is carried by conductors. But in fact conductors only guide electrical (EM wave) energy, which can also flow equally well through insulators like vacuum, air, plastic, etc.

vi) Electrons are naturally forced to flow near the surface of a conductor.

The higher the frequency, the smaller is the thickness below the surface of the conductor, in which they flow.

This is called the ***Skin Effect***, so these currents are called ***Surface Currents***.

Skin effect reduces the proportion of an AC current that can flow *directly through* a conductive material (i.e. from one side to the other).

So, for example, metal sheets are poor conductors in their *thickness* dimension. The higher the frequency – the poorer they are. Skin effect shows us how to assemble shielding and filtering so that they work best, see 2.6.2, 3.3.2 and 4.4.3 of [1].

vii) Return currents automatically take the path (or paths) that minimise the overall amount of EM energy.

This is rather like the way that a falling drop of water automatically assumes a spherical shape to minimise the energy in its surface tension.

It means that all we have to do to make a poor accidental antenna (see iv) above), is to provide a path for the return current that is very close physically to the send current's path – much closer than the wavelength at the frequency concerned.

To make the antenna effect poorer (to further reduce emissions and improve immunity) we make the return path closer to the send path.

Note that we don't have to *make* the return current flow in the nearby path we have created for it – all we do is create the path and the current <u>*naturally prefers to take it*</u>, even when parallel paths exist that it could take instead.

The resulting EM field patterns become much more compact, have less energy in them, and couple less with other conductors. The accidental antenna behaviour of the conductive structure is therefore less efficient – reducing emissions and improving immunity.

This is a lovely example of how the laws of nature (or laws of physics, if you prefer) actually try to help designers control EM interference (EMI) and achieve EMC.

This understanding makes it possible to visualise where a return current flows, and see how its path can be altered to improve EMC.

viii) Everything presents an impedance to a wave

Every conductor has impedance, so wherever and whenever there is a current in it, there is always a corresponding voltage. And vice-versa.

This applies equally to superconductors, as they only have a zero resistance (just one of the three constituents of impedance: resistance, inductance and capacitance). They still have inductance and capacitance.

Every insulator or dielectric (vacuum, air, plastic, fibreglass, ceramic, etc.) has a

"wave impedance", so that whenever and whenever an H-field exists in it, it always experiences a corresponding E-field. And vice-versa.

These eight basic EM principles arise as a direct result of quantum mechanics and quantum electro-dynamics – the laws of nature that give rise to Maxwell's equations.

I learnt this when I bumped into my old electromagnetic lecturer at a conference about 20 years after graduating, and asked him why Maxwell's equations were like they were. (I felt like a 3-year old asking "But *why*, Daddy, *why*?")

He simply pointed me to textbooks on quantum electrodynamics written by Richard Feynman, who discovered how to use quantum mechanics to calculate exactly how EM fields interact with the free electrons in metals and other conductors.

So forget kindergarten ideas of electrical currents and voltages, e.g. that electrical current consists of electrons bobbling along, and that waves and fields are side-effects, and use the above principles instead. They make it possible to easily visualise any/all linear EM phenomena. (I'll deal with the non-linear EM phenomena, which occur in semiconductors and corroded metal joints, later on.)

The remaining few thousand words and all of the graphics in this article go into more detail on the above principles, and also discuss useful associated issues.

2 Wave and Field theory

Design for EMC at RF is mostly about controlling the shapes of E and H-fields so that they are at their most intense where we want power or signals to occur in conductors, and are weak elsewhere so that emissions are low and – because of reciprocity – the conductors do not pick up much noise from other EM fields.

This sounds very difficult, and might be assumed to use a great deal of mathematics, but in fact all we need to help us design are the basic principles mentioned in 1.3, a few other principles, and some simple sums.

2.1 E and H-fields

Fluctuating voltages, whose magnitudes are measured in Volts (V), are associated with correspondingly fluctuating Electric fields (E), whose magnitudes are measured in Volts/metre (V/m).

Fluctuating currents, whose magnitudes are measured in Amps (A), are associated with correspondingly fluctuating Magnetic fields (H), whose magnitudes are measured in Amps/metre (A/m).

2.2 Wavelength, velocity and frequency

A wave has – of course – different amplitudes at different points along its path, which fluctuate as time progresses. Figure 1 shows a way of visualising an EM wave that is propagating along a particular vector, such as along a long thin wire. However, a wave travelling through the air will propagate over a solid angle, not along a single vector like that shown in Figure 1.

P (the EM energy), E and H are all orthogonal to each other – indicated in Figure 1 by drawing them along the mutually perpendicular x, y and z axes.

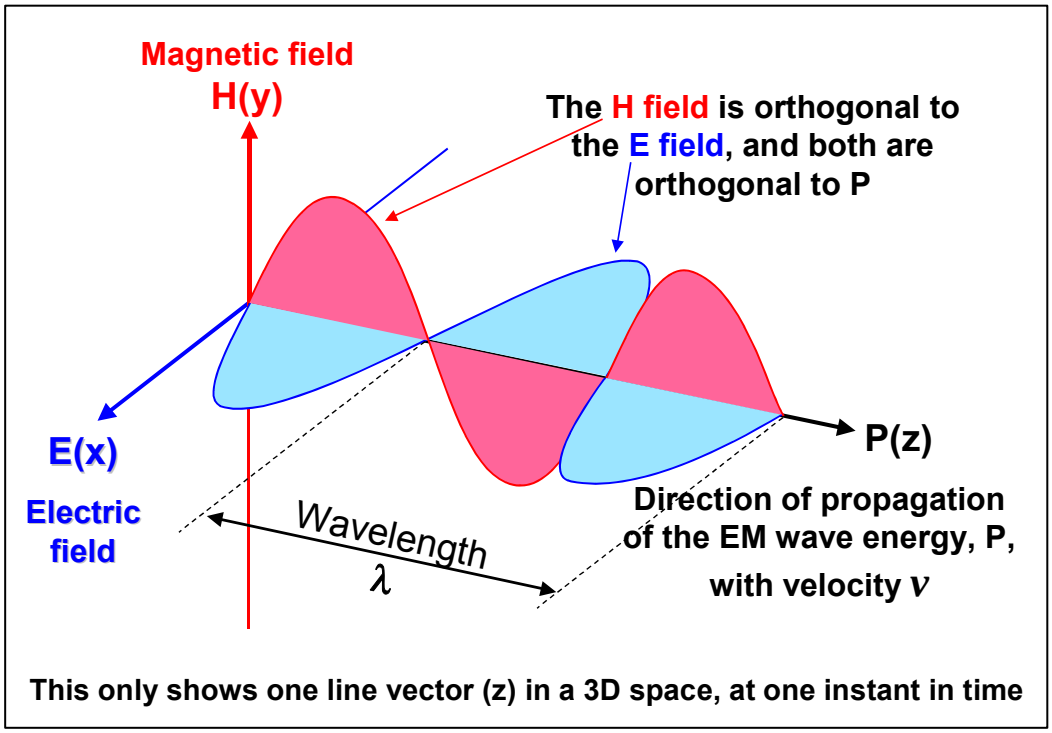


Figure 1 The traditional way of representing an EM wave

Figure 2 shows how a common EM simulator for PCBs represents the EM fields associated with a differential pair of traces on a PCB.

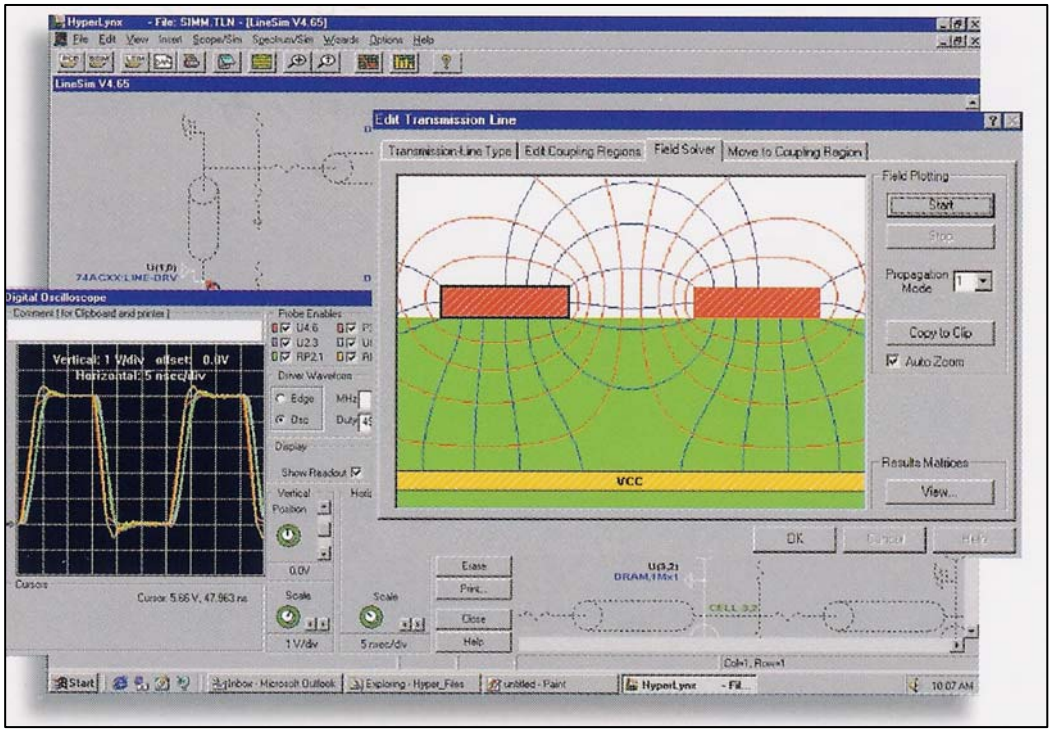


Figure 2 Example of an EM PCB simulator (Hyperlynx) showing plots of E and H-fields

Because EM energy (signals, power, etc.) travel as waves, when a conductor is long enough it cannot experience the same voltage or current, at any instant in time, along its whole length. This “wave effect” is what causes the apparently weird effects that can occur at radio frequencies, or with fast digital signal edges, and makes matched transmission line techniques necessary for good signal integrity and low-cost EMC, see 2.7 and 5.6 of [1].

The ratio between the wavelength (λ) of the frequency (f) we are concerned with, and the dimensions of our conductors, is therefore very important indeed. For the EMC requirements of ordinary commercial or industrial designs we can usually ignore weird "wave effects" when conductor dimensions are < 1/100th of the λ (equivalent to approximately 3/f metres when f is given in MHz. GHz gives the result in millimetres).

For example, at 1GHz this means < 3mm for EM waves on conductors surrounded by vacuum or air; or < 1.5mm for waves propagating on traces embedded inside PCBs made of the common FR4 dielectric. Few if any conductors outside of the silicon metallisation within Integrated Circuits (ICs) are as short as this, and able to have their 'wave effects' ignored during design.

This crude guide will probably not be adequate where EMC requirements are higher, or the EM environment is harsher.

We easily convert between λ into f by using the relationship: $v = f \cdot \lambda$
– where v is the velocity of the wave's propagation in metres/sec, f is in Hz and λ is in metres.

But to understand v, we must first understand permeability, μ, and permittivity, ε.

2.3 Permeability (μ) and permittivity(ε)

All media or materials have the characteristics of conductivity (resistivity), permeability and permittivity. The characteristics of the vacuum (and air) are given the suffix 0, and so:

$\mu_0 = 4\pi \cdot 10^{-7}$ Henries/metre

$\varepsilon_0 = 1/(36\pi) \cdot 10^{-9}$ Farads/metre

Other media and materials are characterised by their *relative* permeability and permittivity, which is simply a numerical multiplier given the suffix R (for relative), and so their absolute permeability is: $\mu_0 \cdot \mu_R$, and their absolute permittivity is: $\varepsilon_0 \cdot \varepsilon_R$.

Conductivity (resistivity) is not associated with any fields, and is simply a characteristic that converts EM energy into heat. Resistance turns Watts and Joules of electrical power and energy into Watts and Watt-seconds of heat.

2.3.1 Impedance (Z)

In conductors: μ and ε cause inductance (L) and capacitance (C) to arise, respectively. L creates an impedance of $2\pi f$ L, and C creates an impedance of $1/2\pi f$ C, and the overall impedance of the conductor, Z, is $\sqrt{(L/C)}$, in Ω.

Since this means that L and C are always present, creating impedance in any conductor no matter how small, it means that <u>*whenever*</u> there is a fluctuating voltage in a conductor there is <u>*always*</u> an associated fluctuating current. And vice-versa.

Some digital designers have been known to believe that because the gates of CMOS integrated circuits have an almost infinite resistance, therefore digital signals have no send or return currents. But of course the fluctuating voltages that are the signal have to charge up the capacitance of the PCB trace and of the gate of the CMOS device, so need a correspondingly fluctuating current.

Since current always flows in loops (another law of nature, known as Kirchoff's Law), and since there is a send current in the trace and CMOS gate, there has to be an equal and opposite return current.

The practical meaning of this for signal integrity is that if the return current is impeded, for instance by having to flow around a slot or gap in a 0V plane, or having to flow in a path that encloses a large area (with respect to the send path), the inductance that is thereby added develops a voltage that distorts the signal's voltage.

And, of course, not having a return path that is close to the send path at all times, means that the circuit behaves as a more efficient accidental antenna, so the digital signal has excessive emissions and is more likely to suffer interference.

So a lack of appreciation of μ and ε and the inevitable impedances that they give rise to, causes some digital designers' PCB layouts to suffer from signal integrity and EMC problems.

In insulators (dielectrics): μ and ε cause effects similar to inductance and capacitance, meaning that *whenever* there is a fluctuating E-field there is *always* an associated fluctuating H-field. And vice-versa. As mentioned earlier, this is why we call the subject electromagnetism.

The impedance of a wave in the "far field" (see later) is given by the ratio of its E and H-fields:

$$Z_{WAVE} = E/H = V/m \div A/m = \sqrt{(\mu_0 \cdot \mu_R / \varepsilon_0 \cdot \varepsilon_R)} \quad \Omega$$

In air or vacuum, when μ_R and ε_R are both 1:

$$Z_{WAVE} = 120\pi \ \Omega, \text{ say } 377\Omega$$

But in a medium other than air or vacuum, so μ_R and/or ε_R are greater than 1:

$$Z_{WAVE} = 120\pi(\mu_R/\varepsilon_R)\ \Omega$$

Notice that the units of wave impedance are simply ohms, Ω, and in fact the Z_{WAVE} = E/H formula is sometimes called "Ohms Law for Fields".

2.3.2 Velocity of EM wave propagation

μ and ε also govern the velocity of propagation, v:

$$v = 1/\sqrt{(\mu_0 \cdot \mu_R \cdot \varepsilon_0 \cdot \varepsilon_R)} \quad \text{metres/second}$$

In air or vacuum, when μ_R and ε_R are both 1:

$$v = 3 \cdot 10^8 \text{ m/s}$$

It is actually a little less than the above figure of 300 million metres/second, and we assume it is equivalent to 3ns/metre, or 3ps/millimetre.

But in a medium other than air or vacuum, so μ_R and/or ε_R are greater than 1:

$$v = 3.10^8/\sqrt{(\mu_R \cdot \varepsilon_R)} \text{ m/s}$$

So when μ_R and/or ε_R are greater than 1, v is slower so the wavelength (λ) at a given f is shorter. For example, for a PCB's FR4 dielectric, μ_R is 1 but ε_R is 4.2, so for a wave propagating along a trace on an inner layer, its v is approximately half of what it would be in air, so its λ is about half too.

For example, a 1GHz wave in the air has a λ of about 300mm, but if travelling instead on a trace on an inner layer of a PCB, λ is about 150mm.

2.3.3 The effect of changing the impedance along the path of a wave

EM waves propagating through space, or along conductors, are reflected when they experience a change in impedance. Examples include a wave propagating through the air and passing into a block wood, plastic, ceramic, etc.; or a trace on a PCB that increases its width (hence has higher C and lower L, therefore a lower impedance).

To have a significant effect on a wave, the change in impedance has to persist for longer than one-tenth of a wavelength, λ/10, so to analyse the path of a signal on in a trace in a PCB to make sure it is experiencing the same Z, called its "characteristic impedance", Z_0, all along its path to maintain good signal integrity, we have to divide the trace's length into segments that are no longer than λ/10, and make sure that each segment has the same value of Z_0.

This is known as transmission line design, and when the impedances of the circuit's source and load equal the Z_0 of the trace, it is called a "matched transmission line". Such lines are very poor accidental antennas and so have very low emissions and pick-up very little noise from ambient EM fields.

This phenomenon of wave reflections where there are changes in impedance, also tells us how to maximise reflections to design shields and filters for greatest attenuation, see 3.2.1

and 4.3.2 in [1].

2.4 Near-field and Far-field

Near to a circuit that has fluctuating voltages or currents, a "source", the corresponding E and H-fields have complex patterns in space: their field strengths vary as a function of $1/r^3$, $1/r^2$ and $1/r$, where r is the radial distance from the source. They are called "near fields", because they have not yet travelled far enough to settle down into the E/H ratio that matches the wave impedance of their medium (usually air).

This is sometimes called the "induction region", because the effects of the fields can be described in terms of stray capacitance and stray mutual inductance effects (i.e. coupling that is dominated by E and H-fields).

Figure 3 shows an example of the near-fields around a heatsink (which experiences fluctuating voltages and currents as the result of having significant levels of stray capacitance to the conductors in the device it is cooling). These were simulated using Microstripes – a computer EM field simulator that used to be called Flo-EMC, and is now owned by Computer Simulation Technology (www.cst.com).

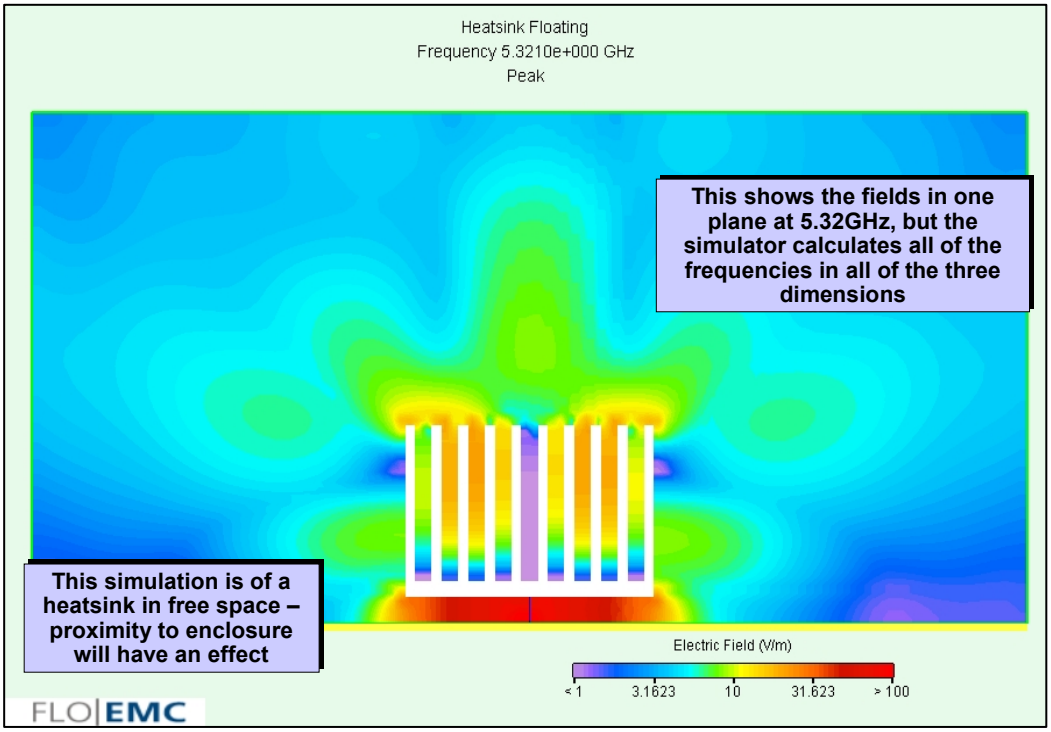


Figure 3 Near fields around a heatsink

When we get far enough away from a source, its propagating E and H waves turn into EM waves, which means that they have E and H-fields in the ratio of the wave impedance of the medium: Z_{WAVE}.

This is now the "far field" region, and the distribution of the fields in space follows a simple "plane wave" with the wave energy spreading in space as a simple expanding sphere. Now, its field strengths vary as a function of only 1/r.

For sources with longest dimensions $<<\lambda$, the boundary between the near and far field regions occurs when r (the distance from the source) is $\lambda/2\pi$.

But for sources with dimensions $>\lambda$, the near/far field boundary is calculated as $r = 2D^2/\lambda$, where D is the largest dimension of the source.

(There is a third expression for when the source's longest dimension is inbetween these two size ranges, but it only gives an value for r that is a little larger than the largest value calculated using both of the above.)

Figure 4 graphs the near-field / far-field boundary for a product that has a longest dimension shorter than one-sixth of a wavelength over the range of frequencies of interest.

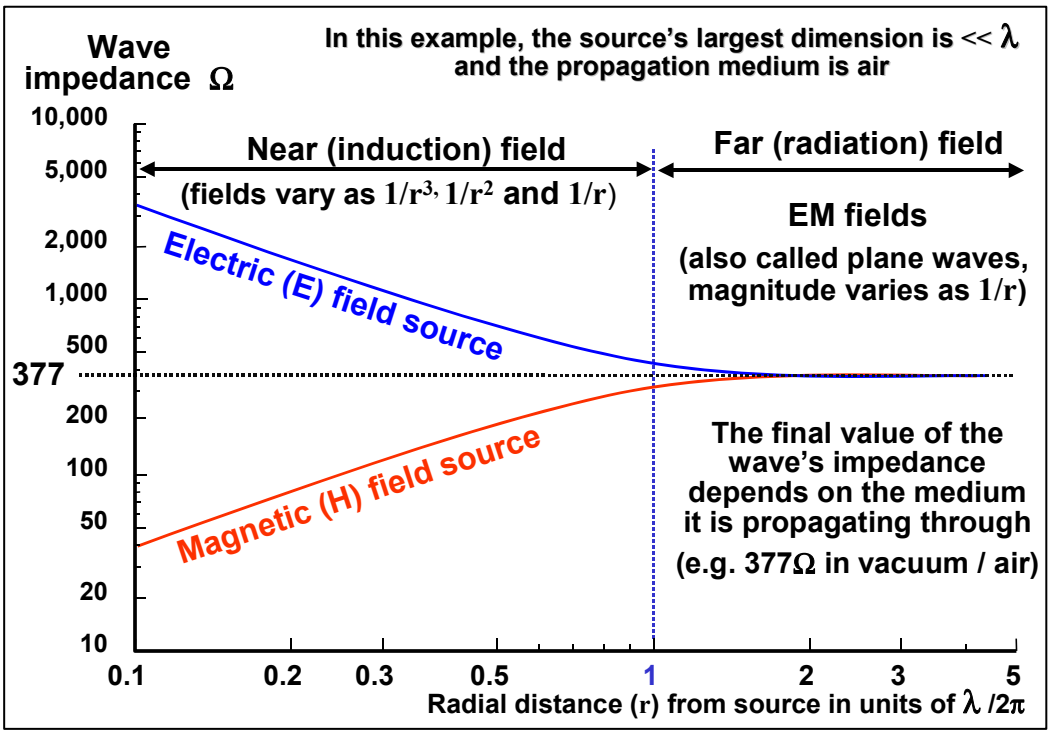


Figure 4 Near fields become Far fields with distance

Fluctuating voltages within Figure 4's circuits are initially associated with E-fields only, which have very high impedance. As the waves propagate further and further away from their E-field source, the impedance of the medium has more and more effect, causing an increasing proportion of the wave's energy to be converted into H-field. So the ratio of E-field to H-field decreases, the further away from the source, and the wave impedance therefore decreases.

Eventually, when the wave has travelled more than $\lambda/2\pi$ from its source, its E/H ratio matches the wave impedance of the medium and does not change any more – it has reached the 'far field' region.

A similar effect occurs to the magnetic fields emitted by the fluctuating currents in the circuit. Initially they have a very low impedance, because their E/H ratio has such a small numerator, but as the wave propagates further from the source the impedance of the medium it is travelling in has more effect, causing E-fields to be generated until the E/H ratio matches that of the medium itself – when the wave can be said to be in its 'far field' region.

3 EMC uses three types of analysis

Because of the effects of wave propagation, discussed in 2 above, we have to use three different kinds of analysis when analysing the EMC characteristics of electrical/electronic circuits and the rest of the structures that go to make a product or item of equipment. These are:

a) "Lumped Element" analysis when dimensions are much less than $\lambda/2\pi$.

b) "Transmission Line" analysis when one dimension is longer than $\lambda/2\pi$ (e.g. a long thin wire).

c) "Full Wave" analysis in 2 or 3 dimensions, when two or three dimensions are longer than $\lambda/2\pi$.

All circuits have RF resonant modes, sometimes called "eigentones", where their currents or voltages experience a resonant gain, called their 'Q factor'. Qs of 10 or more are common (i.e. gains of 20dB or more) in ordinary electrical and electronic circuits, with gains of 100

(40dB) not being unusual and up to 1,000 (60dB) being seen on occasion.

Higher Qs are associated with circuits that have low resistance. High values of resistance cause so much loss that only low Q values are possible.

Accidental antenna structures (that is: all conductors) are most efficient at their resonant frequencies, causing high levels of emissions and poor immunity. So it is very important to control resonances when designing to achieve EMC.

3.1 Lumped Element analysis

For conductor dimensions << $\lambda/2\pi$ we can use 'lumped element' analysis methods, which are based upon resistance (R), inductance (L) and capacitance (C). It is, in fact, the normal circuit analysis we all use in circuit design, but to be useful for EMC design purposes we have to take into account all the stray Ls and Cs that exist in our product. Stray L and C are sometimes called "parasitic" L and C.

Everything has R, L and C characteristics, including all components, wires, cables, PCB traces, connectors, silicon metallisation, bond wires, chassis, shields, mounting pillars, metal brackets, etc.

And everything also has stray (parasitic) R, L, and C, which can be *intrinsic* (e.g. the self-inductance of a wire lead) or *extrinsic* (e.g. stray C or L coupling due to proximity to other objects).

Let's look at these lumped element characteristics one at a time, from an EMC engineering perspective.

3.1.1 Resistance

Resistance increases with increasing frequency, f, due to the Skin Effect.

DC currents travel through the whole cross-sectional area of a conductor, but AC currents are forced to flow close to the surface, which is why it is called the skin effect.

This means that RF currents only penetrate weakly into the depth (thickness) of a conductor, decreasing the cross-sectional area of the copper they flow through and therefore increasing the resistance in their path, as shown by Figure 5.

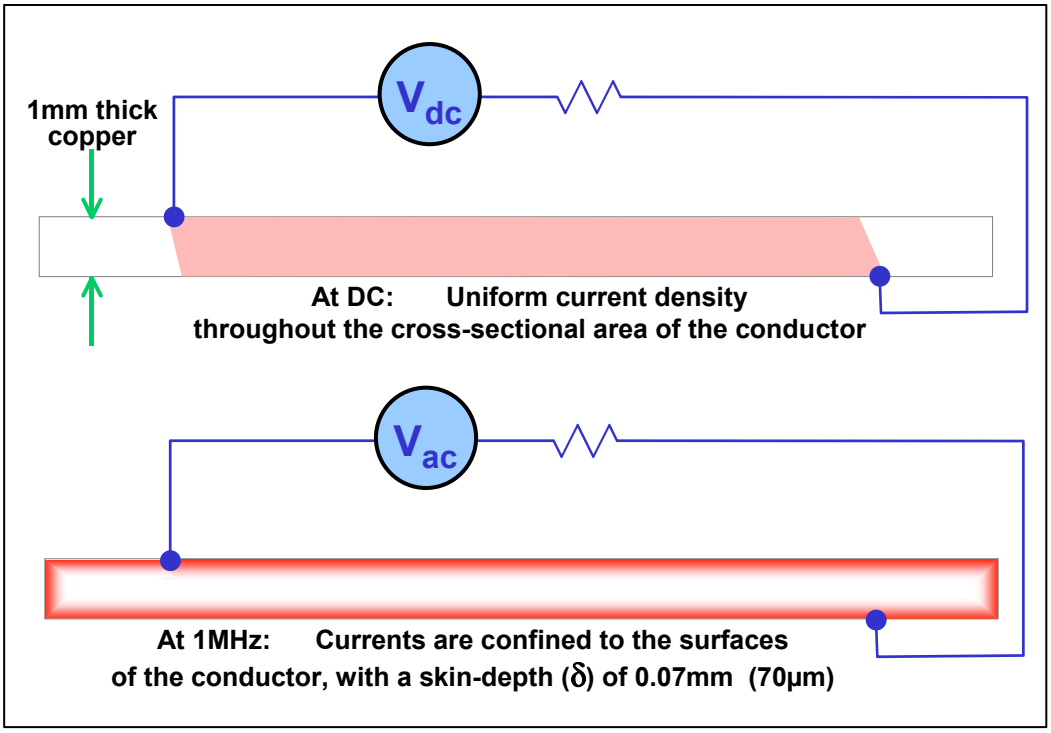


Figure 5 Comparison of current densities for currents at DC and 1MHz

Figure 5 shows quite clearly that at DC the current can take the shortest path *inside* the conductor, but at 1MHz it is forced to travel near the surface – which means that to get from one side of the conductor to the other it must flow around the edges. In a wide conductor, or sheet of metal, this can hugely increase the length of the current path.

One skin depth (δ) is the depth into the thickness of a conductor by which the current density has reduced to $1/e$, and is calculated as $\delta = (\sqrt{\pi \cdot f \cdot \mu_0 \cdot \mu_R \cdot \sigma})^{-1}$ metres, where σ is the conductivity of the conductor material. Each skin depth further below the surface, the current density reduces by a further $1/e$.

For copper conductors: $\delta = 66/\sqrt{f}$ (f in Hz gives δ in millimetres), for example at 160MHz δ = 0.005mm, so at 0.05mm below the surface (10 skin depths) the current density is $(1/e)^{10}$ – which means it is negligible. At this frequency, the resistance of a 1mm diameter wire is increased to about 50 times its DC resistance, due to the approximately 50 times smaller cross-sectional area that is carrying the current.

Volume II of [2] gives the characteristics for a wide variety of metals, enabling their skin depths to be calculated; [3] lists the skin depths of many metals, and Figure 6 graphs δ against frequency for copper, aluminium and a typical grade of mild steel.

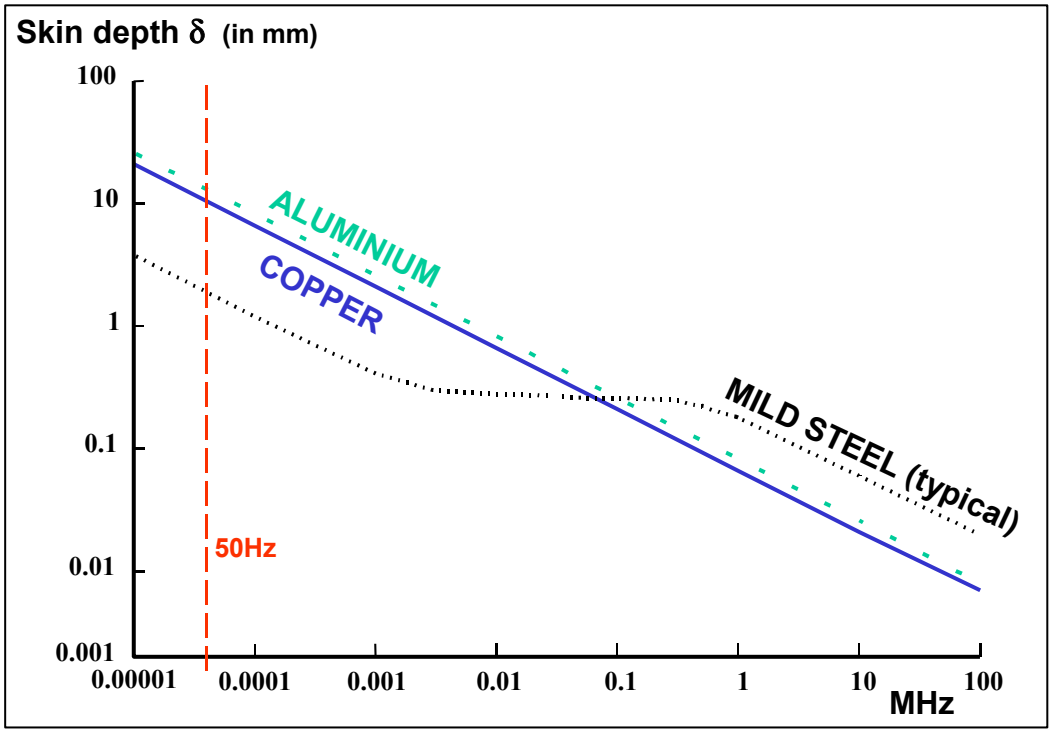


Figure 6 Graph of skin depth (d) for copper, aluminium, and a mild steel

As Figure 6 shows, mild steel has a skin depth that is a complex function of frequency, because of its μ_R of about 300, caused by its ferromagnetism. At lower frequencies its skin depth is reduced by its high value of μ_R, but between 1kHz and 1MHz it loses this characteristic and its μ_R decreases to 0. Since its conductivity is significantly less than aluminium or copper, its skin depth above 1MHz is larger.

There are many hundreds of grades of mild steel, and of course many other kinds of steels, including stainless types, and they all have different values of μ_R with different frequency variations. It can be quite difficult to find the μ_R for a particular type of steel, but it is usually impossible to discover its frequency characteristics. If you need to know this data, you will almost certainly have to measure it for yourself.

Nickel is also ferromagnetic and behaves like the mild steel in Figure 6, and there are also special high-permeability metals available with values of μ_R from about 10,000 to 50,000. These “hi-perm” metals may lose their permeability at lower frequencies than mild steel.

Skin depth has a very important part to play in shielding (see 4.3.3 in [1]), since any current that makes it to the other side of a shield radiated fields and therefore represents worse shielding. This is why Figure 6 has a line indicating 50Hz – to show how large the skin depths are at mains frequency.

Since one skin depth represents about 9dB of absorption, we can see that copper and aluminium sheets of about 10mm thick, and mild steel about 2.5mm thick, only provide a very small amount of shielding against magnetic fields at 50Hz. However, mild steel that is 10mm thick should give about 36dB attenuation at that frequency. This indicates that the best way to deal with magnetic fields at powerline frequency is to make sure that their source is so far away that the field levels are negligible and don't need shielding!

As mentioned earlier, resistance turns Watts and Joules of electrical power and energy into Watts and Watt-seconds of heat. It does not transfer (couple) or store electrical power or energy, like inductance and capacitance do.

3.1.2 Stray Inductance

A component purchased as an inductor will of course have a certain value of inductance, which will probably vary with frequency and temperature, and will have certain tolerances. But what we are concerned with here, is stray inductance. For a review of how stray inductances affect components themselves, see 1.8.1 of [1].

A thin wire has in *intrinsic* self-inductance of about 1μH per metre (1nH per mm), assuming the return path for its current is very far away.

If the return current path is a nearby conductor, it will have an *extrinsic* stray mutual inductance with its send path, which will cancel-out a proportion of the self-inductance and so reduce the overall inductance experienced by a current flowing around the send/return conductor path.

