EMAC services

Interference coupling fundamentals

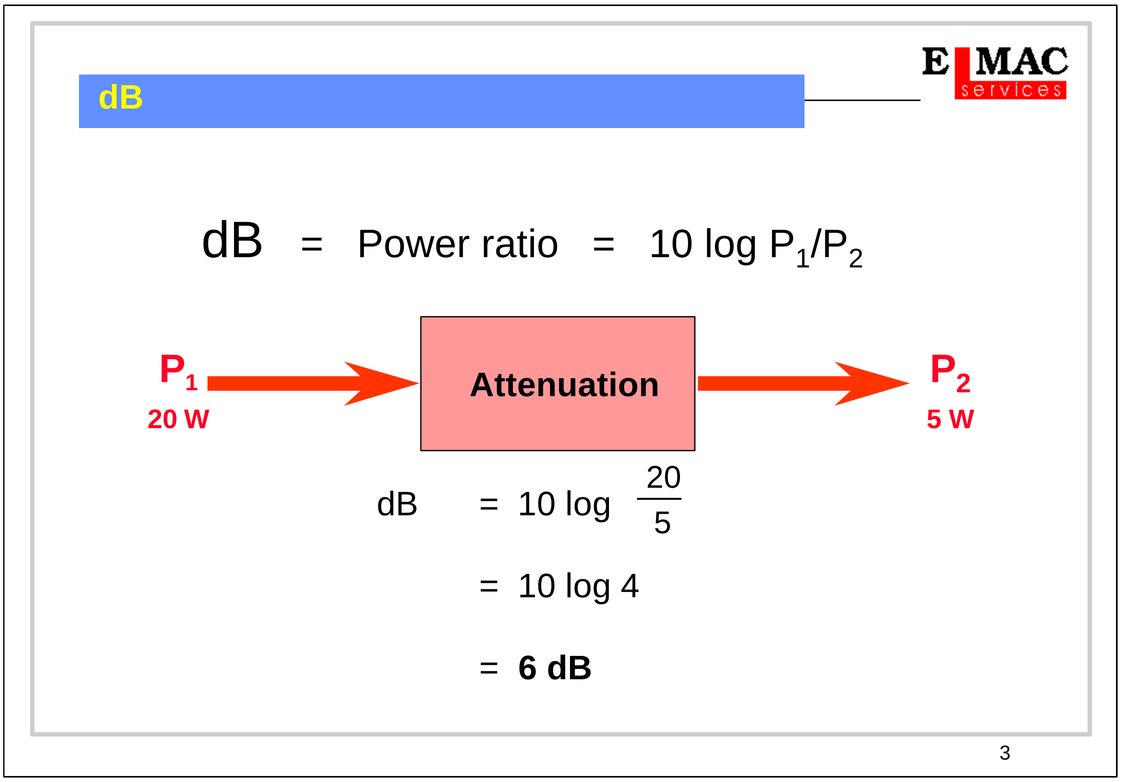
Section 3

1

Outline

- the dB
- the frequency domain
- coupling modes
- electromagnetic fields
- emission and immunity considerations

2



A dB or deciBel is a convenient way of expressing large ratios in a logarithmic fashion.

A Bel is simply the log (to the base 10) of the ration of two numbers and a deciBel is 1/10th of a Bel. So a power ratio P_1/P_2 can be expressed in dB as shown. Notice that if the ratio were inverted the result would be the same dB value but negative; it is reasonable to say that an *attenuation* of 6dB is equivalent to a *gain* of -6dB.

Some simple relationships are evident:-

3dB is a ratio of 2

6dB is a ratio of 4

9db is a ratio of 8

etc.

10dB is a ratio of 10

20dB is a ratio of 100

30dB is a ratio of 1000

Combining these 23dB is a ratio of 100 and 2 i.e. 200

17dB is a ratio of 100 and 1/2 i.e. 50

dBm

dBm = Power level = comparison relative to 1 mW

$$\text{dBm} = 10 \log \frac{P\ (watts)}{0.001}$$

P_1 1 W → $P_{1'}$ 30 dBm

P_1 10 W → $P_{1'}$ 40 dBm

4

If one of the numbers in the ratio is fixed i.e. to 1mW.

We get the dBm which is a ratio of power compared to a mW.

Therefore 3 dBm is 3dB above a mW i.e. 2 mW

6 dBm is 6 dB above a mW i.e. 4 mw

9 dBm is 9 dB above a mW i.e. 8 mw

10 dBm is 10 dB above a mW i.e. 10 mw

20 dBm is 20 dB above a mW i.e. 100 mw

30dBm is 30 dB above a mW i.e. 1 w

Voltage Ratio

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Power (W) = $\frac{V^2}{R}$ therefore for P_1 and P_2 ...

$$P_1 = \frac{(V_1)^2}{R_1} \quad \text{and} \quad P_2 = \frac{(V_2)^2}{R_2}$$

Usually R_1 and R_2 are equal and very often 50 Ohm

$$dB = 10 \log \frac{(V_1)^2}{(V_2)^2}$$

$$dB = 20 \log \frac{(V_1)}{(V_2)}$$

5

dB can also be used to express voltage, current and resistance ratios.

However, since the dB was originally intended as a power ratio and power is related to current and voltage squared, we need to multiply the logarithmic number by 2, so we now have 20 times the log of the ratio instead of 10 times. This changes all the relationships previously mentioned.

For Voltage and Current

6dB is a ratio of 2

12dB is a ratio of 4

etc.

20 dB is a ratio of 10

40dB is a ratio of 100

60dB is a ratio of 1000

Voltage Levels

Voltage can be expressed in deciBel notation, as with the dBm

$$\text{dB}\mu\text{V} = 20 \log \frac{V\,(volts)}{10^{-6}}$$

V_1 1 V	→	**$V_{1'}$** 120 dBµV
V_2 3 V	→	**$V_{2'}$** ~130 dBµV
V_3 10 V	→	**$V_{3'}$** 140 dBµV

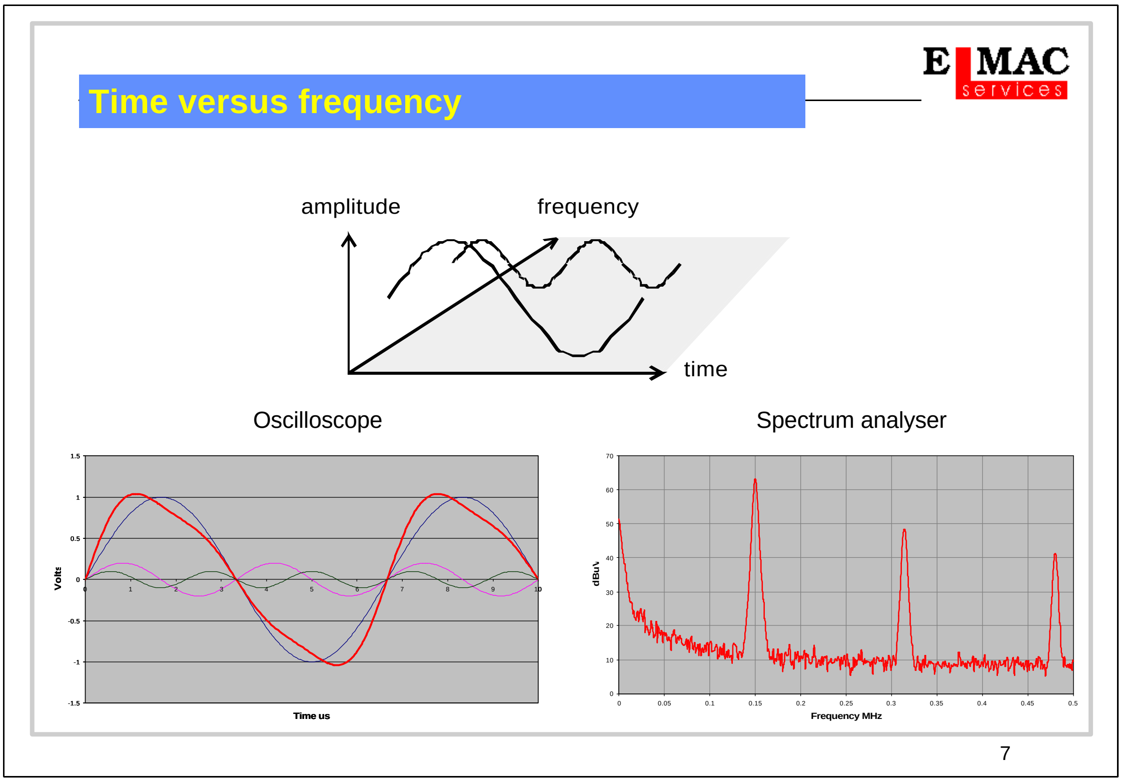
6

As for the dBm we can create other useful units such as :-

dB µV	dB above or below 1µV
dBmV	dB above or below 1mV
dBV	dB above or below 1V
dB µA	dB above or below 1 µA
dBmA	dB above or below 1mA
	etc

