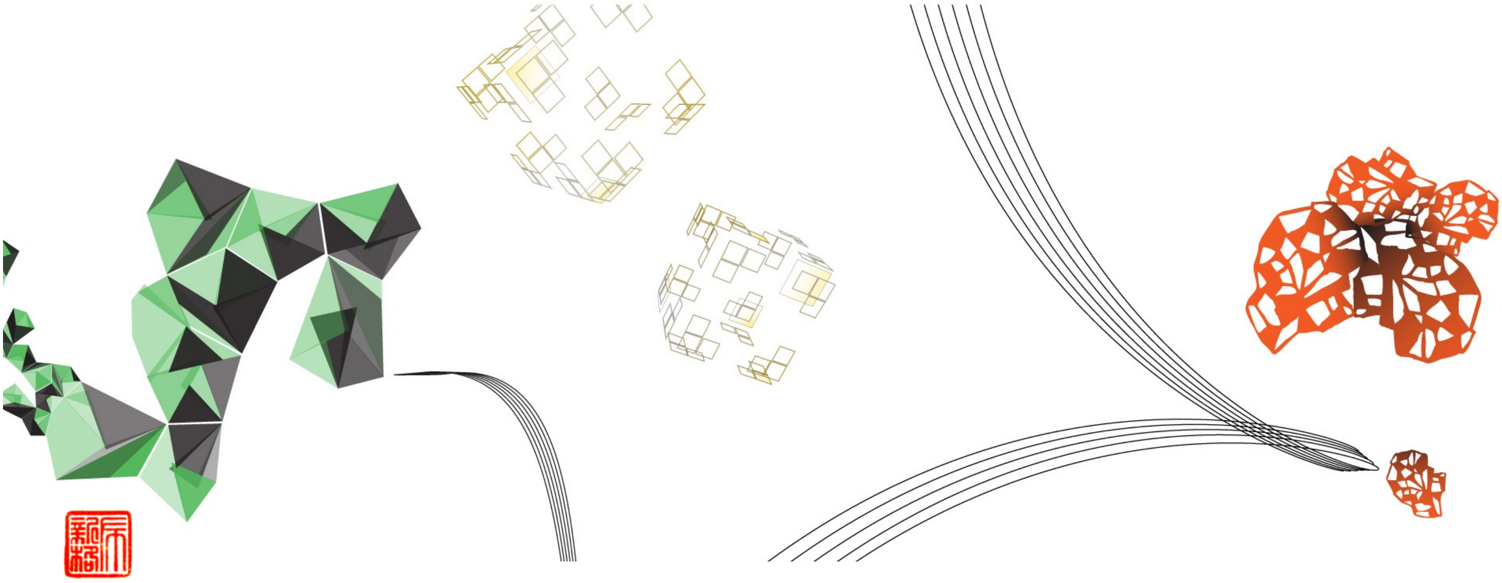

EMC for Engineers

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2023-11-29

EMC for Engineers

This presentation on experiments to educate for electromagnetic compatibility (EMC) awareness addresses engineers in an organization involved with electronics systems who are not primarily involved with electromagnetic (EM)–engineering.

Introduction

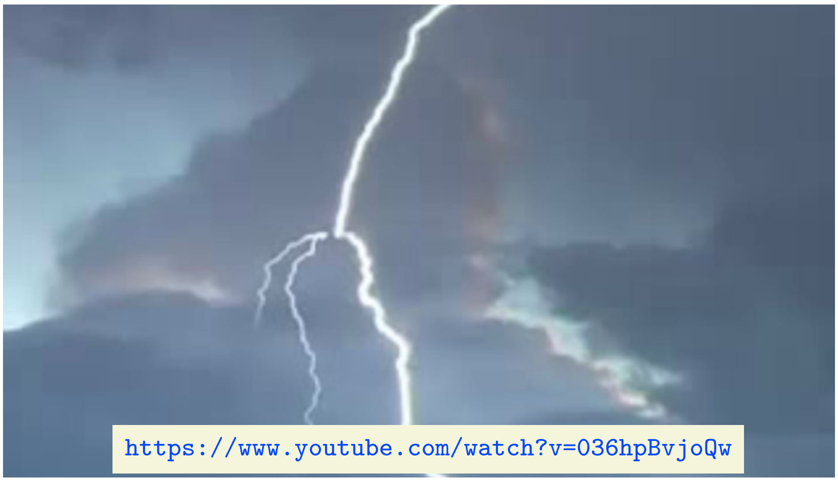
A more detailed investigation of EMC
Introduction

This slide has no notes.

Definition of EMC: emission and susceptibility

Equipment can both disturb and be disturbed

Immunity: "The ability of the system to operate according to its specifications in its intended electromagnetic environment"



https://www.youtube.com/watch?v=036hpBvjoQw

Emission: "Without generating unacceptable electromagnetic disturbances into that environment"

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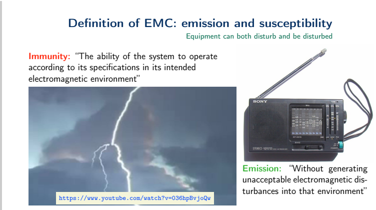

Emission and Immunity are environment related. There are EMC standards for different environments.
When addressing EMC the first step is to find out what the requirements are for the product under development.

In this tutorial, the US military standards will be referred to occasionally since the target audience is involved with military equipment.
A second reason is that they are available for free on the internet[1].

[1] See e.g. http://everyspec.com/MIL-STD/MIL-STD-0300-0499/MIL-STD-464C_28312/

Systems perspective: performance criteria

depending on criticality of failure, a criterion is selected (often included in the standard)

A System continues to work according to specification
Degradation not acceptable
Generally applies to all interference with a continuous nature

B Temporary degradation acceptable, auto recovery.
Usually applies to sporadic interference
to a non-critical function.

C Degradation acceptable. Recovery after manual RESET.
e.g. at mains interruptions. Only for non-critical functions.

An <u>UNSAFE</u> situation is never acceptable!