This is why, for good signal integrity, we need to keep the send and return current paths in intimate proximity all along their length, ideally twisting them together (see 2.2 of [1]). And for good EMC, we notice that such conductor structures have very compact, small, field patterns and so are poor accidental antennas.

Close proximity to ferromagnetic materials (e.g. steel, nickel, ferrite) with a $\mu_R > 1$ will increase both self-inductance and mutual inductance, but close proximity to conductive materials (e.g. cables, metalwork, etc.) will decrease them both.

As mentioned in 1 earlier, the true nature of electrical power and electronic signals is propagating EM waves, which carry the power or energy by means of electrical and magnetic fields. Inductance is associated with magnetic fields, and hence with transferring (coupling) and/or storing H-field power and energy.

The energy present in the H-field of an inductor, whether a component or a stray, is $\frac{1}{2}LI^2$ Joules (where I is the current in Amps and L is in Henries).

3.1.3 Stray Capacitance

A component purchased as a capacitor will of course have a certain value of capacitance, which will probably vary with frequency and temperature, and maybe with time, and will have certain tolerances. But what we are concerned with here, is stray capacitance. For a review of how stray capacitances affect components themselves, see 1.8.1 of [1].

A thin conductor on its own in free space has about 40pF of *intrinsic* stray "space charge" capacitance per metre of length (approximately 0.04pF per mm). The greater the surface area of the conductor, the greater will be its space-charge capacitance.

Close proximity to dielectrics with $\varepsilon_R > 1$, will increase all stray capacitances, both *intrinsic* and *extrinsic*. Whereas close proximity to other conductors will increase only *extrinsic* stray capacitances.

As mentioned in 1 earlier, the true nature of electrical power and electronic signals is propagating EM waves, which carry the power or energy by means of electrical and magnetic fields. Capacitance is associated with electric fields, and hence with transferring

(coupling) and/or storing E-field power and energy.

The energy present in the E-field of a capacitor, whether a component or a stray, is $\frac{1}{2}CV^2$ Joules (where V is the voltage and C is in Farads).

3.1.4 Lumped Analysis of Resonances

3.1.3 and 3.1.4 mentioned the electrical energy associated with the H and E-fields in capacitors and inductors – whether they are capacitor or inductor components, or stray (parasitic) capacitances and inductances.

Normally, the electrical energy present in a given circuit (or stray circuit) is a ratio of E and H-fields, but at the resonant frequency the energy oscillates between being all H-field and all E-field, causing amplification of the circuit voltage and currents.

All types of circuits have L and C (even if they are only strays) and these cause resonant frequencies to occur at : $f_{RES} = 1/(2\pi\sqrt{LC})$ Farads and Henries give the frequency in Hz.

These resonances are 'damped' by the resistance in the circuit, which is governed by skin effect as discussed in 3.1.1. More resistance means more power loss means a lower Q value. The above resonance formula is simplified and does not include the resistance term, but a more complete formula shows that resistance has a small effect on the resonance frequency, causing it to be a little lower than predicted by the simple formula above.

3.2 Transmission Line analysis

3.2.1 Analysing impedance section by section

3.1 showed that all conductors have R, L and C, and the L and C are involved with the storage and transfer of electrical power and energy (the EM wave).

2.4 discussed the near and far field regions of a propagating wave, and mentioned that in the far field the ratio of E and H-fields in the wave is the same as the wave impedance (Z_{WAVE}) of the medium it is propagating through.

The same effect is present in EM waves propagating along conductors, but with the wave impedance replaced by the "characteristic impedance", Z_0, of the send/return conductor structure, which is given by $\sqrt{(L/C)}$ (L in Henries and C in Farads gives Z in Ohms).

When an EM wave (what we think of as our power or signal) is "launched" into a conductor by an active electrical/electronic device, it has the impedance of its driver. And when it is "received" in an electrical/electronic load, it has the impedance of that load. But beyond a certain distance from either the driver or the load, the impedance of the EM wave is governed by the medium it is propagating in, the Z_0 of the send/return conductor structure.

It was mentioned in the initial review of basic principles that changes in the impedance reflect a proportion of the propagating EM wave, with larger changes over longer distances reflecting more. Reflected wave energy, is energy that does not travel from driver to load, which has the effect of distorting the waveform received by the load – a signal integrity or signal quality problem.

In the case of electrical power, reflected waves mean that at certain frequencies the power distribution cannot deliver as much current, and so appears to have a higher impedance – a power integrity (sometimes called power quality) problem.

And reflected EM wave energy makes conductive structures behave as more effective accidental antennas – causing EMC problems by increasing emissions and decreasing immunity.

Changes in impedance that occur for a length of less than about $\lambda/6$ have little effect on reflection, so transmission-line analysis consists of analysing the entire length of conductor as $\lambda/10$ sections, including its driver and its load, and determining the impedance of each. This then makes it possible to determine how a wave will travel from the driver to the load, and what will be the effect on its waveshape, frequency response, and EMC characteristics.

The L and C associated with a $\lambda/10$ section governs the velocity (v) with which EM waves

travel through that section: $v = 1/\sqrt{(LC)}$, and the ratio of L to C governs the Z_0 of the section: $Z_0 = \sqrt{(L/C)}$.

The L and C values used in the above analyses are 'per unit length' (e.g. 1μH/metre, 100pF/metre) where the unit lengths used are equal to or shorter than λ/10. Remember that as the velocity decreases due to relative permeability and/or permittivity, the wavelength for a given frequency gets shorter, so the length of a λ/10 section depend upon the medium surrounding the conductors – their insulators (e.g. PVC insulation on wires) or dielectrics (e.g. FR4 fibreglass on PCB traces).

3.2.2 The effects of keeping Z_0 constant

If the same value is maintained for the Z_0s of each λ/10 section of the conductor from the source to the load, *and for the impedances of the driver and the load*, 100% of the EM wave (the electrical power or signal) is communicated from source to load (ignoring the loss associated with the resistance). This is called matched transmission line design

Since very little energy is lost in such a design, the integrity of the waveform must be maintained, and the circuit must have very poor efficiency as an accidental antenna, and so must have low emissions and good immunity.

An example of matched transmission-line design is general purpose RF test equipment, connectors and cables, which has 50Ω impedances for all sources and loads, and uses connectors and cables with characteristic impedances of 50Ω over their entire lengths.

3.2.3 The effect of changing impedances over dimensions greater than λ/6

As mentioned earlier, such changes cause propagating EM waves to be reflected (whether they are signals or power), rather like the way that ripples spreading across a pool of water are reflecting from a floating object.

This is the technique that is used to achieve "EMC filtering" (see 3.2.1 of [1]) – we deliberately create changes in the characteristic impedance of a conducted EM wave, to reflect unwanted noise away from the circuit that our filter is intended to protect.

3.2.4 Transmission-line analysis of resonances

When a wave hits an impedance that is higher than the Z_0 it is propagating along, the reflection is in-phase with the wave at that point. The reflected wave thus adds to the original (impinging) wave and increases its amplitude at that point.

But when the change in impedance is lower than Z_0s, the reflected wave is in antiphase, and subtracts from the impinging wave, decreasing its amplitude at that point.

So when a portion of a conductor with a constant value of Z_0 along its length experiences the same *type* of impedance change (i.e. either higher or lower than Z_0) at <u>*both*</u> of its ends, the reflected wave "ricochets" backwards and forwards along that portion of the line.

A special situation occurs at the frequency at which the length of the portion of conductor is a whole number of half-wavelengths – a resonance occurs. Instead of the reflected waves decaying away quickly with successive ricochets, they maintain their original amplitudes.

This results in what is called a "standing wave" along that portion of the conductor. It is called that, because – if you probe at a particular location along the conductor (e.g. with a close-field probe and oscilloscope) – instead of seeing peaks and troughs in voltage and/or current occurring over time, you see a fixed amplitude all the time. Varying the position of the probe along the length of the conductor varies the amplitude and sign that is measured.

Of course, it is not really a fixed (standing) wave – it is a wave that is constantly being reflected back and forth between the changed impedances at each end of that portion of conductor. But the point is that its amplitude and polarity at a given point is a function of its position.

The creation of a standing wave causes the conductor to behave as a very efficient accidental antenna. In fact, when using conductors as *intentional* antennas, as radio transmitters do, a key factor is the "standing wave ratio" (SWR) and there are instruments

specifically designed to measure it.

The length of the antenna is adjusted, to tune its resonant frequency to the frequency of the radio transmitter. When an SWR of 1 is achieved, it means that all of the signal is being reflected at both ends and the conductor has the maximum value of standing wave on it, optimising its efficiency as a radio transmitting antenna. The same approach is used to tune receiving antennas too.

(Actually, it is not normal to adjust the actual physical length of the conductor being used as an antenna. Usually, the conductor is made a little shorter than is necessary, and its "electrical length" is *increased* to tune it to resonance, at the desired frequency, by varying a capacitive shunt.)

With same type of impedance change at both ends of a portion having a constant Z_0 along its length, the resonant frequencies of a conductor are $v \cdot l/2L$ where l is an integer (1, 2, 3, etc.) that describes the number of half-wavelengths in the standing wave and L is the length of the conductor between the changed impedances.

For a conductor in air (or vacuum), this expression converts to $150 \cdot l/L$ MHz when L is given in metres. So a 1m long conductor in air would resonate at 150, 300, 450MHz, etc.

When the conductor is not in air (or vacuum) its resonant frequencies are $150 \cdot l/L \cdot \mu_R \cdot \varepsilon_R$ MHz, so for 100mm of trace on an inner layer of an FR4 PCB (ε_R = 4.2 above 1MHz, μ_R = 1), the resonant frequencies would be 750, 1500, 2150MHz, etc.

The upper half of Figure 7 shows the concept graphically, for the situation in which the entire length of the conductor has the same Z_0 and the changed impedances at the ends are caused by the driver (source) and the load both being lower than Z_0.

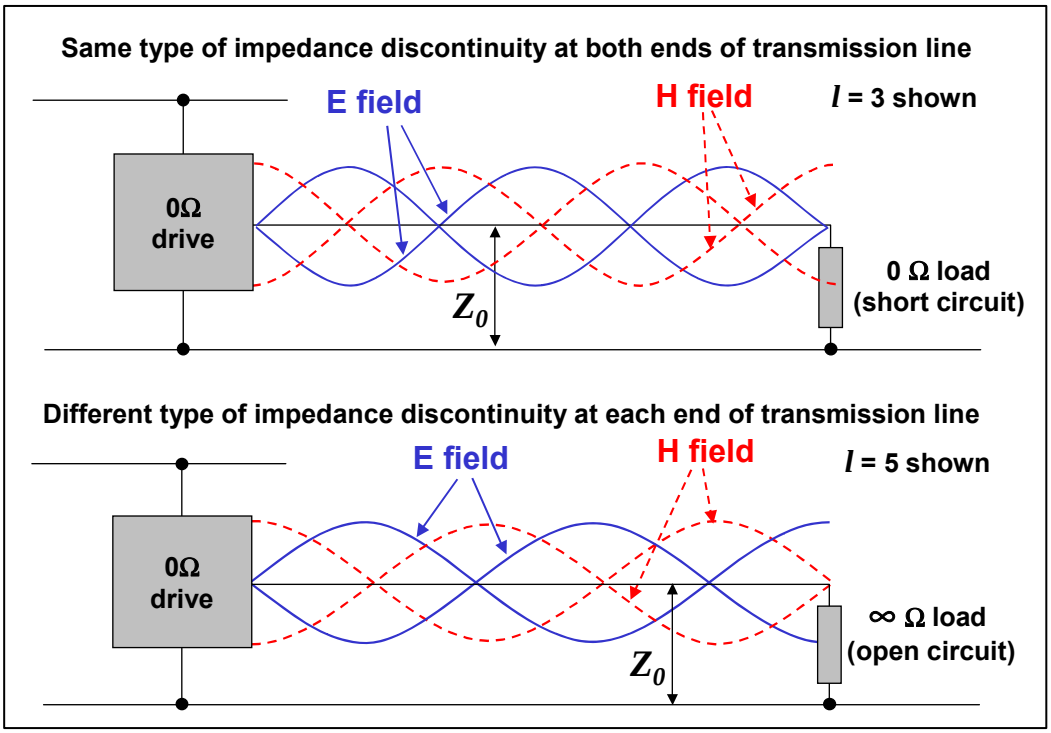


Figure 7 Resonances in mismatched transmission lines

The astute reader will have recognised that I am simplifying things a little. For instance, the velocity of propagation in air or vacuum is not really $3 \cdot 10^8$ m/s, it is a little less, and the root of 4.2 is not 2, but my "roundings-off" are good enough, especially since in real life there are always added complexities which add to inaccuracies (e.g. conductors are not straight, and are affected by other conductors nearby).

Any EMC formulae that can be written down are going to be inaccurate in all but quite simple situations, and so should be treated as guidelines only in real product design.

When a length conductor has different opposing types of Z_0 changes at each end, standing wave resonances occur at the frequencies at which the conductor length is an odd number of quarter-wavelengths, as shown in the lower half of Figure 7. The resonant frequencies are given by $v \cdot m/4L$ where m is an odd numbered integer (1, 3, 5, 7, etc.) that describes the number of quarter-wavelengths in the standing wave.

For a conductor in air (or vacuum), this expression converts to $75 \cdot m/L$ MHz when L is given in metres. So a 1m long conductor in air would resonate at 75, 215, 375MHz, etc.

When the conductor is not in air (or vacuum) its resonant frequencies are $75 \cdot m/L \cdot \mu_R \cdot \varepsilon_R$ MHz, so for 100mm of trace on an inner layer of an FR4 PCB, the resonant frequencies would be 375, 1125, 1725MHz, etc.

There is a complete formula for calculating a transmission line, called the "Telegrapher's Equation" because it was developed by people trying to send telegrams using codes rather like morse, over very long wires in the days of the Pony Express, and well before telephones and radio.

I am not providing all the equations in this article, as it would make it much too long and boring. Instead, I am trying to provide a conceptual understanding, to help designers visualise EM phenomena, and see how to use the various formulae or simulators in their work. The formulas are all available from textbooks like [2] [4] [5] and [6] and/or from the Internet (e.g. rfcafe, see [3]), where some on-line calculators may be available, for example from [7].

3.3 Full Wave analysis

When a conductor is > $\lambda/6$ in two or three dimensions, simple formulae are only practical for very simple situations like the flat plate and empty metal box shown later.

For any reasonable accuracy for real-life products and equipment, we must use 'full-wave analysis' based on simplifications of Maxwell's Equations. This is only practical by using computers to do the analysis, a technique that is called computer-based EMC simulation, or simply EMC simulation.

Figures 8 and 9 show a simple metal plate in air, a conductor which is larger than $\lambda/6$ in just two dimensions. This presents a consistent (very low) impedance to propagating EMC waves, but at its edges it meets the air with its wave impedance of 377 Ω. So it has the same kind of impedance change along all edges and experiences standing wave resonances when a whole number of half-wavelengths are fitted into its area.

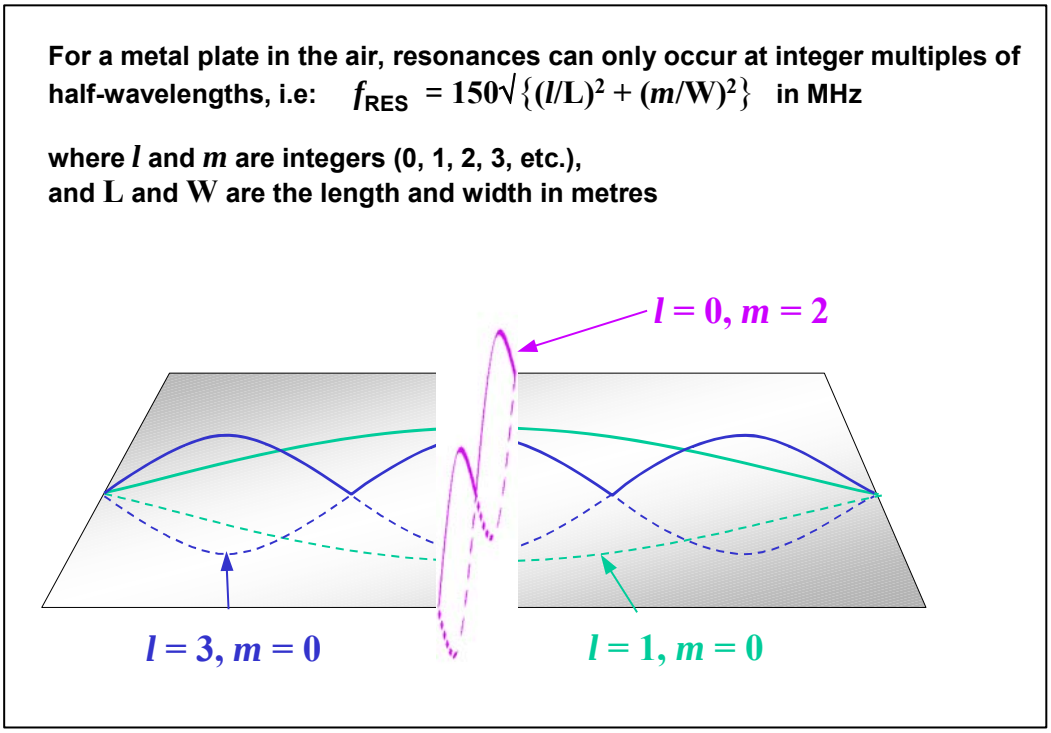


Figure 8 Standing H-fields caused by impedance discontinuities at edges of a metal plate

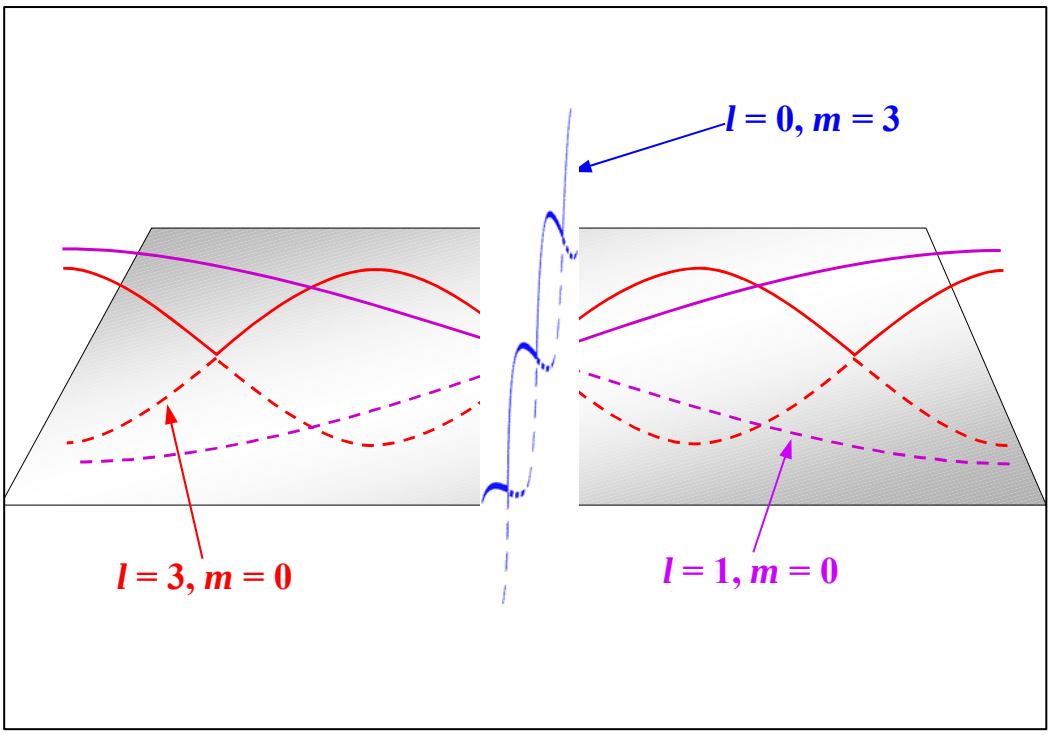


Figure 9 Standing E-fields caused by impedance discontinuities at edges of a metal plate

The simple expression for the resonant frequencies for this simple structure is $150\sqrt{\{(l/L)^2 + (m/W)^2\}}$ where: l and m are integers (0, 1, 2, 3, etc.) and L and W are the plate's length and width (if specified in metres, gives the resonant frequencies in MHz).

The first few, lowest-frequency resonances and the easiest to visualise can easily be found by setting l to 1, 2, and 3 whilst setting m to zero; repeating with m set to 1, 2, and 3 and l to zero. Figure 8 shows the $l = 3$, $m = 0$ mode.

If the metal plate was surrounded by a liquid or solid, its resonant frequencies would be given by $(v/2)\sqrt{\{(l/L)^2 + (m/W)^2\}}$, i.e. $(150/\mu_R \cdot \varepsilon_R)\sqrt{\{(l/L)^2 + (m/W)^2\}}$.

The three-dimensional situation is shown in Figure 10. This time it is the medium inside the empty box that is supporting the standing waves. The air has a wave impedance of 377Ω but the walls of the box have a very low characteristic impedance, so once again we have the same type of impedance change at both ends of a portion of the wave's propagation.

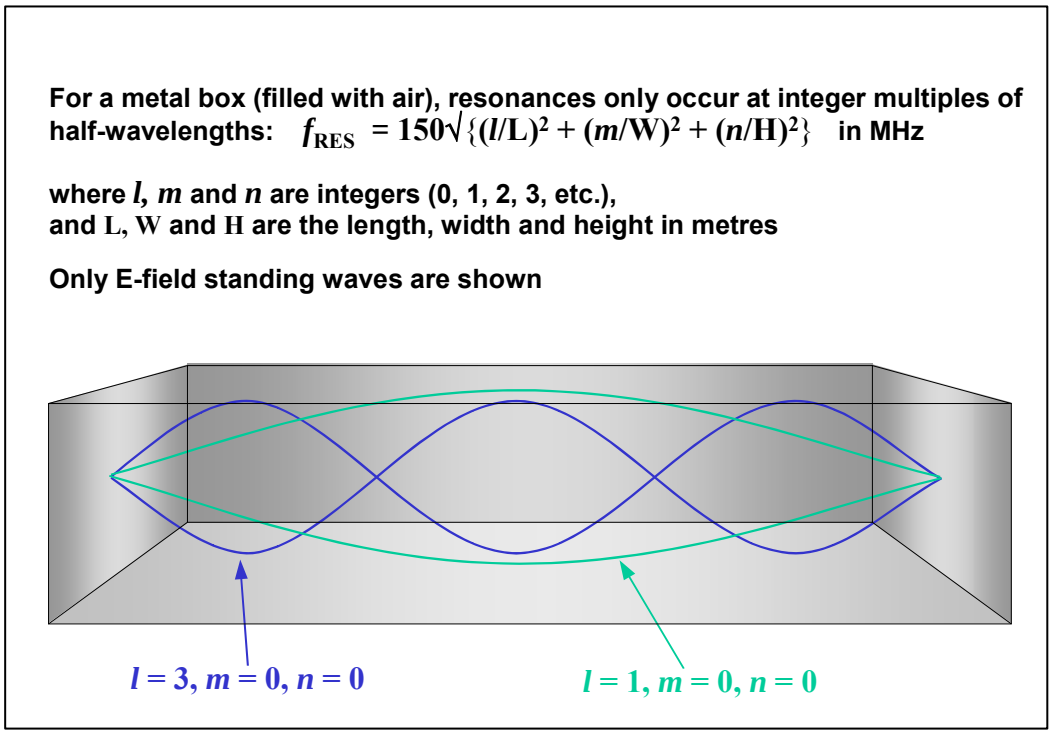


Figure 10 Standing waves caused by impedance discontinuities inside a metal box

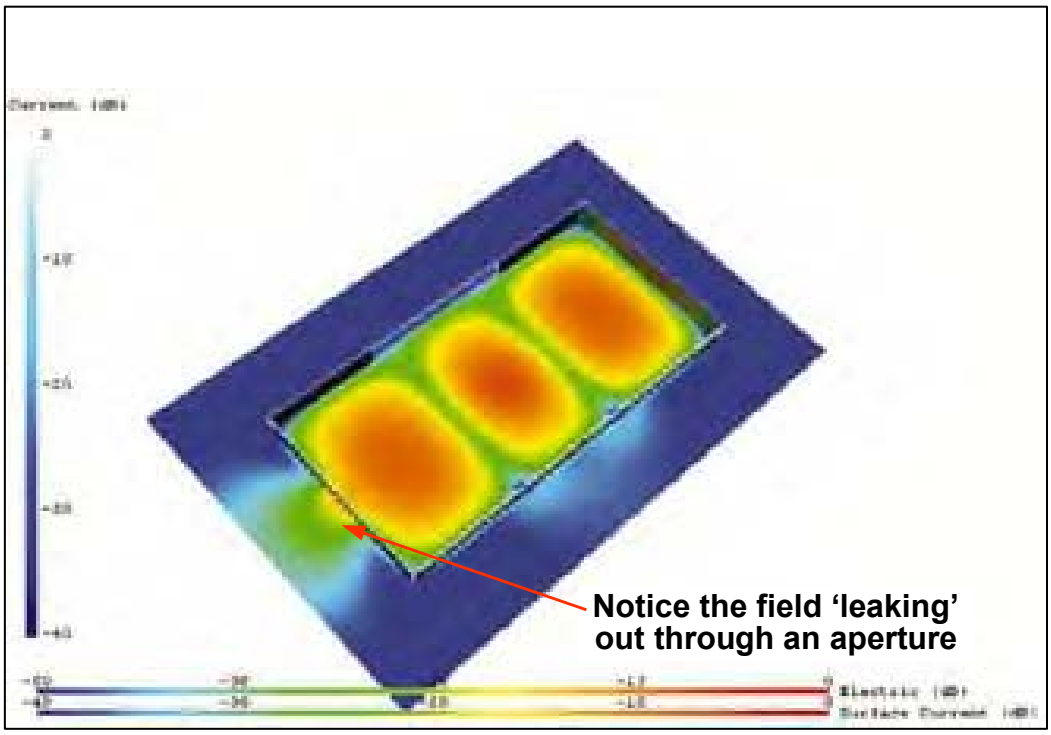
Standing wave resonances can only occur at integer multiples of half-wavelengths, at $(v/2)\sqrt{\{(l/L)^2 + (m/W)^2 + (n/H)^2\}}$ MHz, i.e. $(150/\mu_R \cdot \varepsilon_R)\sqrt{\{(l/L)^2 + (m/W)^2 + (n/H)^2\}}$ MHz, where: l, m and n are integers (0, 1, 2, 3, etc.) and L, W, H are the box's length, width, height (in metres).

For a box filled with nothing but air, the expression simplifies to $150\sqrt{\{(l/L)^2 + (m/W)^2 + (n/H)^2\}}$ MHz. So an empty air-filled metal box 400 x 300 x 200 mm will have standing wave resonances at 375MHz (its 1,0,0 mode, when the standing wave exists along its length), 500MHz (its 0,1,0 mode, when the standing wave exists along its width) and 750MHz (its 0,0,1 mode, when the standing wave exists along its height).

Obviously, we could calculate the 5, 3, 7 mode (it is 5,773GHz) but we can't easily visualise it!

Figure 11 shows a computer simulation of the electric field distribution inside an empty shielded metal box, with some small apertures in its sides for ventilation. It was done using Microstripes, now part of Computer Simulation Technology Ltd's range of simulation products.

The simulator works its way through a range of frequencies, in three dimensions, and its result can be viewed as an animation. To see what is going on inside the box, it allows us to make a "slice" through the box, and Figure 11 is one frame from the animation, at one of the box's internal resonant frequencies – showing the standing waves in its 3,0,0 mode.



Compare it with Figure 10's crude sketch of a box's 3,0,0 mode.

Figure 11 Simulation of the standing waves inside a metal box

4 Waveforms and Spectra

4.1 Analysing in the time and frequency domains

On/off and discontinuous waveforms (e.g. digital, PWM, switch-mode, etc.) are rich in harmonics, which appear as narrow spectral lines when viewed on a spectrum analyser or EMC receiver display – see the example of an idealised 16MHz squarewave with 2ns rise/fall times in Figure 12 (time domain) and Figure 13 (frequency domain).

The famous Mr Fourier developed the mathematics by which a waveform in the time domain can be seen as the sum of a lot of individual frequencies in the frequency domain, so Figure 13 is a "Fourier Transform" of the waveform in Figure 12. (I am not providing references for basic mathematical techniques like Fourier, because everyone can search the Internet.)

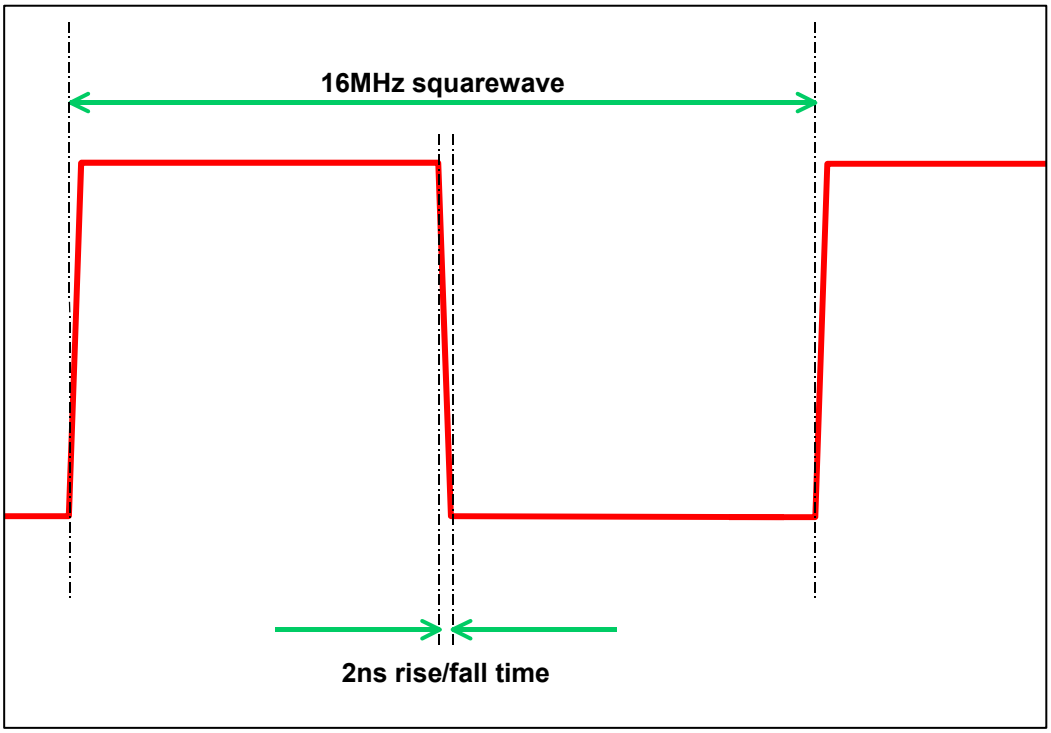


Figure 12 The (idealised) waveform of a 16 MHz squarewave with 2ns rise/falltimes

It is not unusual for a switch-mode power converter operating at 50kHz to cause high levels of emissions at every 50MHz spacing up to 50MHz (the 1,000th harmonic!) – or for 50MHz digital clocks to cause high levels of emissions at 50MHz spacings up to 900MHz.

Some years ago I met an engineer who designed in-ear hearing aids. They used naked silicon chips bonded directly to very tiny PCBs (to fit in the ear) and had no 0V plane area to speak of. They had a microprocessor that was clocked at 1kHz to save battery power, and the poor guy was failing the radiated emissions test at every 1kHz harmonic from 30MHz to 1GHz (the 1 millionth harmonic!).

When resonant frequencies (see Section 3 in Part 1 of this article [8]) happen to lie at the same frequencies as a harmonic, they 'amplify' the common-impedance, E, H and EM coupling effects and can dramatically increase the conducted or radiated emissions as a result. (Coupling effects are discussed later on, in section 5.)

So this is why we find most EMC textbooks and Guides recommending that we don't use voltages or currents that change faster than they really need to. This is good general advice, and I heartily endorse it – but the problem is that very few ICs are available with edge-rate control, usually we just get whatever the silicon happens to switch at. As Moore's Law grinds silicon features inexorably smaller – switching speeds inexorably rise, whether we need the

faster edges or not.

(Solutions to the problem of too-fast devices, where we don't need such fast edges were discussed in Section 1.1.2 of Part 1 of [1] and Section 5.6 of Part 5 of [1]. Section 5.6 also discussed how to control the characteristic impedance of a PCB trace route in cases where the fast edges were not controlled, or were even needed for fast data rates.)

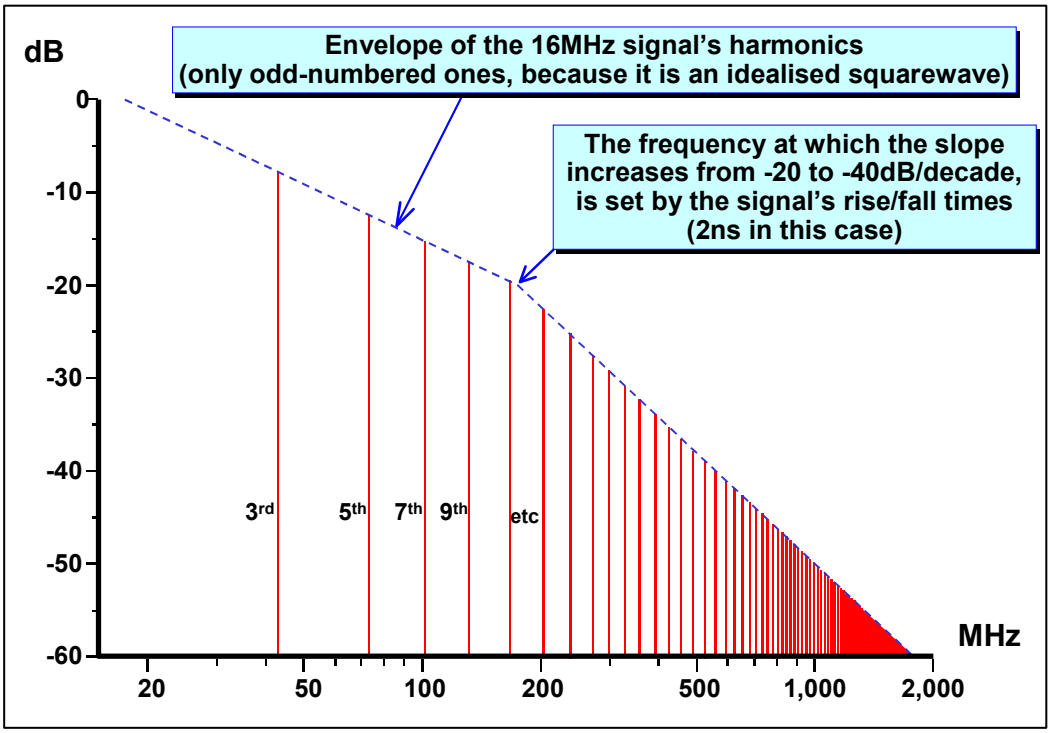


Figure 13 The (idealised) spectrum of a 16 MHz squarewave with 2ns rise/falltimes

Figure 14 shows the (idealised) radiation efficiency (what we might call the "accidental antenna efficiency") of a 200mm PCB trace on an FR4 dielectric, with no 0V plane, a very low-impedance drive, and a very high-impedance load. This is the efficiency (0dB = 100%) with which EM energy conducting along the trace (that we call signals, control, data, power, etc.) is converted to EM energy radiating out of the trace, through the PCB's dielectric and the air. The efficiency of the trace as a transmitting antenna.