1mV is therefore equal to 60 dBµV

1 V is equal to 120 dBµV

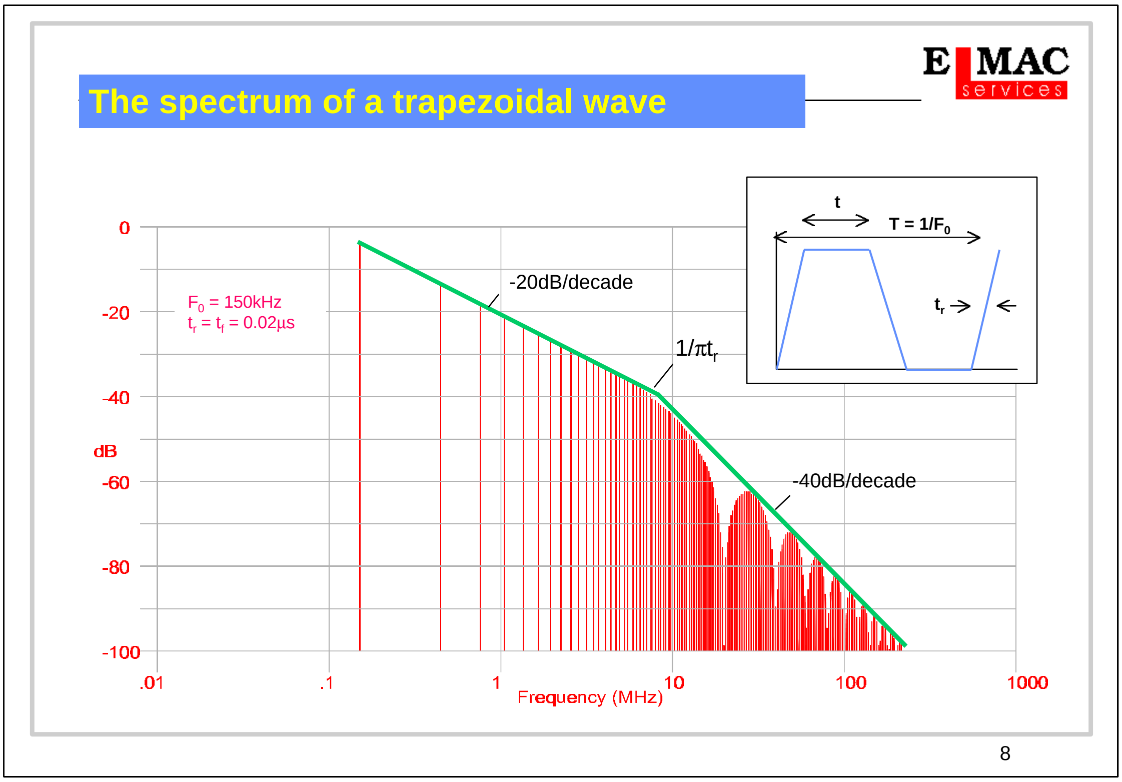


Much EMC work is conducted in the frequency domain. The respective views in the time domain and the frequency domain of a sinewave with its second harmonic illustrate the difference between the two.

The time domain oscilloscope display is of a single amplitude-versus-time waveform, despite the fact that the signal can actually be separated into two sinusoidal components. The oscilloscope does not separate these components. However, displaying amplitude versus frequency reveals them. Their amplitudes can be measured separately.

The spectrum analyser is most useful when many signals of different frequencies are present. Displaying the composite of such signals in the time domain would be meaningless.

Strictly speaking, a full conversion between time and frequency domains involves knowing the relative phases of the various components. This information is of no use for EMC purposes and is invariably ignored.



A sinusoid consists of a single frequency component. In contrast, a finite transition-time square wave, or trapezoidal wave, is rich in odd harmonics (that is, odd multiples of the fundamental frequency). It can be thought of as the sum of an infinite series of sine waves of appropriate amplitudes and frequencies. Trapezoidal waveforms are of course very common in electronic circuits that use switching or digital techniques.

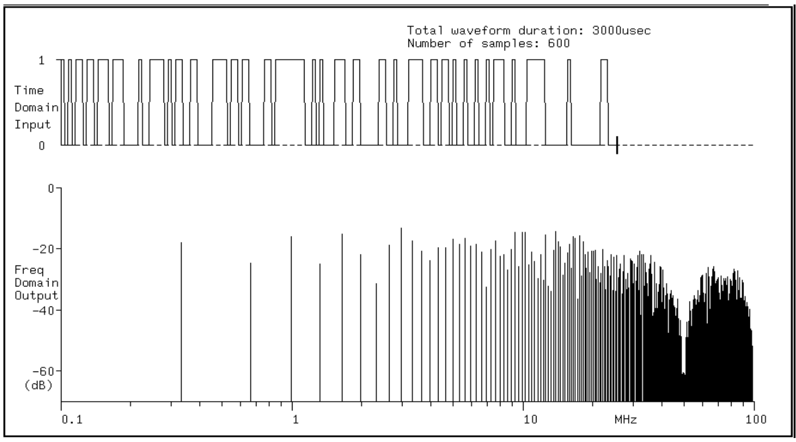
The amplitude envelope of the harmonics decreases at a rate of 20dB per frequency decade until a breakpoint is reached at $1/\pi t_r$, a frequency which is determined by the rise and fall times of the trapezoid. Beyond this frequency the amplitude decreases at 40dB/decade. The actual amplitude of each harmonic follows a (sinx)/x law:

$$V_n = 2Ad \cdot \{\sin(n\pi f_0 T)/(n\pi f_0 T)\} \cdot \{\sin(n\pi f_0 t_r)/(n\pi f_0 t_r)\}$$

This spectrum can readily be observed by using a spectrum analyser to monitor the clock waveform in a digital circuit. A pure symmetrical trapezoidal waveform does not contain any even harmonics, and varying the duty cycle of the square wave away from 1:1 does not add any. Instead it reduces the amplitude of the lower-order harmonics and leaves the upper-order ones untouched.

Differing rise and fall times, in a real, asymmetrical, trapezoid, do add even harmonics. The limit in this case is the sawtooth waveform which exhibits an equal mix of odd and even harmonics. The triangle wave contains only even harmonics.

Broadband noise



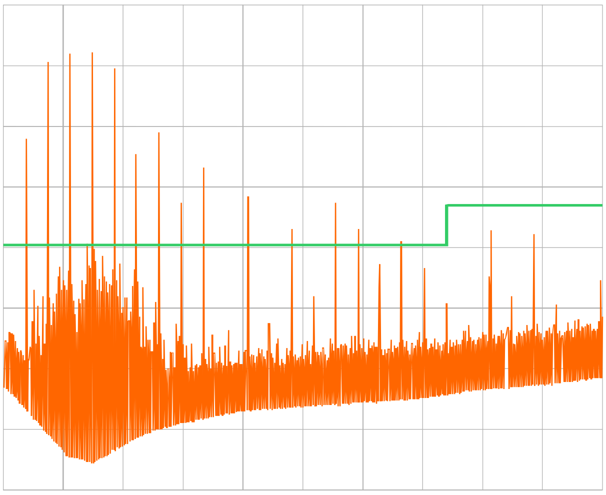
9

Digital data waveforms can be approximated by a random bit sequence. The spectrum of such a sequence contrasts with that of clock signals in that rather than being concentrated at a single frequency and its harmonics, it is spread nearly continuously over a wide frequency range. The spacing of the individual spectrum lines is given by the reciprocal of the repetition rate of the sequence, and the first null in the spectrum is at the reciprocal of the bit period. A truly random sequence, i.e. one which does not repeat, has a continuous spectrum. Many digital processes repeat with a period measured in milliseconds and their spectra are quasi-continuous with line spacings less than 1kHz, which is essentially broadband for EMC measurement purposes.