Systems perspective: performance criteria
depending on criticality of failure, a criterion is selected (often included in the standard)
A System continues to work according to specification Degradation not acceptable Generally applies to all interference with a continuous nature
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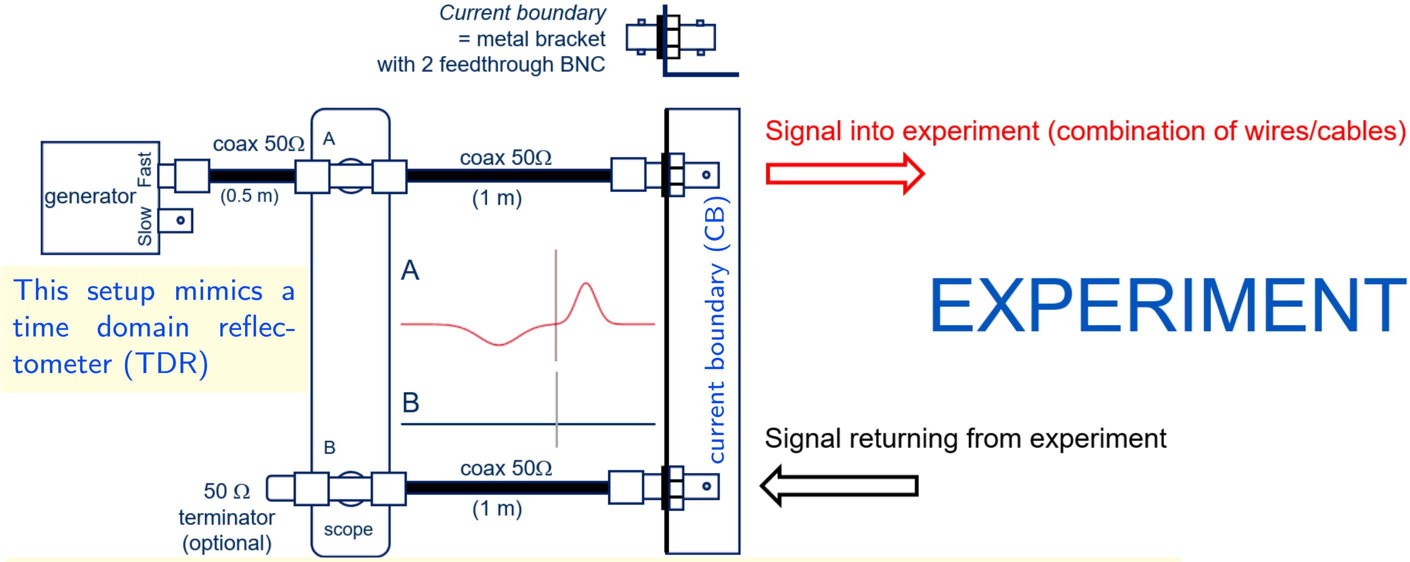
An UNSAFE situation is never acceptable!

Viewed from the systems perspective, the immunity aspect can be translated into performance criteria. In the example of e.g. a plane struck by lightning, the choice is definitely 'A'. This is the only performance that is safe for the passengers and crew of the plane.
In less critical situations criterion 'B' or even'C' can be acceptable. B could be interpreted as lines of interference in the old fashioned television receivers when a vehicle with noisy ignition signals would pass. After the vehicle passed, the interference would stop by itself. No user actions necessary.

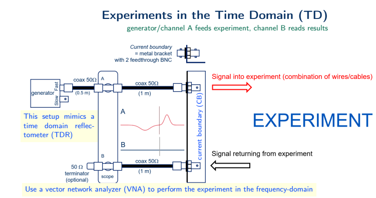
An example of performance criterion 'C' would be a power failure of a desktop computer. The computer would stop working and user intervention is required to get it to work again. After that intervention the computer should work as specified again.
An aspect that is especially stressed under the European law (CE marking) is that a system shall never pose a threat to people, animals or infrastructure under any of the criteria (Aspect 49). So even under the (not shown) criterion 'D': failure, the machine will never work again but it is not allowed to explode or catch fire.

Experiments in the Time Domain (TD)

generator/channel A feeds experiment, channel B reads results



Use a vector network analyzer (VNA) to perform the experiment in the frequency-domain



All experiments used the setup shown in this slide: a generator connected to the A-channel of an oscilloscope and feeding a feed-through connector on a current boundary (CB).
The CB has a second feed-through connector which is connected to the B-channel of the oscilloscope. There it is usually terminated in a 50Ω resistor.
On the right hand side of the CB the current experiment is connected.
Channel-A shows whatever signal in the time-domain goes into that experiment.
Channel-B shows whatever comes back out of the experiment.

This is actually the circuitry found in a time domain reflectometer (TDR).
See e.g. https://en.wikipedia.org/wiki/Time-domain_reflectometer
This is a more natural presentation of electrical phenomena for people that are less familiar with the option to show the frequency domain using a vector network analyzer (VNA).
See e.g. https://en.wikipedia.org/wiki/Network_analyzer_(electrical)
But if desired, a VNA can replace the circuitry to the left of the CB.
All experiments can be performed in the time- or frequency-domain (Aspect 50).

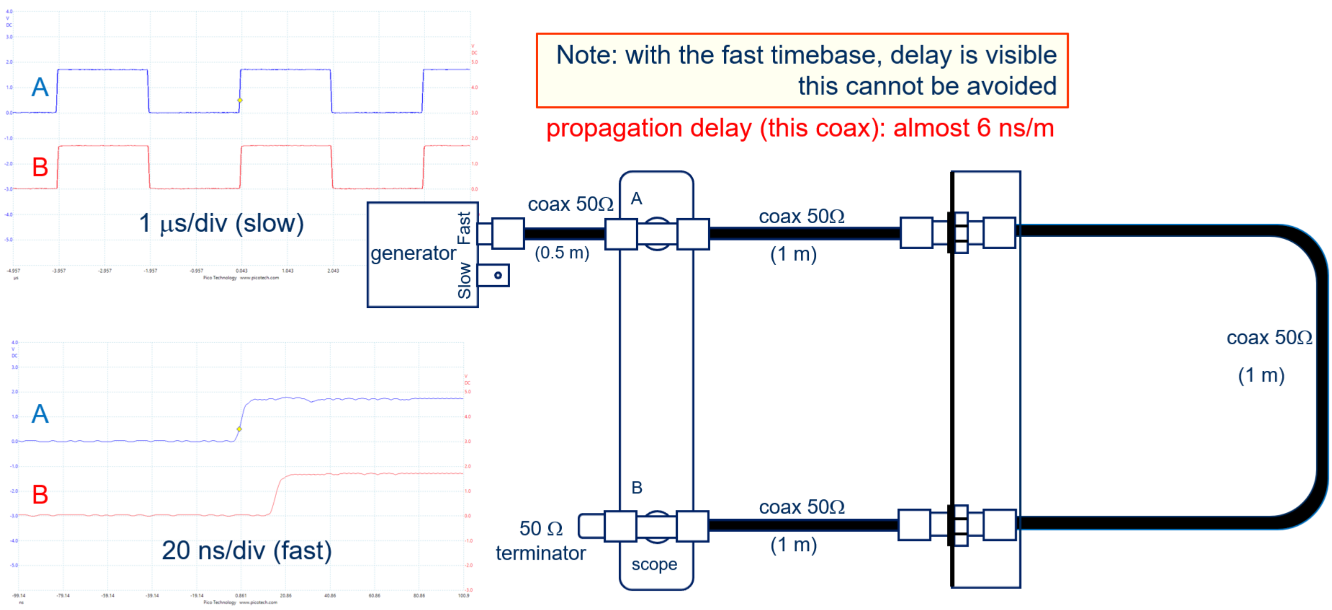
Part 1: Signal Integrity

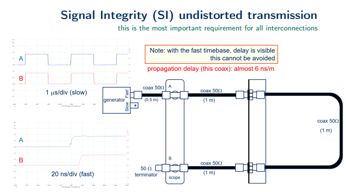
Focus at the quality of transmitted signals
Part 1: Signal Integrity

An important aspect of EMC is the flawless operation of the system itself.
As EMC is all about interconnections, we look at what happens to the signals that are transmitted over them.
The system benefits from undisturbed signals.
That situation is labeled signal integrity.
If a signal arrives *in excellent shape* this implies nothing is lost underway that could disturb others.
Signal Integrity helps the achievement of EMC