The lower sketch in Figure 7 of [8] shows the resonance mechanism due to mismatches between the source and load impedances and the Z_0 of the trace, which is responsible for the trace behaving as an accidental antenna, with Figure 14 its result.

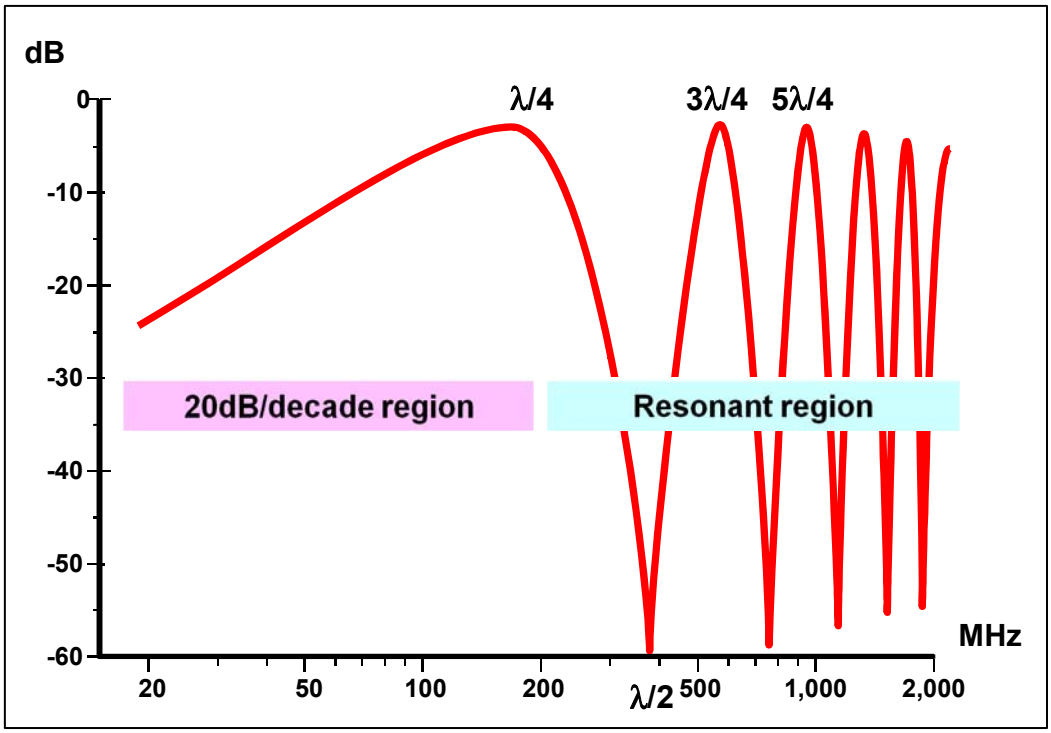


Figure 14 Example of the antenna efficiency of a 200mm long PCB trace

Figure 15 shows an idealised sketch of the sort of waveform that would be expected to occur at the end of the 200mm PCB trace of Figure 14, when the 16MHz clock signal of Figure 12 was conducted along it.

Everyone who has ever measured voltages around a digital circuit (usually to find out why it wasn't working properly) with an oscilloscope probe has seen waveforms that look like this – with overshoots and ringing – but probably didn't realise that they indicated EMC emissions problems.

In fact, a circuit designer who is used to measuring circuit waveforms and EMC emissions will usually be able to make a good prediction of which frequencies will exceed the emissions limit line, and even by how much, from the amount and shape of the overshoot and ringing.

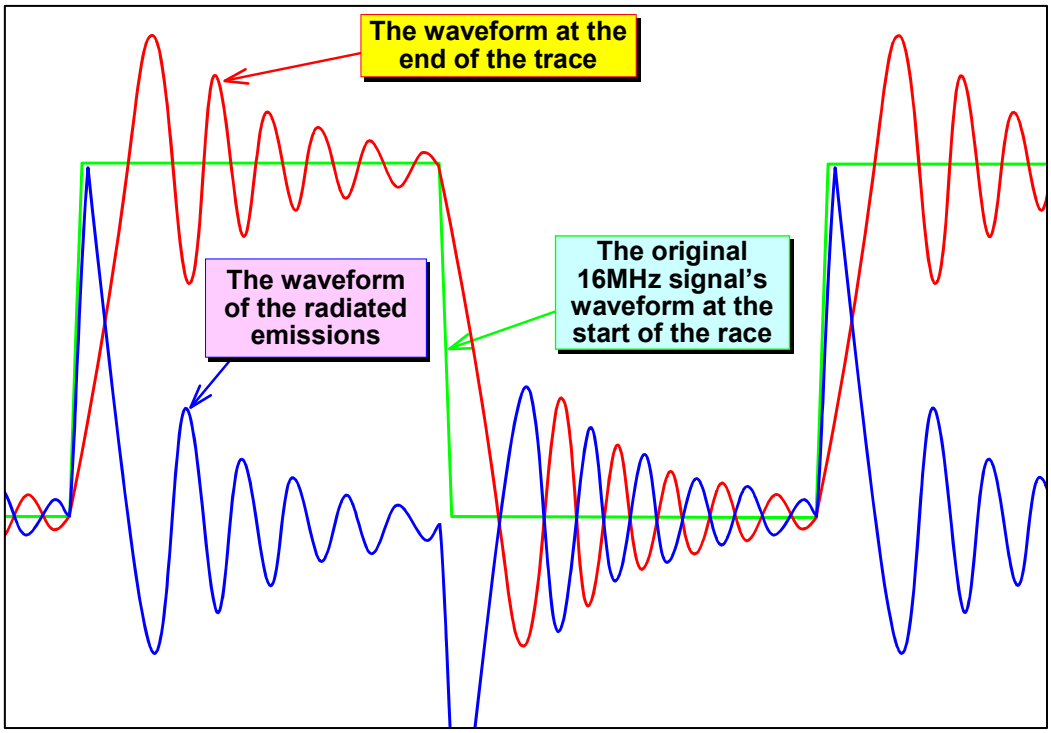


Figure 15 Figure 12's 16MHz squarewave after passing along Figure 13's PCB trace

What has happened is that some of the harmonics (Figure 13) of the 16MHz squarewave (Figure 12) have been radiated into the air by the 200mm PCB trace (Figure 14) behaving as an accidental antenna. These unlucky harmonics have lost a significant amount of their energy to the air, so that very little energy at their frequencies is now available at the far end of the PCB trace.

(In real life, the PCB trace would also lose energy to any conductors unlucky enough to be nearby, causing crosstalk and noise in other PCB traces, and in its power rails, possibly even causing failure to meet conducted emissions tests if they 'couple' energy into the mains lead. But such real-life details are not considered in this simplified and idealised analysis.)

Performing a Reverse Fourier Transform (sounds impressive, doesn't it!) on the spectrum of energy remaining at the far end of the 200mm trace, results in Figure 15. The frequencies that have lost most of their energy to the air are the cause of the ringing in the waveform.

It just happens that in this simplified and idealised example, the harmonic that loses the most energy to the air is the one that is closest to the first resonance frequency ($\lambda/4$) of the trace, around 187MHz. This is the 11th harmonic of the 16MHz signal, 176MHz, and so the dominant ringing frequency of the waveform at the end of the trace (Figure 15) is 176MHz.

> (Unfortunately, my choice of 2ns for the rise/fall time of the 16MHz squarewave in Figure 12 resulted in the corner frequency of its spectral envelope in Figure 13 being at 160MHz. My choice of a 200mm trace placed its first resonance at 187 MHz in Figure 14, and my choice of a 16MHz squarewave (Figure 12 again) put its 11th harmonic at 176MHz in Figures 12, 14 and 15. The closeness of these three frequencies, especially when crudely sketched in my figures should not be taken to mean that there are any relationships between any of them. They just ended up being so close together by accident and bad drawing. Sorry.)

Figure 15 also sketches what I have called the waveform of the radiated emissions. These are what you would see (simplified, idealised, of course) if you connected the EMC test lab's antenna to an oscilloscope instead of to the usual spectrum analyser or EMC receiver.

As sketched in Figure 15, it appears that if we could add this radiated waveform to the conducted waveform that remains when the signal reaches the end of the trace, we could reconstruct the original squarewave! (This is not always exactly true in real life, because a

mismatched PCB trace can have other waveform distortion mechanisms that might not be associated with radiated energy. But its generally close enough.)

Figure 16 sketches the simplified radiated emissions spectrum from our 200mm trace excited by a 16MHz squarewave, when the test lab antenna is connected to their spectrum analyser or receiver as usual.

It shows that the harmonics that have frequencies closest to the trace's resonant frequencies (odd numbers of quarter wavelengths, see Figure 7 of [8]) are being emitted the most, and causing the most problems during emissions testing.

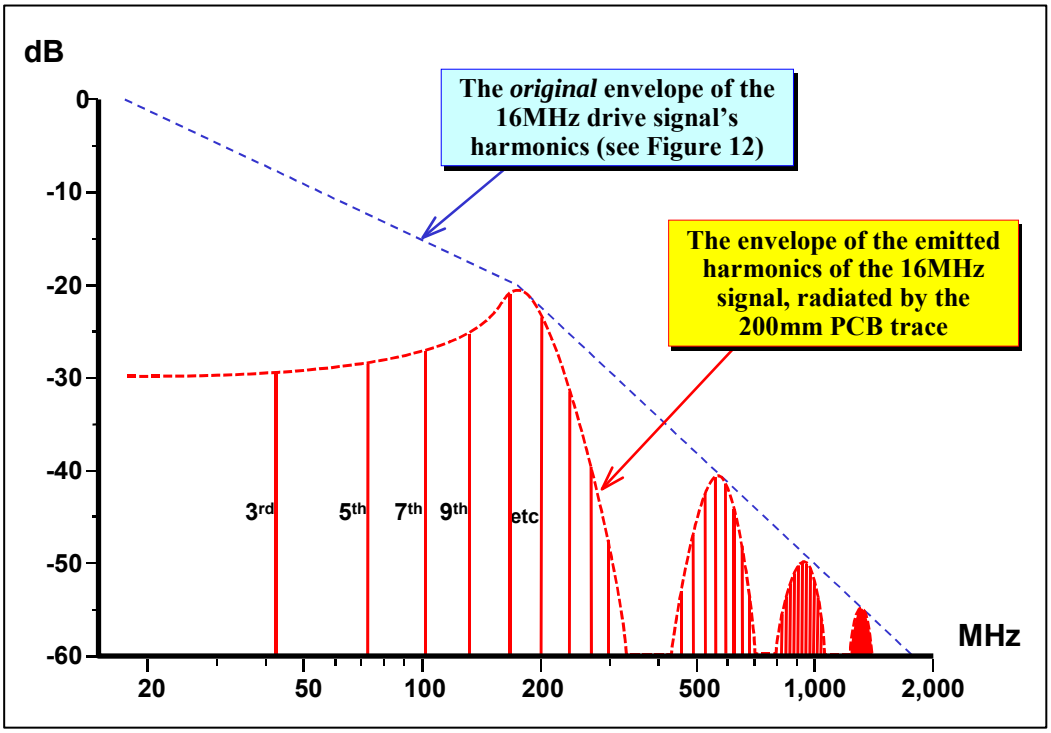


Figure 16 The (idealised) spectrum of the radiated emissions from the 200mm PCB trace

Design engineers who are experienced in designing for good EMC performance are usually adept at converting in their "mind's eye" between emissions spectrum plots and the waveforms that caused them, and between actual waveforms and their likely emissions.

Also, from the spacing between the groups of harmonics in an emissions plot (Figure 16) they can get a good feeling for the likely lengths of traces, cables and/or shielding gaps in the product that might have caused them. (Then out come the ferrite beads and copper tape, to apply to said conductors and shield gaps, to see which ones are being naughty.)

4.2 EMC and Signal Integrity

I find the one-to-one relationship between the ringing frequencies of the waveform at the end of the trace – which is a signal integrity (SI) issue – and the frequencies at which there are significant radiated emissions, very interesting indeed, and it is very useful in practice too.

Even if you only have an oscilloscope and voltage probes, you can still get a good idea of whether a digital circuit design/layout is going to be good for EMI. A high level of ringing that is good enough SI is not going to be good enough for EMC compliance, unless the PCB concerned is enclosed in a shielded housing with all of its interconnections shielded and/or filtered (as discussed in Part 4 of [1]).

When we want to design low-cost products that are nevertheless legally complaint and work well in real life, we design to ensure that all the digital waveforms on the PCB have very little ringing and overshoot. Typically we would aim for levels that are one-tenth of those that are

acceptable for SI.

So, there is a strong relationship between EMC and SI. If we don't "lose" significant amounts of harmonic energy from our signals' spectra, we won't suffer waveform distortion, overshoot, undershoot, double clocking, etc., etc.

Testing with a high-speed oscilloscope (with high-speed probes, properly used) will reveal that our waveforms remain undistorted, and testing in an EMC lab will show that emissions are low, so it may be possible to pass the EMC tests without having to add costly shielding and filtering.

Designing a PCB and digital circuit to have low emissions without the cost of enclosure shielding means that the waveforms are undistorted – nice clean rectangular waveforms with low overshoot and ringing – and we also get truly wonderful SI, making software more reliable and saving time during product development.

Because of the principle of reciprocity (see note iii) in Section 1.3 of [8]) – a trace that does not radiate very much at a circuit's critical frequencies (i.e. its waveform's fundamental and harmonic frequencies) and so has low emissions, will also not pick up much noise from the EM environment, and so will have good immunity too.

For analogue circuits: although different parameters are used to characterise their SI, exactly the same considerations apply as for digital circuits. In this case, it is the frequency response of the trace that is the issue. If a trace interconnecting two analogue devices together has a flat frequency response across the entire bandwidth that is used by the signals, then this is one measure of good analogue SI, and it also means that emissions will be low and immunity high. The "flatness" of the frequency response of the analogue circuit interconnection corresponds directly with the ringing and overshoot of a digital circuit's rectangular waveforms.

4.2.1 Implications for computer-aided EMC simulation

The above understanding has significant implications for simulation. Good EMC simulators are still quite costly, especially when we want to model the emissions from a complex PCB and all of its devices.

However, many of us already have SI tools in our PCB CAD workstations. Using such tools, and setting the maximum levels of all SI parameters (overshoot height, ringing decay time, ground/power rail bounce, etc.) to be five or (better still) ten times less than the minimum needed for SI, and then designing the circuits and PCB layouts to achieve them, will generally result in PCBs with very low RF emissions and good RF immunity.

Obviously, not all immunity concerns are addressed by this approach – for instance it won't help improve immunity to power quality disturbances (dips, dropouts, distorted mains waveforms, etc.) or the over-voltages or over-currents caused by surges, fast transient bursts or ESD. But it will help cope with the fields created by fast transient bursts and ESD, which are RF disturbances.

4.3 Conductors as accidental antennas

Figure 17 shows the frequencies in common use by ordinary people over the frequency range 1Hz to 3GHz. In fact, apart from a few small frequency bands kept clear for radio astronomers, all of the frequencies are allocated to some sort of radiocommunications activities.

A few small bands are allocated to what is called "Industrial, Scientific and Medical" use (ISM for short) – where radio energy is used to process materials (melt metal; weld plastic; dry paint and glue; relieve sport injuries by heating muscles; cut human tissue with coagulation, etc.). Microwave cookers are ISM equipment and operate in the 2.5GHz ISM band.

The ISM bands are very well harmonised worldwide and have been for years, because just one site operating powerful ISM equipment can easily ruin radio reception at that frequency in neighbouring countries. So it was in everyone's best interests to cooperate internationally

and agree the ISM bands.

Radiocommunications are allowed to use the ISM bands without requiring much, if any, regulatory approvals, as long as their users don't complain if they are interfered with by ISM equipment. And because the ISM bands are pretty much agreed worldwide they are attractive for manufacturers of short-range radiocommunications products with global ambitions. This is why Wi-Fi, ZigBee and Bluetooth all operate in the same ISM band as microwave cookers, and why your wireless Internet connection may go slow or stop altogether when someone starts to microwave their meal nearby (see Banana Skin No. 391 in [9]) – and there's nothing you can do about it.

The microwave cooker is allowed to emit lots of noise in that band, and if your Wi-Fi (or your neighbour's) doesn't work as a result, that's just tough for the Wi-Fi. Which makes it all the more interesting when people try to use Wi-Fi (because it is cheap technology and widely available) in industrial control, or in desperately-safety-critical automotive systems [16], especially when even in the absence of ISM noise the 2.5GHz band can be overcrowded with nothing but Wi-Fi in any case [10].

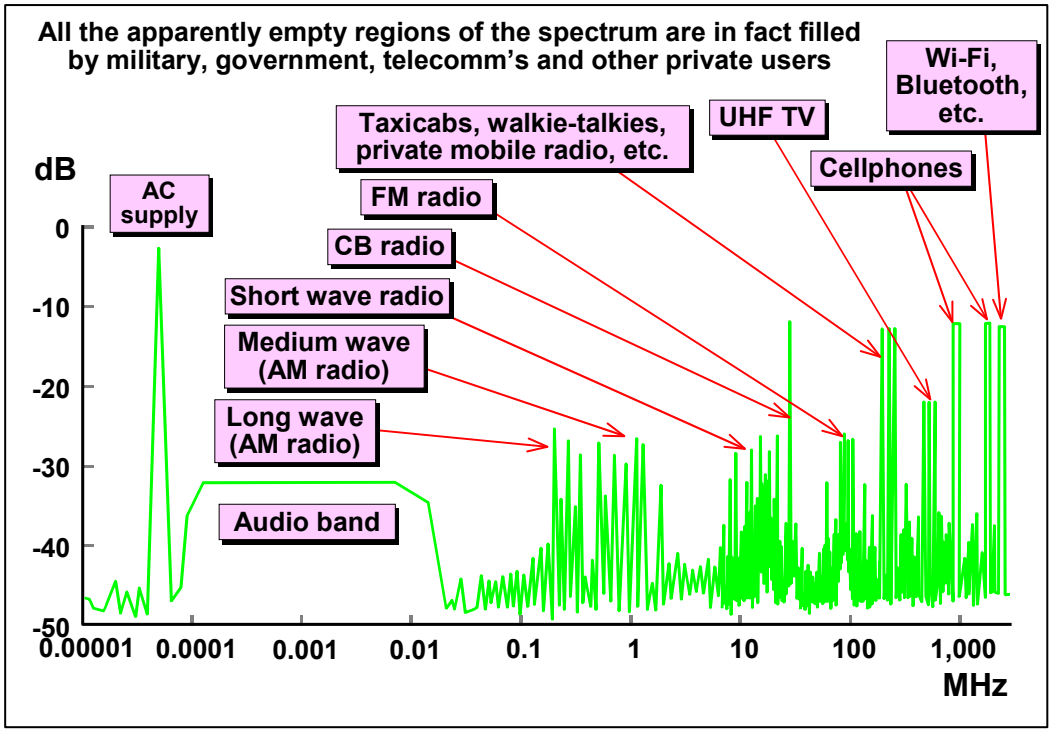


Figure 17 The frequencies we use (1Hz to 3GHz)

Figure 18 shows the same graph as Figure 17, but with the typical harmonic-spectral emissions created by 50/60Hz AC rectifiers, switch-mode DC power converters (the example assumed to be switching at 70kHz) and ordinary microprocessors (the example assumed to have a clock rate of 32MHz, hardly blisteringly fast these days).

It shows that the noise from these very commonplace circuits is all over the frequency bands we want to use for radio and TV broadcasting and communications. Clearly, it is very important that these internally generated noises are not radiated outside of products into the air. This is exactly why we have CISPR [11], and their emissions standards and limits.

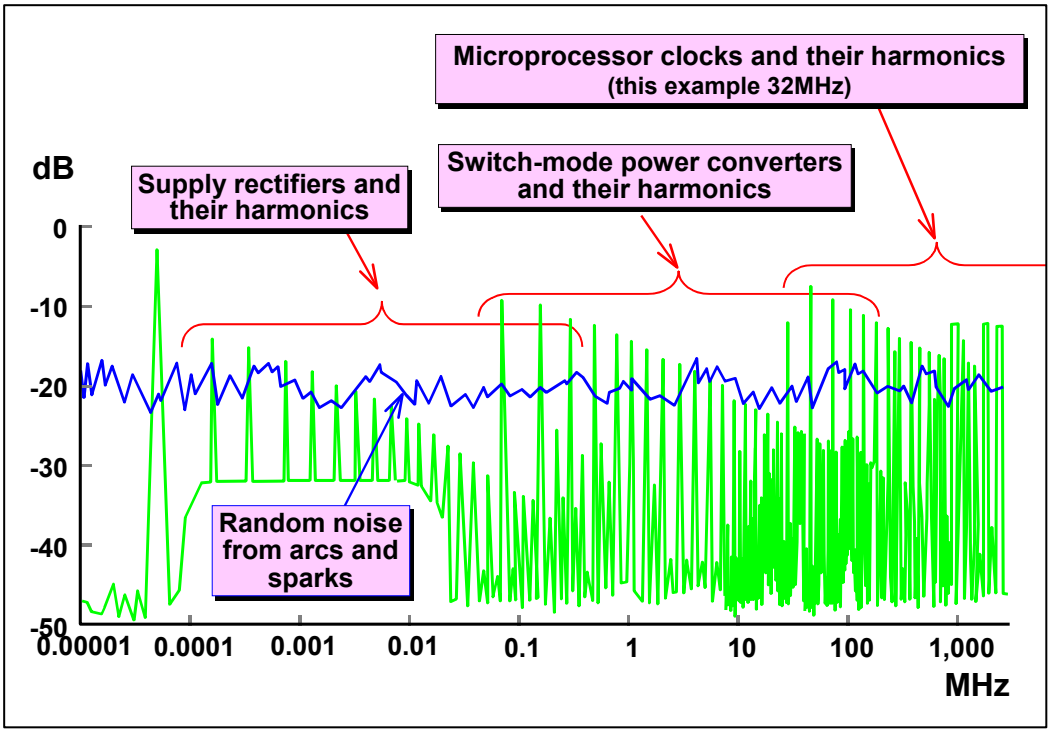


Figure 18 ....plus the noises emitted by electrical and electronic activities

But the problem is that any/all conductors make very efficient accidental antennas, unless they are designed and constructed to be well-matched transmission lines. (Section 2.7 of Part 2 of [1] describes transmission lines made with wires and cables, and Section 5.6 of Part 5 of [1] or Chapter 6 of [12] describes transmission lines made with PCB traces.)

As Figure 19 shows, any practical length of conductor (wire, cable, PCB trace, connector pin, bracket, heatsink, etc.) is a good accidental antenna at frequencies below 3GHz. Even if the conductors are not carrying noisy signals/data/power/etc. in their own right, they are usually carrying common-mode (CM) RF noises (e.g. ground/power bounce noise) from a wide variety of unintended but often unavoidable sources. So, all practical wires and PCB traces can cause us to fail an emission test.

(What CM noise is, and where it arises, is discussed in section 5.5 later.)

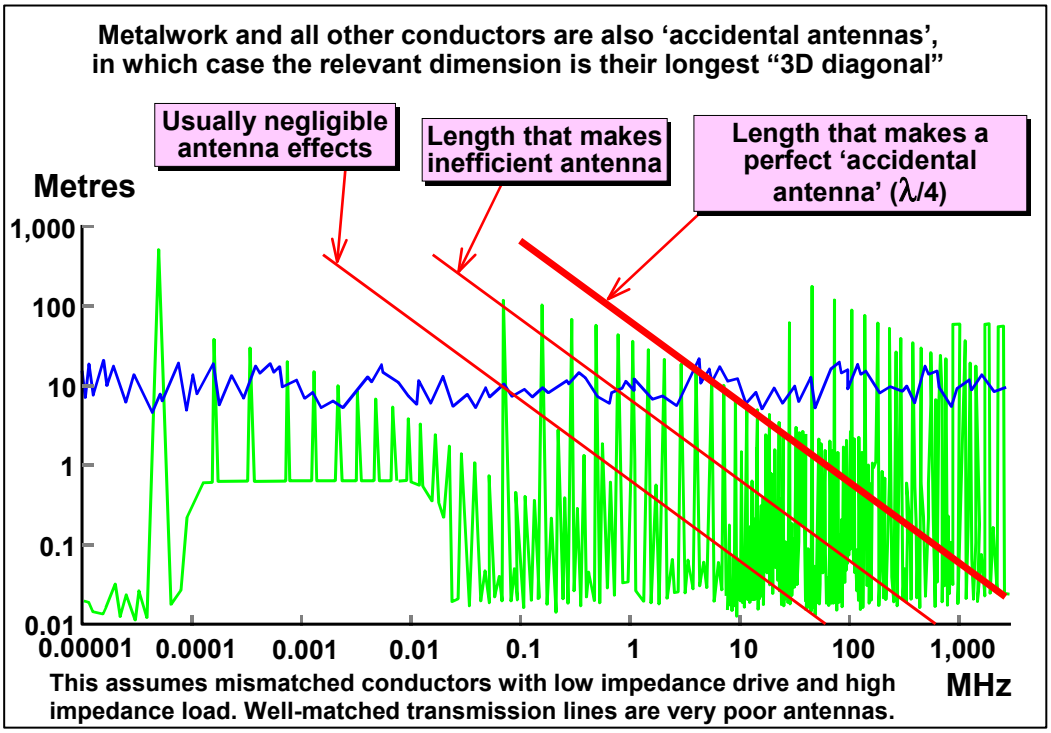


Figure 19 All conductors behave as accidental antennas

All conductors behave as 'accidental antennas', like the 200mm trace example in Figure 14. Figure 19 assumes mismatched conductors with low impedance drive and high impedance load when compared with the Z_0 of the conductor – typical of most (non-RF) analogue and digital circuit interconnections. Well-matched transmission lines are very poor accidental antennas, perhaps up to 40dB less efficient than an unmatched conductor – but they are still antennas.

Since wires, cables and other conductors behave as accidental antennas, they always "leak" the signals or CM noises they are carrying as electric (E) and/or magnetic (H) fields, possibly causing emissions problems. They also pick-up interfering noises as voltages and/or currents, from electric (E) and/or magnetic (H) and/or electromagnetic (EM) fields, possibly causing immunity problems for our circuits.

This is not what we want our electrical and electronic interconnections to do, never mind other metal items such as support brackets or decorative trim, but it is unavoidable, and dealing with it affects much of our EMC design.

If we don't control it at device or PCB level, we are stuck with using costly enclosure-level shielding, which also tends to restrict the aesthetic design of the product, add weight, and increase assembly time.

Figure 19 includes two lines that indicate lower levels of accidental antenna efficiency over certain frequency ranges – with the left-most line showing what lengths would generally be negligible for ordinary consumer, commercial and light industrial applications. But such conductors would probably behave unacceptably as accidental antennas over those same frequency ranges if tougher emissions and/or immunity specifications were applied, e.g. for motor car manufacturers, flight-critical avionics, etc.

It was the case not long ago that the cables attached to a portable product were its only significant causes of radiated emissions. In fact, the IEE (now called the IET) had an event in 2003 entitled: "EMC: It's (nearly) all about the cabling". But those days are long gone, and most EMC remedial work to convert EMC test failures to passes these days involves working on the PCBs.

So far, the figures in this article have shown simple harmonic structures based on square waves and the like, which makes for easy analysis and helps to communicate the basic

concepts. But in the real world of ICs and PCBs, signals from microprocessors and similar ICs have all sorts of additional noises on them, as Figure 20 attempts to show.

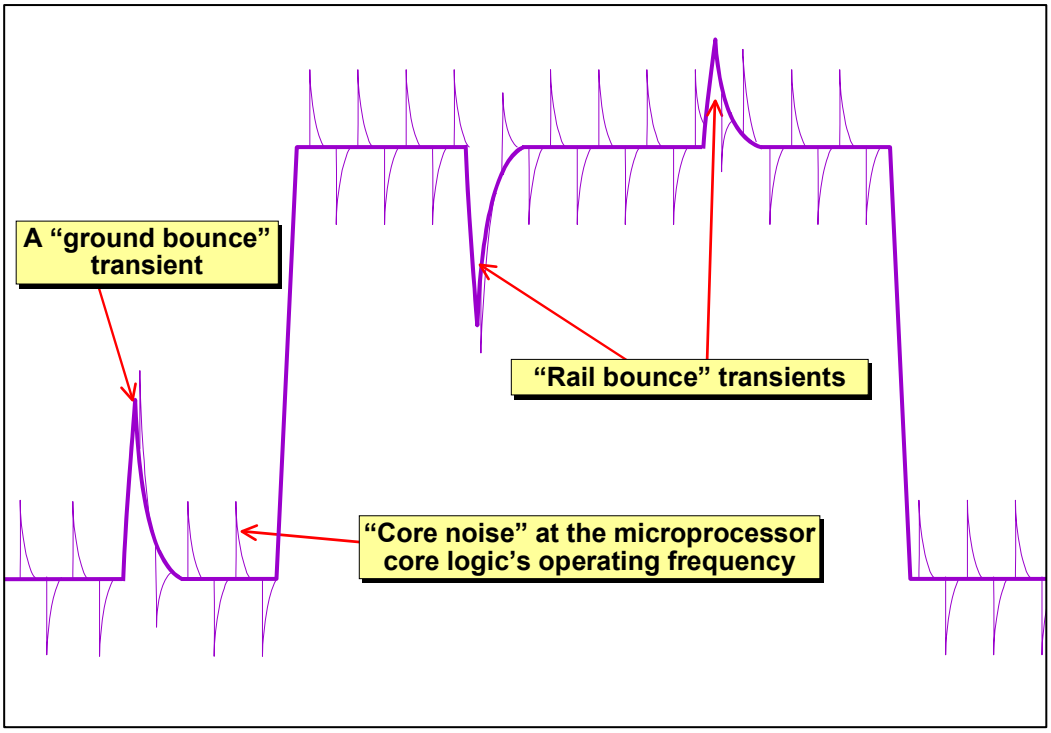


Figure 20 Real noises on processor I/O pins

Ground Bounce and Power Bounce are also called simultaneous switching output (SSO) noise, or simultaneous switching noise (SSN). They arise because of the inevitable inductances between the transistors in the IC's silicon chip itself, and the ground and power supply rails on the PCB.

Even if we could achieve perfect power supply decoupling on the PCB, the inductance in the bond wires and lead frames within the IC's package is enough to cause the silicon chip's internal ground and power rail voltages to "twitch" with respect to the PCB's ground and power rails, every time an output transistor switches from high to low or vice-versa.

When several outputs change state at the same time, in the same direction (e.g. High to Low) their ground/power bounce noises add up linearly – e.g. ten outputs means ten times the noise level. Sometimes this effect can cause the ground/power bounce noise to be so high as to exceed the logic thresholds of other devices, causing what appears to be a software bug (but is really software-related internal interference within a product, one kind of what we might call "Internal EMI").

The adding-up of output switching noises is one reason why fixing a software bug, or making a small improvement to the software, can sometimes have a surprisingly large effect on emissions – maybe even making them better.

I understand that some clever software writers can program their devices so as to limit the number of outputs that change state the same way at the same time, to reduce emissions. And some FPGAs and ASICs are available with internal clocks that are used to switch outputs sequentially, fast enough to be within the skew limits for the external circuit (maybe just a few picoseconds apart). So instead of a (for example) ten times larger ground/power bounce transient, there is a fast burst of ten small transients – which prevents their levels from adding up.

This "twitching" of the IC's internal ground/power rails with respect to the PCB's rails as the outputs change state puts ground/power bounce noise onto <u>all</u> of its I/O and voltage reference pins – as CM noise. Section 5.5 will show that CM noise is often the dominant contributor to emissions below 1GHz.

Core logic noise (see Figure 20) is caused by crosstalk inside a microprocessor, between the very fast core logic circuits and the much slower I/O, voltage reference and power supply circuits and their pins. Also, the switching of the core logic transistors causes ground/power bounce in the same way as the much larger but slower currents in the output transistors. Even low-cost microprocessors these days can have cores that operate at GHz. Core noise is also CM.

And of course the complexity of the signals and noises that can contribute to a product's emissions in real life is not simply limited to the differential-mode (DM) signals and CM noises associated with a single IC – there are usually many ICs, DC/DC converters, etc., all contributing to noise on the DC power rail(s) and coupling noise into traces, wires and cables, and the enclosure, where it is not wanted.

(DM signals are our wanted signals, control, data and power, discussed in more detail in section 5.5 later.)

4.4 Apertures and metal structures as accidental antennas

4.4.1 Accidental slot antennas

Just as any conductor in an insulator like FR4 PCB material or air, behaves as an accidental antenna, with strong antenna characteristics at/near its resonant frequencies – an aperture (gap, seam, hole, etc.) in an area of conductor also behaves as an accidental antenna, and has stronger antenna characteristics at/near its resonant frequencies.

Apertures in metal areas can be intentionally used as "slot antennas", so we can say that all other apertures, gaps, etc. that are not intentional antennas, are "accidental slot antennas".

Figure 21 follows on from Figure 17 and Figure 18 but instead of dealing with accidental conductor antennas (e.g. Figure 19) it describes the simplified, idealised behaviour of accidental slot antennas. Slots resonate at half-wavelengths, so Figure 21 is simply Figure 19 with the maximum dimensions of the accidental antennas doubled.

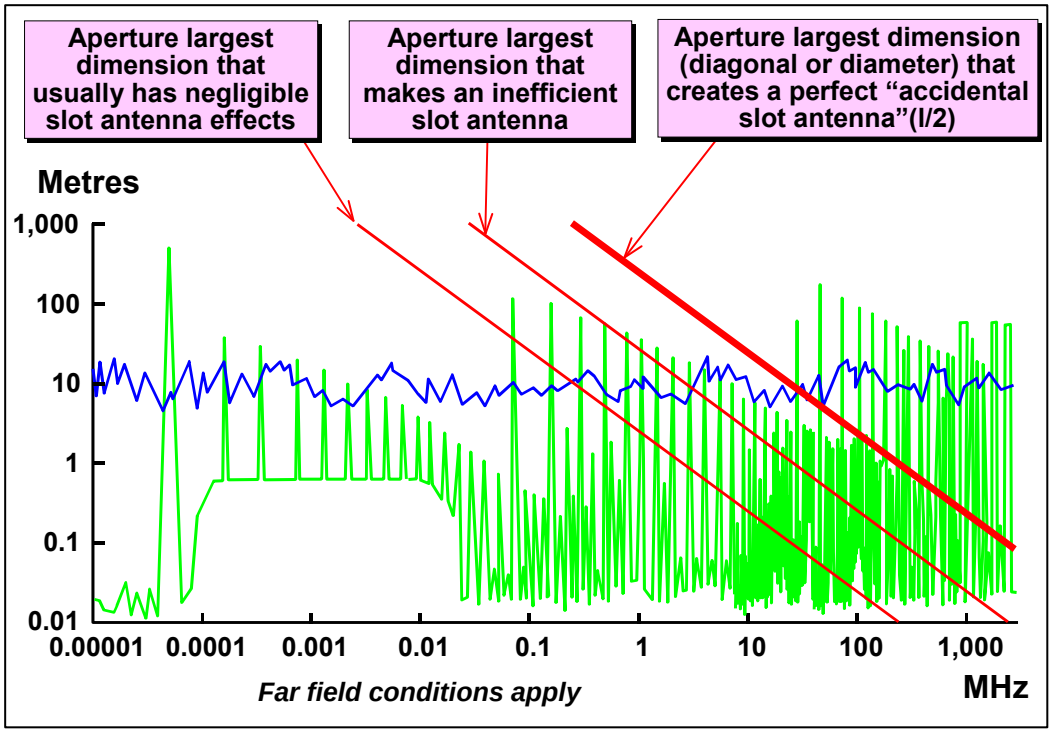


Figure 21 All apertures behave as accidental slot antennas

Like Figure 19, Figure 21 includes two lines that indicate lower levels of accidental slot antenna efficiency over certain frequency ranges. And – just as for Figure 19 – the left-most line shows what dimensions would generally be negligible for ordinary consumer,

commercial and light industrial applications. But, again like Figure 19, these might not be good enough assumptions to meet tougher emissions or immunity test standards.