Because the energy from a given circuit carrying a data signal is spread nearly evenly across the spectrum, its amplitude at any one frequency is many times lower than a clock signal from a circuit with equivalent radiating parameters.

A further very important source of broadband noise is that due to motors and other repetitive switching operations. In this case each interruption of current causes a brief, fast-rise-time transient which has a very broad, continuous spectrum. Repetition of the interruptions results in noise which is also quasi-continuous in time.

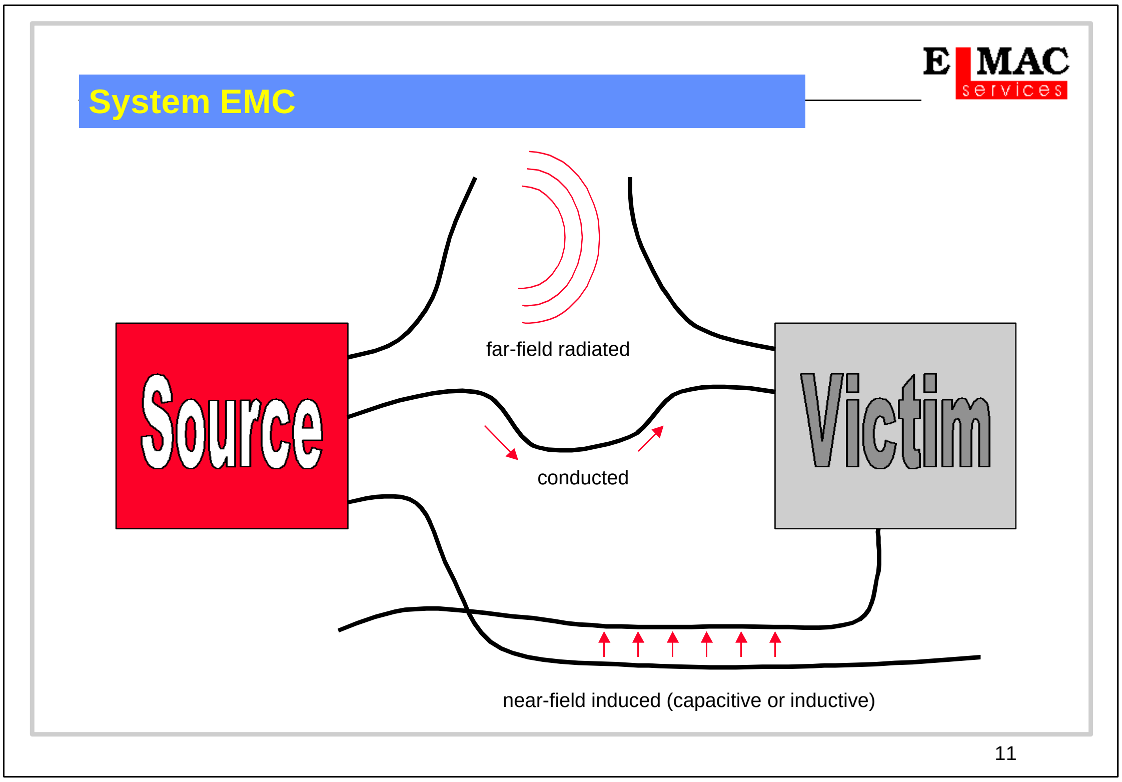
Clock and data emissions



10

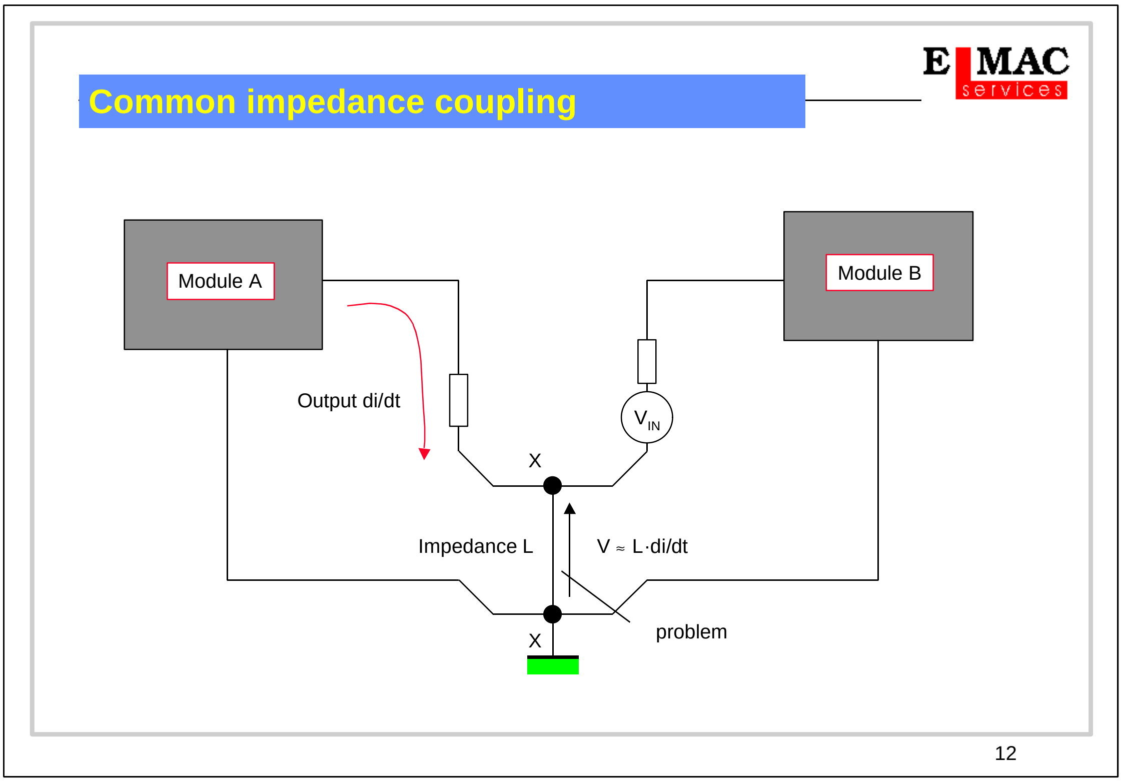
The main source of radiation in digital circuits is the processor clock (or other system clocks) and its harmonics. All the energy in these signals is concentrated at a few specific frequencies. Since the radiated emissions standards do not distinguish between narrowband and broadband, these narrowband emissions should be minimized first, by proper layout and grounding of clock lines. Then pay attention to other broadband sources, especially data/address buses and backplanes, and video or high-speed data links. The least significant data/address bit usually has the highest frequency component of a bus and should be run closest to its ground return. Because backplanes carry all buses and clocks in the system, and carry a heavy capacitive loading, they should always use a multilayer board with a ground plane, and daughter board connectors should have a ground pin next to every clock or data pin.

The diagram shows a typical radiated spectrum from a microprocessor-based product. Note as well as the clock harmonics the tell-tale “hump” in the broadband noise spectrum due to low-Q cable resonances.



Putting source and victim together shows the potential interference routes that exist from one to the other. When systems are being built, it is necessary to know the emissions signature and susceptibility of the component equipment, to determine whether problems are likely to be experienced with close coupling. Adherence to published emission and susceptibility standards does not guarantee freedom from systems EMC problems.