Signal Integrity (SI) undistorted transmission

this is the most important requirement for all interconnections





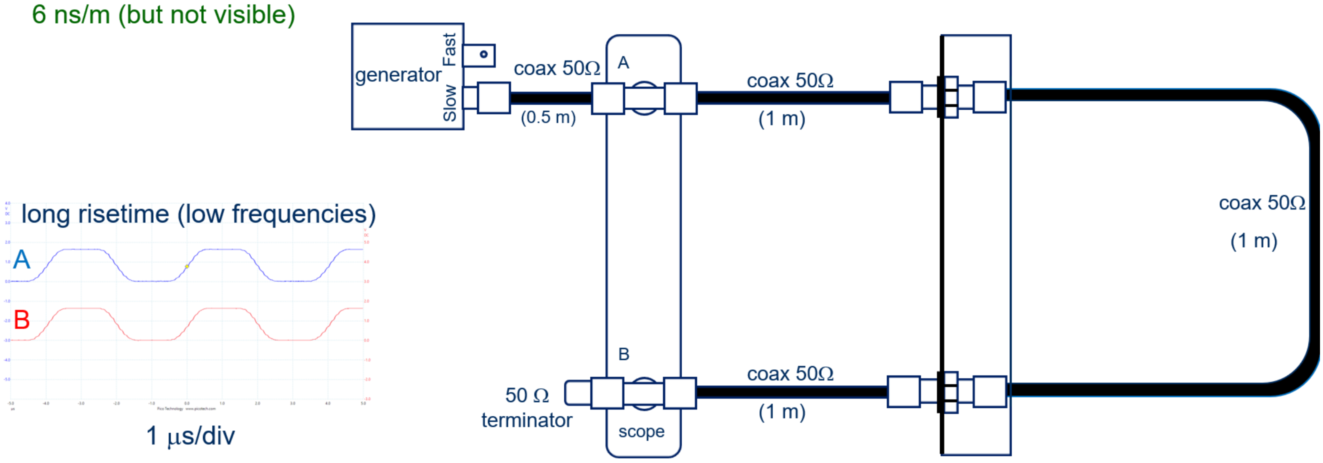
Signal integrity means transmitted signals over an interconnection are received intact, without any distortion.
The experiment shown in this slide uses a fast risetime square wave which is sent from one side to the other over a 3 m coaxial interconnection.
The arriving signal at channel-B of the oscilloscope looks exactly like the signal sent out at channel-A.
The only difference is the arrival time of the transitions.
They arrive about 18 ns later at channel-B.
This effect is called: propagation delay, τ_{PD}, dimension s/m (Aspect 51).

Signal Integrity (SI) undistorted transmission

this is the most important requirement for all interconnections

Note: the slow output has a very long risetime (1μs)

propagation delay same 6 ns/m (but not visible)



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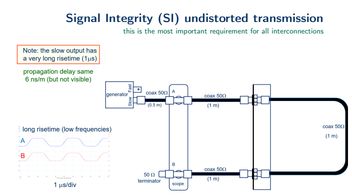
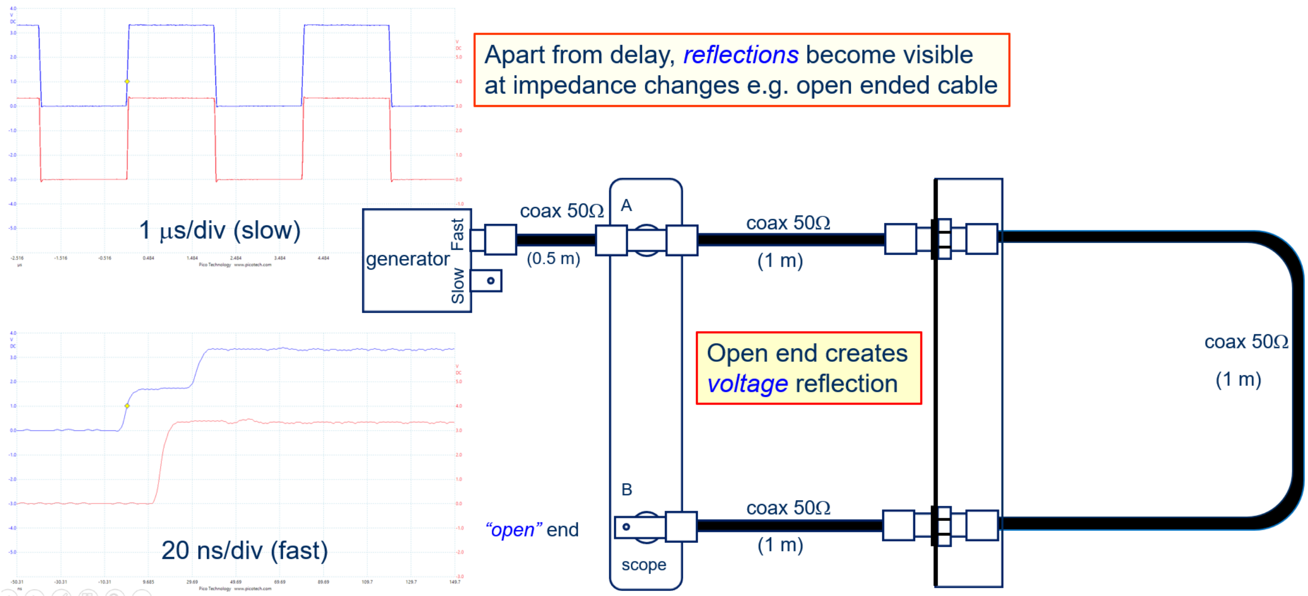


If we use a slow rise time signal, the propagation delay is the same but no longer visible on the oscilloscope.

Signal integrity: distorted signal transmission

unterminated (open) transmission line distorts by generating reflections



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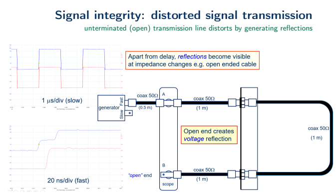


Returning to the fast risetime signal, we do the same experiment after removal of the 50Ω load resistor at channel-B. When viewed at 20 ns/div, the signal at channel-B looks O.K. but has twice the amplitude compared to the signal sent out at channel-A. Another 18 ns after the transition reached channel-B, the signal at channel-A also rises to the double amplitude seen at channel-B.

The phenomenon seen is called reflection. The signal transition at channel-A is a combination of voltage and current departing in the direction of channel-B. The ratio of this voltage to this current is $\frac{V}{I} = 50\Omega$, the, so called, characteristic impedance of the coax transmission-line. When arriving at channel-B, the current can no longer flow as the load has been removed. What happens to the energy in this signal flow? It bounces back into the direction of channel-A. This bounced signal transition has the same amplitude as the original but the opposite current. At channel-B we see the double amplitude voltage immediately.