Section 4.3 of Part 4 of [1] describes how the diversion of surface currents by shield apertures is the cause of their E and H field radiation (i.e. their accidental antenna behaviour, including its resonances) so it will not be repeated here.

4.4.2 Structural resonances

So far, we have only considered the effects of self-resonance on the antenna efficiency of accidental antennas and accidental slot antennas. But structural resonances, such as shown in Figures 8a, 8b above, and Figures 9 and 10 in [8], can also influence the efficiency of accidental antennas, whether they are conductors or slots. Similar figures to those above could be drawn showing how accidental antenna and slot antenna efficiencies can be boosted (or suppressed) by 2-dimensional structures (e.g. printed-circuit boards) and 3-dimensional structures (e.g. metal boxes), possibly resulting in high levels of emissions (or poorer immunity) at those frequencies.

For example, Figure 11 of [8] shows a 3-D field simulation of the (3, 0, 0) mode of the cavity resonance inside a metal box, and the apertures (e.g. for ventilation) that are in the higher field strength areas due to the cavity resonances will have increased levels of surface currents, and so will radiate more EM energy than those located in areas of low field strength at that resonance.

Section 4.3.8 of Part 4 of [1]) also discussed the effects of structural resonances on accidental slot antenna behaviour.

When a structure resonates, the four coupling effects (described in Section 5, following) are dramatically affected. The coupling of EM energy, say from the near-fields associated with an IC or trace on a PCB, to nearby cables or other conductors or conductive structures and shield apertures can be increased by the structural resonance, often by up to 100 times (+40dB), possibly even more.

So structural resonances can have very significant effects for emissions and immunity – especially so when a signal has its fundamental frequency or one or more harmonics close to one or more resonant frequencies.

4.5 Some very simplified formulae for accidental antennas

Many of the following equations are from Appendix D of [13].

4.5.1 Fields emitted by DM or CM currents

For a small wire loop (maximum dimension <<λ/6 at the highest frequency of interest) the maximum possible far-field E-field emission (maximised by varying antenna height over a groundplane using the normal CISPR emissions-testing method) is given by:

$$\mathbf{E = (263)\cdot 10^{-16}(f^{2}\cdot A\cdot I_{DM})/R \quad V/m}$$ (divide by 2 for free-space emissions, i.e. no groundplane)

For a short monopole (a wire perpendicular to large 0V plane, max dimension <<λ/6 at the highest frequency of interest) the maximum possible far-field E-field emission (maximised by varying antenna height over a groundplane using the normal CISPR emissions-testing method) is given by:

$$\mathbf{E = (1.26)\cdot 10^{-6}(f\cdot L\cdot I_{CM})/R \quad V/m}$$ (divide by 2 for free-space emissions, i.e. no groundplane)

Where: E = electric field in Volts/metre at the measurement distance R
f = the frequency in Hz
λ = the wavelength in metres
A = the loop area in square metres
L = the length of monopole in metres
I_{DM} = the DM (loop) current in Amps
I_{CM} = the CM (monopole) current in Amps

R = the measurement distance from the loop or monopole in metres

4.5.2 DM voltage noise picked up from external far-fields

For a small wire loop (max dimension <λ/2) the maximum possible DM voltage induced in it by an external far-field H-field is:

$$V_{DM} = (8)\cdot 10^{-6}(f\cdot H\cdot A) \quad \text{Volts}$$

But if the maximum DM current loop dimension is longer than λ/2, $A=\lambda^2/4\pi$ gives the highest voltage possible in *any* size of loop, hence:

$V_{DM}(max) = (188.5)\cdot H\cdot\lambda$ alternatively: $(5.73)\cdot 10^{10}\cdot H/f$ **Volts** (no control over loop area)

For a small wire loop (max dimension <λ/2) the maximum possible DM voltage induced in it by an external far-field E-field is same as the above equation divided by 377 (the impedance of free space, in Ohms):

$$V_{DM} = (2.1)\cdot 10^{-8}(f\cdot E\cdot A) \quad \text{Volts}$$

But if the DM current loop's maximum dimension is larger than λ/2, $A=\lambda^2/4\pi$ gives the highest voltage possible in any size of loop, hence:

$V_{DM}(max) = E\cdot\lambda/2$ alternatively: $(1.5)\cdot 10^{8}\cdot E/f$ **Volts** (no control over loop area)

Where: V_{DM} = the DM voltage induced around the loop by the external field, in Volts
f = the frequency in Hz
λ = the wavelength in metres
H = the external magnetic field in Amps/metre
E = the external electric field in Volts/metre
A = the loop's area in square metres

4.5.3 CM voltage noise picked-up from external far-field E-fields

For a short wire monopole (perpendicular to reference plane, maximum length λ/4) the maximum possible CM voltage induced by an external far-field E-field is:

$$V_{CM} = E\cdot L \quad \text{Volts}$$

But if the wire length is longer than λ/4, L = λ/4 gives the highest voltage possible in any length of wire, so:

$V_{CM}(max) = E\cdot\lambda/4$ alternatively: $(0.75)\cdot 10^{8}\cdot E/f$ **Volts** (no control over loop area)

For a small wire loop (max dimension <λ/4) the maximum possible CM voltage induced in it by an external far-field E-field is:

$V_{CM} = E\cdot 2\pi\cdot A/\lambda$ **Volts** (for a given loop, this gives the same V_{CM} in Volts as I_{DM} in Amps)

But if the wire loop's maximum dimension is larger than λ/4, $A=\lambda^2/4\pi$ gives the highest voltage possible in *any* size of loop, hence:

$V_{CM}(max) = E\cdot\lambda/4$ alternatively: $(0.75)\cdot 10^{8}\cdot E/f$ **Volts** (no control over loop area)

For the induced CM current, divide the CM voltage by the (complex) CM impedance of the affected circuit (vector calculation finds the phase angle between the induced current and the induced voltage).

Where: V_{CM} = the induced CM voltage in Volts
E = the external electric field in Volts/metre
A = the loop area in square metres
λ = the wavelength of the external electric field in metres
L = the length of the wire in metres

For a *very* great deal more information on antennas, whether intentional or accidental, including lots of mathematics, see Chapter 7 of [14]. And for even more maths, see [15].

5 Coupling of EM energy

As Figure 22 shows, there are three parts to every EMC issue:

a) A source of EM energy, which could be a natural phenomenon like ESD or lightning, or an electrical or electronic circuit with fluctuating currents and/or voltages. (In fact, we can't have one without the other, see number viii) in the eight basic principles that introduced this series, in Section 1.3 of [8]). EM energy sources are sometimes referred to as EM threats.

b) A path by which EM energy can couple from Source (threat) to Victim

c) A "victim", an electrical/electronic circuit that picks-up the EM energy as a noise that could possibly interfere with its operation. For example, when testing emissions, the victim circuit is a calibrated antenna connected to a spectrum analyser or EMC receiver.

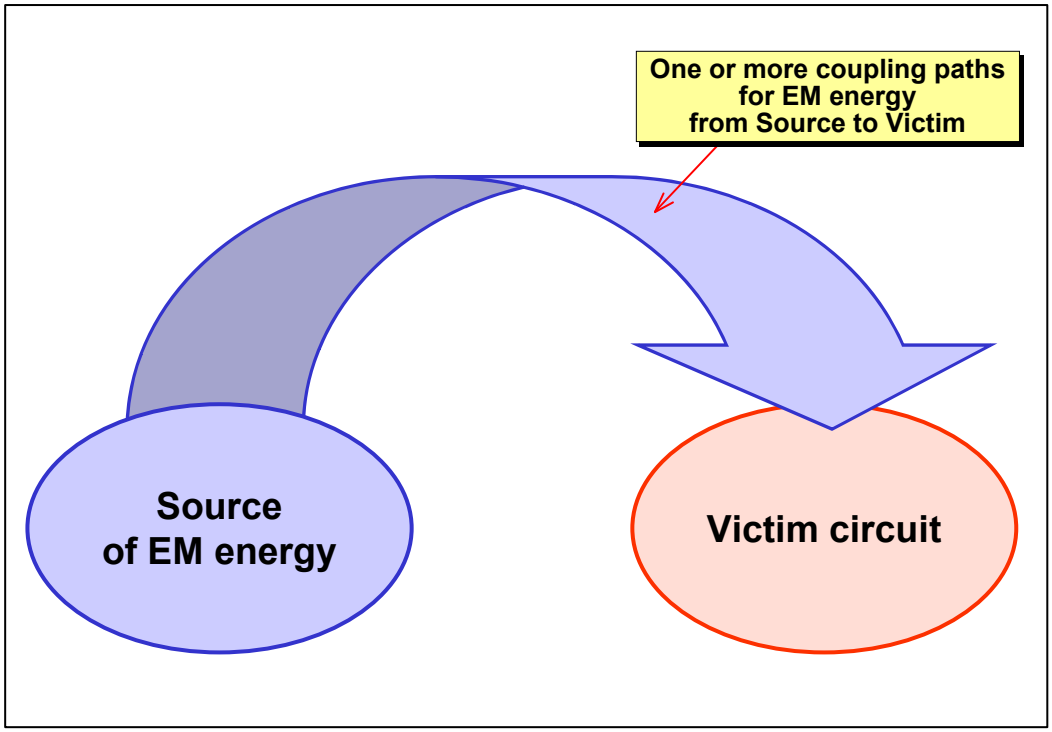


Figure 22 Three parts to every EMC issue

There are four types of EM coupling, which will be discussed in following subsections:

1) Common impedances
2) Electric (E) fields
3) Magnetic (H) fields
4) Electromagnetic (EM) fields

5.1 Common-impedance coupling

5.1.1 How it arises

Section 3.1 showed that every conductor has intrinsic ("self") resistance, R, inductance, L, and capacitance, C. So it doesn't matter if it is a PCB trace; a wire of any gauge; a solid metal bar with any diameter or cross-sectional area; an infinitely large sheet of metal (only found in equations) of any thickness; even a superconductor – it always *always* <u>*always*</u> presents impedance to the flow of AC current.

(A superconductor just has a zero resistance, it still has the same L and C as if it were a normal conductor of the same dimensions and shape.)

And it doesn't matter what *name* we give our conductors: e.g. 115/230V mains supply;

14Vdc automotive battery supply; "ground" or "earth"; 0V; connector pin; cable shield; chassis, frame; enclosure shield; "Faraday cage"; reference plane; support bracket; I-beam girder; hull; armour; superconductor, etc., etc., it *always* presents impedance to the flow of AC current, as before.

So whenever we share a conductor between two or more circuits, the currents that circuit 1 causes to flow in the impedance of the common conductor generate voltages that appear as noise to circuit 2. And vice-versa.

We call this effect "common-impedance coupling", and it always happens. We strive to reduce it to negligible amounts, for signal integrity as well as for EMC, but we can never eliminate it.

The impedance of PCB traces, wires and similar long conductors generally increases at higher frequencies, because skin effect increases the resistance, R, see Section 3.1.1 in [8], and because of increasing inductive reactance: $X_L = 2 \cdot \pi \cdot f \cdot L \quad \Omega$ (L in Henries, f in Hz).

The noise voltages that arise in the impedance of a shared conductor are CM, because they are common to all the circuits that share it.

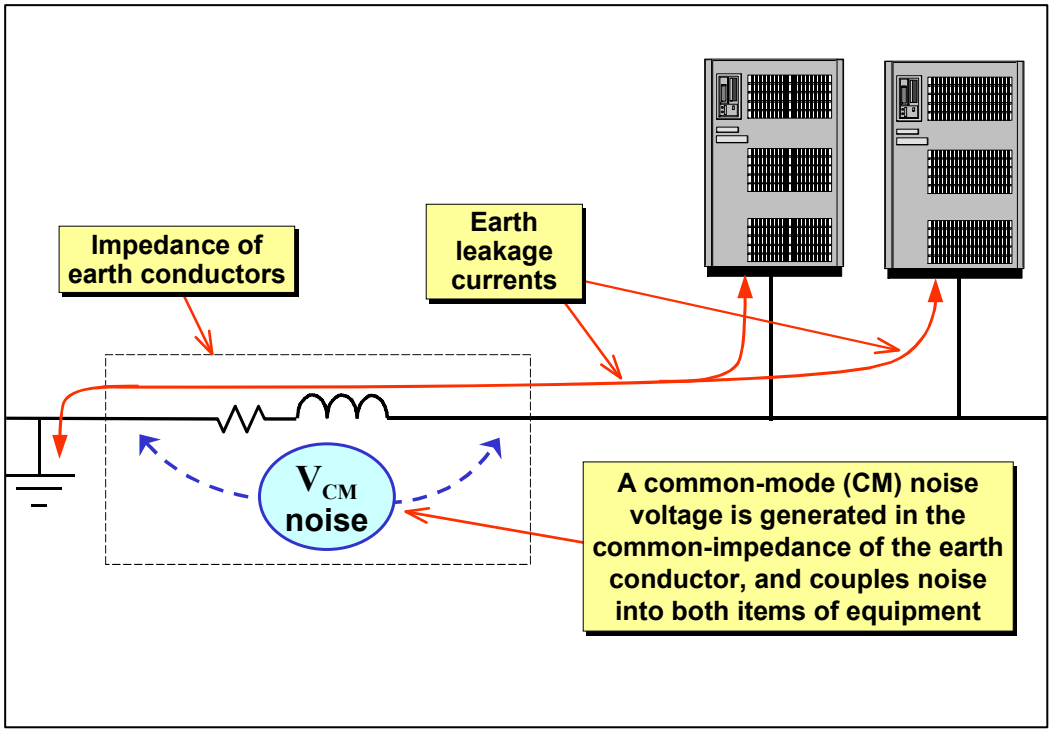


Figure 23 Simple examples of common-impedance coupling

Figure 23 shows two different items of equipment sharing a safety earth wire. The earth leakage currents of one, the source, cause a CM noise voltage as they flow in the impedance of the earth wire, which then couples EM energy into the other item of equipment, the victim.

Of course, the victim is also a source of earth leakage current that creates CM noise in the earth wire that couples to the first equipment, so they are both sources and victims of EM energy according to the simple scheme of Figure 22.

Figure 23 shows items of equipment connected to a building's earthing system, but similar sketches could be drawn for different circuits sharing a power or 0V rail – for example an input amplifier and digitiser, digital signal processor, display driver, I/O processor, etc. They all act as sources and victims for common-impedance coupling through their shared power and 0V rails.

Cable shields also carry currents, and if the chassis, frame or enclosure shield they connect to does not have a low-enough impedance at the highest frequency of interest, their currents will not be well controlled. Currents that should have stayed within the shields get onto the

outside, causing excessive emissions, and currents that should have stayed on the outside get into the inside, causing problems for immunity. All because of common-impedance coupling through the chassis, frame or enclosure shield.

A typical example is the card cage – multiple vertical or horizontal cards slid into a metal-sided cage, each with a narrow L-shaped piece of metal to act as a front panel that is fixed to the card cage by quick-release captive screws at top and bottom. Where there are one or more cable connectors on a front panel, it is often necessary to fit conductive EMC gaskets along both sides of each card's panel so that they all bond together – and to the card cage end-cheeks at both sides – to create a low-enough common-impedance for the front panels so that their cable shields (or filters of unshielded cables) work as well as intended.

5.1.2 Why single-point earthing/grounding is no longer a solution

The single-point earth/grounding/bonding technique (sometimes called "star-earthing/grounding") was originally developed to overcome the problem of common-impedance coupling at 50/60Hz in badly-designed electronics, in the 1920's or thereabouts. Since then it has become some sort of unquestionable dogma for some designers and (it seems) almost the entire electrical installation industry worldwide.

But the problem is that this old-fashioned technique *can't* control current flow at high frequencies, and causes huge and very costly problems above a few kHz, never mind tens or hundreds of MHz. At these frequencies the propagating EM energy that we call signals, control, data and power can more easily flow through the air into nearby conductors via E, H and EM field coupling, than flow in the high impedances of long PCB traces or lengths of green or green/yellow wire.

So the only references to single-point earthing/grounding techniques you will find in anything written by me is to deprecate them. It was fine up to the 1950's when radio broadcasting was AM and no-one had heard of digital circuits or switch-mode power conversion, much less cellphones, but is not at all a good technique these days (unless you are designing baseband analogue equipment with linear power supplies for use in EMC screened rooms with other baseband analogue equipment with linear power supplies). (But audiophiles who are prepared to shield their listening rooms and keep all other equipment outside of them are rather a limited market.)

Incidentally, I have a copy of Wireless World magazine dated 1947, which contains an article questioning whether there is any future in television broadcasting. Nothing to do with this article, but such perspectives are always fun!

5.1.3 Circuit design education is badly flawed

Circuit design is taught in schools, colleges and universities as if power rails and 0V returns have zero impedance. Not just zero R, but zero L and C too, hence no E or H fields can exist. Just look at any circuit design textbook – the power rails and signal returns are not drawn on its schematics. Often even the device power and return pins are not shown.

RF circuit design is an exception, because it teaches people to use transmission lines which have a send and a return conductor, but even RF design textbooks don't bother to draw in the power rails, or the return conductors for signals that are no transmission lines.

So when we learn to do circuit design, we also *assume* that it does not matter how the power and signal currents return, to complete Kirchoff's Law (paraphrased: all currents flow in closed loops). They obviously manage to do it somehow, but it can't be at all important how or when they do. Otherwise the textbooks authors would have said something about it, wouldn't they!

This leads circuit designers completely astray, in turn leading to the common practice of them leaving it up to their PCB designers to figure out which pins on the devices are supposed to be used for the boring unimportant power and signal returns, and connecting them up somehow, anyhow, because their routes don't matter at all.

If there is a better way of misleading students to ensure that their circuits will suffer from poor signal quality / signal integrity, and suffer and/or cause awful levels of EMI – I have yet

to see it. The first time a designer discovers non-flat frequency responses or overshoot and ringing where his graduate-level circuit theory says there should be none, is the start of another long learning curve.

This approach to teaching people how to do circuit design almost *guarantees* that circuits will not function as well as required in real life (and of course legal EMC compliance has no hope).

The result is that most electronic engineers have a great deal to learn when they first start work, and they learn by making the same mistakes that everyone else has made, on real projects, iterating their designs and delaying market introduction until they somehow get them to both function and comply well enough as they climb their learning curve, at huge cost to their employers.

(There are some exceptions, of course. Some educational establishments teach budding designers what they *really need to know*, but they are lighthouse beacons of hope in a storm-racked sea of darkness.)

Modern digital devices generate frequencies that are so high in both level and frequency (see Figures 12, 17 and 19) that having low common impedances in their power rails and 0V "grounds" at the highest frequencies, and taking good care that the paths taken by all return currents are very close indeed to the path taken by their "send" conductors (see 5.2 and 5.3 following), are *absolutely vital* for signal integrity, low emissions and good immunity of digital circuits; the signal quality and immunity of analogue circuits – even "DC" and low-frequency instrumentation/audio circuits and the unintentional emissions of RF circuits.

Electric (E) field coupling

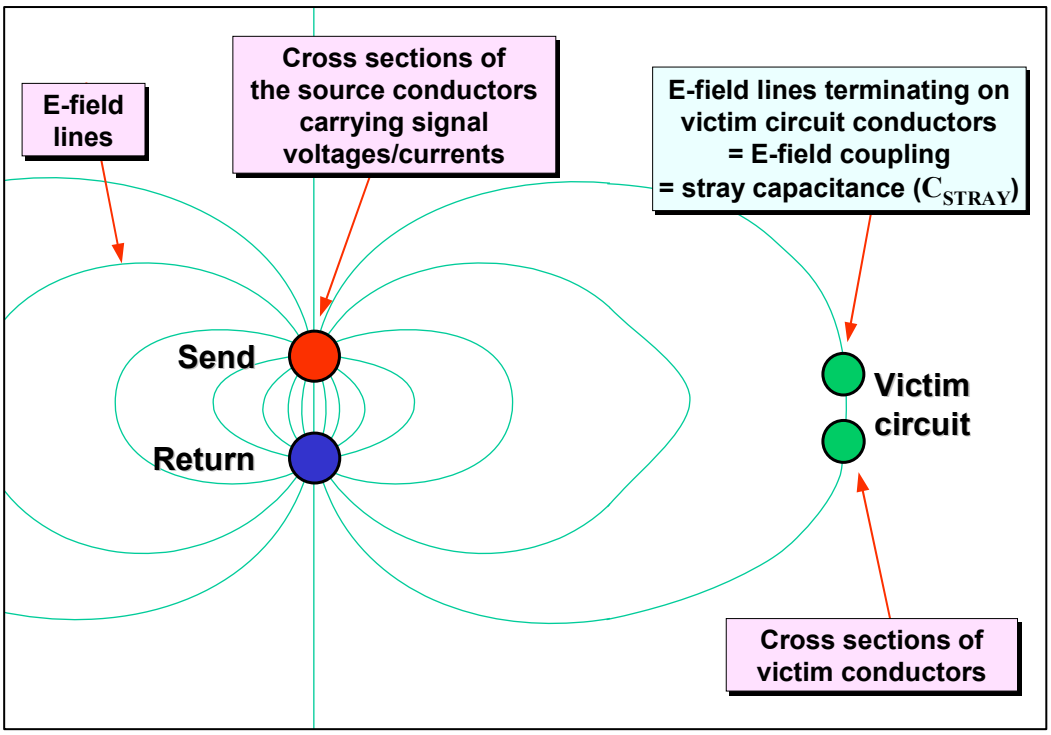


Figure 24 Example of E-field coupling

E-field lines are drawn so that they always terminate on conductors, at 90° to their surfaces. Of course these lines are not real, they are similar to the contour lines on a geographical map. The density of the lines tells us how strong the E-field is in Volts/metre, just like the density of geographical contour lines tells us how step a slope is. So we can see from Figure 24 that the highest E-field strength is exactly between the Source circuit's send and return conductors.

But we also see that the fields spread out in the space around the conductors, getting

weaker the further away they spread. If another conductor is nearby, the local field pattern (distribution of field lines) will be distorted as some of the lines terminate on those other conductors.

When E-field lines terminate on other conductors, this represents E-field (EM energy) coupling from the source to the victim as sketched in Figure 22. Another term for E-field coupling is stray capacitance, and Figure 25 shows the equivalent circuit.

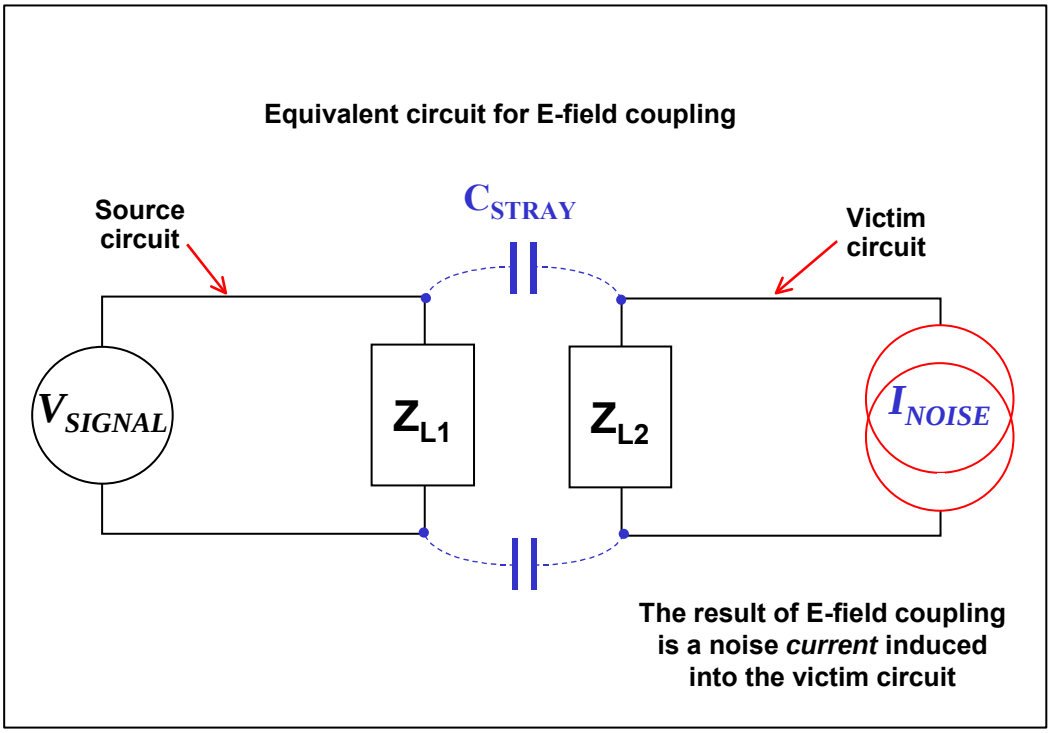


Figure 25 Equivalent circuit for E-field coupling

E-field coupling causes noise currents to be injected into victim circuits.

Basically: $I_{COUPLED} = C_{STRAY}\,(dV_{SIGNAL}/dt)$

This seems like a nice simple time-domain equation (which can also be simply expressed in frequency domain terms too) – but like all simple EMC equations it makes certain assumptions that often limit its use in real life. The assumptions in this case are that the victim circuit's impedance is << $1/(2\pi \cdot f \cdot C_{STRAY})$, and that the maximum dimension of the victim circuit's current loop is << $\lambda/6$.

If the above assumptions hold, a stray capacitance of just 0.1pF between a victim circuit and a 100MHz 5V peak-peak squarewave signal would couple 310µA of peak-peak noise current into the victim's circuit at the 100MHz fundamental frequency, also 310µA at each of its harmonics (3rd, 5th, 7th, etc.).

As we will see when we come to consider CM noise in section 5.5 later on, it only needs about 5µA of CM noise in an external cable to fail to meet the CISPR 22 Class B emissions limits at some frequency. So just 0.1pF of stray coupling, say between two short lengths of PCB traces, could cause a radiated emission test to be failed by as much as 36dB!

The reactive impedance of C_{STRAY} reduces as frequency increases:

$X_{CSTRAY} = 1/(2\pi \cdot f \cdot C_{STRAY})\ \Omega$ (C_{STRAY} in Farads, f in Hz)

— so reducing a signal's rate of change of voltage (dV/dt) reduces its high frequency content (which can be seen by a Fourier transform) and therefore reduces its E-field emissions, and the noise current that it injects (couples) into victim circuits.

So when we are designing, we want to have a low stray capacitance (low E-field coupling) between the source and victim circuits to reduce their noise coupling (interference).

But for a single signal's own send/return paths, we generally want a high capacitance

between its send and return current paths, because this makes its E-fields more compact, reducing its stray capacitance to other circuits, thereby reducing the amount of noise coupling (interference).

Figure 26 shows how this works, with two circuits Source A and Source B that are identical in every way and carry identical signals, except that Source B's send and return conductors are closer together. (E-field pattern distortion by the victim circuit is not included in this sketch.)

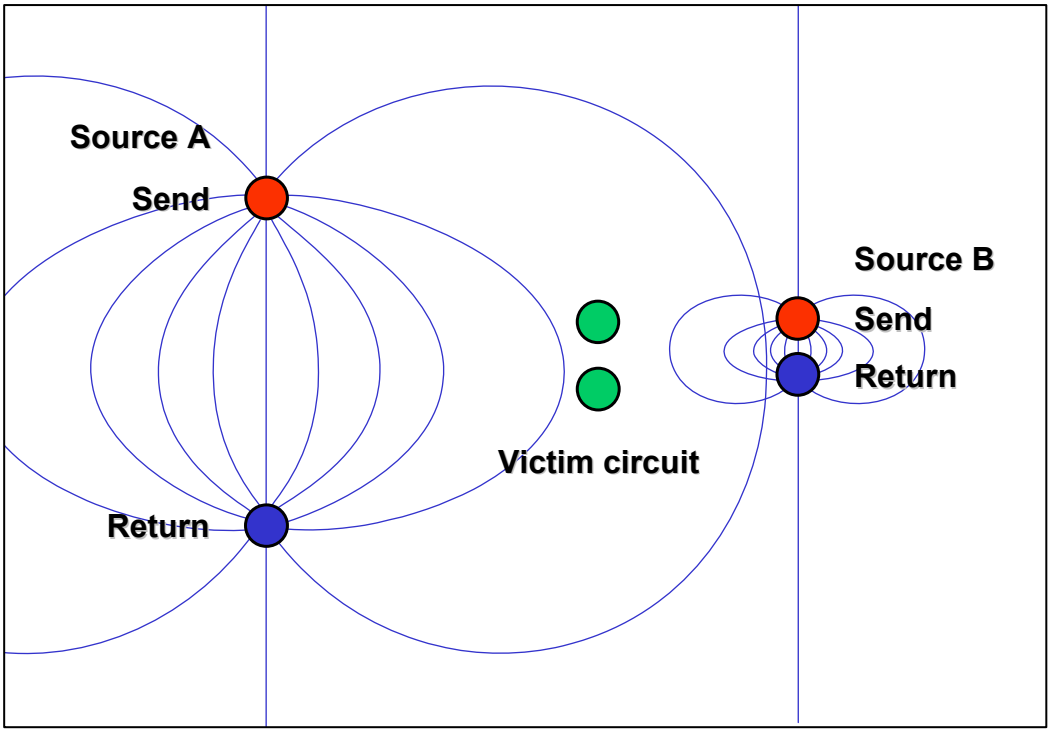


Figure 26 Reduce C_{STRAY} coupling by routing send/return conductors closer together

Figure 26 shows that source A couples to the victim circuit more strongly than source B, because source B has its send/return conductors closer together, creating more capacitance between them and making its E-field more compact overall. Being more compact, it has less stray E-field coupling (less stray capacitance, lower induced currents) to the victim circuit.

Figure 26 is in fact a graphical representation of number vii) in the eight basic principles that introduced Part 1 of this article, in Section 1.3 of [8].

Routing victim circuit send/return conductors closer together also reduces stray capacitance (E-field) coupling from sources of EM energy.

Stray capacitance is also a function of conductor surface area, so reducing the surface area of the source and/or victim circuits is another way to reduce E-field coupling. Make source and victim circuits smaller.

We must not forget one of the best E-field suppression methods of all – distance. Place the source and victim circuits far apart from each other. This doesn't help pass EMC tests – the receiving or transmitting antennas are placed at fixed distances from our products – but it is a very good way to reduce crosstalk and improve "Internal EMI" within a product, and to reduce EMI between different items of equipment in a system or installation.

Also, this little section has assumed that the conductors are surrounded by air (or vacuum). But if they were embedded in a dielectric such as glass, plastic or the FR4 substrate of a PCB – or surrounded by oil or some heat-exchange fluid – the relative permeability (i.e. dielectric constant) of the insulating medium would increase the stray capacitance between source and victim.

Metal areas and volumes can be used to decrease E-field coupling – this is called E-field

shielding and is described in Part 4 of [1].

5.2 Magnetic (H) field coupling

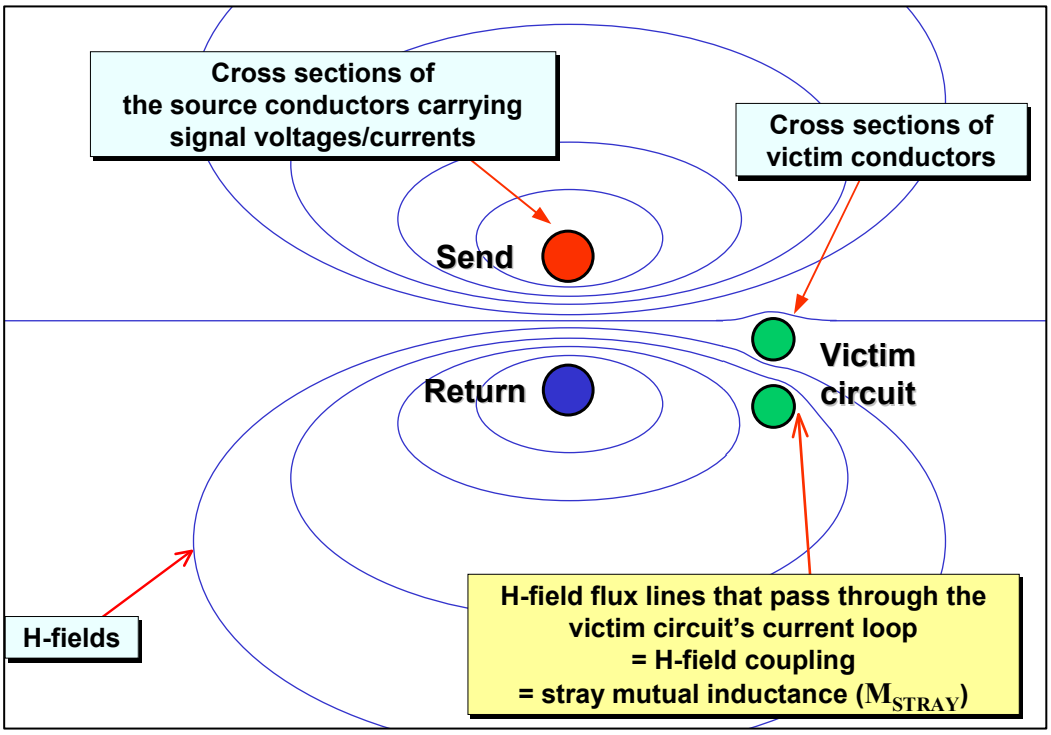


Figure 27 Example of H-field coupling

As Figure 27 shows, H flux lines never terminate on conductors. Just as for E-field lines, it is the density of the lines that tells us the H-field strength, and we can see from Figure 27 that the field is highest exactly in-between the source's send and return conductors. However (and just as for E-fields again) – the H-field lines spread out in the space around the conductors, showing that the H-fields weaken with increasing distance.

When the H-field lines encounter the area enclosed by a victim circuit's current loop, they may be distorted slightly from their free-space patterns, and some lines will pass outside the area of the victim circuit's current loop, whilst others will pass through it.

The lines that pass through the victim circuit's current loop area represent H-field coupling, which can also be described as a stray mutual inductance, M_{STRAY}, and Figure 28 shows the equivalent circuit. Essentially, it causes an accidental transformer to be connected between the source and victim circuits.