In practical situations, intra-system and external coupling between equipment is modified by the presence of screening and dielectric materials, and by the layout and proximity of interfering and victim equipment and especially their respective cables. Ground or screening planes may enhance an interfering signal by reflection or attenuate it by absorption. Cable-to-cable coupling can be either capacitive or inductive and depends on orientation, length and proximity to other materials. Dielectric materials may also reduce the field by absorption. Each component has a complex frequency-dependent behaviour and may include or introduce harmonic and intermodulation components due to non-sinusoidal waveforms and nonlinearities.



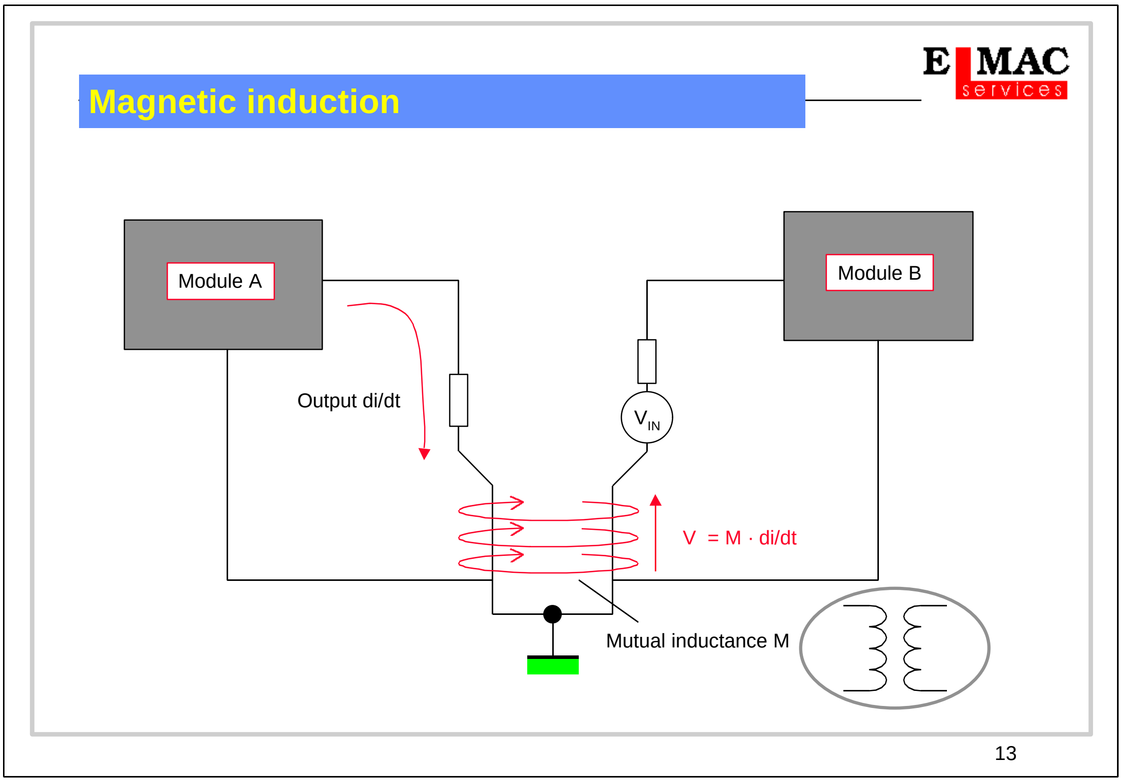
When an interference source (output of system A) shares a ground connection with a victim (input of system B) then any current due to A's output flowing through the common impedance section X-X develops a voltage in series with B's input. The common impedance need be no more than a length of wire. This has an increasing impedance with frequency because it is inductive, so high frequency or high di/dt components in the output will couple efficiently.

The solution is to separate the connections so that there is no common current path, and hence no common impedance, between the two circuits. This applies to any circuit which may include a common impedance, such as power rail connections. Grounds are the most usual source of common impedance because the ground connection is taken for granted.

The inductance of a straight length of wire is

$L = 0.0051 \cdot l \cdot (2.3 \log (4l/d) - 1)$ µH, where *l* and *d* are length and diameter in inches.

A useful rule of thumb is 20nH/inch.



Even with no direct connection, the fields resulting from circuit operation will allow coupling between two adjacent circuits. This is known as *reactive* or near-field coupling.

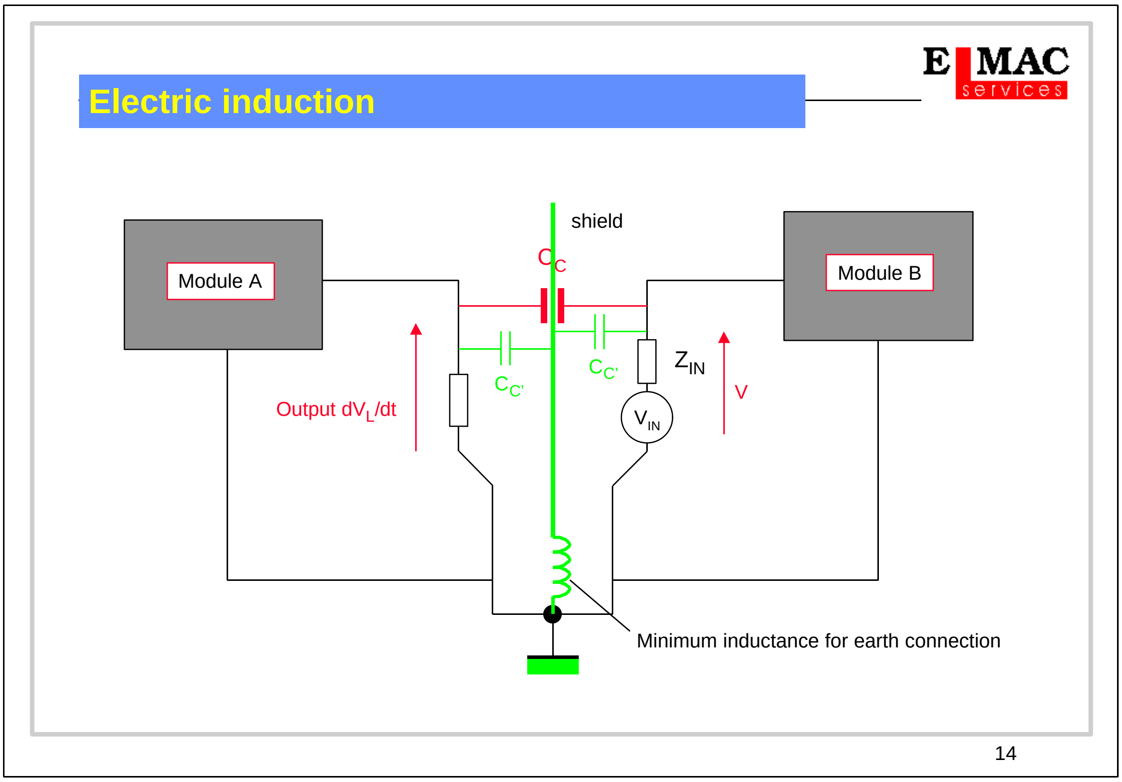
Changing current flowing in one conductor creates a magnetic field which may couple with a nearby conductor and induce a voltage in it. The voltage induced in the victim conductor is

$$V = -M \cdot dI_L/dt,$$

where M is the mutual inductance in henries.

M depends on the areas of the source and victim current loops, their orientation and separation, and the presence of any magnetic screening (note that this is not the same as electric field screening).

Typical values for short lengths of cable loomed together lie in the range of from 0.1 to 3μH.