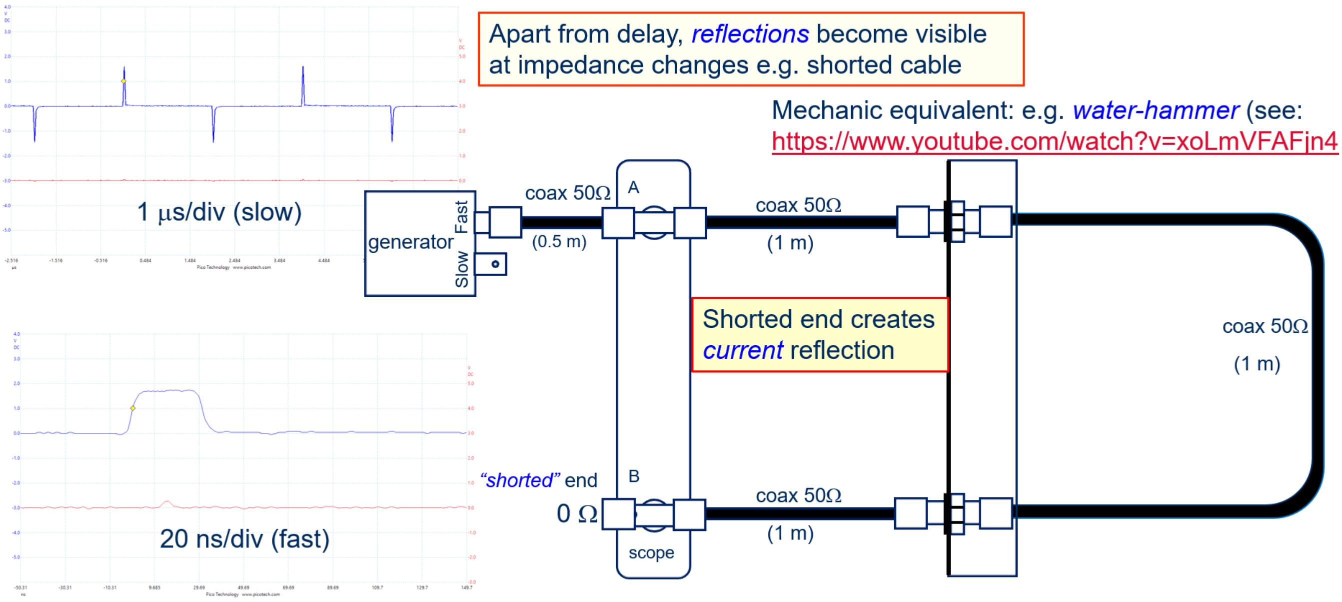
18 ns later, the reflected signal arrives back at channel-A. This is visible as a rise of the voltage to twice its original value (Aspect 52).

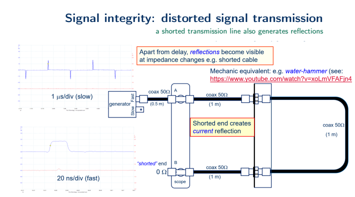
The reflected signal continues on towards the generator.

But as the generator has 50Ω impedance, there are no more reflections.

Signal integrity: distorted signal transmission

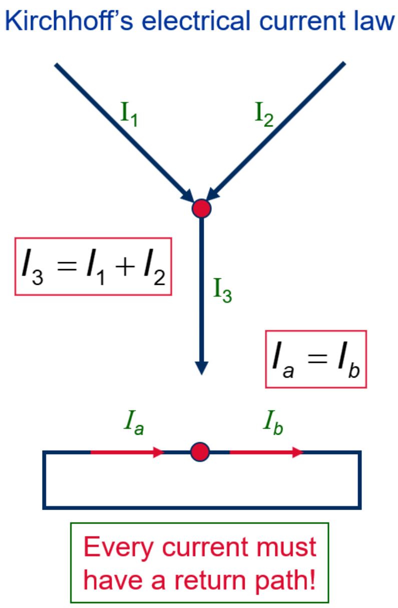
a shorted transmission line also generates reflections



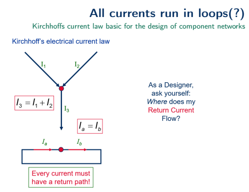


An open ended transmission-line reflects the incoming current.
In the same way, a short-circuited transmission-line reflects the incoming voltage.

At channel-B we see 0 V.
At channel-A we see the departing transition as before but this time, after a delay of $2\tau_{PD}$, the signal voltage returns to zero (Aspect 53).



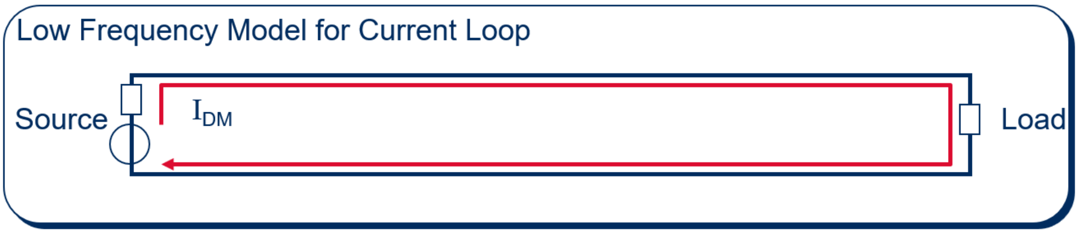
As a Designer, ask yourself: *Where* does my Return Current Flow?

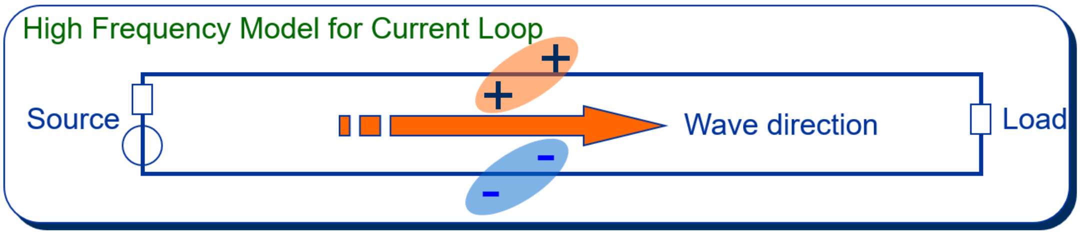


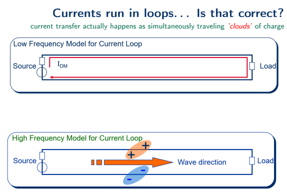
This is all very well, but what about our initial observation that currents run in loops?
The loop story, i.e. the idea that electrons leave the source to flow to the load and then return to the source via the return conductor can never explain the observed reflection effects.

Currents run in loops. . . Is that correct?

current transfer actually happens as simultaneously traveling *'clouds'* of charge







So, we need a different metaphor for our statement that *currents run in loops*, depicted in the top picture on the slide.
What appears to happen is depicted in the bottom picture: a signal moves over the –transmission– line as two “clouds” of opposite charges moving from source to load together as a wave, in the same direction (Aspect 54).