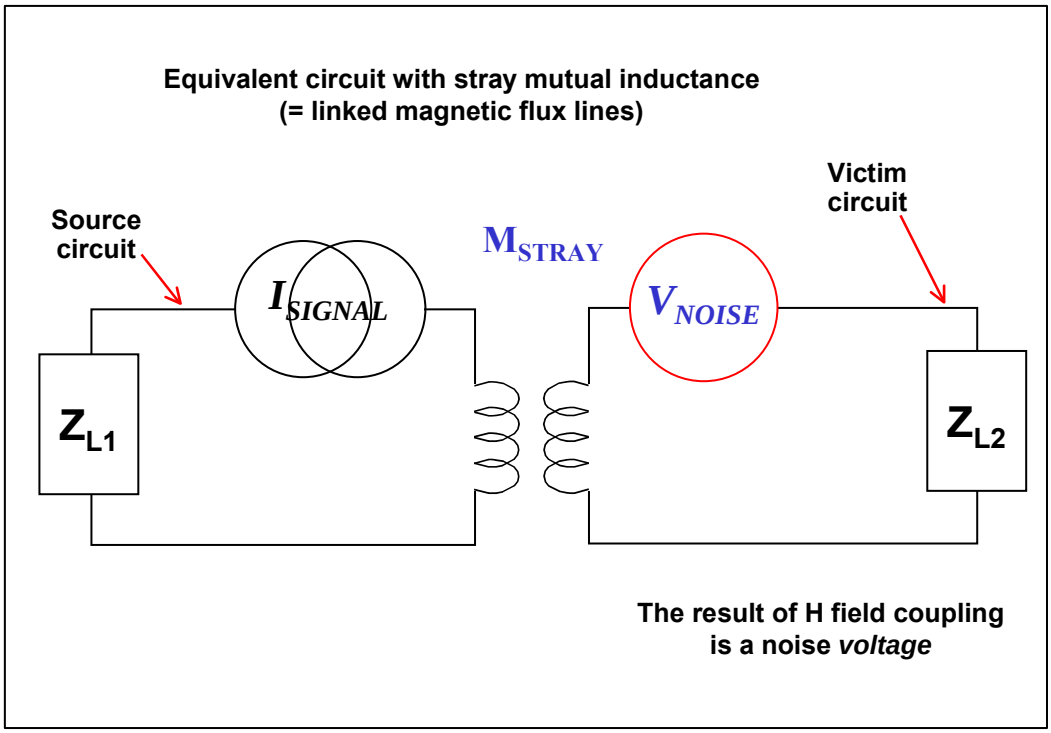


Figure 28 Equivalent circuit for H-field coupling

H-field coupling cause noise voltages to be injected into the victim circuit's loop.

Basically: $V_{COUPLED} = -M_{STRAY}\,(dI_{SIGNAL}/dt)$

This simple time-domain expression is only valid when the victim circuit's impedance >> $1/(2\pi \cdot f \cdot M_{STRAY})$ and the maximum dimension of the victim circuit's loop is << $\lambda/6$.

Given these restrictions, we can see that a stray mutual inductance of only 10nH between a victim circuit and a circuit carrying a 100MHz squarewave at 20mA, would couple 126mV of noise into the victim at the 100MHz fundamental frequency, also 126mV at each of its 3rd, 5th, 7th, etc. harmonics. That's a lot of noise coupling between what might only be two adjacent pins in a DIN 41612 PCB-PCB connector!

Just like the reactive impedance of stray capacitance – the reactive impedance of M_{STRAY} also decreases with increasing frequency: $X_{MSTRAY} = 1/(2\pi \cdot f \cdot M_{STRAY})\ \Omega$ (M_{STRAY} in Henries, f in Hz). So reducing a signal's rate of change of current (dI/dt) reduces its high frequency content (by Fourier transform) and therefore reduces its H-field emissions, and the noise voltage that it injects (couples) into victim circuits.

We want to have low mutual inductance (M_{STRAY}) coupling between different circuits, to reduce their noise coupling (interference), but for a single signal's own send/return paths we generally want a high mutual inductance, because it makes its H-fields more compact – reducing the stray mutual inductance with other circuits, and thereby reducing the amount of noise coupling (interference).

Figure 29 shows how this works, with two sources, A and B, that are identical in every way and carry identical signals – *except* that B's send and return conductors are routed closer together. (H-field line distortion by the victim circuit is not included in this sketch.)

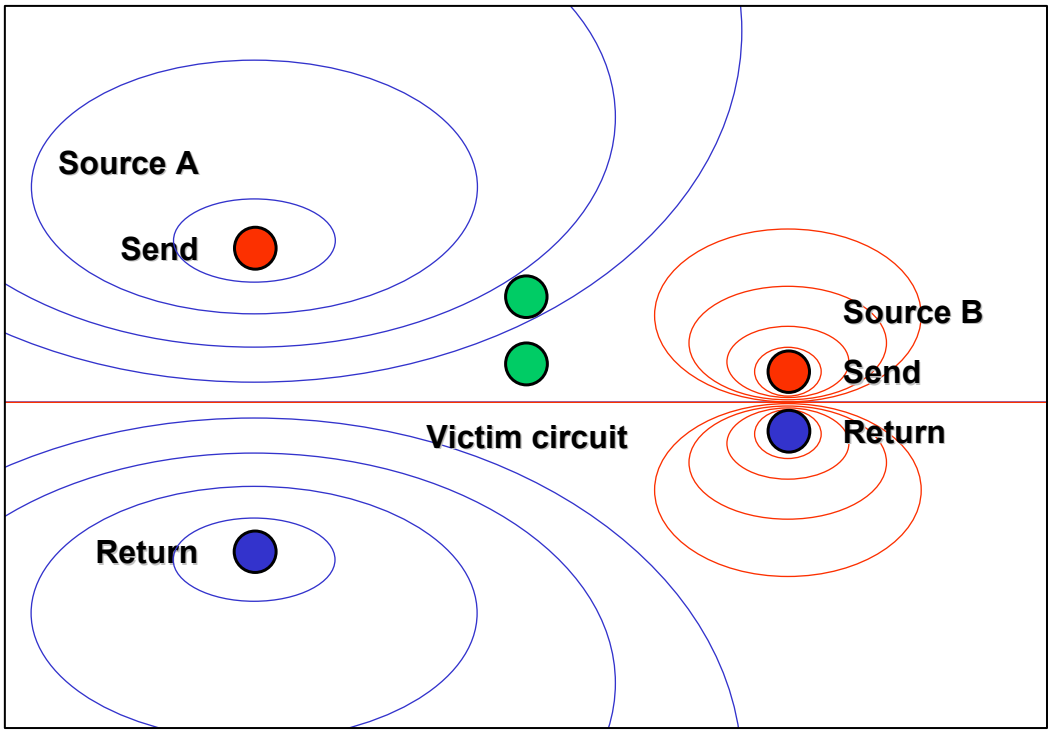


Figure 29 Reduce M_{STRAY} coupling by routing send/return conductors closer together

Figure 29 shows that source A couples more noise into the victim circuit than does source B, because source B's send/return conductors are closer together, creating more mutual inductance between them and making its H-field more compact. It therefore couples less noise (less stray H-field coupling, smaller stray mutual inductance, smaller induced voltages) with the victim circuit. Figure 29 is another graphical representation of number vii) in the eight basic principles that introduced this series, in Section 1.3 of [8].

Routing victim circuit send/return conductors closer together also reduces stray mutual inductance (H-field) coupling from sources of EM energy.

Stray mutual inductance is also a function of current loop area, so reducing the loop areas of the source and/or victim circuits is another way to reduce H-field coupling. Make source and victim circuits smaller.

Just as for E-fields in the previous section, we must not forget one of the best H-field suppression methods of all – distance. Place the source and victim circuits far apart from each other. As before, it doesn't help pass EMC tests, but is a very good way to reduce crosstalk and improve "Internal EMI" within a product, and to reduce EMI between different items of equipment in a system or installation.

Also, this brief section has assumed that the conductors are surrounded by air (or vacuum). But the proximity of metals, and also materials with a relative permeability greater than 1, e.g. iron, steel, nickel, ferrite, "ferrofluid", etc., can significantly increase or decrease the stray mutual inductance between source and victim circuits, depending on the construction.

So, for example, transformers are wound on steel or ferrite cores, to *increase* the mutual H-field coupling between source (the primary winding) and victim (the secondary winding). And metal and/or magnetic materials can be constructed so as to decrease the mutual inductance between source and victim – this is called H-field shielding and is described in Part 4 of [1].

5.3 EM-field coupling

This is the last of our four coupling methods, and it is a bit of a disappointment really because it is simply a development of E and H-field coupling.

Figure 30 shows that in the far field, the E and H near-fields from the source have each become fully-fledged EM fields (plane waves), which have both E and H-field components propagating at the same time.

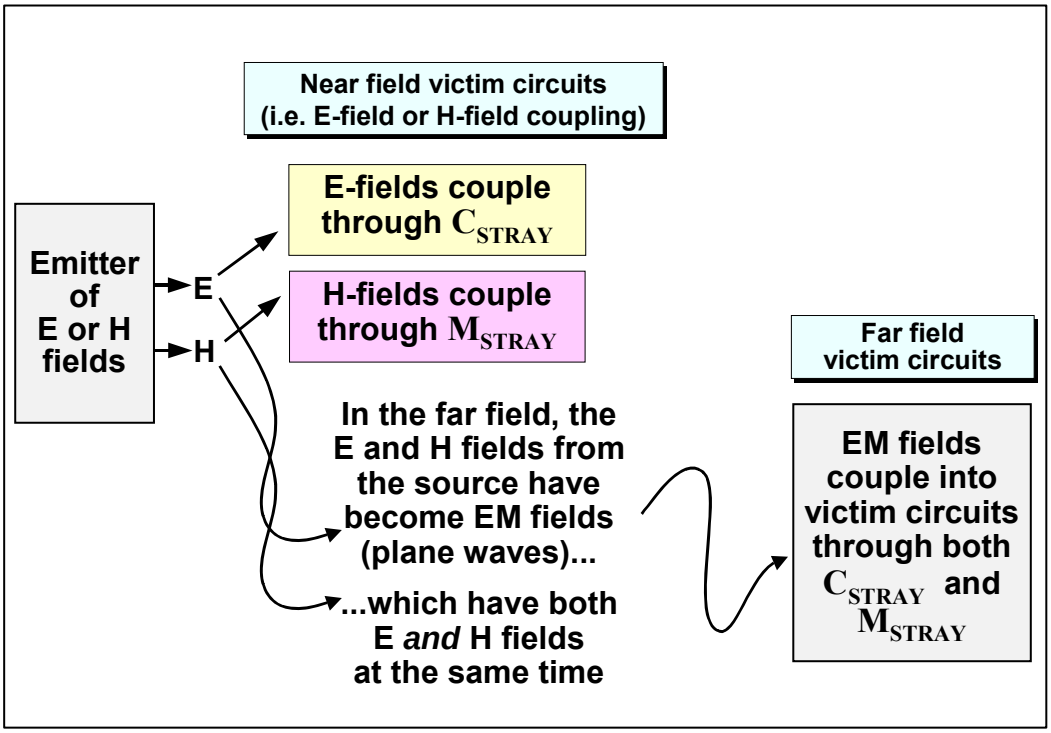


Figure 30 Far-field EM field coupling

So when these plane waves meet a victim circuit, they couple with its M_{STRAY} and C_{STRAY} at the same time. So it doesn't matter whether the source emits an E or an H wave at a given frequency, if the amplitude of the resulting far-field EM wave is exactly the same, it will couple with the victim circuit in exactly the same way.

Note that there is no way for the victim circuit to know whether the origin of the coupled noise was an E-field or an H-field.

The levels of EM fields that are emitted by a source circuit depend upon the levels of the E *and* H fields that it emits, so the same techniques that reduce emissions of E and H fields also reduce emissions of EM fields.

The amount of noise current and voltage picked up by a victim circuit depends on its propensity for coupling E and H fields, so the same design techniques that reduce E and H field coupling also reduce EM field coupling.

5.4 Differential Mode (DM) and Common Mode (CM)

Differential Mode (DM) is where the send and return conductors carry opposing voltages or currents. Wanted (intentionally created) signals and power are always DM, sometimes called 'transverse' because their voltages appear between the send and the return conductors.

But unbalanced 'stray' coupling converts some of the DM signal or power into Common Mode (CM) currents and voltages, which we call "noise" because they are unwanted.

Part 2 of [1] described how conductors in cables attempt to balance their strays, for example by using twisted-pairs, or contain them – for example by using shielding. [12] describes similar techniques for traces on printed circuit boards (PCBs). But nothing is ever perfect, so

there is always a difference between the stray currents and voltages caused by the route taken by the send conductor, and the strays caused by the route taken by its return conductor, resulting in CM noise.

These accidental, unintentional, but nevertheless real and always present CM currents and voltages also have 'stray' couplings into victim circuits. This is called DM to CM conversion, and is a feature of all electronic hardware, active and passive.

CM is sometimes called 'longitudinal', when it appears along the length of a cable. The longer the cable, the more CM is created from the DM signals it carries. DM to CM conversion in cables is generally specified by cable manufacturers as Longitudinal Conversion Loss (LCL) – in dB/metre. It varies with frequency (generally becoming worse as frequency increases) so is specified at certain spot frequencies, or as a graph of dB/metre versus frequency.

LCL is not merely a concern for generating EMI (or picking up EMI from EM fields in the air, since DM-CM conversion works either way around, giving a reciprocal CM-DM conversion). Energy that is lost to the DM signal by conversion into CM noise causes distortion of the signal (see Section 4.1 and 4.2 of [17], and limits the distance the signal can be carried before it becomes unusable. For example, one of the main differences between the various categories of Ethernet cables (e.g. Cat 5, 5e, 6, 6a, 7, 7a etc.) is their LCL at high frequencies.

Figure 31 shows an example of DM (wanted) signals causing CM (unwanted) noises, for a 'floating' load. It doesn't matter if the electronic unit shown in the figure is floating or connected to the chassis or earth, or if the load is floating or bonded to chassis or earth, radio-frequency (RF) energy flows quite happily through quite small values of stray capacitance (see Section 5.2 and 5.3 in [17]) so CM current loops form anyway.

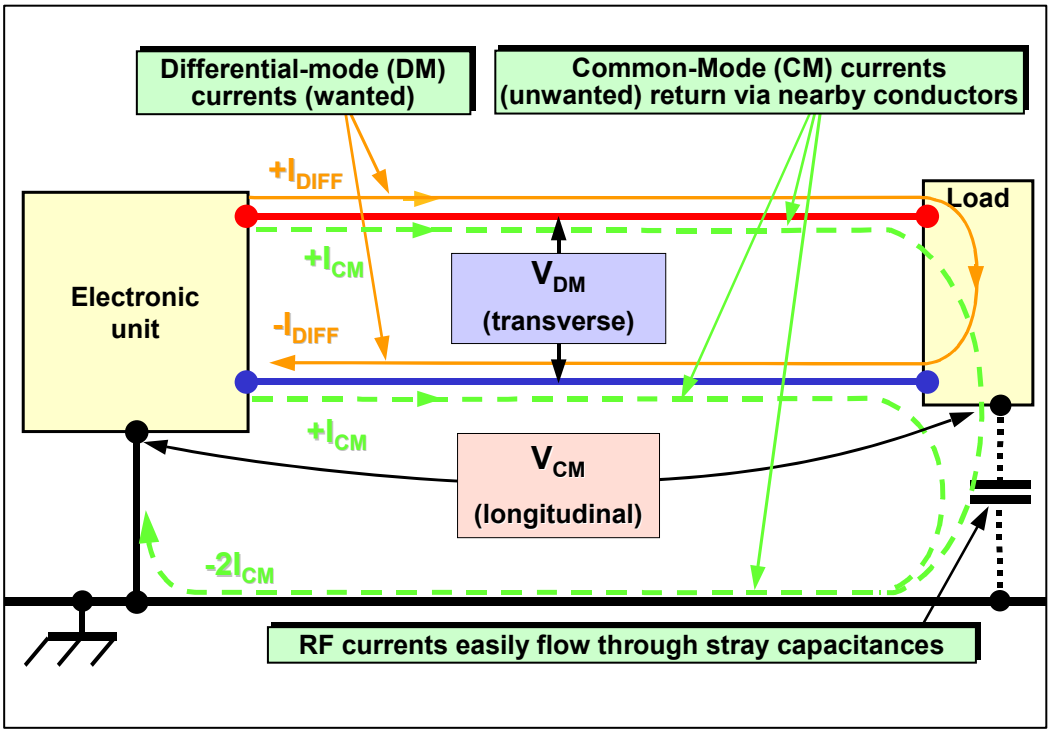


Figure 31 The loops enclosed by DM wanted signals/power and by unwanted CM

Remember – the four types of 'stray' coupling are: Common-impedance; E-field; H-field, and EM-field, (see Sections 5.1 to 5.4 in Part 2 [17]) and they all couple stray CM current and voltage noises just as well as they couple stray DM currents and voltages.

Figure 32 shows an example of CM H-field coupling, between a pair of send/return conductors, and some local conductor (e.g. a metal structure in a wall, floor or ceiling).

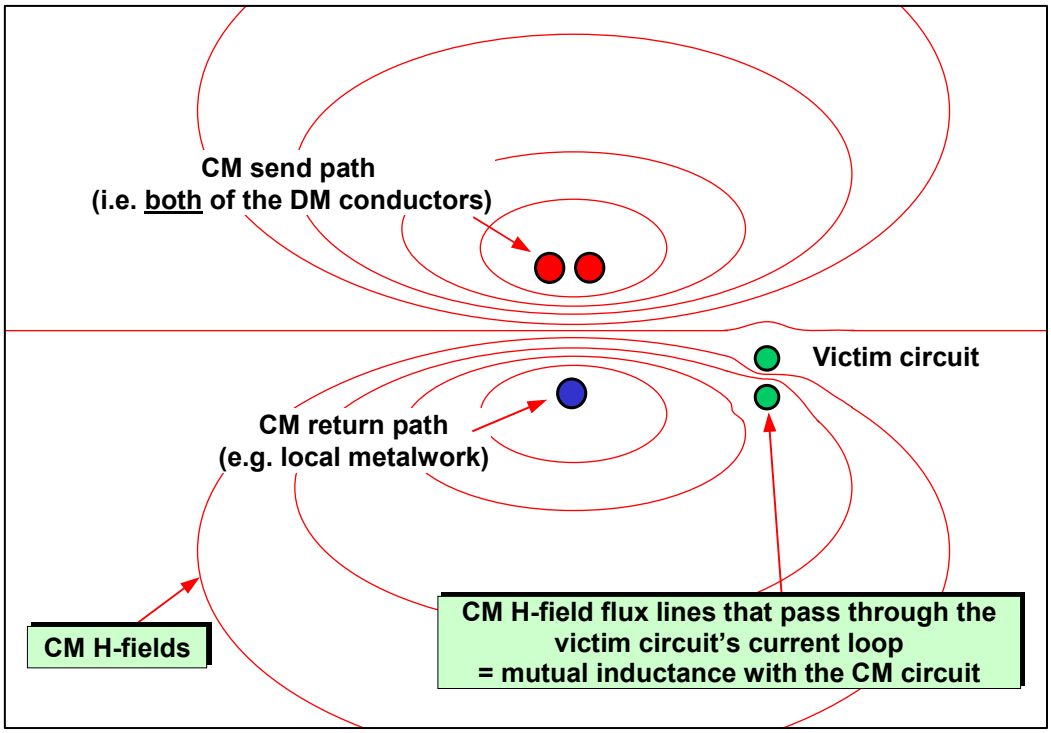


Figure 32 Example of H-field CM coupling

Figure 33 shows an example of a cause of CM. Stray capacitances (E-field couplings) from a logic signal cause stray (CM) currents to flow in all nearby conductors, in this case a metal water pipe.

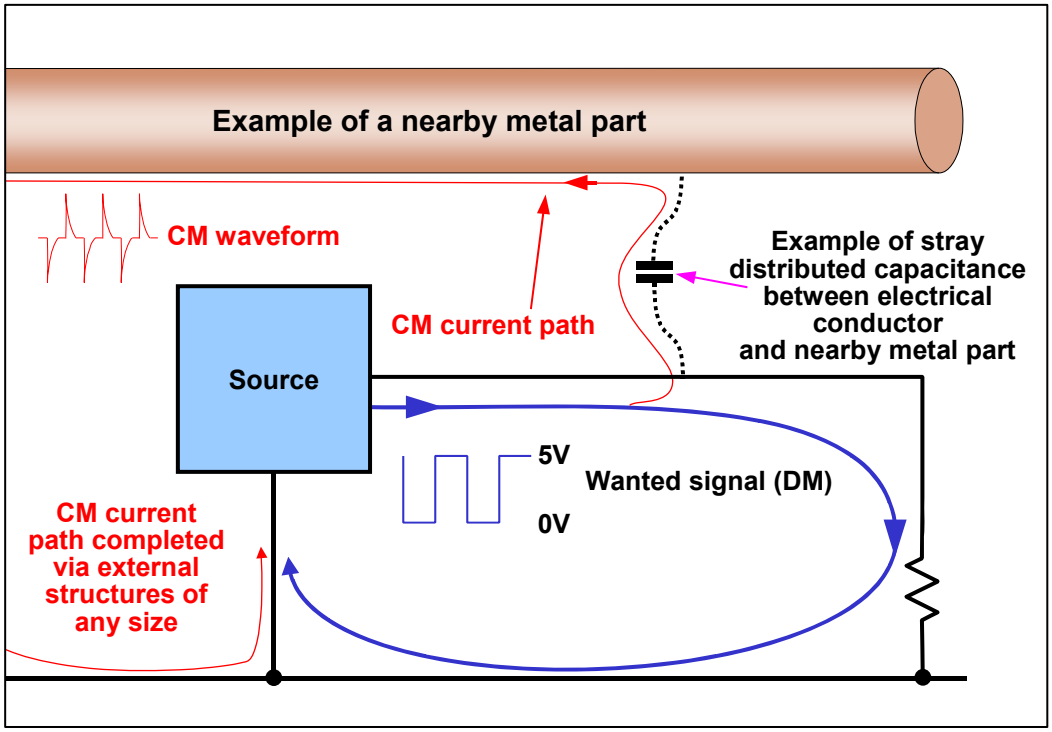


Figure 33 Example of stray capacitance coupling causing CM currents

The best way to reduce the generation of (and susceptibility to) CM, is to use "balanced" or "differential" communications, as shown by Figure 34. These use symmetrical constructions (including PCB layout) and signals/power that are the inverse of each other, to help ensure that stray couplings cancel themselves out as far as practicable.

"Balanced audio" has been in use for many decades, to reduce the power frequency hum

noise that intrudes into microphone and other audio signals when long cables are used. This hum is caused by stray powerline currents flowing in the common impedance of the earth/ground structure, creating significant levels of voltage differences and/or currents at the powerline frequency between different parts of the structure.

The CM to DM conversion ratio (called the Common Mode Rejection Ratio, CMRR) of balanced audio circuits can be as good as -100dB at 50Hz, but maintaining such high levels of balance up to (say) 10kHz is very difficult, and above 100kHz it is virtually impossible.

In recent years, LVDS (low-voltage differential signalling) has become a popular technique to reduce emissions (and improve immunity) for high-speed data transfer. If the balance can be maintained to a high-enough frequency, it can avoid the need for costly shielded cables.

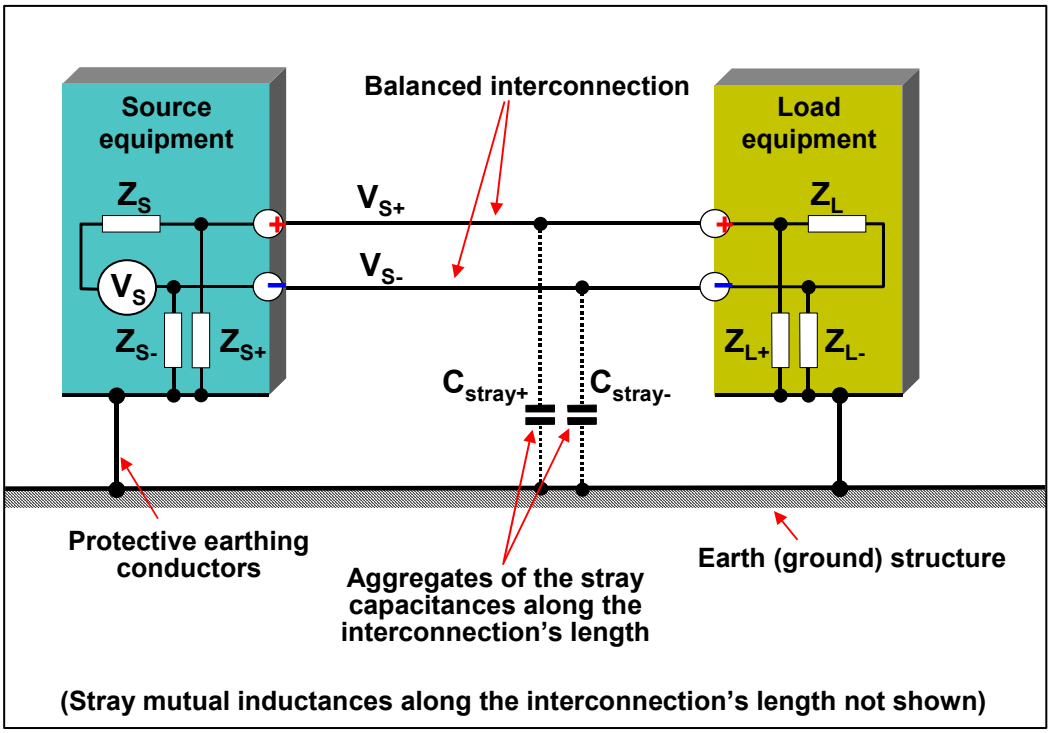


Figure 34 Using balanced interconnections to reduce DM-CM (and CM-DM) conversion

But nothing is ever perfect in the real world, so despite our best efforts, even when we design taking tolerances, soldering shocks, temperature coefficients and ageing into account, we still find that (referring to Figure 34) $Z_{S-} \neq Z_{S+}$, $Z_{L-} \neq Z_{L+}$, $C_{stray-} \neq C_{stray+}$, and also the differential signals are never *exactly* balanced in phase or amplitude, so $V_{S+} \neq V_{S-}$.

The inevitable result is that CM currents and voltages will still arise even when using balanced interconnection techniques, although they can be reduced very significantly compared with using single-ended signalling.

Single-ended interconnections can be made to behave as balanced/differential interconnections very easily at RF, by using CM chokes. Over their effective frequency range, they act as non-isolating "baluns" (balanced to unbalanced converters), and I've solved many EMC problems by adding CM chokes, either soldered onto the PCB or clipped around a whole cable or cable bundle.

Because CM voltages tend to appear across large areas, and CM currents tend to flow in very large loops, CM can cause much higher emissions than a DM signal of the same amplitude and frequency. In fact, the accidental conversion of DM into CM is often the main cause of excessive emissions from 1 – 1,000MHz.

The corresponding conversion of CM signals in the environment (e.g. due to radio frequency fields being picked-up by conductors acting as accidental antennas) into DM noise in electronic circuits is the main cause of poor immunity 1 – 1,000MHz.

From page 460 in Appendix D of [13], we learn that a small wire loop or monopole ("small" means <<λ/4) will emit the following worst-case E field strengths at 10m over a ground plane (the typical CISPR radiated emissions test method) of:

For DM currents: $E = 26.3 \cdot 10^{-16}(f^2 \cdot A \cdot I)$ Volts/metre

For CM currents: $E = 1.26.10^{-7}(f \cdot L \cdot I)$ Volts/metre

— where f is the frequency in Hz, A is the loop area in m^2, L is the monopole length in metres, and I is the DM or CM current in Amps. See Section 4.5 of [17] for other simplified formulas for common conductor structures.

A few sums using typical values will soon reveal that because of the huge attenuation factor of 10^{-16} in the DM formula, at frequencies below 1GHz it is more common for radiated emissions to be caused by CM currents, even though they may be measured in µA, than typical DM (signal, power) currents measured in mA.

In fact, a handy guide is that just 5µA of CM current on just one cable connected to a product can be enough to cause failure to comply with the CISPR 22 (= EN 55022) Class B limits at frequencies around 100MHz, and 15µA can be enough to fail Class A.

This means that we can very quickly and easily get a good idea of whether our product will pass emissions tests, purely by clipping a (calibrated) RF current monitoring probe over each cable in turn, and measuring its CM noise on a spectrum analyser. And we can do this at our usual development bench, we don't need a screened room or open area test site, as Figure 35 shows.

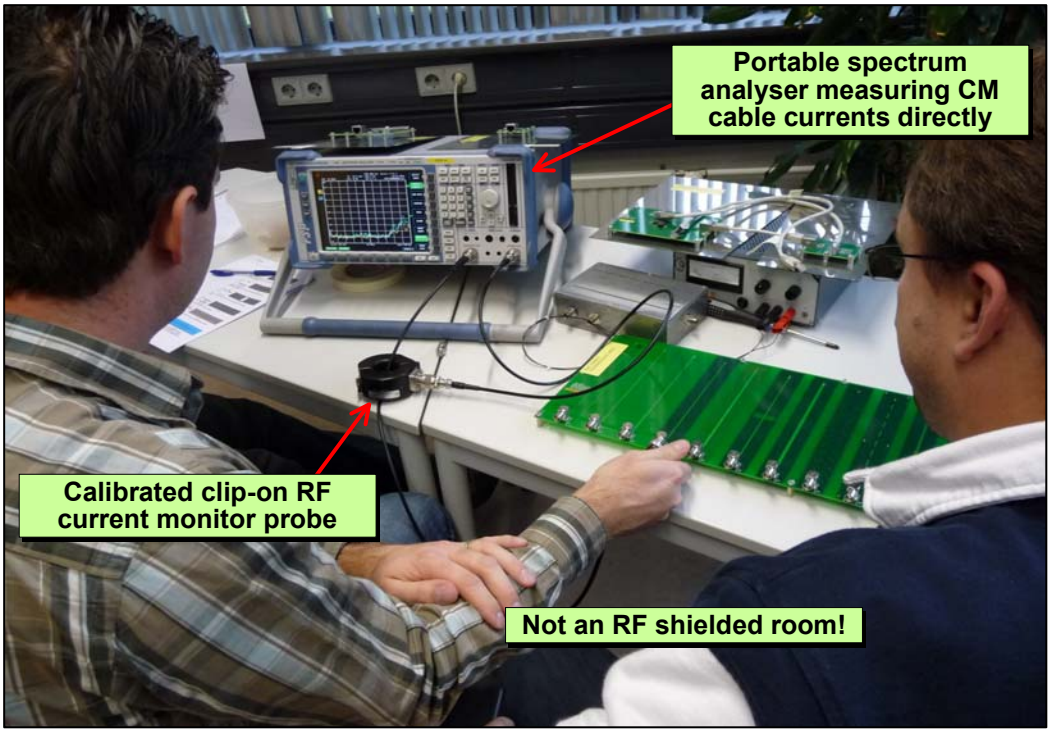


Figure 35 Quickly checking CM noise emissions without an EMC test chamber or site

Of course, nothing is guaranteed in real-life, and sometimes emissions problems are caused by DM signals rather than CM noise. I have had to reduce excessive radiated emissions using DM suppression techniques, when CM techniques did not work.

As the tested frequencies rise to 1GHz and above, the f^2 term in the DM equation above means that it becomes easy for DM signal currents to exceed the emissions caused by CM noise currents, even when their currents are similar.

So much experience has been gained below 1GHz, over the last few decades, and written-up in textbooks and articles like this, that there is a real danger that people will assume they

can solve all emissions problems above 1GHz by using CM noise suppression techniques – when what they may need to address instead is the emissions from the DM (wanted, signal or power) currents. You have been warned!

5.5 Controlling CM return currents

For the reasons discussed above, controlling the paths taken by each CM return current, relative to its send current's path, is very important indeed in EMC design.

Firstly, we reduce the generation of CM by:

- Reducing the RF impedance in shared conductors
- Providing DM send/return paths in close proximity for both signals and power (e.g. twisted-pair conductors)
- Using balanced/differential signalling techniques, or at least CM chokes

Secondly: we reduce the effect of the resulting CM voltages and currents by:

- Where practical, providing a path for CM current to return in very close proximity to each CM send path (i.e. each DM circuit)
- Electrically bonding all 'floating' circuits to the CM return current path, designing the bond to have the lowest practical impedance at the frequencies concerned

Remember, one of the key issues in EMC design is that currents always take the path that uses the least energy, which is also the path that emits the least E or H-fields. So by providing a CM return path close to its send path, and helping ensure that most of the CM current takes this path by low-impedance bonding, we cause the least CM stray coupling.

Mobile and portable equipment cannot take advantage of the techniques listed under "Secondly" above. But on the other hand, they are not often connected to large numbers of long cables and so their CM generation is more limited than permanently installed equipment that often can control the CM return path, e.g. using cable trays (see [18]).

5.6 RF "Grounding"

Safety earthing (grounding) does not help EMC at RF. I haven't mentioned safety earthing/grounding yet in this series of articles, because the terms "earthing" and "grounding" are so widely abused that it is best to use them only for electrical safety issues, and not for circuit design or EMC.

I've seen many projects suffer huge delays, because the different teams working on different parts of the equipment took different views on what was their RF Reference (that they simply called earth or ground, hence the confusion) and so had huge "internal EMC" problems (see Section 8).

But anyway, wired connections to the protective (safety) earth/ground have little effect at frequencies above 100kHz, because...

- They have far too much inductance (e.g. a 2 metre length has 188Ω at 30MHz, when a good "RF Ground" needs <<1Ω)
- Like all other conductors, they behave as accidental antennas (e.g. a 2 metre length makes a perfect antenna at a variety of frequencies at or above 75MHz; green insulation striped with yellow has no magical anti-antenna effects!)

So what must we use for our RF Ground and how should we connect to it?

The only effective RF Ground is what we should learn to call an "**RF Reference**", and it provides a low impedance, <<1Ω (preferably <10mΩ, with no lower limit), over the range of frequencies that need to be controlled to achieve EMC.

An RF Reference is a highly-conductive area (i.e. metal) that is as large as possible. It could be a 0V plane in a PCB, a chassis or frame, an enclosure, even (for frequencies below 100MHz) a grid of cross-connected metal structures in a room or a building. The larger the

area of the RF Reference, the better it is, with no upper limit on size.

Also – to be able to be used effectively – the RF Reference must be very close to the item that is to be "RF grounded": <<λ/10 at the highest frequency of concern (equivalent to <<30/f_{max} when surrounded by vacuum or air. f_{max} in MHz gives spacing in metres; GHz gives millimetres).

Much closer spacing is better: <<λ/100 at the highest frequency of concern (equivalent to <<3/f_{max} when surrounded by vacuum or air).

RF grounding to an RF Reference Plane is more correctly (and less ambiguously) called "**RF Bonding**". Direct metal-to-metal connections give the best RF bonds (i.e. the lowest impedances at up to f_{max}).

Where two conductive parts are to be joined that are not just a circuit connection, for example a metal filter body to the RF Reference, there should be 'RF Bonds' at multiple points equally spaced <<λ/10 apart along the perimeter of the seam or joint (equivalent to <<30/f_{max} when surrounded by vacuum or air). A single-point connection cannot work, the RF energy will just flow through the stray capacitance or stray mutual inductance instead, and will resonate at certain frequencies causing accidental antenna behaviour of the part.

Ideally, instead of multiple RF bonds – seam-weld, seam-solder, or apply a continuous conductive gasket all around the perimeter of the joint.

Some mystique surrounds the use of metal braid straps, especially if they have a length/width ratio of no more than 5. They so obviously have a very low resistance that it might appear that they must make a good bond at any frequency – but I'm afraid this is not so. As mentioned in Part 1 [8], everything that has physical existence in this universe has inductance and capacitance, so at RF can have considerable *impedance* even if its *resistance* is negligible.