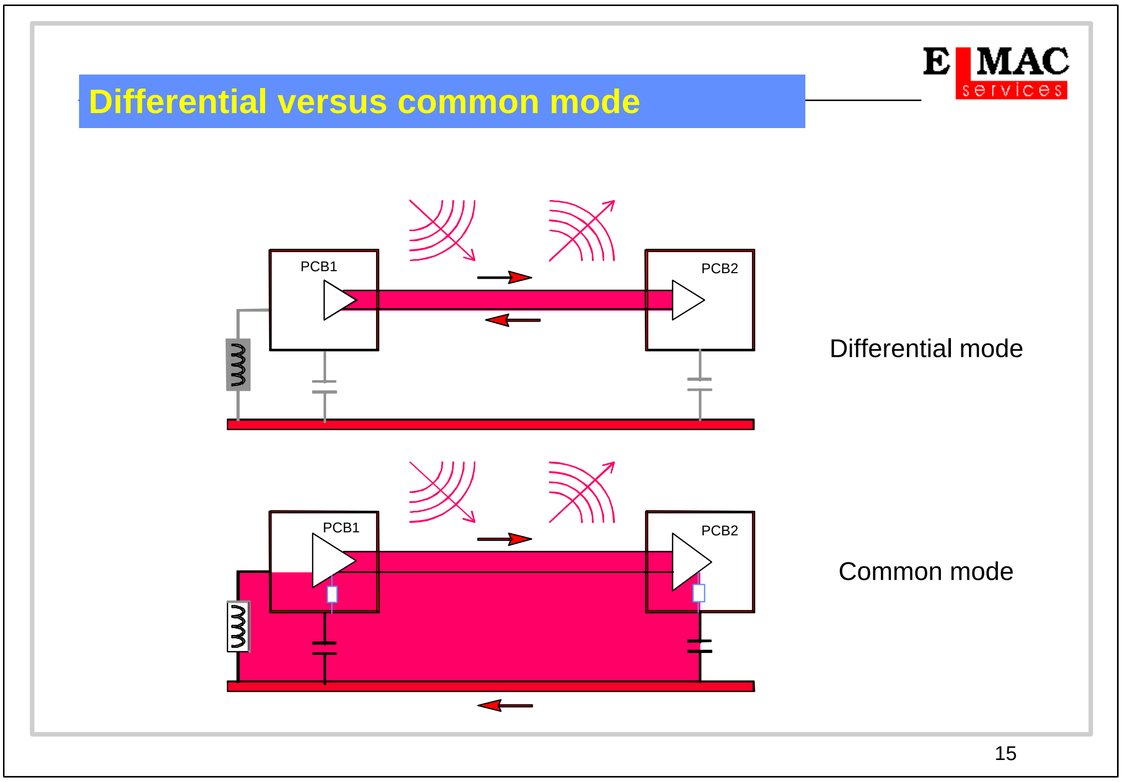


Changing voltage on one conductor creates an electric field which may couple with a nearby conductor and induce a voltage on it. The voltage induced on the victim conductor in this manner is

$$V = C_C \cdot dV_L/dt \cdot Z_{IN},$$

where C_C is the coupling capacitance and Z_{IN} is the impedance (to ground) of the victim circuit (given that the impedance of the coupling capacitance is much higher than that of the circuit impedances).

The value of C_C is a function of the distance between the conductors, their effective areas and the presence and point of connection of any electric screening material. Typically, 2 parallel wires 0.1" apart show a coupling capacitance of about 50pF per metre; the primary-to-secondary capacitance of an unscreened medium-power mains transformer is 100-1000pF. Note that stray capacitance will complete the coupling path even if the two circuits are not directly referenced to each other.



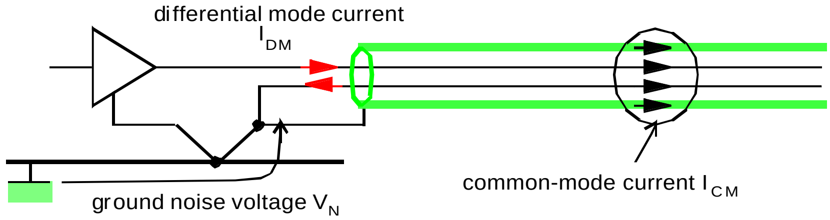
The concept of differential mode versus common mode radiated field coupling is fundamental to an understanding of EMC. Consider two PCB assemblies interconnected by a cable. The cable carries signal currents in <u>differential</u> mode (go and return) down the two wires in close proximity. A radiated field can couple to this system and induce differential mode interference between the two wires; similarly, the differential current will induce a radiated field of its own.

The cable also carries currents in <u>common</u> mode, that is, all flowing in the same direction on each wire. These currents may have *nothing at all to do with the signal currents*. They may be induced by an external field coupling between the loop formed by the cable, the ground plane and the various impedances connecting the assemblies to ground. Alternatively, they may be generated by internal noise voltages between the ground reference point and the cable connection, and be responsible for radiated emissions.

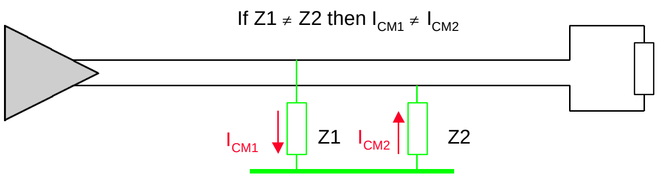
Incoming interference generates current flow principally in common mode; differential voltages are created within the circuit, and hence cause susceptibilities, when this current takes different paths through different impedances within the circuit structurs. Good circuit layout is a matter of miniizing these differentials.

The same equivalent circuit applies to a single PCB assembly. Here, each circuit track pair (signal and return) takes the place of the cable between two assemblies, and the external ground return becomes the chassis on which the PCB is mounted.

Common-mode cable noise



A common but mistaken cable screen connection couples ground noise out



Longitudinal (differential to common mode) conversion

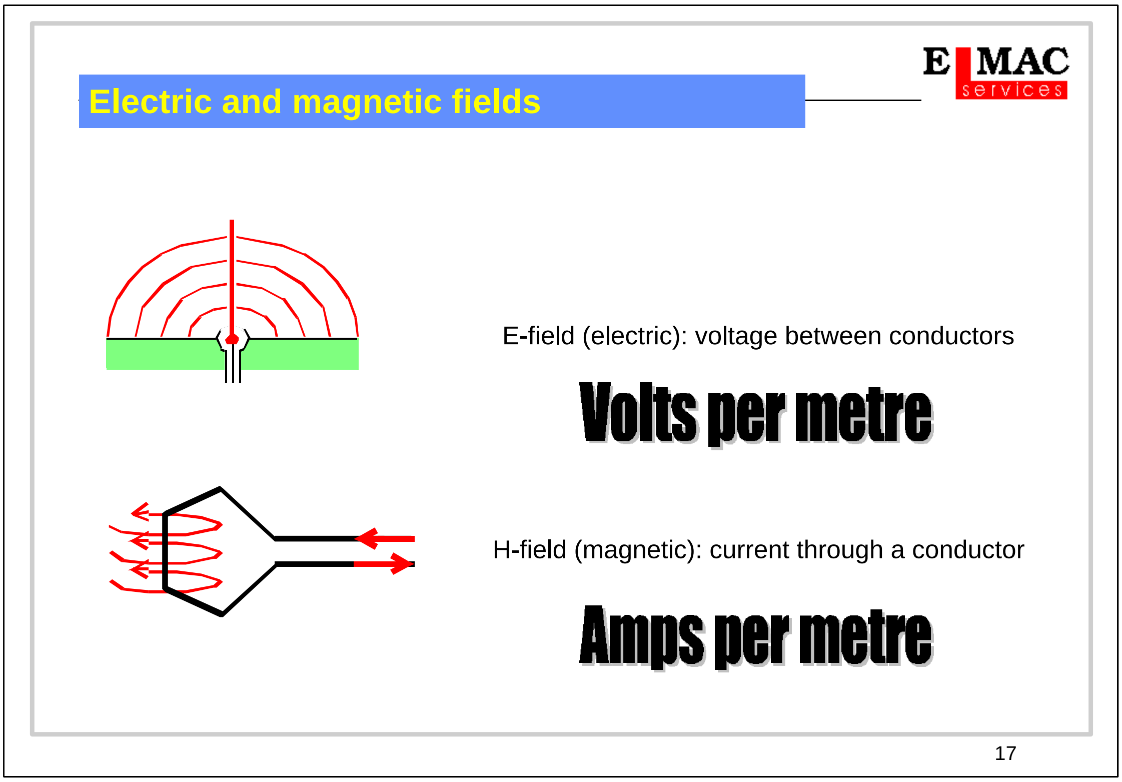
16

Because, nevertheless, cables form an important route for interference coupling, it is important to appreciate the difference between common-mode and differential-mode cable currents.