Part 2: Measuring electric fields

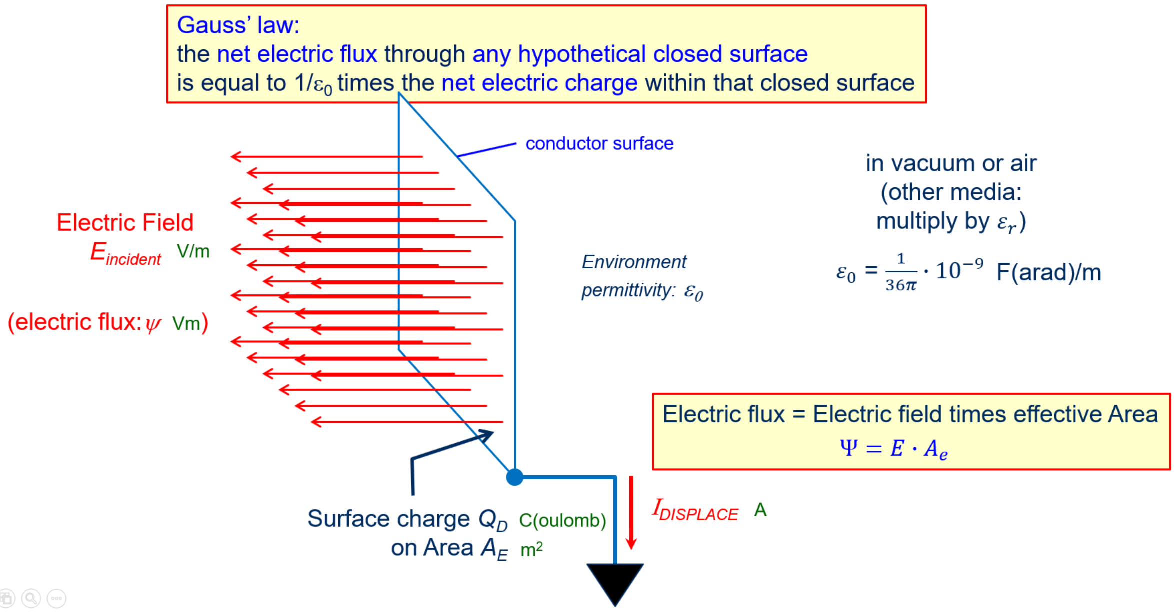
Quantify EM-phenomena

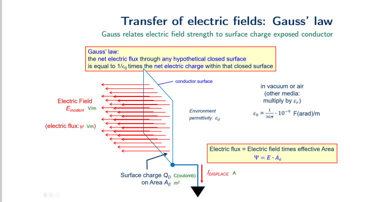
Part 2: Measuring electric fields

In order to get a better handle on EM-phenomena, we need to be able to quantify, measure them.
First we focus on electric fields.

Transfer of electric fields: Gauss' law

Gauss relates electric field strength to surface charge exposed conductor

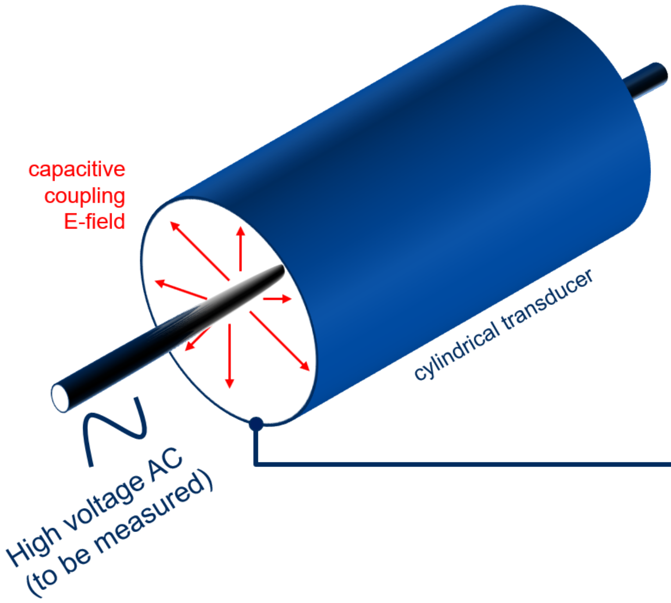




We will now look at how cross-talk works. Starting with capacitive cross-talk i.e. the exchange of energy between conductors based on electric fields.
Electric fields exist between two conductors with a voltage difference. Hence the dimension: volts per meter (V/m).
In the cross-talk situation, the *victim* conductor is exposed to the electric field of the source. The effect is that a surface charge appears on the exposed surface corresponding to the existing electric field level. Gauss' law describes the amount of charge, Q_D, based on the strength of this electric field, E, the area exposed, A_e, and the permittivity or dielectric constant of the medium between the conductors, ε_0, in free space, to be multiplied by ε_r, the relative permittivity with respect to air or vacuum if in another medium.
If the victim conductor is insulated, charge is pulled towards or pushed away from the exposed surface. This is called charge-*displacement*.
If the victim conductor is connected to the outside world, the displaced charge will give rise to a *displacement current* into or from the outside world. (Aspect 55)
The displacement current can be used to measure (alternating current (AC)) electric fields.

Changing electric field induces conductor *current*

mutual capacitance gives rise to substitute current source



$Q = A_e \cdot \varepsilon_o \cdot E$ (Gauss' law AC or DC)

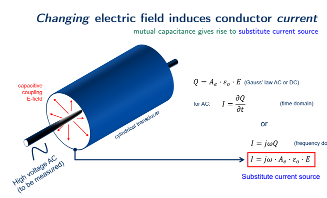
for AC: $I = \dfrac{\partial Q}{\partial t}$ (time domain)

or

$I = j\omega Q$ (frequency domain)

$\boxed{I = j\omega \cdot A_e \cdot \varepsilon_o \cdot E}$

Substitute current source

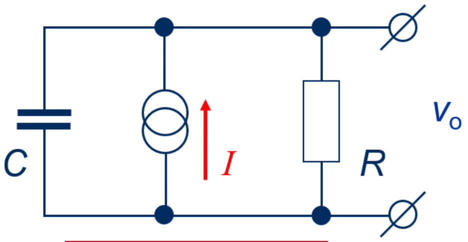


This surface charge occurs even if the electric field, E, is static –the direct current (DC) situation–: $Q = A_e \cdot \varepsilon_0 \cdot E$, possibly multiplied by ε_r for a medium other than air.
For a changing electric field –the AC situation– the induced surface charge is rising and falling over time, t, with the value of the electric field, $E(t)$.
This implies a current, $I(t) = \frac{\partial Q}{\partial t}$. In the frequency domain this can be written as $I(j\omega) = j\omega \cdot Q$.
Here, ω, is the angular frequency in radians per second.
It relates to the frequency $f = \dfrac{\omega}{2\pi}$ in Hertz.
For an insulated victim conductor, this implies that the charges moves back and forth between the surface facing the field-source and the opposite side.
But usually this victim conductor is connected to e.g. a measuring instrument in order to measure the AC-voltage on the source conductor.
As such, a plate or rod of metal can be used as a transducer to measure electric fields or even voltages on a wire that cannot be touched e.g. for safety reasons, as shown in the slide.
This transducer generates a current based on the changing electric field: $I(j\omega) = j\omega \cdot A_e \cdot \varepsilon_0 \cdot E(j\omega)$. Gauss' law hence leads to what is called a *substitute* current source (Aspect 56).

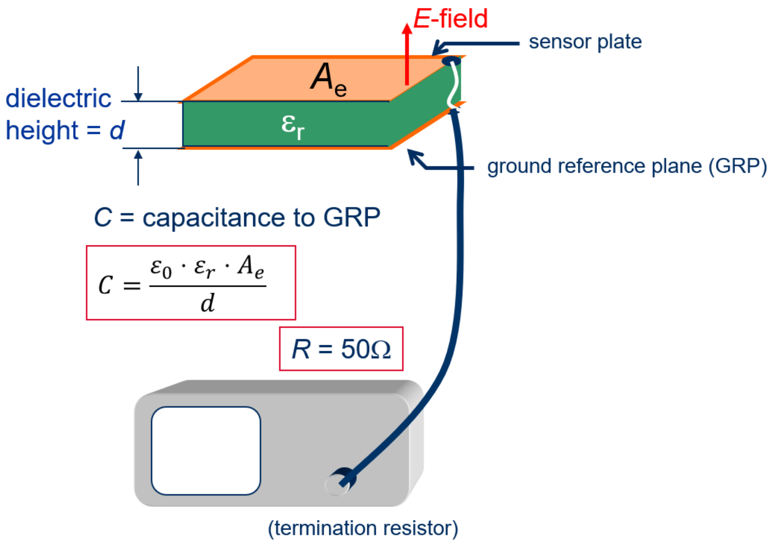
Use Gauss’ law to measure electric fields