In fact, the use of braid straps with a length/width ratio of 5 appears to come from a practical recommendation in a very early military "EMC good installation practices" concerning short-wave communications, with frequencies below 50MHz. These days we have to deal with frequencies that are at least 20 times higher, where braid straps have too much inductance.

I have seen the results of a test conducted in the 1980s that showed that a metal cabinet, RF bonded to the deck of a ship with a single 9.5 inch long braid strap, had lower emissions up to about 10MHz then higher emissions above 20MHz, when compared with no connection to the deck at all.

Around 20MHz, the stray capacitance of the cabinet was resonating with the series inductance of the strap, making the cabinet into a more efficient accidental antenna than without the strap.

However, a number of wide braid straps <150mm long equally-spaced <<λ/10 apart around the perimeter of a metal cabinet might have been effective at RF bonding it to the metal deckplate and reducing its emissions up to about 100MHz.

5.7 Metal planes bring many EMC benefits

Planes have very much lower RF impedance than conductors such as wires, cables or PCB traces, so when used in a shared circuit they cause very much lower common-impedance coupling (see Section 5 of [17]).

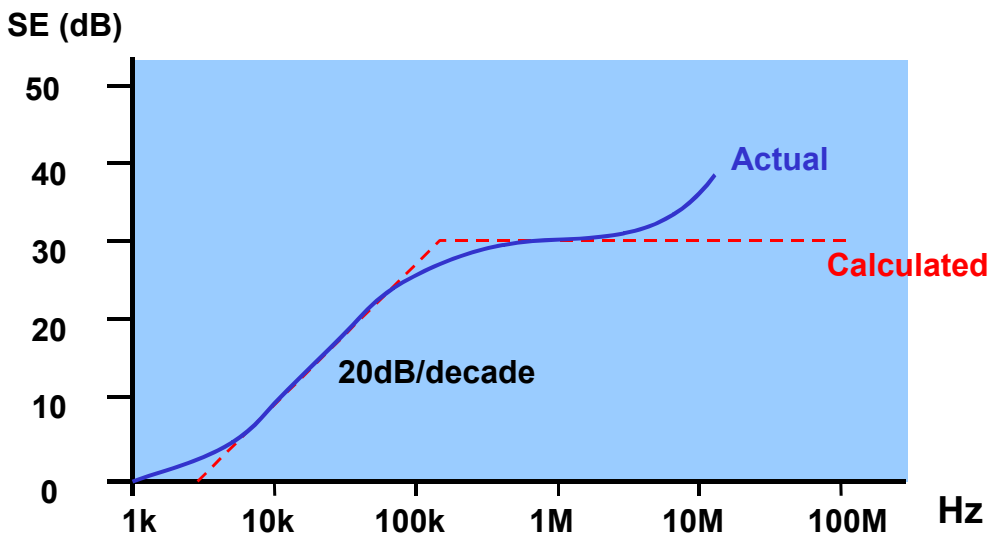
Figures 3 and 4 in [19] show that replacing a 300m long 4mm wide copper PCB trace by 300mm-spaced connections to a copper PCB plane, reduces the impedance at 50Hz from 35mΩ to 0.8mΩ, an improvement of about 43 times, or 33dB lower noise levels due to common-impedance coupling. This is entirely due to the plane having a much lower *resistance* than the trace (at 50Hz, the inductive component of the impedance is negligible).

At frequencies above a few kHz, the inductive component of a conductor's impedance starts to dominate its resistance, and the difference between the trace and plane will be much greater than it is at DC. The higher the frequency, the greater the improvement, for instance,

at 160MHz (and taking skin effect into account) the 300mm-spaced plane connections provide an impedance that is about 70dB less than the 300mm long 4mm wide trace (a 3,000-fold improvement).

For a source or victim circuit that is closer than $\lambda/10$ at the highest frequency of concern (equivalent to $30/f_{max}$ in vacuum or air), electromagnetic waves that hit a highly-conductive plane are partially cancelled out by their anti-phase reflections from the plane.

So, when a source or victim circuit is very close to a large area of metal plane this 'image plane' effect reduces its coupling with E, H and EM fields in its environment, as if it was being shielded to some degree. Figure 36 shows the shielding effect (SE, in dB) of a copper PCB plane, using the example of a 120mm long trace spaced 1mm above a 150mm square plane that is carrying its return current.



From: Swainson, IEE International EMC Symposium, 1990

Figure 36 Shielding effectiveness of a plane under a PCB trace

The image plane effect for a nearby circuit results in its return current travelling mostly underneath the conductor and following its route very closely, for frequencies above 100kHz. No matter how the conductor wriggles, as long as it stays close to the plane its return current in the plane follows it very closely indeed. Figure 37 gives the example of the return current density in a PCB plane, for a trace routed above it (shown end-on, in cross-section).

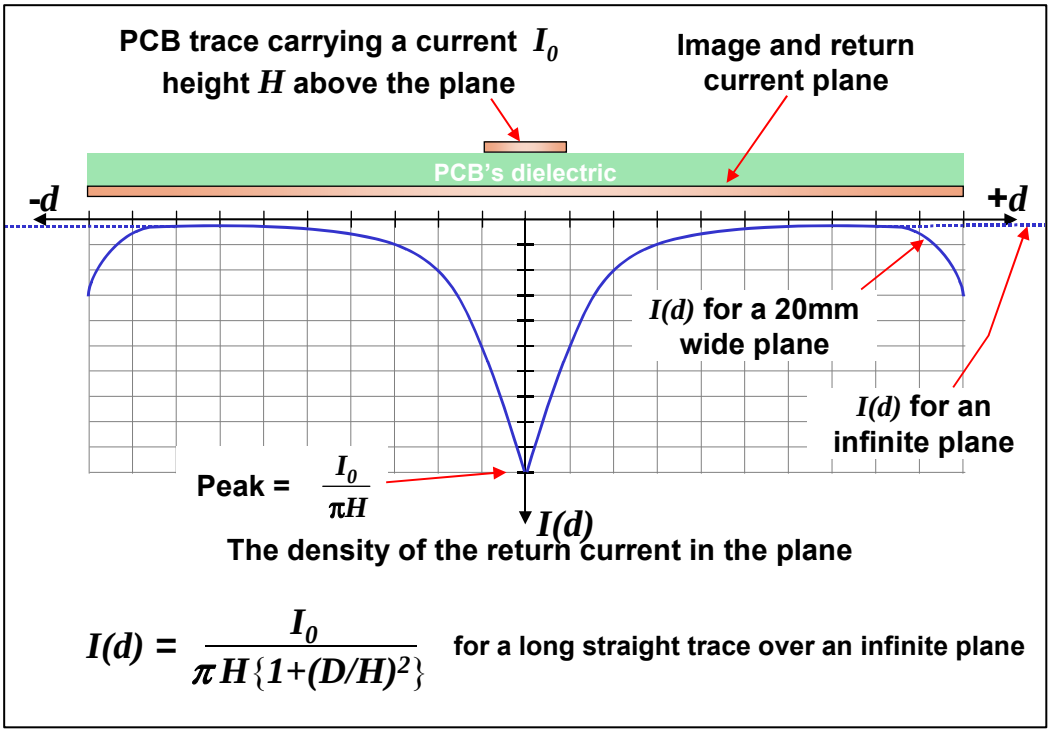


Figure 37 Return current density in a plane under a PCB trace

As we learned earlier (Section 5 of [17]), having send and return current paths close together generates the most compact fields and so couples the least E, H and EM fields into victim circuits (such as an antenna in a test laboratory). When using a plane near to a conductor, we automatically get the most compact fields and least stray coupling. This is another way of explaining why a nearby plane provides shielding effectiveness.

So, metal planes are a powerful tool for EMC, and they are used in some ICs and most PCBs. Large systems sometimes use meshes instead, but unlike sheet metal they are only effective up to a frequency related to the reciprocal of their mesh size (smaller mesh size – effective to higher frequencies).

In fact, their highest effective frequency can be crudely calculated as f_{MAX} = l00/D (f_{MAX} in MHz, D in metres, where D is the diagonal size of the mesh's elements – the largest dimension of the mesh). So, for example, a 1m diagonal mesh would provide some control of RF emissions and immunity up to 100MHz. At 100MHz the mesh will not be *terribly* effective, it will be much better at 10MHz, but at least it will be very much better than the traditional "single-point earthing/grounding" method, which simply cannot control emissions or immunity whatsoever above a few kHz.

6 An overview of emissions

Real-world EMC is often very complex indeed, because of the eight modes of EM coupling (4 each for DM and CM) and the ever-increasing complexity of modern devices, circuits and systems. When bogged down with complexity whilst trying to deal with emissions, it often helps to get a perspective, and overview, by realising that the situation is usually very simple – all electronics can be thought of as many tens of thousands (maybe millions) of noise sources, connected to thousands of accidental antennas.

The noise sources are the transistors, either in integrated circuits (ICs) or power transistors. The accidental antennas are all the conductors, e.g. IC leadframes, PCB traces, wires and cables, metal boxes, etc., all of which have resonant frequencies that depend on their length, exact build conditions, terminations, routing, gaps and slots, and proximity to other conductors and materials.

There are many enjoyable aspects to EMC, and one of them is determining the "accidental

antenna behaviour" of components, conductors, packages, boxes, assemblies, installations, etc. For example, a heatsink might have its first resonance at the GPS L1 frequency (1.6GHz), and respond to the 16th harmonic of a 50MHz clock – radiating so much noise at that frequency that a nearby GPS antenna cannot "see" any satellites.

An example I had recently in an installation, that used a number of variable-speed motor drives for pumping gasses and liquids. The noises the drives created (harmonics of their power switching frequency) could be found all over the installation, particularly at frequencies around 1MHz, where they were easily 20dB worse than at other frequencies and interfering with measurements. The installation's metal structure was circular with a diameter of about 150 metres, giving it a strong self-resonance at around 1MHz in all directions, which was selectively amplifying the noise from the drives around that frequency.

We often estimate the accidental antenna effects from simple dimensions, but Doug Smith [20] describes a very useful method of finding the resonant frequencies of anything, using a spectrum analyser with tracking generator and applying a pair of current probes to the object in question.

7 Immunity issues

7.1 Issues not covered so far

All of the previous discussions are equally valid for emissions and immunity, because they are all concerned with how conductors interact with the propagation of the E, H and EM fields that that we generally call electrical signals and power. Because of the principle of reciprocity, those discussions are equally valid when we want to control RF emissions, and/or RF immunity.

Now we have to discuss some additional topics, which are generally only of concern for immunity.

7.2 Non-linearity, demodulation, and baseband noise

In a linear material the output is linearly proportional to the input, but all semiconductors are non-linear (as are some oxidised electrical connections) so they tend to rectify AC waveforms. Rectification results in a DC signal plus harmonics, just like a mains AC rectifier.

In a radio receiver this rectification characteristic is combined with a low-pass filter to obtain the modulation signal that is carried by the transmitted radio wave, and is called demodulation or detection. In this case, the DC signal output by the rectifier fluctuates in accordance with the modulation, and is called the "baseband".

One result of this is that all transistors will demodulate radio signals that are allowed to get into their terminals, acting just like radio receivers with accidentally-tuned antennas.

Figure 38 shows the example of a 'slow' opamp rectifying (demodulating) the 1kHz modulation of a radio frequency immunity test, at frequencies up to 1,000MHz, before I modified it to make it pass the test. It is taken from an actual product test that I did in the early 90s. The product was an analogue signal converter in an unshielded plastic box for DIN-rail mounting in industrial cabinets. Inside, it was little more than a quad opamp, type LM324, on a small PCB. It had three cables connected to it: 24Vdc supply, input and output, all of different lengths.

As the tested frequency increased, I saw the classical signs of rectification – the product's output error showed sharp peaks at frequencies that could easily be correlated with the resonant lengths of the cables connected to it (see Figure 7 in Part 1 of this series [8]). So far, so humdrum, but at frequencies above 400MHz I was surprised to see the error increasing linearly with frequency – and still rising at 1GHz, where the test stopped.

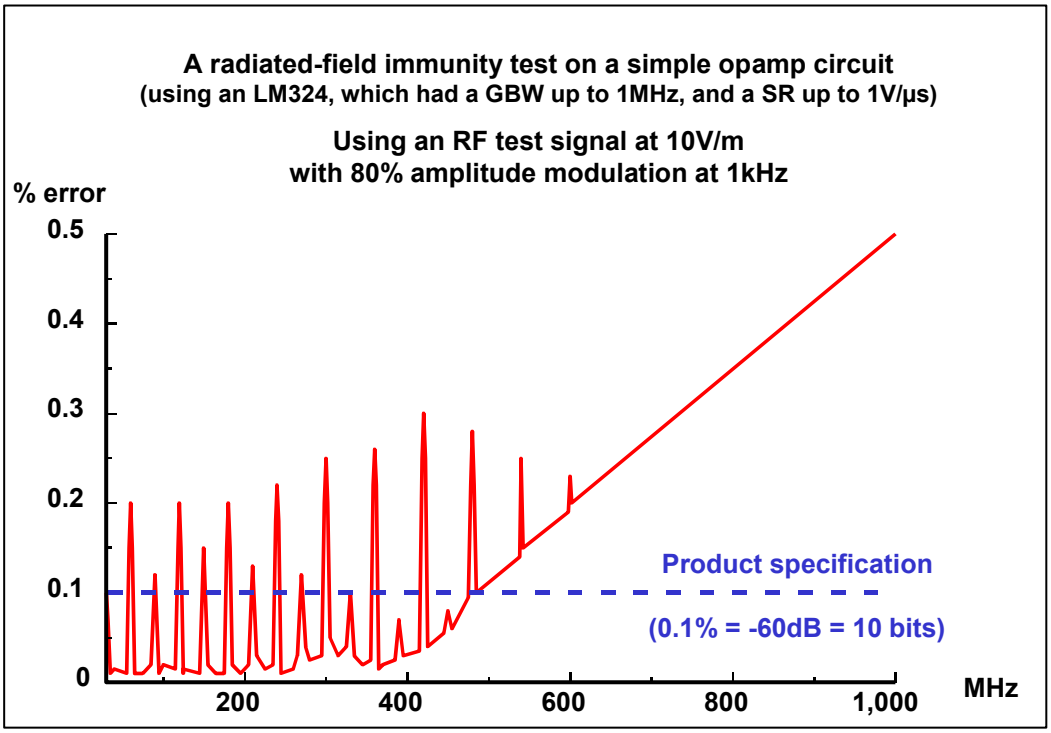


Figure 38 Demodulation in a "slow" opamp

It is very common to hear designers of audio and low-frequency instrumentation say that because they are using "very slow" opamps, they will not respond to an RF immunity test. They use this as an excuse for not bothering to protect their designs against RF interference, and also for not testing them for it. In the early 90s the LM324's main claim to fame was that it was the cheapest quad opamp available, and with a slew rate of up to 1V/µs and a GBW (gain-bandwidth product) of up to 1MHz, it was definitely a "slow" opamp!

And yet, in the test reported in Figure 38, it demodulates at 1000 MHz about twice as effectively as it does at 500MHz. The datasheet figures for an opamp are for its *linear* behaviour, not for its non-linear responses, and it is these that we see causing the error to rise considerably over the specification, in Figure 38.

Even a cheap, low-performance opamp like the LM324, used very tiny transistors that had correspondingly low values of collector-base and base-emitter capacitance. So when hundreds of MHz were applied to their terminals (courtesy of the accidental antennas that we call DC, input and output cables, and PCB traces) there was too little capacitance to prevent the non-linear semiconductor junctions from being exposed to the RF, which they promptly demodulated, producing an error in the opamp's output.

Figure 38 shows us that no opamp is ever "too slow to see RF", and all analogue (and digital) circuit designs are susceptible to RF interference, whatever the application. An on-chip capacitor is used for dominant-pole compensation inside the opamp's IC, but as far as RF is concerned that it just a nice low impedance for coupling RF noise inside the device.

It is tempting to think that it is just an IC's input pins that are susceptible to demodulating RF, but in fact all the pins are much the same in this regard, as Figure 39 indicates. The output impedance of feedback amplifiers may be designed to be 0.1Ω or less, but this is only for the bandwidth for which there is at least 60dB of excess open-loop gain above the closed-loop gain requirement.

At frequencies beyond a few 10s of MHz, when opamp open-loop gain is 0dB or less, the output impedance is typically around 100Ω and makes a good impedance match for the CM impedance of typical cables. So at such frequencies the RF noise picked up by the cables acting as accidental antennae pours straight into feedback amplifier outputs, to get demodulated in whatever silicon junctions it can find inside the IC.

And the power supply rejection ratio of opamps might be specified as 120dB, but that is measured at 60Hz (the US mains power frequency) and typically falls at 20dB/decade to below 0dB at or above 1MHz. Some types of analogue ICs have no power supply rejection at all, at any frequency, as I found in the early 80s when working with some of the first switched-capacitor filter chips.

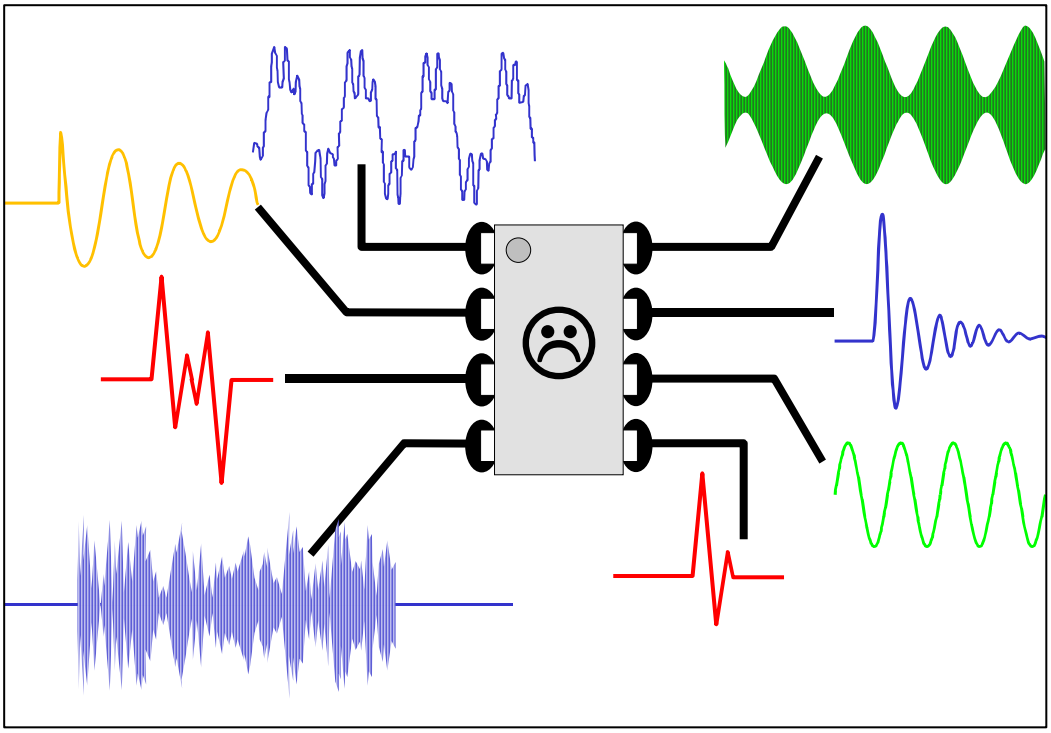


Figure 39 All IC pins are susceptible to RF demodulation

Interference is a bigger problem for well-designed analogue circuits than for digital, because a well-designed digital circuit has a "noise margin" – the peak noise level in the circuit (created by its own digital activities) is less than the threshold at which the devices make decisions about whether the signal is a 1 or a 0. External EMI that adds to the noise in the circuit has to be larger than the noise margin, before a false logic decision is made.

It is a similar issue with Analogue to Digital converters – they resolve their input signals to 1 least-significant-bit (LSB), and so must have an internal or background noise level that is half an LSB or less. But even an 8-bit A/D converter has an LSB that is much smaller than the noise margin of a well-designed digital circuit, so we can say that analogue circuits are always more susceptible than well-designed digital circuits.

Whether devices are analogue or digital, once RF gets inside their package, it can couple through stray capacitances to any/all of the semiconductors inside the device, be rectified by them, then amplified by others. The result of a rectification is a shift in the DC operating point, the DC bias, of a transistor. Modulation of the level of the RF results in modulation of the DC bias point of the transistor, hence what we call baseband noise (or, in a radio receiver, demodulation or detection).

Figure 40 shows the effect of an amplitude-modulated RF broadcast transmission being coupled into an opamp by some means – the opamp's output noise being the demodulated radio signal. Increasingly, radio transmitters are changing to use digital modulation techniques instead, and in typical cellphones the digital data is sent as several bursts a second, each burst containing packets of data at the rate of 217Hz. This is demodulated in opamps in our car radios, landline telephone handsets, etc. as the familiar "blippety-blip" sound we get when making or receiving a call on a cellphone that is too close to the car radio or landline.

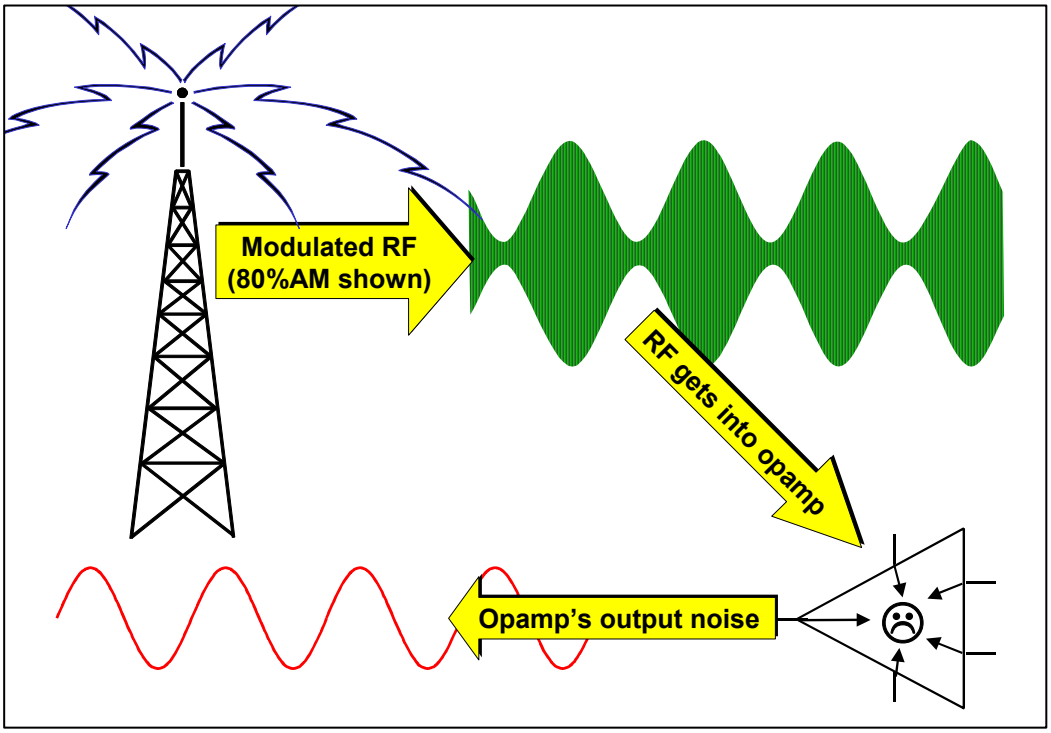


Figure 40 Example of RF demodulation of a radio broadcast

All sparks create noise at radio frequencies – sparks in switch and relay contacts, thermostats, commutators in DC motors, the sliding contacts used by electric trains, etc. This noise is effectively randomly-modulated RF, although it can have strong components at the frequency of the AC mains power line and its harmonics. When it couples with opamps the result of a switching operation is a burst of random noise, that might sound like a "pop", "fizz" or "splat". In the case of a DC commutator – the noise is a whine that varies as the speed of the motor varies. Figure 41 sketches this situation.

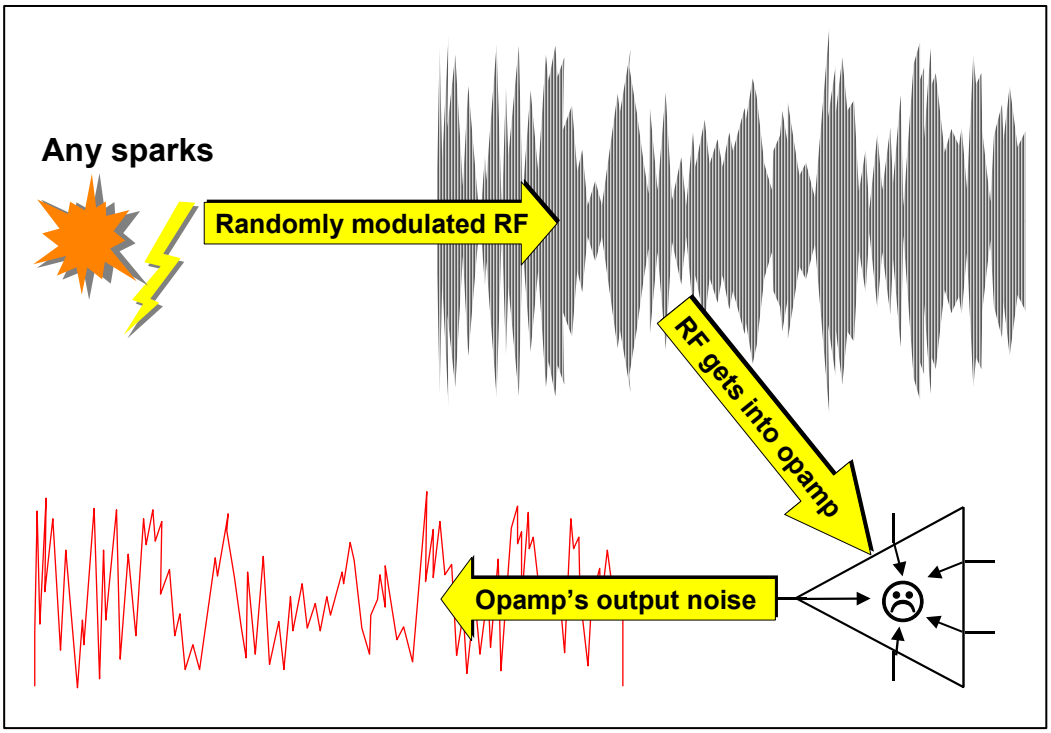


Figure 41 Example of RF demodulation of random RF noise

If the RF signal level is high enough, the DC bias points of the transistors can be moved

more than just a few millivolts, altering the way that the semiconductor affects the wanted signal that is passing through it. If the DC bias is moved too far by the RF noise demodulation, the wanted signal might become severely distorted, or even not be amplified at all. This happens in just the same way for digital transistors as analogue ones, but has different consequences for digital circuits.

Once an analogue signal is corrupted by EMI there's no way back (without sophisticated digital processing that "knows" what the signal is supposed to look like). The amplitudes of digital signals are supposed to be always either maximum (1) or zero (0) so RF demodulation doesn't affect them, but of course a sufficiently high level of noise can create a signal that is not full-scale, but nevertheless is enough to cross the logic threshold cause a "bit flip" – turning what should have been a 1 into an erroneous 0 (or what should have been a 0 into an erroneous 1).

Software consists of instructions and the data on which the instructions operate, with both instructions and data stored in memory and accessed when required. Obviously, "bit-flips" can corrupt data to varying degrees, a bit-flip in the LSB being of little importance – but very important indeed if it occurs in the most significant bit, the MSB. So data on the desired speed of a machine might be corrupted to turn the speed up or down by a little more than was actually required, or by a more significant amount, even up to making it stop dead or race away at full speed in forward or reverse.

If a bit-flip occurs in an instruction, it will be corrupted, resulting in a nonsense instruction that could do nothing (perhaps making the software process "hang") or do something very unexpected (but always undesirable). It might, for example, cause the wrong memory location to be accessed, fetching the wrong instruction or data, again with undesirable and possibly calamitous results if anything powerful is being controlled.

Software techniques are available to help correct bit-flips caused by EMI, but are never complete solutions on their own because sufficiently long burst of RF noise or continuous interference can cause critical data or instructions to be delayed by too long, and could possibly even cause the rate of either to fall to zero. Hardware design techniques are still required for EMC.

Another problem for digital circuits is that the rectification of RF noise within the semiconductors causes their logic threshold to vary, and this causes timing jitter on the edges of the digital signals and clocks. When the timing jitter exceeds a certain margin, the digital process will perform other than expected. This is not an error in the data or an instruction, it is an error in the basic operation, and it is very hard to predict what might happen.

Analogue circuits tend to "fail gracefully" when interfered with – higher levels of EM disturbances cause higher levels of signal degradation. This makes it relatively easy to estimate the reliability of a design when exposed to real-life EM disturbances.

But the problem with testing digital circuits for RF immunity is that they might pass the test at a given test level with no degradation in performance, but a very slight increase in the level it experiences in real-life could result in a total failure to operate, or some extreme misbehaviour.

York University have proposed a way of determining how close a digital circuit is to failing, by measuring the emissions from its intermodulation (see 7.3) with the RF frequency it is exposed to, [21]. This can help provide more confidence in the reliability of operation in real-life.

7.3 Demodulation, intermodulation, and the creation of new frequencies

Rectification creates even and odd-numbered harmonics, which were mentioned in the previous section but then ignored. However, where two or more RF signals are simultaneously present in a non-linear device, new frequencies are created from their sums and differences, and from the sums and differences of their harmonics. One of the signals

might be the one the circuit is meant to be processing, the wanted signal, whilst the other might be noise, or both frequencies might be noise.

Figure 42 shows the first few "intermodulation products" with two frequencies, one (f_1) at 400MHz and the other (f_2) at 500MHz in this example.

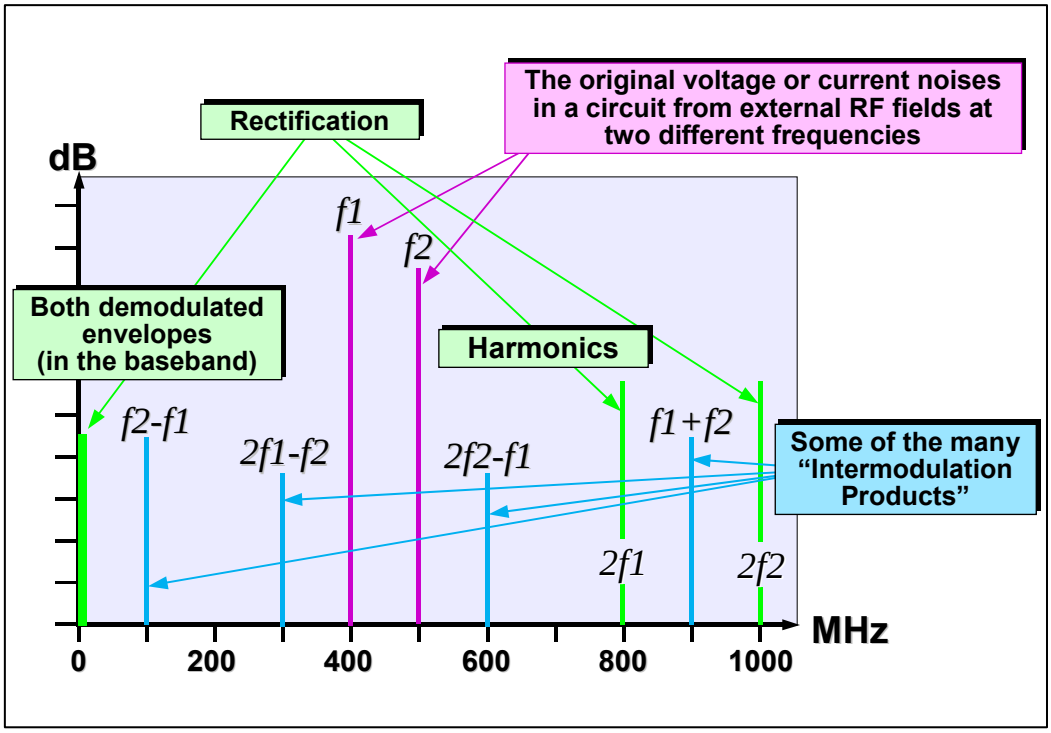


Figure 42 Example of the intermodulation of two frequencies in a semiconductor

Firstly we notice that the baseband noise, caused by rectification, is the sum of the modulation envelopes of the two RF signals. If f_1 was pure unmodulated carrier wave and f_2 was transmitting, say, music, the baseband noise would be a DC offset plus the music. If f_1 was transmitting speech and f_2 music, the baseband noise would be a smaller DC offset plus the speech and music mixed together.

Secondly, we see the second harmonics of both f_1 and f_2, at 800MHz and 1000MHz respectively, the scale of the figure not allowing their 3rd, 4th, 5th, etc., harmonics to be shown.

Thirdly, Figure 42 shows us that we have what are called the "first-order intermodulation (IM) products". There are two of them, at the sum and difference of the two frequencies: $f_1 - f_1$ = 100MHz and $f_1 + f_1$ = 900MHz.

Finally, we see the "second-order IM products", between each of the 2nd harmonics and all of the other frequencies. There are five of them, at $2f_1 - f_2$ = 300MHz and $2f_2 - f_1$ = 600MHz (with $2f_1 + f_2$ = 1300MHz, $2f_2 + f_1$ = 1400MHz and $2f_1 + 2f_2$ = 1800MHz being off the scale).

The figure does not show the 3rd-order IM products, between each of the 3rd harmonics and all of the other frequencies (nine of them, at $3f_1 - 2f_2$; $3f_1 - f_2$; $3f_2 - 2f_1$; $3f_2 - f_1$; $3f_1 + 2f_2$; $3f_1 + f_2$; $3f_2 + 2f_1$; $3f_2 + f_1$ and $3f_2 + 3f_1$), or the 4th, 5th 6th, etc. orders of IM products.

And the above is just with two original frequencies. It starts to become really complicated when there are three or more frequencies present in a semiconductor at the same time. Various IM calculators are available (for a price), such as the one from www.telecomengineering.com/software-download1.htm.

As the order of the harmonics and IM products increases, their levels decrease, so, on a graph like Figure 42, the end result looks like an increased noise floor over much of the frequency range

Figure 43 shows how the three mechanisms by which EMI causes immunity problems: direct interference, demodulation (rectification), and intermodulation, typically cause problems for

electronic circuits. Section 9 will go into the practical results of this in more detail.