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

Common-mode current I_{CM} flows equally in the same direction along all conductors in the cable, including the shield if it is present, and may or may not be quite unrelated to the signal currents. That part of the signal current which does not return via the cable because of stray leakage paths, does appear as a common mode component. In this particular case the conversion mechanism is known as “Longitudinal Conversion Loss” (LCL). It is particularly significant for telecom ports which carry signals within the interference frequency range being measured,such as LAN and other high-speed data networks.

Common mode current returns via the associated ground network and therefore the radiating loop area is large and uncontrolled. As a result, even a small I_{CM} can result in large emitted signals.

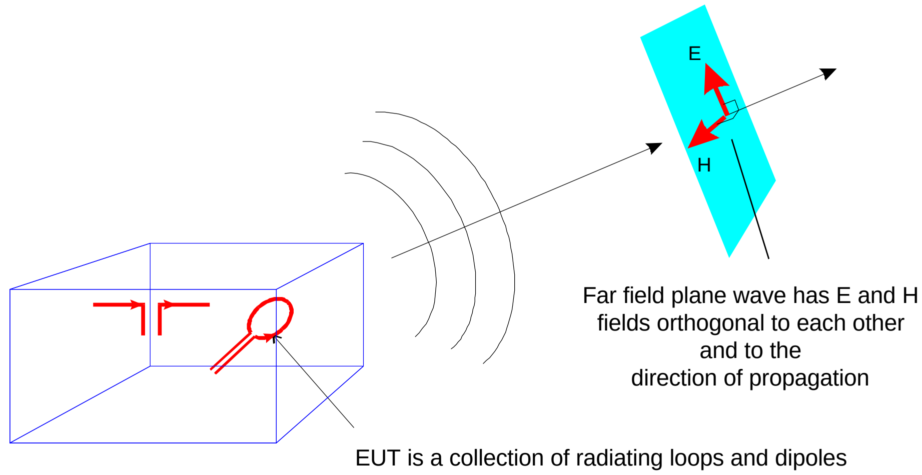


An electric field (E field) is generated between two conductors at different potentials. The field is measured in volts per metre and is proportional to the applied voltage divided by the distance between the conductors.

A magnetic field (H field) is generated around a conductor carrying a current, is measured in amps per metre and is proportional to the current divided by the distance from the conductor.

When an alternating voltage generates an alternating current through a network of conductors, an electromagnetic (EM) wave is generated which propagates as a combination of E and H fields at right angles. Near to the radiating source the geometry and strength of the fields depend on the characteristics of the source; further away only the orthogonal fields remain.

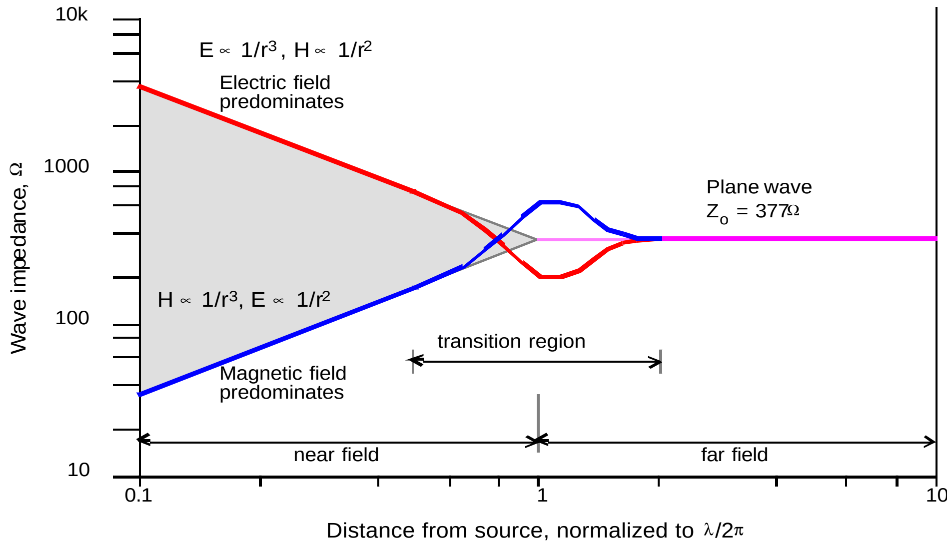
Electromagnetic field



18

When an alternating voltage generates an alternating current through a network of conductors, an electromagnetic (EM) wave is generated which propagates as a combination of E and H fields at right angles. Near to the radiating source the geometry and strength of the fields depend on the characteristics of the source; further away only the orthogonal fields remain.

The wave impedance



19

The ratio of the electric to magnetic field strengths (E/H) is called the wave impedance. In the far field, d > $\lambda/2\pi$, the wave is known as a plane wave and its impedance is equal to the impedance of free space given by

$$Z_o = (\mu_o/\varepsilon_o)^{0.5} = 120\pi = 377\Omega$$

where μ_o is $4\pi \cdot 10^{-7}$ H/m

and ε_o is $8.84 \cdot 10^{-12}$ F/m

In the near field, d < $\lambda/2\pi$, the wave impedance is determined by the characteristics of the source. A low current, high voltage radiator (such as a rod) will generate mainly an electric field of high impedance, while a high current, low voltage radiator (such as a loop) will generate mainly a magnetic field of low impedance. Measurements made in the near field will observe either E or H fields, but do not accurately represent the total interfering capability of a product.

The region around $\lambda/2\pi$, or approximately one sixth of a wavelength, is the transition region between near and far fields.

The Rayleigh criterion

Frequency	Maximum dimension D (m)	Rayleigh $d = 2D^2/\lambda$ (m)	Maxwell $d = \lambda/2\pi$ (m)
10 MHz	2	0.267	4.77
30 MHz	2	0.8	1.59
100 MHz	0.5	0.167	0.477
	2	2.67	
300 MHz	0.5	0.5	0.159
	2	8.0	
1 GHz	0.5	1.67	0.0477

20

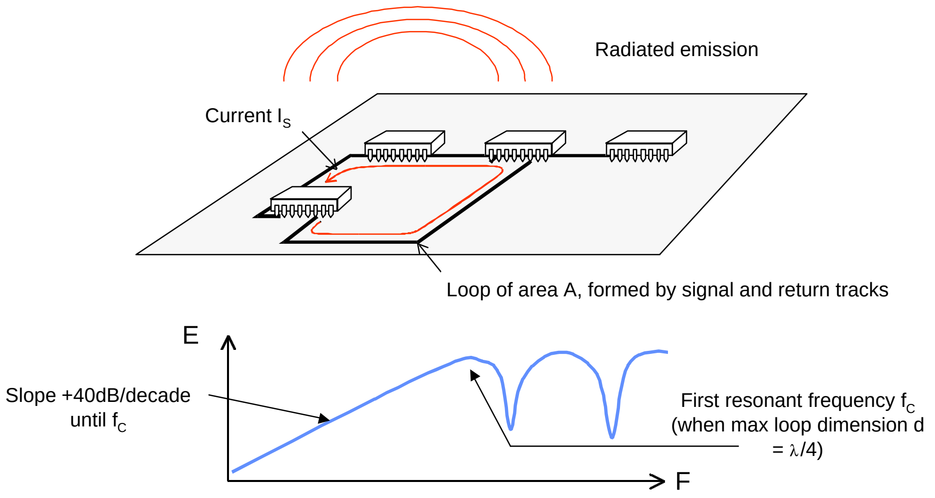
There is another definition of the transition between near and far fields, determined by the Rayleigh range. This has to do not with the field structure according to Maxwell's equations, but with the nature of the radiation pattern from any physical antenna (or equipment under test) which is too large to be a point source. For the far field assumption to hold, the phase difference between the field components radiated from the extremities of the antenna must be small and therefore the path differences to these extremities must also be small in comparison to a wavelength. This produces a criterion that relates the wavelength and the maximum dimension of the antenna (or EUT) to the distance from it. Using the Rayleigh criterion, the far field is defined as beyond a distance:

$$d > 2D^2/\lambda$$

where D is the maximum dimension of the antenna

The table shows a comparison of the distances for the two criteria for the near field/far field transition for various frequencies and EUT dimensions. Note how for typical EUT dimensions the Rayleigh range determines the far field condition above 100–200MHz.