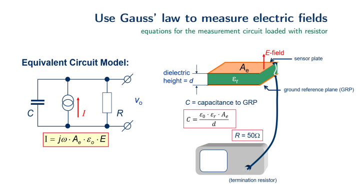
equations for the measurement circuit loaded with resistor

Equivalent Circuit Model:



$$\boxed{\mathrm{I} = j\omega \cdot A_e \cdot \varepsilon_o \cdot E}$$



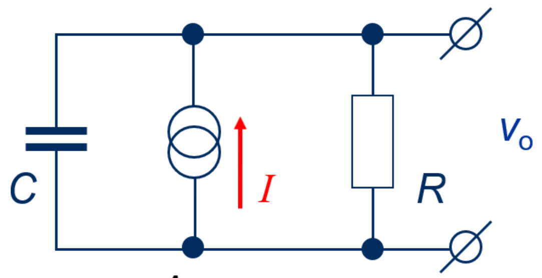


Intentional electric field transducers usually have a second metal plate forming a capacitor with the measurement plate electrode. This second or ground reference plane (GRP) is often shaped as a cup, shielding the measurement plate from other sources in the environment.
Why? For the simple reason that two wires are required to measure the generated substitute current source. Remember: currents run in loops, slide 12!
This is a coax transmission-line that connects the transducer to the measurement instrument.
The instrument terminates this transmission-line with a load resistor equal to the characteristic impedance, Z_0, of the line to avoid reflections.
The equivalent circuit to measure electric fields hence has three components:

- The substitute current source, I
- A parallel capacitor, C
- A parallel resistor, R

Use Gauss' law to measure electric fields

evaluate behavior of transducer when loaded with a resistor



$$C = \frac{\varepsilon_0 \cdot \varepsilon_r \cdot A_e}{d}$$

$$v_o = I \cdot \frac{R}{j\omega RC + 1}$$

$$I = j\omega \cdot A_e \cdot \varepsilon_o \cdot E$$

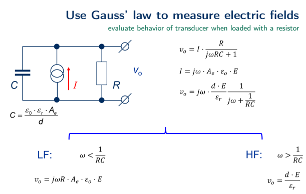
$$v_o = j\omega \cdot \frac{d \cdot E}{\varepsilon_r} \frac{1}{j\omega + \frac{1}{RC}}$$

LF: $\omega < \frac{1}{RC}$

$$v_o = j\omega R \cdot A_e \cdot \varepsilon_o \cdot E$$

HF: $\omega > \frac{1}{RC}$

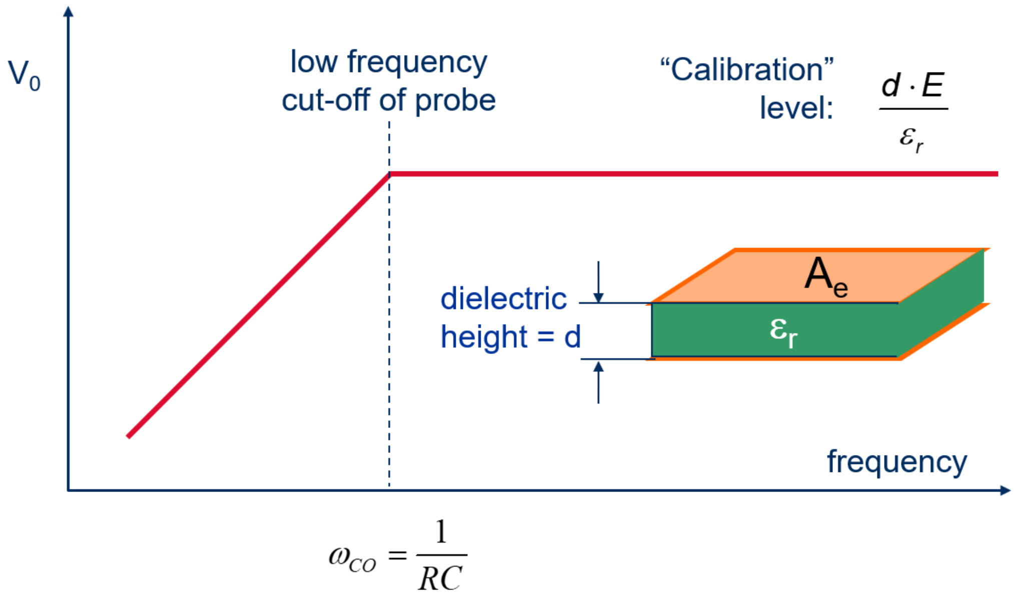
$$v_o = \frac{d \cdot E}{\varepsilon_r}$$

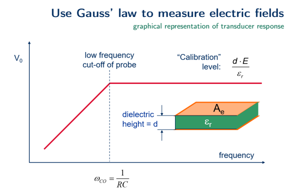


We are interested in the relation of the voltage that appears across the measurement instrument, V_0, and the electric field to which the sensor is exposed.
The slide shows the network math involved.
Two frequency regions can be recognized:
1. A low frequency region for frequencies below $\omega = \frac{1}{RC}$.
Here V_0 increases proportional to frequency;
2. A high frequency region above that corner frequency.
From this point on, V_0 is constant over frequency.
This capacitive probe-model assumes the probe's diameter is much smaller than a quarter wavelength of the highest frequency to be measured. This is called the low-frequency approximation.
More detail can be found starting on slide 39.

Use Gauss' law to measure electric fields

graphical representation of transducer response

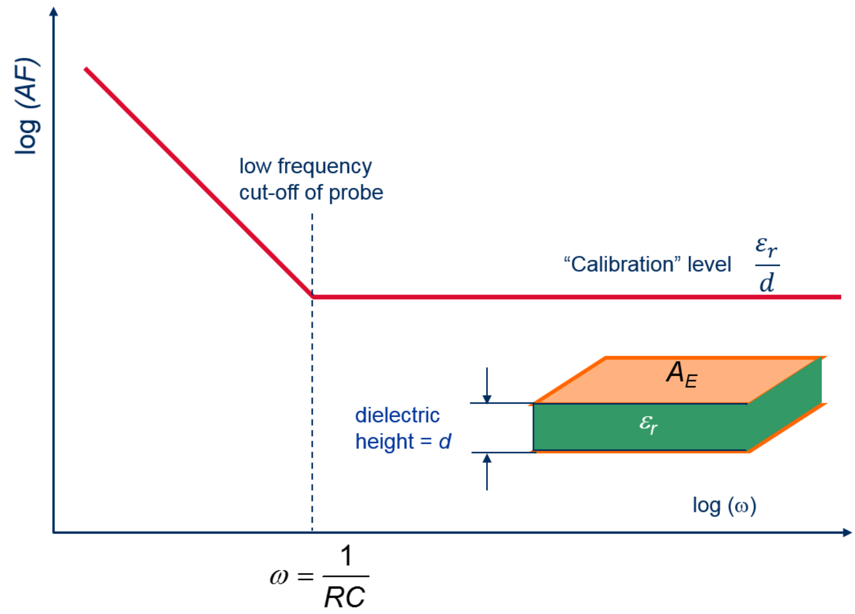


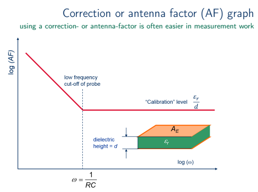


The graph, or "Bode-plot", in this slide shows, what is called, the *transfer-function* of our electric field "probe".
The corner frequency is often labeled as the cut-off frequency, ω_{CO}, of the transducer:
$\omega_{CO} = \frac{1}{RC}$.