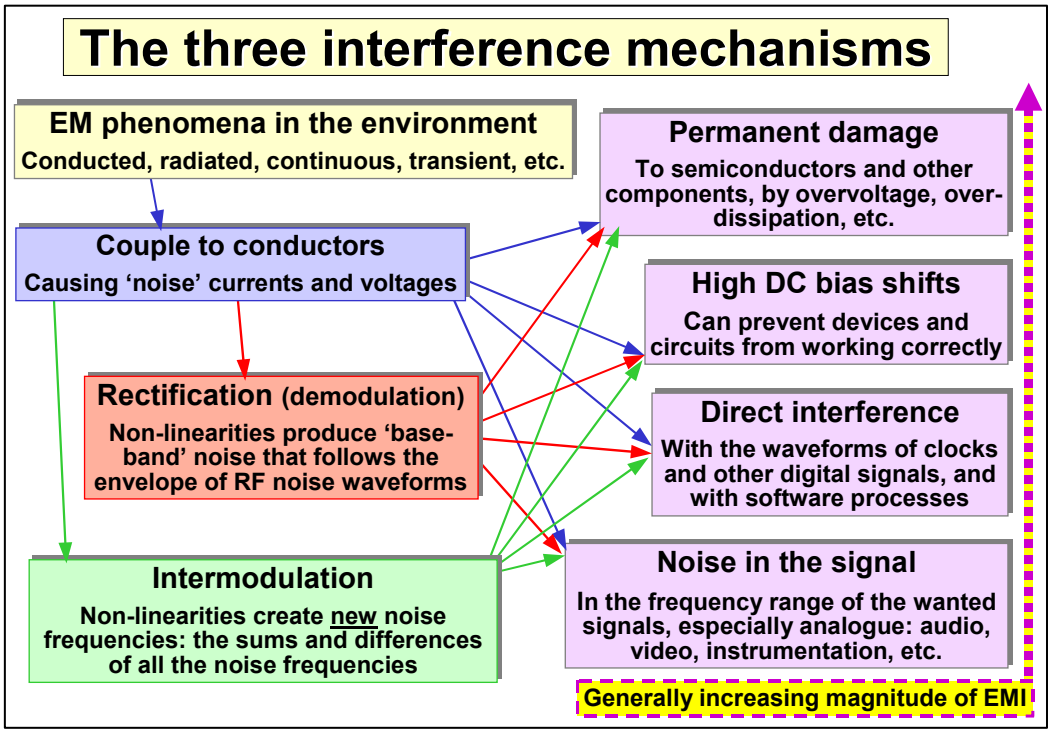


Figure 43 Overview of the three ways in which EMI can interfere

Now we can discuss an example of how the new frequencies created within a circuit by demodulation and intermodulation, can cause immunity problems in real-life operation.

Conventional RF immunity testing applies a single frequency, with modulation, over the range 150kHz to 1GHz and discovers that the product is too susceptible over the range 50 to 200MHz. It would not be unusual to find that they are 20 to 30dB too susceptible.

Being good EMC engineers, we fit shielding and filtering that is effective over the 50 – 200MHz range, and the equipment now passes the test. We pat ourselves on the back and fondly imagine that because our product passes the continuous RF immunity test at levels well above those that it will experience in real-life operation, it will be totally immune and not suffer failures due to that type of EMI. But we would be wrong.

It is quite common to hear people saying that because their products pass some set of EMC immunity tests, maybe in expensive anechoic test chambers, even at levels that are much higher than they will be exposed to on real life, their products are *therefore totally immune to all EMI*. But they are wrong too.

When we added shielding and filtering to our product to pass the test, we didn't try to make it effective over the frequency ranges below 50MHz and above 200MHz. There was no point, and anyway it would have taken longer and increased the cost of the BOM (bill of materials) by more than was necessary, since all we were interested in was passing the test.

But the real-life electromagnetic environment does not consist of just one radio frequency at a time. Simultaneous radio frequencies can and do exist, indeed they are more typical than just one frequency. For example, if you can receive FM radio channels, you are subjected to several radio frequencies (the different channels) at once. Of course, they do not generally have very high levels, but fields in the range 1V/m – 10V/m at each transmitted frequency are not untypical on public roads near to a broadcast radio or TV transmitter in a city.

Near a clinic, hospital, factory or beauty parlour where they are using RF energy to heat human tissues, plastic or metal; dry glue or paint, depilate or remove warts from human skin, there can be many frequencies present at quite high levels at the same time. And then there is the plethora of mobile transmitters we have now, including CB radio, cellphones at (in Europe) 900MHz, 1800MHz and 2100MHz, and Wi-Fi and Bluetooth both hopping rapidly

around over a great many frequencies with the 2.5GHz "ISM band".

Frequencies outside the band we filtered and shielded can enter our product and intermodulate in its semiconductor junctions, creating new frequencies in the very susceptible range 50 - 200MHz, and causing interference. Since the new frequencies are generated *inside the very circuits that we protected* with our shielding and filtering, our efforts at protecting our product from 50 to 200MHz are made ineffective.

Intermodulation cannot be created by testing at a single frequency at <u>any</u> RF power level, so passing such simple tests creates a false impression of the likely reliability of any electronics or software.

Of course, these "simple" single frequency RF immunity tests take long enough to do, for example covering the specified frequency range in 0.1% frequency steps and "dwelling" at each frequency step for three or four seconds to have time to determine if the product is being interfered with, takes about an hour. But then there is horizontal and vertical polarisation to consider, plus at least one other antenna position (usually three more), so a full test will take between a half and one day. If we were to test with two frequencies in order to simulate intermodulation that might occur in real life, we would probably want to vary f_1 over the range in 0.1% steps as before, and at each step in f_1 vary f_2 over the entire frequency range in 0.1% steps, dwelling 3 or 4 seconds at each f_2 step.

The test time would then be 1,000 times longer than for a single frequency test, at between 500 and 1,000 days.

The solution to this problem of making electronics fit for the real world, especially important for safety-critical applications, is not to rely solely on immunity testing. The IET's 2008 guide to EMC for Functional Safety [22] describes how to do this, and also provides 26 pages of design techniques that can be used. It can also be used to improve the reliability of electronics used in high-reliability and mission-critical systems, and legal metrology.

And while we are on the subject of EMC immunity testing, [23] describes many reasons why the automobile industry's EMC test programmes cannot ensure the achievement of tolerable levels of safety, intermodulation being just one of them.

In case readers who work in other industries (e.g. medical, rail, aerospace, military, machinery, robotics, etc.) are feeling superior at this point, I should point out that exactly the same arguments apply to their standardised EMC immunity test programmes, see [24].

[25] describes some of the many reasons, not just intermodulation, why simply increasing the immunity test level cannot provide additional confidence in the reliability of the tested electronics, and hence the safety of the applications they are controlling.

I find it very interesting that most EMC test engineers and test lab managers worldwide believe that applying the regular immunity tests at ever-increasing levels is all that needs to be done for safety, and that higher test levels correlate directly with increased "safety margins". When I sat down to write [25] it only took me about half an hour to find several very simple and obvious reasons why this assumption *could not possibly be true*.

But this article is supposed to be about the physical basis of EMC, not the impossibility of ever doing enough immunity testing to ensure high reliability – so let's get back on track.

7.4 Overview of immunity

When bogged down with complexity whilst trying to deal with immunity, it often helps to get a perspective, and overview, by realising that the situation is usually very simple – all electronics can be thought of as many tens of thousands (maybe millions) of "accidental demodulators" (i.e. rectifiers) and "accidental superheterodynes" (i.e. intermodulators). These are every silicon junction in every diode and every transistor, whether hidden inside analogue or digital ICs, or power devices.

These accidental radio tuners are connected to thousands of accidental antennas – all of their PCB traces, wires and cables, and coupled to other accidental antennas created by nearby metal structures (e.g. gaps and slots in metal boxes).

These accidental antennas all have resonant frequencies that depend on their length, build conditions, terminations, routing, and proximity to other conductors and materials.

Another cause of accidental superheterodyne behaviour is instability in feedback amplifiers, when just a single RF noise frequency can result in a number of new frequencies as it intermodulates with an amplifier that happens to be self-oscillating at a particular time.

8 Crosstalk and "internal EMC" issues *inside* a product

For EMC compliance we are only concerned with the EM interactions between an item of equipment and its external EM environment.

But EM interactions also occur between devices, components, traces and wires *inside* a product or item of equipment, and we care about these because they affect the number of design iterations and time-to-market. We also care about them because they can easily affect reliability and hence warranty costs, and customer perceptions hence future sales.

Banana Skins [9] includes many examples of companies that lost a lot of money, maybe even their whole company, because of poor EMC engineering. These examples can help engineers encourage their managers to take the subject seriously.

We might call this issue "Internal EMC". EMC consultants worldwide have been trying for decades to get designers to use good EMC engineering from the start of any project, to save money and time overall and reduce financial risk, and they have usually had a very poor response – causing them to stop trying.

Well, I have been banging this particular drum for 20 years now, and still haven't given up [26]. Maybe Figures 44, 45 and 46 will add to [26] and help get the message across. Like all businesses today, all that matters is money [27] and so – as an engineer – to get anywhere in discussions with your managers you have to bring it down to terms they can understand – money, time, and probability. Strangely enough, it is just as if you were discussing the odds of various gambling methods.

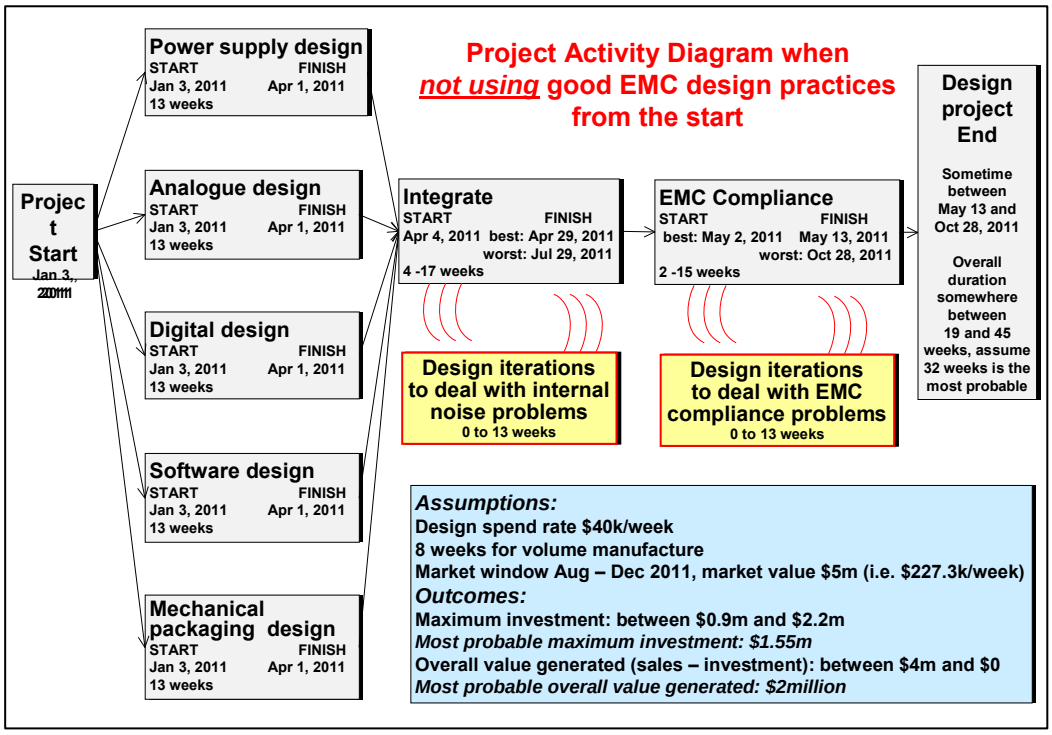


Figure 44 Example Project Activity Diagram when not using good EMC design practices

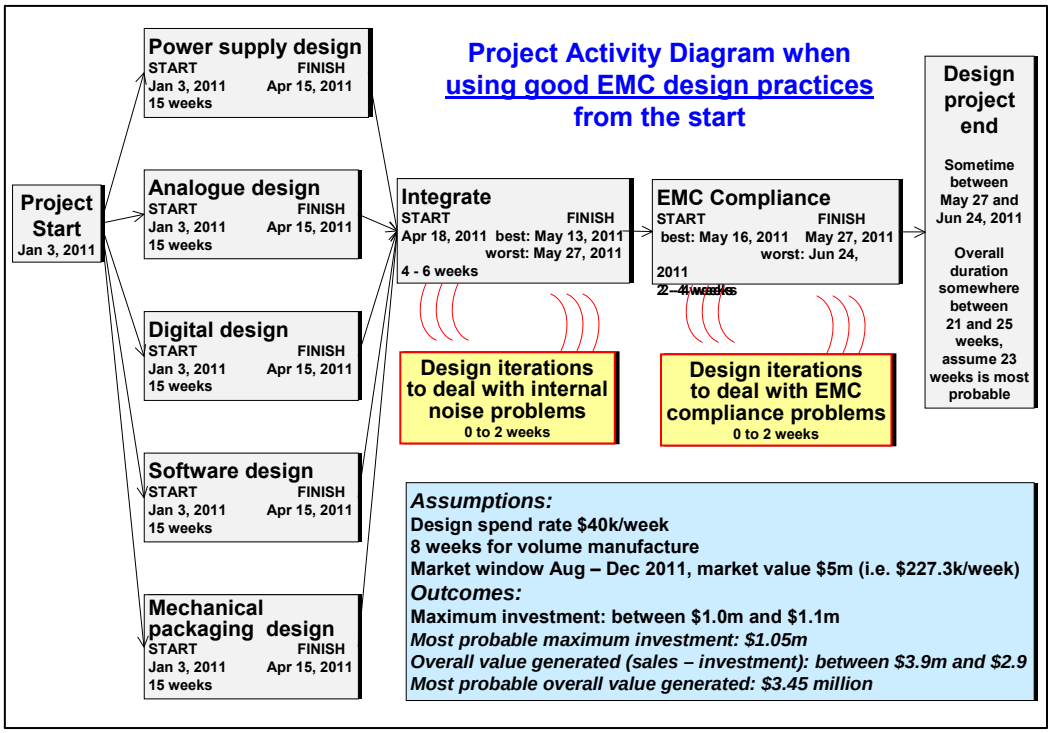


Figure 45 Example Project Activity Diagram when using good EMC design practices

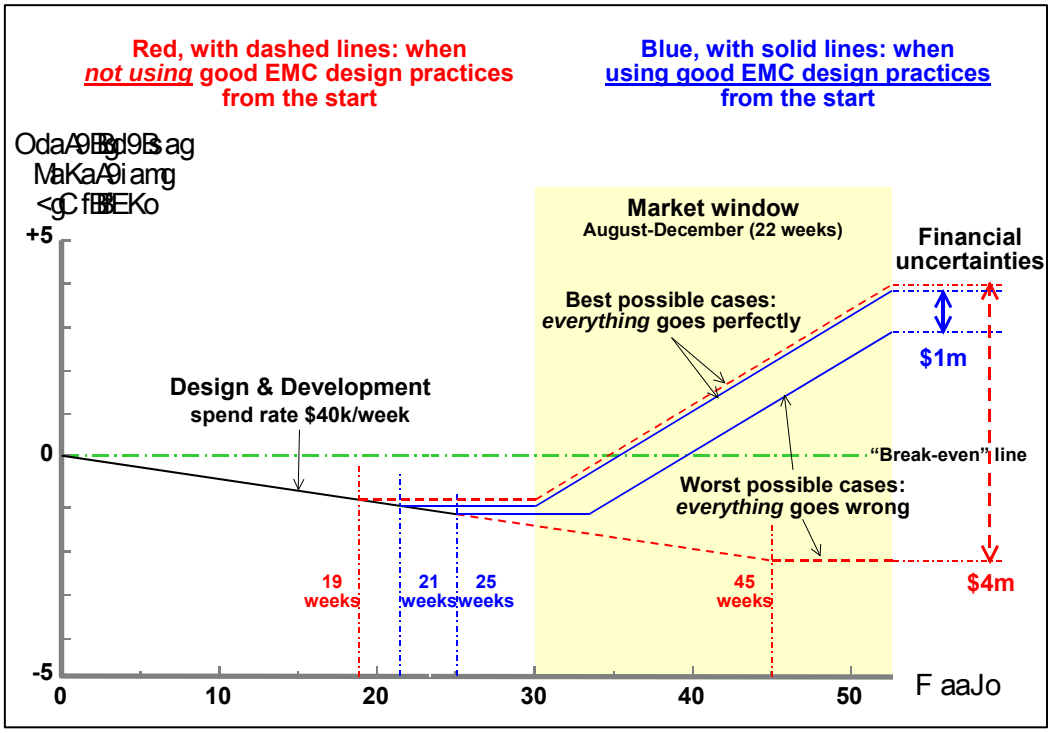


Figure 46 Comparison of financial risks from the projects of Figures 43 and 44

Now I must get back on course again and discuss the physics of EMC. The material in this little series of articles applies equally well whether the issue is "external" or "internal" EMC.

Internal EM interactions are traditionally called crosstalk and analysed in terms of stray capacitance and stray mutual inductance, i.e. a Lumped Analysis approach (see Section 3 of Part 1 [8]). But this only works when the victim is in the near-field of the E of H-field emissions from the noise source.

Traditional crosstalk is often inadequate for modern designs, because the high frequencies we now employ (e.g. clock harmonics) have such short wavelengths that parts of the inside

of the equipment we are designing are in their far field, and the wires and cables inside an equipment; PCB traces; heatsinks and even devices themselves, can behave as very efficient accidental antennas.

Remember that far-field and resonant EM interactions cannot be estimated by lumped analysis methods.

But there is more to internal EMI than crosstalk: noise, signal-to-noise ratio; noise margin; eye closure; overshoot; ringing; double-clocking, etc. are also aspects of it, as sketched by Figure 47.

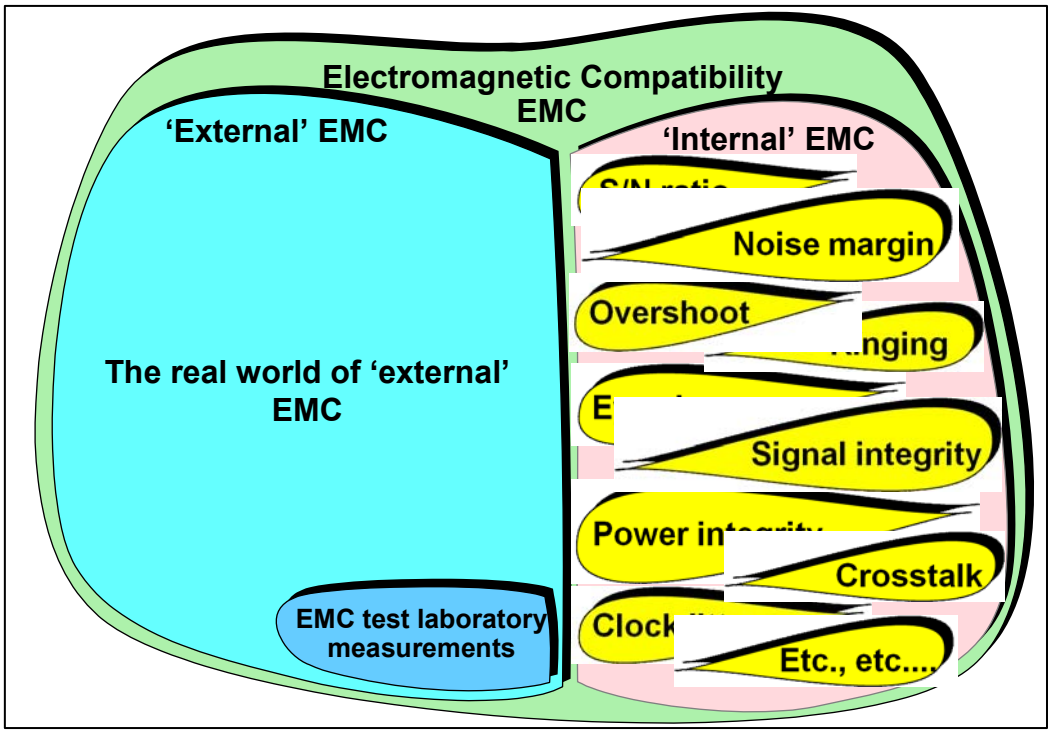


Figure 47 Aspects of Internal and External EMC

9 Types of EM phenomena and how they can interfere

A guide to assessing EM phenomena is one of the many very useful Technical Guidance Notes (TGNs) issued by the EMC Test Labs Association, as TGN 47 [28].

9.1 EM phenomena that couple into all metalwork and conductors

When I write the word "conductors" in this series on the Physical Basis of EMC, I always mean anything that can *possibly* conduct electricity, for example, metal structures, brackets, chassis, wires, cables (power, data, signal, control, etc.), PCB traces, water (except when distilled), etc.

9.1.1 Surge transients

All conductors pick-up noise currents and voltages from transient over-voltages, in IEC standards-speak these are called "surges" (which term in the USA means what the IEC calls voltage fluctuations, so be careful) or more colloquially "spikes", sometimes just "transients".

These are generally caused by nearby thunderstorms, even by cloud-to-cloud lightning when there are no thunderstorms, by reactive load switching (e.g. electrical motors, relay and contactor coils, solenoids, capacitors, long power cables (especially HV transmission lines) etc.), fault clearance (e.g. fuses, circuit-breakers, etc.).

Surges can be directly injected into conductors (e.g. when de-energising a relay coil or solenoid) or coupled into them by one or more of the four coupling effects described in section 5 of Part 2 [17]. Figure 48 shows a surge waveform induced into a wire that is

bundled together with a pair of wires that were powering a relay coil, when that coil was switched off.

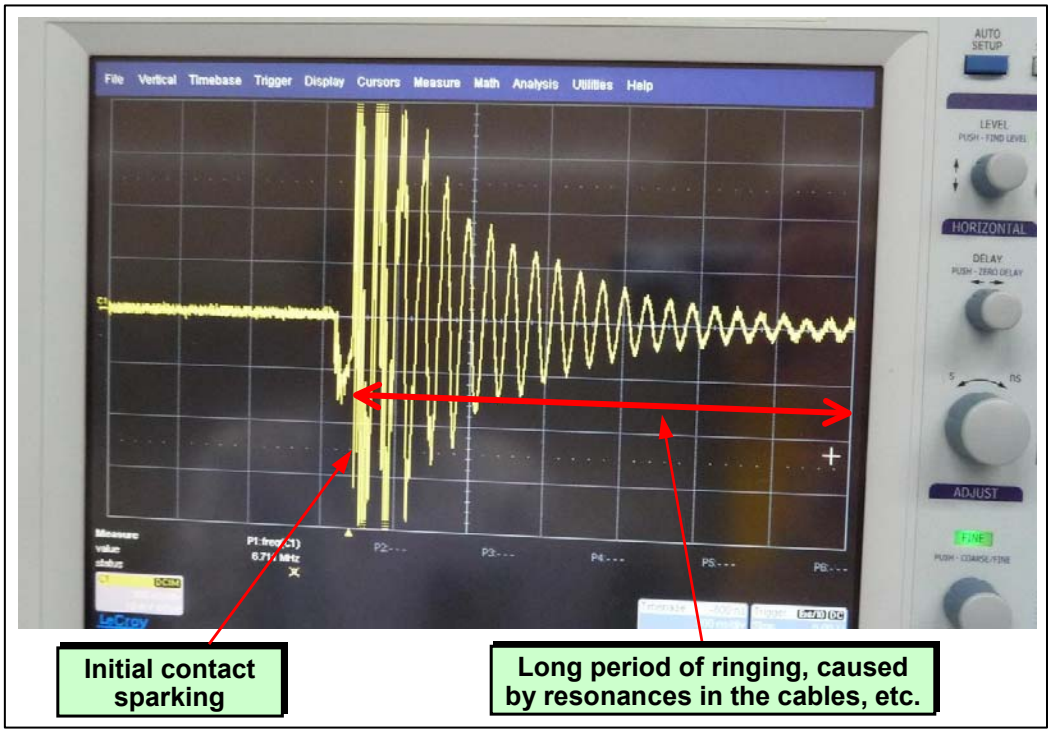


Figure 48 Example of induced surge from wires powering a relay coil

There is an infinite variety of surge waveshapes, all having different effects on electronic circuits, but IEC test standards only use three: Unidirectional; Oscillatory Wave, and Ring Wave, all sketched in Figure 49.

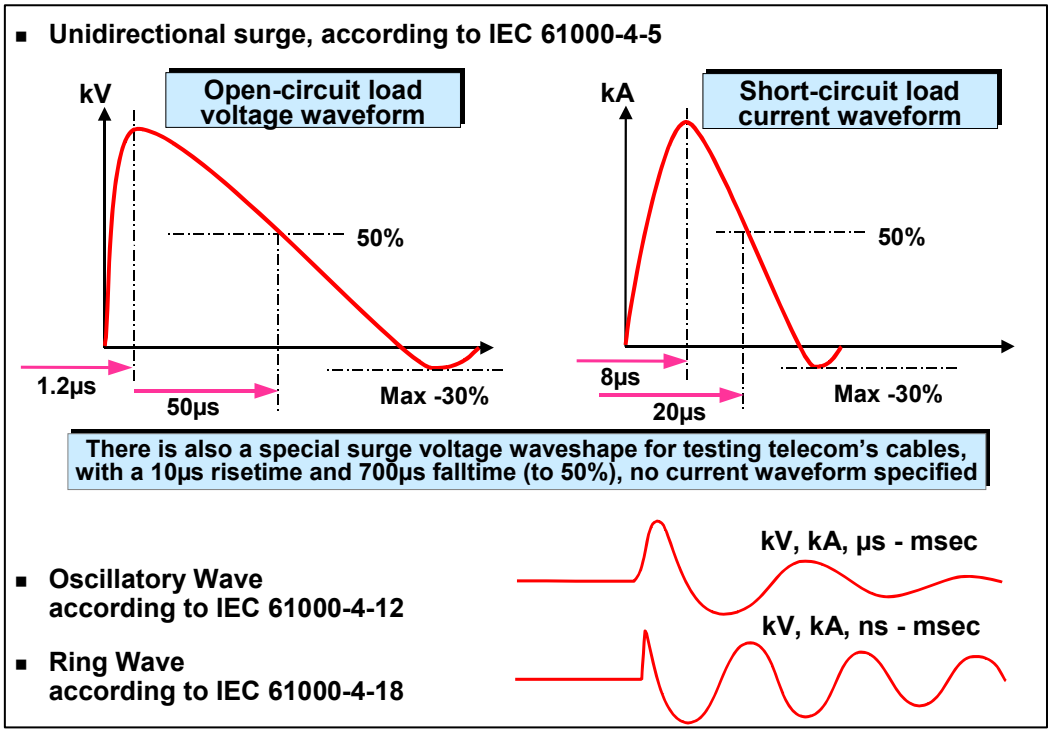


Figure 49 The IEC's three kinds of test surges

All power supply systems are plagued with transients, and some of them are related to the type of equipment that generates the power or is powered from it. So there are special test standards (e.g. ISO 7637, for motor vehicles) that specify a few types of transients that are different from the IEC ones, to try to simulate the infinite variety of transients that they can

experience. Older motor vehicles used to suffer surges of up to ±200V (maybe more) on their 12V dc battery supply systems, but modern vehicles have lower levels due to the large numbers of surge protection devices fitted to help protect all their electronics.

Surges occurring in AC power distributions that are not protected by surge arrestors correctly-installed as part of a lightning protection system (see Section 5.13 in [29]) are only limited in voltage by insulation breakdown, usually spark-over in the rear terminals of power sockets. Single-phase sockets in almost all countries worldwide tend to spark-over at around ±6kV (can be more) but three-phase power networks with only larger three-phase sockets and no single-phase sockets might spark-over at ±12kV or more.

Of course, most product standards only test with mains surges to ±2kV, because of the "economic/technical compromises" that were "agreed upon" by their standards committees.

9.1.2 Electrical fast transient bursts

These are fast random transients occurring in bursts lasting from a few microseconds to several seconds, caused by sparking. Sparks emit noise across the entire electromagnetic spectrum, from almost DC to beyond the visible range, and we can see this broad bandwidth if we measure very close to them with a broadband close-field probe. See Figure 50.

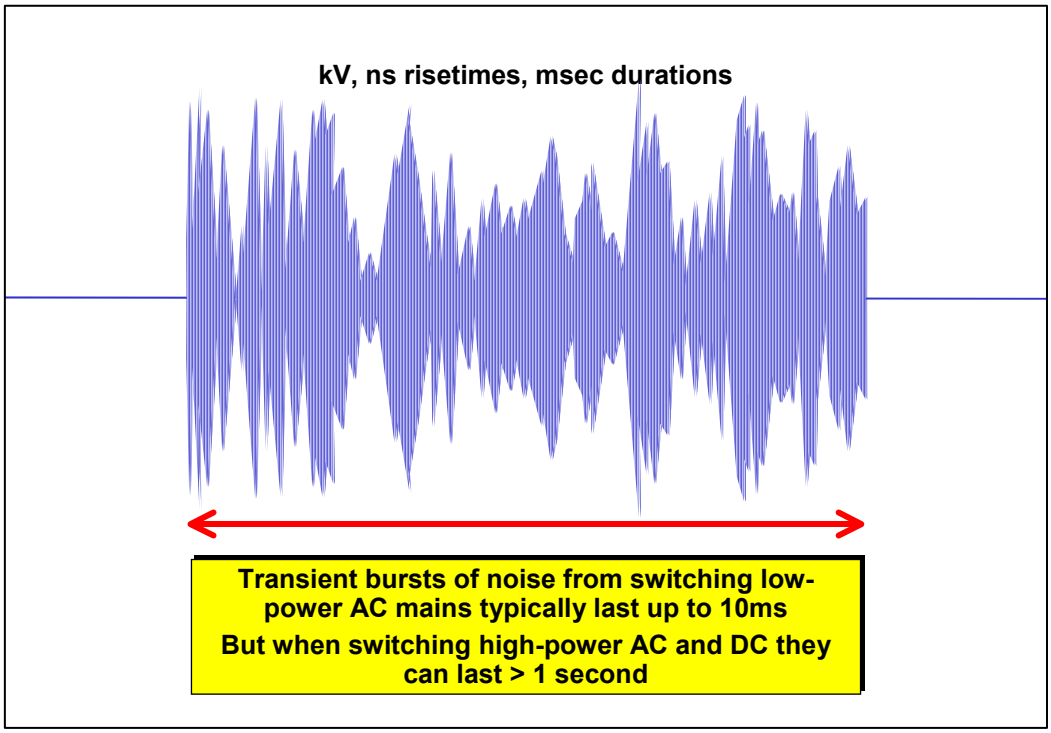


Figure 50 Fast transient bursts from sparking

However, at some distance from the spark, what we tend to see is the accidental antenna behaviour of the cables carrying the current that has been interrupted to cause the inductive flyback that caused the spark, energised by the random noise of the fast transient burst, upon which is superimposed the accidental antenna behaviour of the conductor picking up the noise.

Figure 51 shows an example of a fast transient burst from an unknown sparking event measured with a GHz broadband antenna and 1.5GHz 8Gsa oscilloscope in a larger server room, taken from [30].

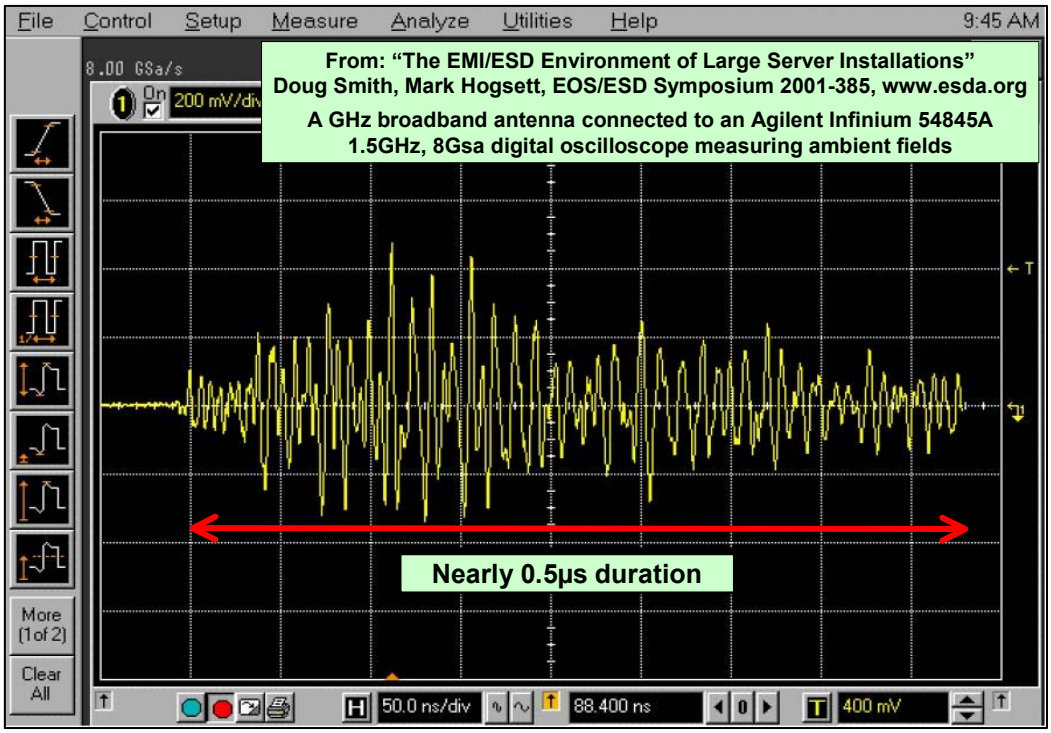


Figure 51 Example of a fast transient burst from an unknown sparking event

Any/all sparks caused by switching electrical power cause such fast transient burst noise, but the power does not have to be very high. I once had a mains indicator light that consumed 15mA from 230V, enough to cause the on/off switch to cause the product to fail the CISPR 14-1 (EN 55014-1) limit for discontinuous emissions.

The pantographs and sliding shoe contacts used in electrical traction vehicles with remote power supplies (e.g. trains, trams, etc.) will spark from time-to-time, causing electrical fast transient bursts. They are usually worse in bad weather, especially when there is icing.

Recently I saw a video of a French TGV travelling at 300mph or some such incredible speed (it was an experiment, with no paying passengers). Apart from the spectacle of all those hundreds of tons travelling at that speed on land, I was particularly struck by the fact that its pantographs were continually spraying huge sparks as it rocketed past. I remember thinking that if they planned to run this as a regular service, they would probably need to improve their technique for picking up power from overhead cables.

Poor connections in power conductors can cause continuous low-level sparking, which may be audible nearby as a slight fizzing noise. This causes *continuous* electrical fast transients, which have been known to interfere with satellite communications at GHz over a very large area (see Banana Skin No. 5 [9]).

9.1.3 Very fast transients

These tend to be single events, often caused by electrostatic discharge from people. But they can also be caused by tribocharging effects during materials processing (a big problem in photocopiers and printers) or when insulating materials that are part of some machine are rubbed (e.g. the plastic bearings of a motor; rubber drive or conveyor belts; plastic pulleys, etc.).

Figure 52 shows the noise emitted by allowing a charged up coin to touch another coin. Notice that the ringing caused by the event lasts very much longer, caused by the resonances/reverberation of the metal cabinets in the server room where this event was measured [30].

Doug Smith, the author of [30] and very much else that truly excellent on EMC and ESD, jingles a plastic bag of coins to create a low-cost near-field broadband GHz noise source. He keeps the cost low by only using low-denomination coins, but you can use gold Sovereigns

or Krugerrands if you prefer.