Radiation from the PCB



21

In most equipment, the primary emission sources are currents flowing in circuits (clocks, video and data drivers, and other oscillators) that are mounted on printed circuit boards.

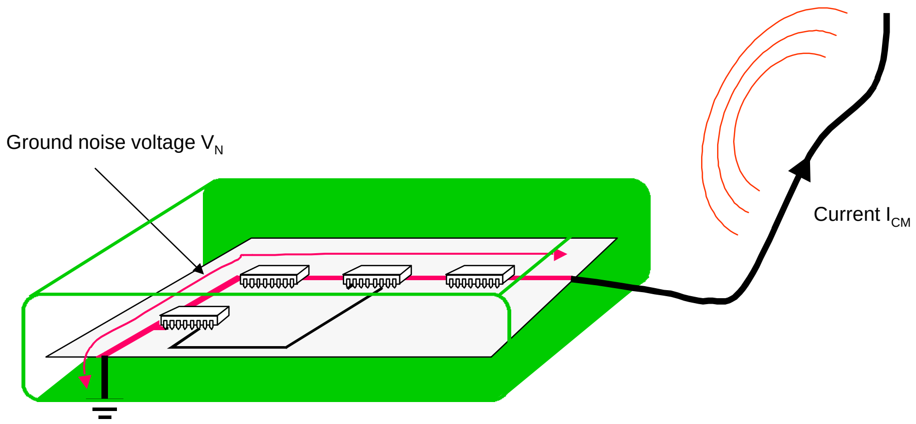
Radiated emission from a pcb can be modelled as a small loop antenna carrying the interference current. A small loop is one whose dimensions are smaller than a quarter wavelength ($\lambda/4$) at the frequency of interest (1 metre at 75MHz). The maximum electric field strength from such a loop over a ground plane at 3 metres distance is proportional to the square of the frequency (in MHz):

$$E = 87.7 \cdot 10^{-8} (f^2 \cdot A \cdot I_S) \text{ volts per metre}$$

where A is the loop area in sq. cm, and I_S is the source current in amps

If A = 10cm^2, I_S must be less than 4.5mA at 50MHz for an equivalent field strength of 40dBμV/m - which is equivalent to the European Class B limit (corrected from 10m to 3m).

Radiation from cables



22

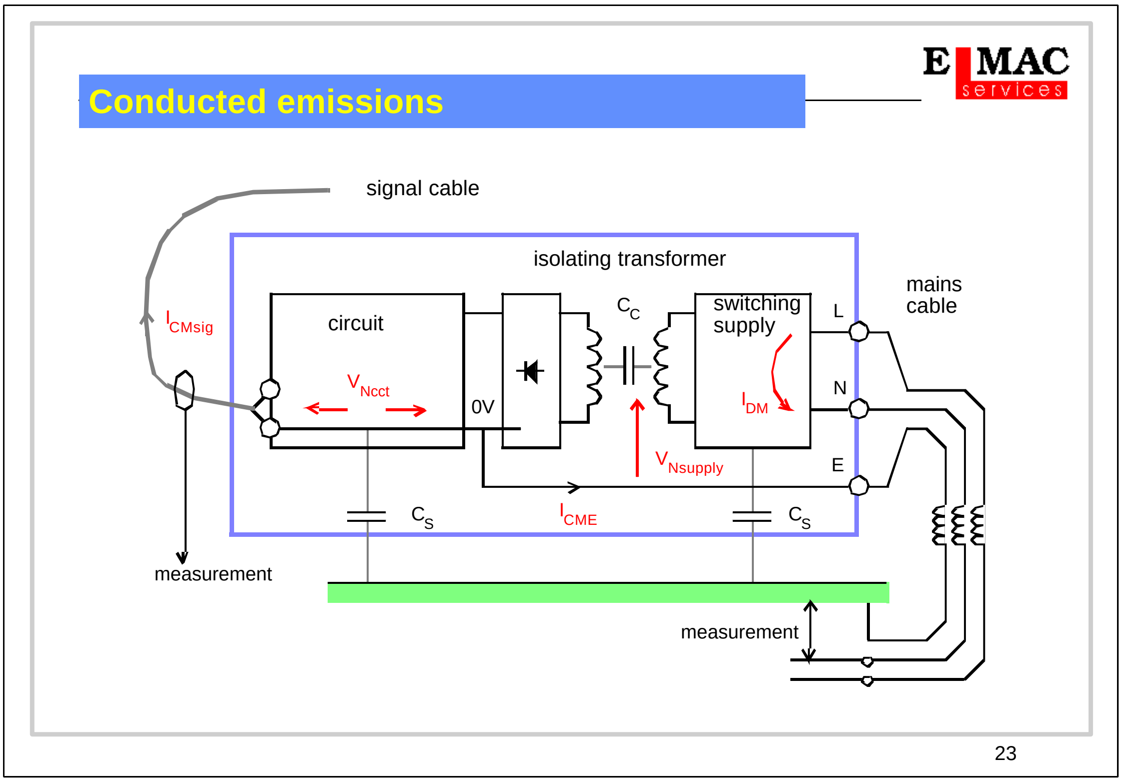
In most products, a PCB is connected to cables and housed within an enclosure. Cable emission is mainly due to common-mode interference current flowing in the cable. The interference current is generated from ground noise developed across the pcb or elsewhere in the equipment and referred to the enclosure, and may flow along the conductors, or along the shield of a shielded cable.

The model for cable radiation at lower frequencies is a short ($L < \lambda/4$) monopole antenna over a ground plane. (When the cable length is resonant the model becomes invalid.) The maximum field strength at 3m over a ground plane due to this radiation is directly proportional to frequency:

$$E = 0.42 \cdot (f \cdot L \cdot I_{CM}) \text{ volts per metre}$$

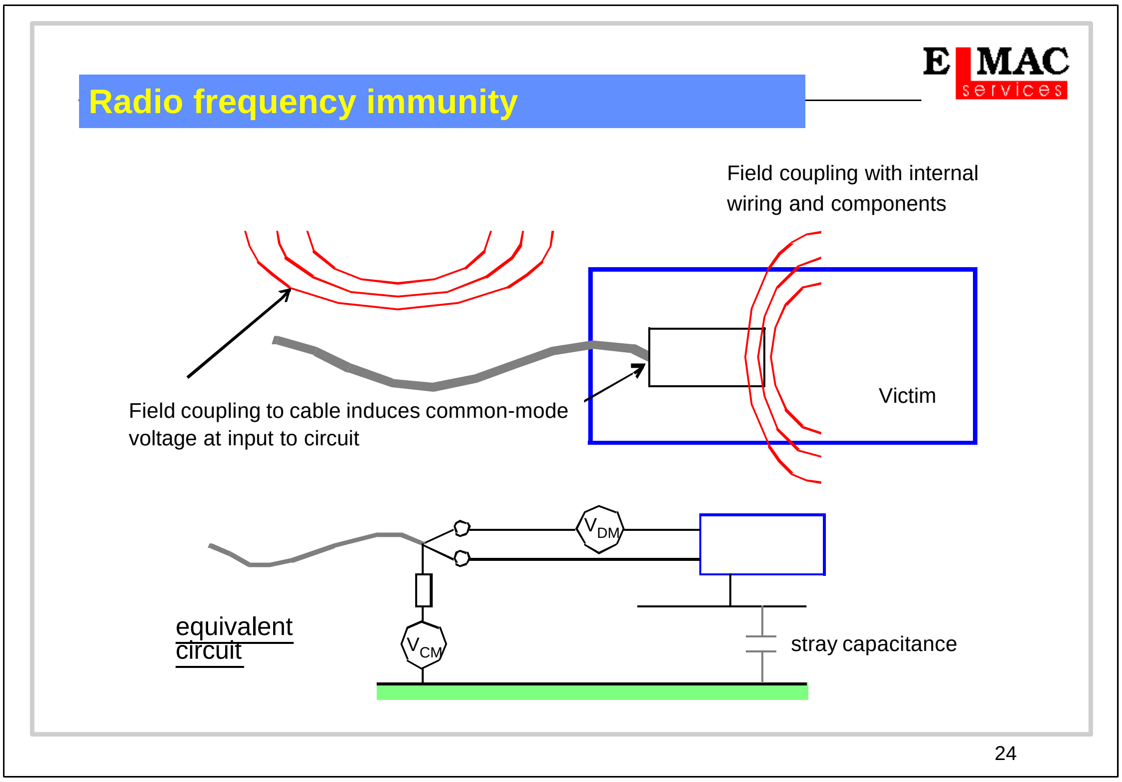
where L is the cable length in m and I_{CM} is the common-mode current in amps at f MHz flowing in the cable

For a 1m cable, I_{CM} must be less than 4.8µA at 50MHz for an equivalent field strength of 40dBµV/m - very much lower than the equivalent I_{DM}.