Cut-off does not mean the transducer cannot be used. It just implies that values below this frequencies have to be corrected (Aspect 57).
Note that the frequency and V_0 axes are both logarithmic.
In the V_0 case here it is linear as the voltages are often expressed in dBμV

using a correction- or antenna-factor is often easier in measurement work



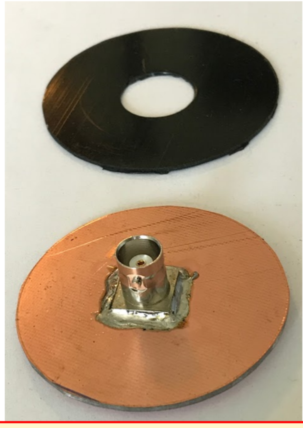
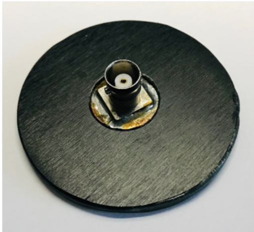


The correction- or antenna-factor is a graph of the ratio $AF = \frac{E}{V_0}$ m^{-1}.
The electric field value, E can then be found by multiplying the measured output voltage, V_0, with the AF yielding a dimension $\frac{V}{m}$.

Note: In decibell calculus, both the voltage measured, e.g. as dBμV, and the logarithm of the antenna factor AF should now be added instead of multiplied!
The treatment of decibel calculus is beyond the scope of this lecture.

Use Gauss’ law to measure electric fields

example capacitive ’D-dot’ probe

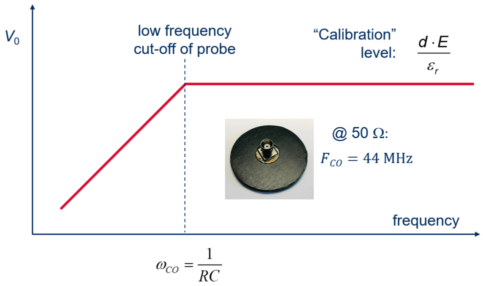
Finished model “D-dot” probe

Note

$D = \varepsilon_0 \cdot E$ (dielectric displacement)

$I = \frac{\partial Q}{\partial t} = A_e \cdot \frac{\partial D}{\partial t}$ or $\boxed{A_e \cdot \dot{D}}$ output proportional to $\frac{\partial D}{\partial t}$ or $\dot{D}$

The MIL-STD-461 hence calls this device a “D-dot probe”



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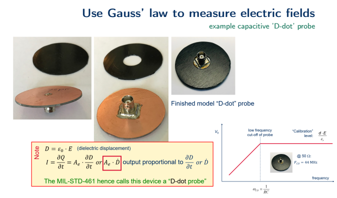


The slide shows a possible E-field probe design: a circular section of double-clad printed circuit board (PCB).
The capacitance of this device, C can be measured with an LCR-meter. Given the termination resistor, usually $R = 50\omega$, the cut-off frequency can then be calculated: $f_{CO} = \frac{1}{2\pi RC}$.
In EMC-standards, e.g. the MIL-STD-461, an electric field probe as shown here is called a $\dot{D}$ or “D-dot” probe.
This is because the electric field, E, is multiplied by ε_0, which is called the dielectric displacement, D.
The equation for the substitute current, I, then uses the derivative $I = A_e \frac{\partial D}{\partial t}$ which can be written as $I = A_e \dot{D}$ (Aspect 58).
The capacitance between the sensor plate and the reference is $C = 72$ pF. Therefore, with $R = 50\ \Omega$, the cut-off frequency is $F_{CO} = \frac{1}{2\pi RC} = 44$ MHz.

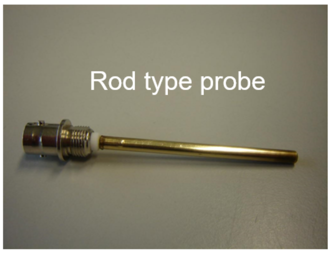
This frequency would come down if a *higher* load resistor is used e.g. $R = 100$ kΩ: theoretically, 22 kHz. A 50Ω connecting cable would disturb this[2]. So, a local high-impedance receiver would have to be placed in/at the transducer in this case.
The flat curve will start to fluctuate at very high frequencies, see slide 39 Aspect 58).

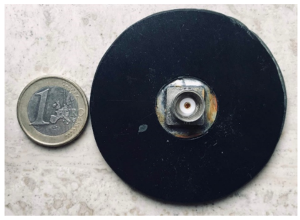
[2] It behaves as an extra *capacitive* load now ≈ 100 pF/m.

Example electric field probes

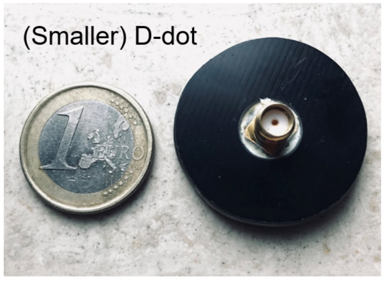
'home made' (+1 commercial) electric field transducers

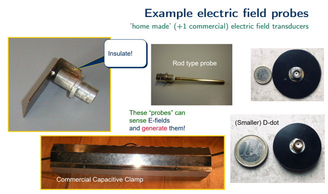
Insulate!





These "probes" can sense E-fields and generate them!





Next to the probe described in slide 20 and its little brother, three more are shown here. A simple copper plate mounted on a BNC-connector is shown at the top left.
A brass rod is shown to the right of it.
A disadvantage for this type of probe is the missing –defined– parallel capacitor.
Which makes it difficult to calibrate: there is always capacitance from the sensor-plate or rod to the GRP –which is only the connector body here– but it is very dependent of the way the probe is handled and the closeness to other metallic objects around –sometimes called "hand-effect"–.

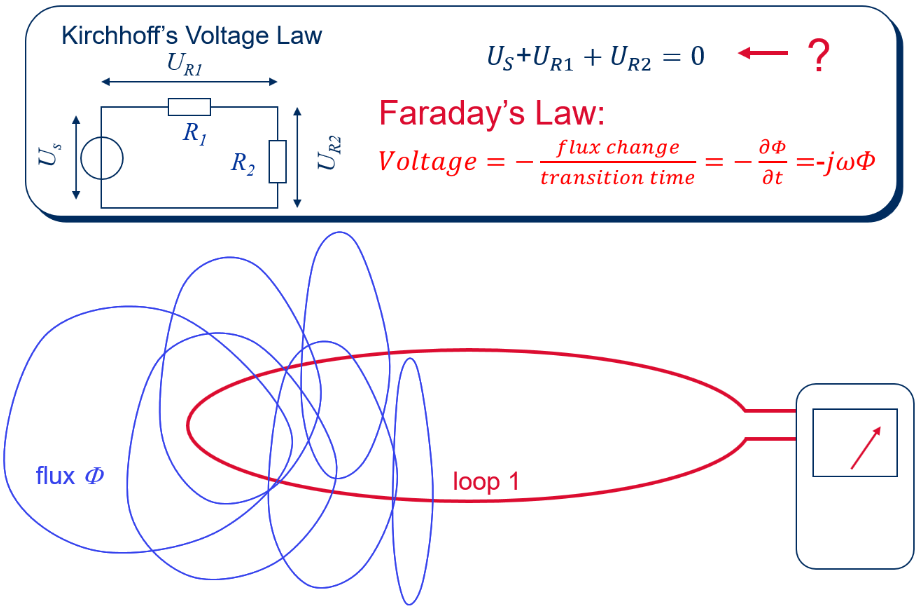
On the other hand, calibration is not common for D-dot probes as they are only used to show E-fields are present. Their amplitudes are not *measured*.
The device shown in the left-bottom corner is a commercial capacitive coupling clamp which is used to measure emission interference from cables and also to inject capacitive currents onto cables to check susceptibility. As it concern cables, these tests are classified as *conducted*.
In principle all probes shown are able to **generate** electric fields towards nearby conductors (Aspect 58).