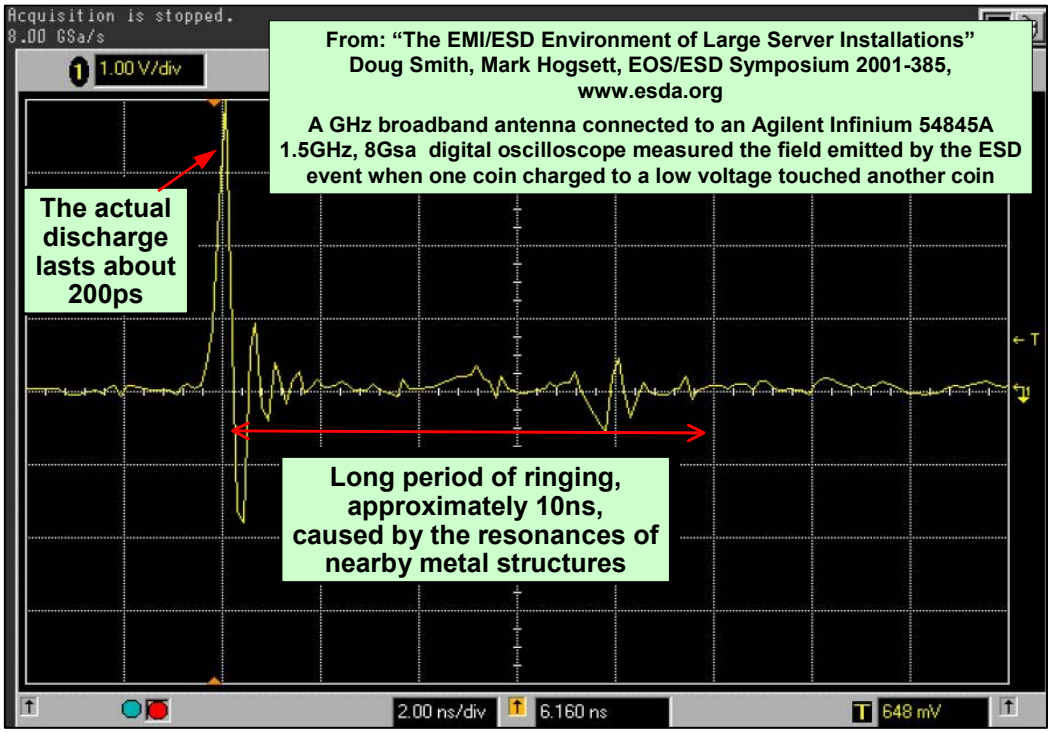


Figure 52 Example of very fast transient from a small metal discharge

Notice that the rise-time of the ESD event in Figure 52 is measured at 200ps – but it is almost certainly shorter than this because of the rise-time of the measuring system (the theoretical rise-time of 1.5GHz oscilloscope is 212ps). Normal ESD test equipment (e.g. to IEC 61000-4-2) have rise-times no faster than 700ps, but it is clear that real-life rise-times can be much less than this.

This is important, because the latest ICs, using sub-130nm silicon processes, can switch in 100ps or so (e.g. PCI Express drivers) and so can respond to ESD events with similar (or longer) risetimes, and at 45nm or less can be upset or even permanently damaged by voltages of less than 100V [31].

ESD events that discharge directly into conductors can have voltages of up to ±35kV (in very dry climates). I remember speaking to one poor designer who was having a problem with ESD punching through a 0.5mm polycarbonate layer on his membrane keyboards, which means it must have been at least ±20kV. And Banana Skin No. 418 [9] describes a popular choice of footwear causing ±25kV in a hospital and interfering with medical equipment.

However, ±8kV is more typical when the relative humidity of the air is more than 25%, with a statistical distribution that favours smaller voltages. Peak ESD currents can be 10s of Amps.

People generally do not notice personnel ESD events (usually from their fingers) when they are less than about ±3.5kV, but larger voltages usually elicit an involuntary muscular response, larger still often causing involuntary vocalisation with possibly embarrassing consequences. This means that we can all be walking around zapping electronic products and equipment, and charging up portable electronics, with up to 3.5kV without even being aware of it.

Computer keyboards are all made of plastic, and ESD tests directly applied to them usually fail to achieve a discharge at up to ±8kV. However, the PCB traces connected to the keyboard's IC have been measured as experiencing ±100V very fast transients, due to the traces coupling with the E and H fields from "indirect" discharges during IEC 61000-4-2 testing to nearby vertical or horizontal metal sheets. This shows us that nearby ESD events that are not direct discharges, can still couple noise into circuits that can cause upsets and even damage.

9.1.4 Continuous low frequency voltage and current noises

In EMC parlance, the term “Low Frequency” usually means everything less than 150kHz, but I have worked with microwave engineers who called everything less than 5MHz “DC”.

All AC mains power supplies now have distorted waveforms and carry harmonic and interharmonic currents. Interharmonic currents are low frequencies that are not related to the mains frequency, and are mostly caused by variable-speed AC motor drives.

These voltages and currents couple into other conductors via the common-impedance of the protective bonding structure (often called the “safety earth”), E and H fields.

Figure 53 shows a noise current waveform I found caused by the lighting scheme in a road tunnel under a major river. This is dominated by the harmonics of the megawatts of fluorescent and discharge lighting equipment being used. Interestingly, although all of the lamps met their harmonic emissions limits in IEC/EC 61000-3-2, the aggregation of their harmonic currents caused overheating in the lighting control switchgear (the reason I was there at all).

This is one reason for why we should never assume that all we have to do for an easy life is to buy products that are CE marked, even if we take the trouble to check the manufacturers compliance documentation and ensure that his products actually do comply with what we might assume from their CE marking (quite a lot of products do not [32]).

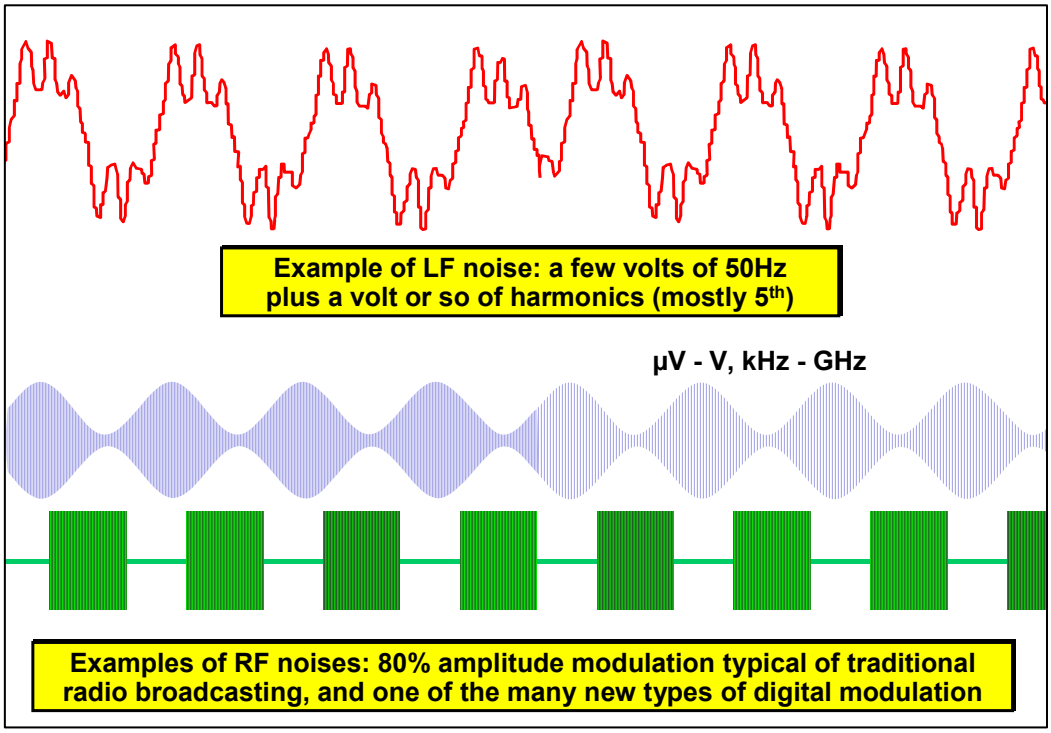


Figure 53 Example of low-frequency and RF noises

Where power is used at other frequencies (e.g. high-power audio in entertainment venues) continuous low-frequency noise will also arise in nearby conductors, at those frequencies.

When there is an insulation failure between a mains conductor and the protective bonding, or during the “follow-on” period of conduction following the discharge in a gas-discharge-tube surge protection device that is connected between an AC supply and the protective bonding, the level of mains-frequency noise induced into conductors can be as high as the mains voltage itself (e.g. up to 240V rms), and last until the fault is cleared (could be several seconds).

If the insulation fault occurs in a high-voltage mains distribution, the low-frequency noise level can reach kV for a few seconds.

[33] provides some more information on the causes of low-frequency noises in conductors, and their effects.

Trams and light rail systems are becoming quite popular in some cities, and it is unfortunate that they insist on using electrical technology that would have been familiar in the late 1800s – I mean electrical power delivered by rails or cables, with sliding connections to the vehicle. Since traction currents for one vehicle can easily reach 1000A, the H fields produced can be very large, and because the send and return conductors are so far apart, they can spread widely.

This is particularly a problem in the case of overhead cable with rail return. The send/return spacing is very large, and the H field sprays out sideways. Any equipment sensitive to low-frequency magnetic fields, such as cathode ray tube (CRT) displays, had better be fitted with hi-permeability metal or "active" H-field screening or be located at least 40m away from the line of the track. More sensitive scientific and medical equipment, like electron microscopes, had better be fitted with appropriate H-field screening (not cheap!) or located at least 200m away.

Since the current return rails for electrified traction systems are usually connected to earthing electrodes in the soil, heavy currents can flow through the ground and interact with other metal structures nearby – usually structures associated with buildings, increasing corrosion and adding to the low-frequency common-impedance noises in their protective bonding and other "earthed" metal structures.

9.1.5 Continuous radio frequency (RF) voltage and current noises

As was discussed earlier in Parts 1 [8] and 2 [17], all conductors (or slots in metal sheets) act as accidental antennas for E, H and EM waves (and their associated fields). So they pick up all the radio and TV broadcasts, mobile phone transmissions, etc., that are present in their environment – many frequencies and types of modulation – all at once.

Any conductor can be connected to a radio receiver and used as an antenna, although it might not be much good over certain frequency ranges. Wire coathangers make good antennas for FM broadcasts (around 100MHz) and it used to be quite common to see them used as car antennas when the old antenna had broken off. (In fact, I was surprised to find that the FM antenna mounted in the loft space of the house I moved into recently, and which works quite well, was a wire coathanger.)

To help save paper and hence the planet, Figure 53 includes some examples of RF noises in conductors.

9.2 EM phenomena associated with the mains power supply (Power Quality)

9.2.1 Waveform distortion

Harmonic distortion is caused by the harmonic currents drawn by non-linear loads, such as fluorescent and discharge lighting, and especially (these days) AC-DC rectifiers. Interharmonic distortion is caused by mains currents that are not related to the mains frequency, and are mostly caused by frequency-changing power converters (e.g. for variable speed AC motor drives).

Total harmonic distortion (THD) can be up to 10%, with 8% considered the limit beyond which electronic products and equipment not specifically designed for highly-distorted mains might malfunction.

It is becoming increasingly common to see THD of up to 30% for ships and offshore installations, because of the recent large increase in use of "thrusters" (variable-speed AC motors driving propellers) and even electrical propulsion. This is made worse than similar loads would cause on land, because marine vessels generate their own electricity, and since the source impedance of a generator is about three times higher than that of a HV-grid supplied distribution transformer of the same VA rating, a given non-linear load causes three times the distortion.

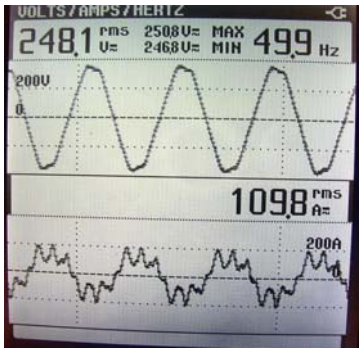
Figure 54 shows some examples of distorted mains waveforms form my own files. The top left hand one is from an industrial lighting installation, and the voltage waveform shows the sort of "flat topping" that is characteristic of supply networks that are heavily loaded by AC-

DC rectifiers.

These sorts of voltage and current waveforms are not untypical of light industrial, commercial and domestic premises. Of course, the currents tend to be lower than the 108A shown in Figure 54 but the mains source impedances tend to be correspondingly higher, so the waveform distortion often stays much the same.

The top right hand waveform was provided by Dr Nick Maroudas, from an oscilloscope measuring the mains in his house in Israel in 2000, see Banana Skin No. 104 [9].

The bottom right waveform is from a generated supply feeding a 1MW AC motor drive switching at about 1.7kHz. The measuring equipment had a bandwidth of only 5kHz (typical of power quality instrumentation) and if it had a wider bandwidth the depth of the notches in the waveforms would have been much deeper than their already very excessive magnitude.

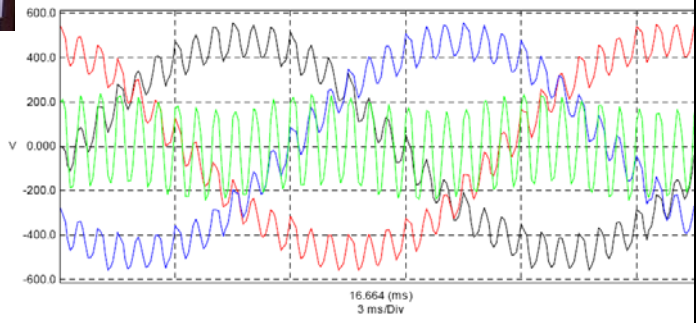


A typical industrial lighting installation

Upper trace: Voltage

Lower trace: current

A domestic house, Israel, 2000



Generator-supplied (off-grid) AC mains with very high distortion caused by a 1MW motor drive

Figure 54 Examples of AC mains distortions

Harmonic and interharmonic currents, and the voltage waveform distortions they cause, are just an EMI problem. They can cause overheating with very costly, if not dangerous consequences. [34] and [35] provide more information on harmonic and interharmonic currents, waveform distortion, and their effects.

Of course, waveform distortion is a problem only for AC supplies, whether 50Hz, 60Hz or 400Hz (aircraft). There are many other power quality problems that afflict both AC and DC supplies (e.g. the 48V supplies used in telecommunications rooms, and by blade servers [36].)

9.2.2 Fluctuations in the supply voltage

Rapid fluctuations in supply voltage are caused by rapid load fluctuations (assuming there is not some instability in the network voltage control system). Up to ±10% is possible; higher levels usually indicating a serious problem with the supply impedance.

Voltage dips and flicker from power distribution network control and fault-clearance can be of any depth and any duration, with a statistical distribution in favour of smaller depths and durations.

Dropouts and interruptions are also caused by power distribution network protection and fault-clearance, and of course interruptions can last for weeks (especially in the case of a natural disaster).

Figure 55 shows some rather clean versions of these three power quality issues. In real life, they are unlikely to look as simple. For example, when an insulation fault causes a fuse to blow or a circuit breaker to open, there will generally be a dip or dropout at first, combined with a significant increase in mains-related noise in the protective bonding structure (also appearing on earthed neutrals), plus a locally intense increase in H-field emissions due to the very high levels of current in the fault. There may also be some continuous broadband RF noise if the fault includes some sparking.

Then when the faulty circuit is eventually disconnected by the overcurrent protection device, there will be a larger burst of RF noise (an electrical fast transient burst) due to the sparking as the large amplitude of the fault current is interrupted. This will be followed by a surge of up to double the nominal mains voltage, caused by the flyback of the energy stored in the mains network's wiring during the gross overcurrents that occurred during the fault.

Of course, when tested in a lab, for all products except critical telecommunication infrastructure (and maybe not even then), they test with just one type of EMI at a time.

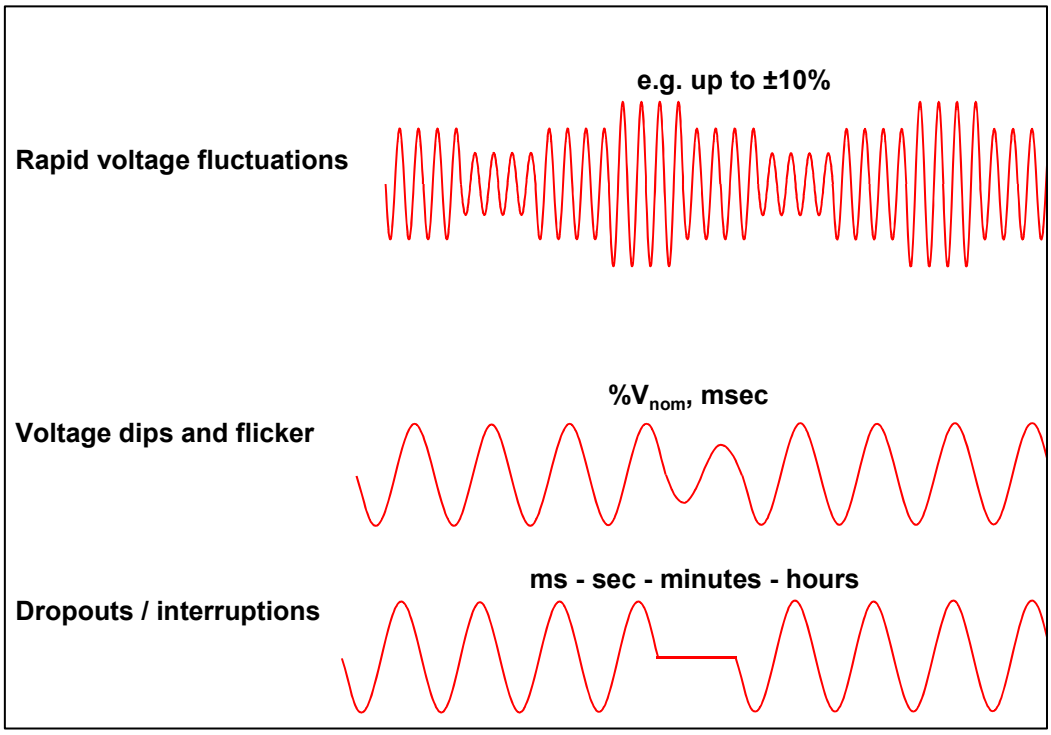


Figure 55 Examples of some more power quality issues

Figure 56 sketches some more power quality issues:

- Slow variations (called sags and swells in IEC-speak, surges and brownouts in the USA) caused by long-term load variations.
- Three-phase voltage unbalance, in voltage and/or phase, caused by unbalanced loads or insulation failures in three-phase systems.
- Frequency variations caused by significant load fluctuations on generators. For national-grid-supplied power, frequency variations usually do not exceed ±0.5% (except for a few seconds before load-shedding to protect the grid causes a whole area to lose power).

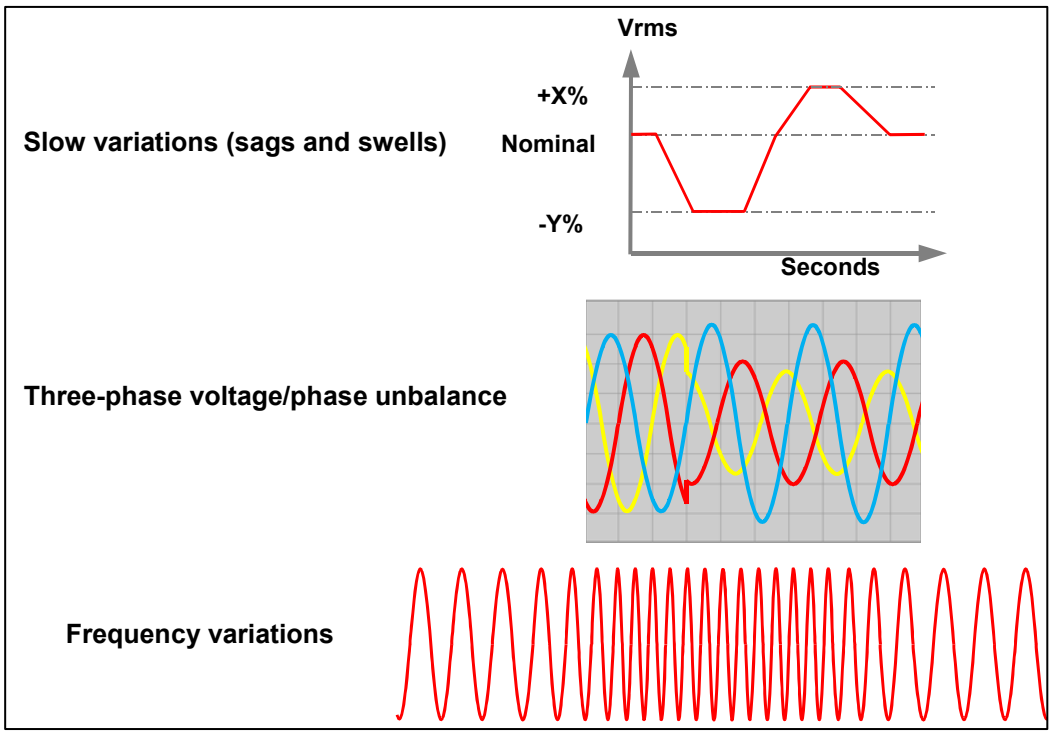


Figure 56 More examples of power quality issues

Don't forget that because AC and DC power distribution uses cables, all of the EMI issues discussed in 9.1 apply – often to a larger extent because many of the electrical noise sources are directly connected, rather than induced, and because mains cables can be very long indeed and so pick up higher levels of lower-frequency noises.

And when running from a generator, for example a hospital running on its emergency mains supply, most/all power quality issues are made much worse.

[37] provides more information on a wide range of power quality issues, including how to deal with them.

9.3 High Power Electromagnetic threats (HPEM)

The topic of HPEM includes:

- Lightning
- Powerful radio and radar transmitters creating High Intensity Radiated Fields (HIRF), for example at airports and harbours.
- Nuclear electromagnetic pulse, which comes in various "flavours": EMP, NEMP and HEMP
- Intentional EMI (IEMI) – using a variety of powerful EM devices often originally intended for military use, some of which can be purchased for private use, or constructed by a reasonably competent engineer, making IEMI a real possibility for some applications.

Figure 57 shows some examples of HPEM, and is taken from IEC 61000-1-5. Figure 58 shows an example of an IEMI weapon in a suitcase.

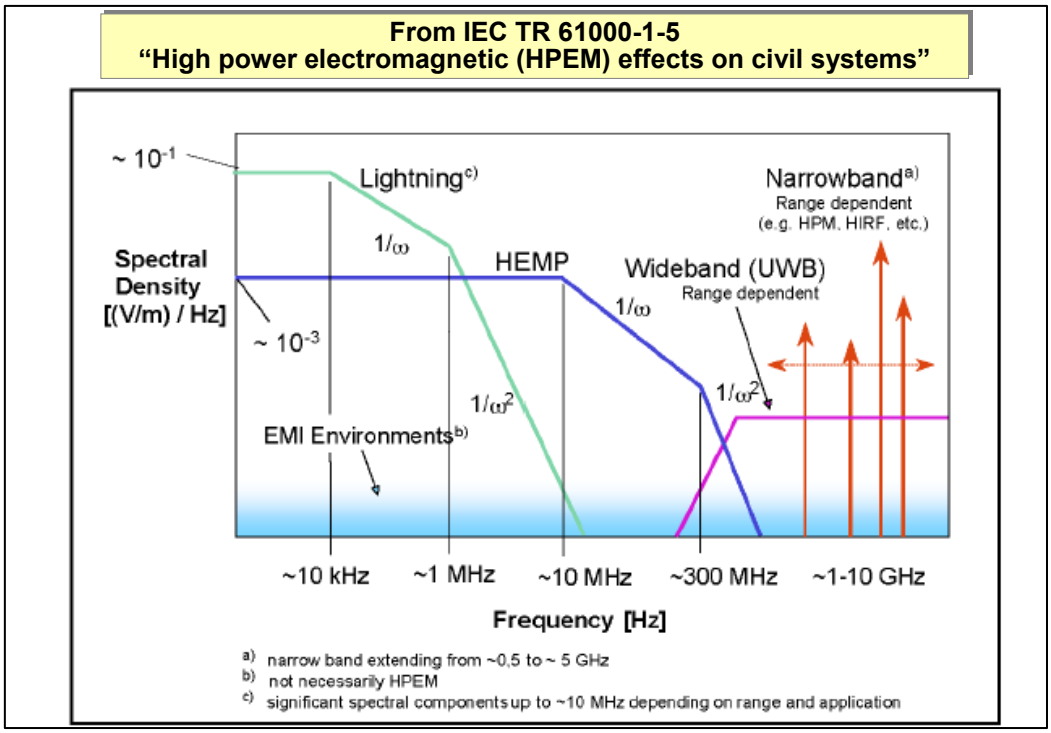


Figure 57 Examples of HPEM threats

This IEMI weapon is available from Diehl Munitions Systeme for €150k (in 2000).

Suitcase sized, it will run off its batteries for 3 hours, creating pulses of 120kV/m (at 1m).

It will prevent operation of non-military hardened electronics for a radius of at least 20m.

Figure 58 Examples of an IEMI weapon

A number of IEC standards have recently been written, describing various HPEM threats, and how to harden systems against them, in the 61000-5 series [38].

9.4 Mechanically or climatically induced EMI

This issue concerns electrical connections, and electrical insulation. Poor-quality connections can have a variable contact resistance, and even make intermittent connections. The quality of their contact resistance and whether they are intermittent (temporarily open-circuit) is frequently affected by mechanical vibration, and temperature

changes (because of coefficients of thermal expansion).

So we can get modulations in supplied voltages and signals, caused by mechanical shocks, vibration, and temperature. Sometimes they can mimic real signals, for example when a engine vibration causes a poor connection in the cable from a vehicle's speed sensor to produce noise that can be mistaken for a valid speed signal [39].

Poor quality electrical insulation can cause corona discharge (broadband RF noise), maybe even spark-over occasionally (causing electrical fast transient bursts) due to shock, vibration, or climatic effects such as condensation or conductive dust.

And all these problems get worse with ageing and corrosion (whether fretting, oxidative and galvanic).

9.5 High voltage power distribution using overhead cables

The high voltages in the conductors cause local corona discharges in the air – continuous broadband noise at the frequency of the mains supply, plus all of its harmonics up to 100s of GHz. In the case of HVDC, it is just random noise all the way from DC to 100s of GHz.

9.6 EMI Victims: Analogue circuits

The effects of EMI on analogue circuits include noise and/or distortion and/or zero-shift errors of up to ±full-scale. This causes problems particularly for measurement and control of physiological parameters; chemical reactions; temperature; pressure; weight; mass; flow; velocity; movement; level; etc., which often use analogue sensors that need signal conditioning, amplification, and analogue-digital conversion.

Analogue semiconductors are easily destroyed by overvoltages and overcurrents, although they are generally not as vulnerable to this problem as digital devices, because they usually larger silicon processes that have thicker insulation layers.

However, there are some analogue devices that use very small features indeed, whether in silicon (e.g. RF power devices) or other materials (e.g. giant magneto-restriction (GMR) devices used for reading data in computer hard drives) – these could be permanently damaged by just a few volts above their nominal supply rating.

9.7 EMI Victims: Digital circuits (including software)

A well-designed digital circuit has a significant noise margin – a voltage gap between the peaks of its self-generated circuit noise (e.g. due to ground/power bounce) and the thresholds at which its devices each decide whether they are being supplied with a signal that is either 1 or 0.

External noise (which could be from other parts of the same product, or from the ambient EM environment) "fits into" this threshold, so that normal levels of EMI do not cause a device to mistake a 1 for a 0, or vice-versa.

So, in a well-designed digital circuit, functional performance errors tend only to occur when the EMI's magnitude passes a threshold, but then a variety of malfunctions can occur – often quite unpredictable. Sometimes, when testing digital circuits with various EMI threats, we see the following escalation in functional errors:

- False key-presses, errors in communications, data and control, which could be quite dangerous, for example changing operational mode (e.g. from crawl speed to full speed). These errors are caused by the accidental antenna behaviour of the long cables or PCB traces usually associated with manual controls, keyboards, and datacommunications.
- As the level of the EMI is increased, smaller cables and PCB traces start to pick up enough noise to cause all sorts of incorrect software operation, not just limited to controls and datacomm's, for example continually repeating an inappropriate activity (stuck endlessly repeating a section of the program) and of course also being able to

change the operational mode.

- At higher levels of EMI still, stopped operation (often called a “freeze” or “crash”) will occur. This is usually detected by a “watchdog” circuit that reboots the software process after some time has passed, resuming normal operation (or some default state). But while the processor is rebooting, its control outputs will assume random combinations of states, which can possibly include those with undesirable or unsafe results for whatever is being controlled.
- At higher levels still, digital devices can be permanently damaged. Modern processors and memory ICs are very easily destroyed by overvoltages, maybe by as little as 10V, because of the very small silicon processes they are made on, e.g. 60nm or less. At these levels, the insulation layers are films of silicon nitride whose thickness is measured as a (small) number of atoms, and incapable of handling more than a few volts.
  In the past, ICs were protected by ESD diodes connected to their I/O and power pads, but with these smaller silicon processes and the huge number of I/Os they support, there simply isn’t the room for such high levels of protection, so as well as the chips being much more susceptible to overvoltages, they are less well protected.

9.8 EMI Victims: Power switching semiconductors

These include IGBTS, SCRs and PowerFETs. They can all be permanently damaged by overvoltages, surges, ESD and overcurrents. Being large devices intended for high voltages and frequencies of up to only a few 10s of kHz, they generally have large internal capacitances, lots of insulation and heavy-duty conductors, so are not easily damaged by most types of EMI.

In most high-power switching applications, very high levels of “internal EMI” are created anyway, and if the devices were not very immune they would not survive. Even so, care must be taken with their design if they are to be reliable and not blow themselves up, and some types of power switching devices use optical fibres instead of conductors to control their gates, to make them more immune to noise.

However, all power switching devices are driven by lower-power circuits, in turn controlled by microprocessors or similar low-power ICs, which can be fairly easily interfered with or damaged. If the control terminals of the high-power switching devices are triggered at the wrong time, cross-conduction can occur, shorting the power supply out through the devices and causing malfunctions, and/or actuation of protective devices, and/or actual damage.

When high-power devices cross-conduct, they can explode like hand-grenades, with similar energies and shrapnel.

9.9 EMI Victims: latchup can affect all semiconductors

Latchup is an extremely important factor in determining product reliability, and affects NMOS, CMOS, Bipolar and all variations and combinations of these technologies, whether they are used for analogue or digital processing [40]. A negative or positive voltage transient on any input or output pin of an IC, that exceeds either supply rail voltage by more than one diode drop, is a common cause of latchup.

Latchup can also be caused by high temperatures, and by ionizing radiation, and the presence of either increases the susceptibility of latchup to voltage transients.

During latchup, parasitic SCR-like structures in the IC turn on and short-out the power rails. If the power current is not limited by external components, the IC will overheat and could be permanently damaged.

Where overheating is prevented by the external power supply circuit (as it may need to be, for safety reasons, to prevent fire) and the IC is not damaged, it can only be restored to normal operation by removing the power, waiting a few seconds for the device to cool, then reapplying the power – when the IC will work normally again.

Clearly, all types of EMI transients can cause latch-up, but they are also often caused by non-EMI effects such as hot-plugging, or switching products on or off when they contain multiple voltage rails, when those activities can cause a voltage that is more than one diode drop above or below the instantaneous values of an IC's power rails to be applied to one or more of its input or output pins.

9.10 EMI Victims: Electromechanical devices

Many designers seem to assume that electromechanical devices are totally immune to all EM threats, but dips and dropouts in their electrical power supply can cause relays, contactors and solenoids to drop out. What it will take to make them drop out is hard to predict because susceptibility varies individually, depending on their make and model, age, and temperature.

Few, if any, "relay logic" designers ever test for susceptibility to power quality issues, which is surprising when the main use of electromechanical devices these days is in "hard-wired" safety systems, possibly protecting the lives of dozens of people.

The US Nuclear Inspectorate discovered, many years ago, that the relays used in safety systems in nuclear power plants, would drop out with mains dips that were rather common when compared with the reliability required of the safety systems. So they made the plant operators fit "coil hold-up" devices to the supply to each relay. These would keep the relay operating as usual even when the mains had dropped to less than 50%, indefinitely, and would hold-up through most dips and dropouts to lower voltages. (Since generating plants create electrical power all the time, a total loss of power, such as the rest of us might expect and have to protect against, is not a problem.)

There are at least two common relay logic practices that can cause particular problems when power quality issues cause relays, contactors or solenoids (e.g. in safety door interlocks) to become de-energised.

One is when relays or contactors are "held-in" by a normally-open contact. They are usually energised by a momentary contact, such as a push-button, and then hold-in by power applied through the normally-open contact, which is now closed. If the relay or contactor drops out through its supply falling momentarily below its hold-in level, it will not recover to normal operation when the dip, dropout or voltage fluctuation, which might only have lasted a few milliseconds, has passed.

The other is when the relays, contactors or solenoids are pulled-in at full voltage, but then kept energised by a reduced "hold-in" voltage, to save power. Depending on the way in which their reduced voltage is supplied, they may drop out during a dip, dropout or voltage fluctuation – and not pull back in again when power is back to nominal.

Other EMI problems with electromechanical devices include:

- High levels of shock and vibration can make electrical switch contacts "chatter", causing sparking that generates broadband RF noise (see 9.1.2) that can interfere with electronic devices. This is an example of mechanically-induced EMI.
- Overvoltages due to surges and fast transients can make open contacts spark-over, which is the same as closing them momentarily – applying power more-or-less at random to circuits which should be off, causing who knows what kind of functional errors or safety problems.
- And surge currents can weld closed contacts together, so that they won't open when they are supposed to.

This is a problem for most types of switches, relays and contactors, because they can change their mechanical state from ON to OFF without their electrical state changing at all (or only partly changing). Once again, power may be supplied to circuits that are meant to be off.

Also, despite instructions to isolate at the mains supply before removing any covers, most operators and maintenance engineers would assume that a circuit was safe because they

can see that the switches, relays or contactors that supply power to it are in the OFF position, not realising that some or all of its electrical contacts might still be ON. Clearly an electric shock hazard.

Switches, relays and contactors are available with “positively-guided” or “forced” contacts, that cannot change state mechanically unless they also change state electrically, but choice is very limited and they are larger and more costly. An alternative might be to use devices fitted with low-current contacts that provide feedback of actual contact position.

One of the problems with electromechanical components, is that their manufacturers often try to improve or add to their functionality by adding electronics to them.

For example, I saw an advertisement some years ago for a motor control contactor (MCC) that said this new model had an “electronic brain” (i.e. a cheap microprocessor) that provided an extra dozen motor protective functions to the few normally provided by the bi-metal strips, coils and magnets used by traditional MCCs.

What the advertiser should have added, was a bold warning that you should not use their new “brainy” MCCs near to any high-power switching converters; on very noisy supplies; within 5 metres of any walkie-talkie or 3 metres of any cellphone– if you wanted your motor to remain protected, and not to stop working at random times. Of course, these are all places where the traditional design of MCC would have been perfectly happy, and in fact were often installed.

So it is a good idea to ask about the electronic content of any electromechanical device, and if it contains even one rectifier, transistor, or IC, treat it as an electronic device and not as an electromechanical one.

9.11 EMI problems are worsening all the time

This is happening because:

- Traditional electrical, electromechanical, pneumatic, hydraulic and mechanical technologies are continually adding (or being replaced by) electronic technologies.
- Electronic technologies are emitting more noise (e.g. wireless communications, switch-mode power conversion, digital processing, etc.) whilst also becoming more susceptible to EMI due to the use of ever-smaller silicon processes
- The complexity of both hardware and software is continually increasing, making it harder to understand and perform risk analysis upon.

It surely can’t be long before the first question asked of any candidate for an electronic design or design management position (hardware, software, PCB, cabling, enclosure design, etc.) is “What do you know about EMI and EMC?” It simply won’t be worth the financial risk of employing anyone who does not understand how to apply good EMC engineering in everything they do.

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