Interference sources within the equipment circuit or its power supply are coupled onto the power cable to the equipment. Interference may also be coupled either inductively or capacitively from another cable onto the power cable. The resulting interference voltage may appear as differential mode (V_{DM} between live and neutral) or as common mode (V_{CM} between live/neutral and earth) or both. Signal cables will themselves also carry noise.

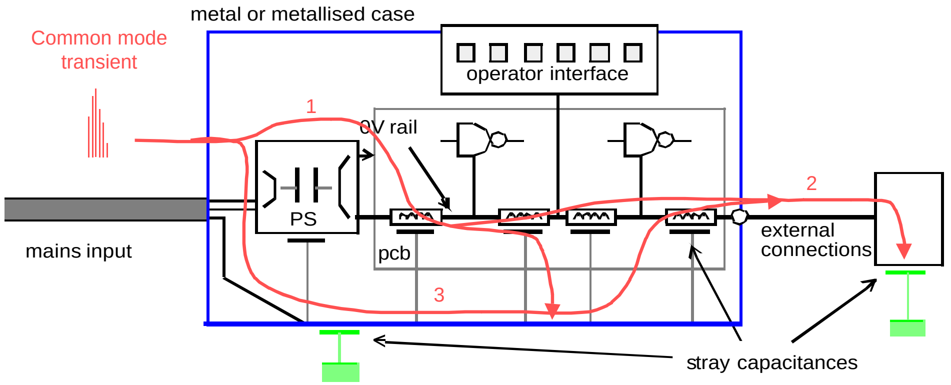
Differential mode current I_{DM} generated at the input of the switching supply is converted by imbalances in stray capacitance, and by the mutual inductance of the conductors in the mains cable, into interference voltages with respect to earth at the measurement point. Higher frequency switching noise components $V_{Nsupply}$ are coupled through C_C to appear between L/N and E on the mains cable, and C_S to appear with respect to the ground plane. Circuit ground noise V_{Ncct} (digital noise and clock harmonics) is referenced to ground by C_S and coupled out via signal cables as I_{CMsig} or via the safety earth as I_{CM}E.



An external field can couple either directly with the internal circuitry and wiring in differential mode or with the cables to induce a common-mode current. Coupling with internal wiring and PCB tracks is most efficient at frequencies above a few hundred MHz, since wiring lengths of a few inches approach resonance at these frequencies and the internal dimensions of partially-screened cases may also form a resonant cavity.

At frequencies below 200-400MHz coupling with the connected cables is more efficient, and this will induce a common mode current in each exposed cable which can be modelled as a common mode voltage appearing at the interface with the enclosure. This in turn will cause common mode currents to flow internally, and differences in internal circuit impedances will convert this to a differential mode interference signal to which the circuit will respond. Typically a radiated field of 1V/m will induce a current of 1–3mA in the cables, and the resultant interface voltage will depend on the common mode impedance Z_{CM} at the interface. A well designed interface will have a low Z_{CM}.

Transient immunity



25

Transient interference is generated predominantly by lightning, supply network switching and fault clearing events, and local load switching events. The resulting voltage transients propagate either by conduction along the power cables, in either differential or common mode, or are radiated from these cables or the source and induce transient voltages in other nearby cables, which may carry signal or control circuits. In the latter case the incoming interference is almost exclusively in common mode.

The characteristics of the transients themselves vary greatly from one to the next and have to be treated statistically, but some broad groupings of characteristics emerge. These can be classified as high energy surges due to lightning and fault clearing, fast but low energy bursts due to local load switching, and ringing or damped oscillatory waves due to high power switching operations in electrical supply networks which are modified by the resonant properties of the transmission lines involved.

An incoming common-mode transient on the mains can travel through the circuit's 0V rail, generating ground differential spikes as it goes, through any or all of several paths:

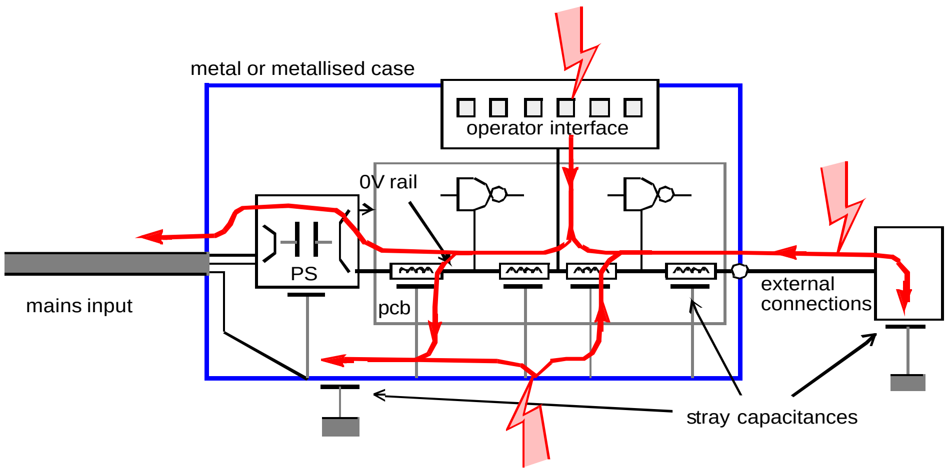
•stray capacitance through the power supply to 0V, through the equipment and then to case

•as above, but then out via an external connection

•direct to case, then via stray capacitance to 0V and out via an external connection

If there are no external connections, (1) is the only problem and can be cured by a mains filter and/or by an electrostatic screen in the mains transformer. (2) arises because the external connection can provide an easier route to ground than case capacitance. With external connections, (3) can actually be caused by a mains filter, since at RF parts of the case can float with respect to true ground.

ESD immunity



26

When an electrostatically charged object is brought close to a grounded target the resultant discharge current consists of a very fast (sub-nanosecond) edge followed by a comparatively slow bulk discharge curve. The human body can be charged to several kV and its equivalent source circuit can be approximated by a 150pF capacitor in series with a 330Ω resistor.

The resultant sub-nanosecond transient of several tens of amps follows a complex route to ground through the equipment and is very likely to upset digital circuit operation if it passes through the circuit tracks. The paths are defined more by stray capacitance and track or wiring inductance than by the designer's intended circuit. The high magnetic and electric fields associated with the very high rate of change of current and voltage can induce transient voltages in nearby conductors that are not actually in the path of the current. Even if not discharged directly to the equipment, a nearby discharge can induce sufficient upset current in it.

An electrostatic discharge can occur to any exposed part of the equipment. Common trouble spots are keyboards and controls (1), external cables (2) and accessible metalwork (3). A discharge to a nearby conductive object can also induce currents within the equipment. Because there are many potential points of discharge, the possible routes to ground that the discharge current can take are widespread. Many of them will include part of the pcb ground layout, via stray capacitance, external equipment or exposed circuitry, and the induced transient ground differentials will cause maloperation.

The discharge current will take the route of least inductance. If the case is well bonded to ground then this will be the natural sink point. If it is not, or if it is non-conductive, then the routes of least inductance will be via the connecting cables. The discharge edge has an extremely fast risetime (sub-nanosecond) and so stray capacitive coupling is essentially transparent to it, whilst even short ground connectors of a few nH will present a high impedance.

End of this section

27