Part 3: Measuring currents and magnetic fields

Quantify EM-phenomena
Part 3: Measuring currents and magnetic fields

In order to get a better handle on EM-phenomena, we need to be able to quantify, measure them.
Here we focus on magnetic fields and currents.

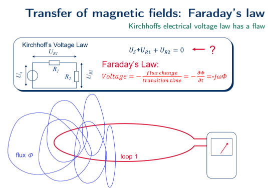
Kirchhoffs electrical voltage law has a flaw



EMC for Engineers

2023-11-29

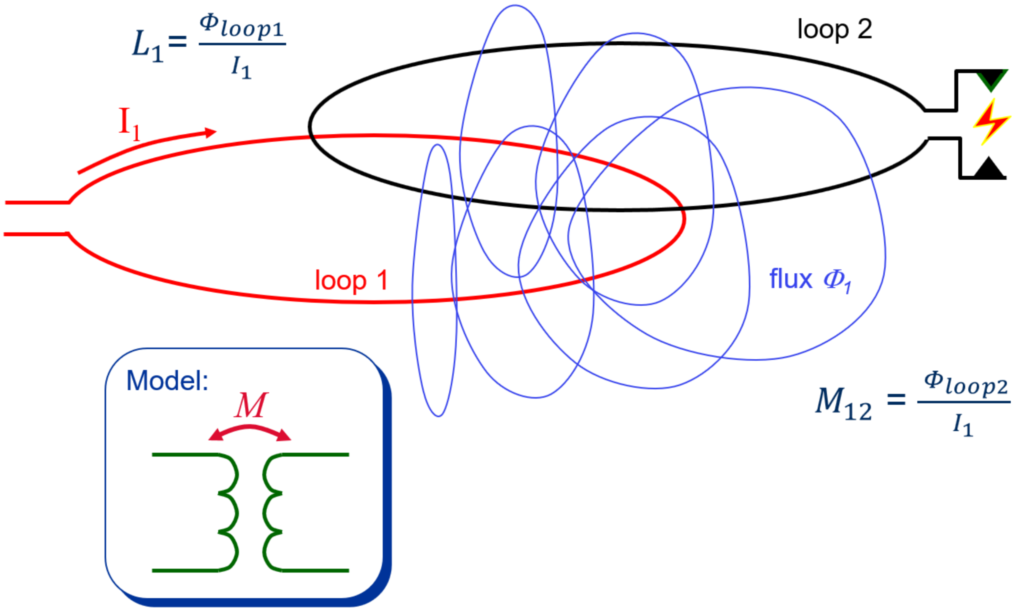
Transfer of magnetic fields: Faraday's law

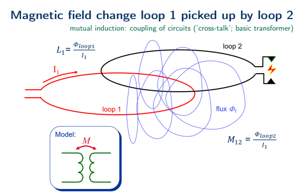


Kirchhoff's electrical voltage law helps to calculate components in network analyses.
It states that the sum of voltages in a network loop must be zero.
But this is only true if there are no external magnetic fields (Aspect 59).
As soon as the network loop is exposed to alternating magnetic fields, extra voltages are induced described by Faraday's law.
Of course, electric fields may also induce additional currents. The focus here is on magnetic fields and the phenomenon of mutual inductance.

Magnetic field change loop 1 picked up by loop 2

mutual induction: coupling of circuits ('cross-talk'; basic transformer)



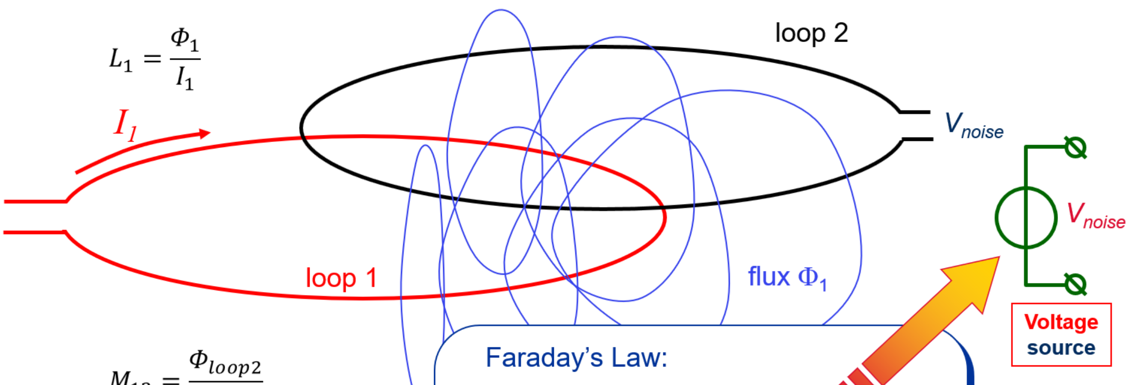


In $loop_1$, which has an alternating current, I_1, generating the magnetic flux, Φ_{loop1}, the process creating the flux Phi_{loop2} in $loop_2$ is called mutual-induction.
The magnetic flux produced by a current, I_1, in a wire loop, L_1, is $\Phi_{loop1} = L_1 I_1$.
Therefore, the *self-induction* of the loop is $L_1 = \frac{\Phi_{loop1}}{I_1}$.
Some of this flux, Φ_{loop2} is picked up by the nearby second loop and induces a *voltage* based on Faraday's law.
The ratio $\frac{\Phi_{loop2}}{I_1} = M12$ is called the mutual induction between $loop_1$ and $loop_2$.
This principle is used e.g. in transformers.

Current *change* loop 1 induces *voltage* in loop 2

substitute voltage source = model for inductive cross-talk

$L_1 = \frac{\Phi_1}{I_1}$



$M_{12} = \frac{\Phi_{loop2}}{I_{loop1}}$

$\Phi_{loop2} = M_{12} \cdot I_{loop1}$

Faraday's Law:

$$V_{noise} = -\frac{d\Phi_{loop2}}{dt} = -j\omega\Phi_{loop2}$$

or

$$V_{noise} = -\frac{dM_{12} \cdot I_{loop1}}{dt} = -j\omega M_{12} \cdot I_{loop1}$$

minus sign associates with **Lenz' law**: induced field opposes changes in the initial magnetic field

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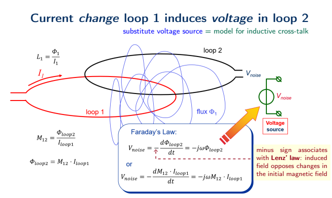
└Current *change* loop 1 induces *voltage* in loop 2



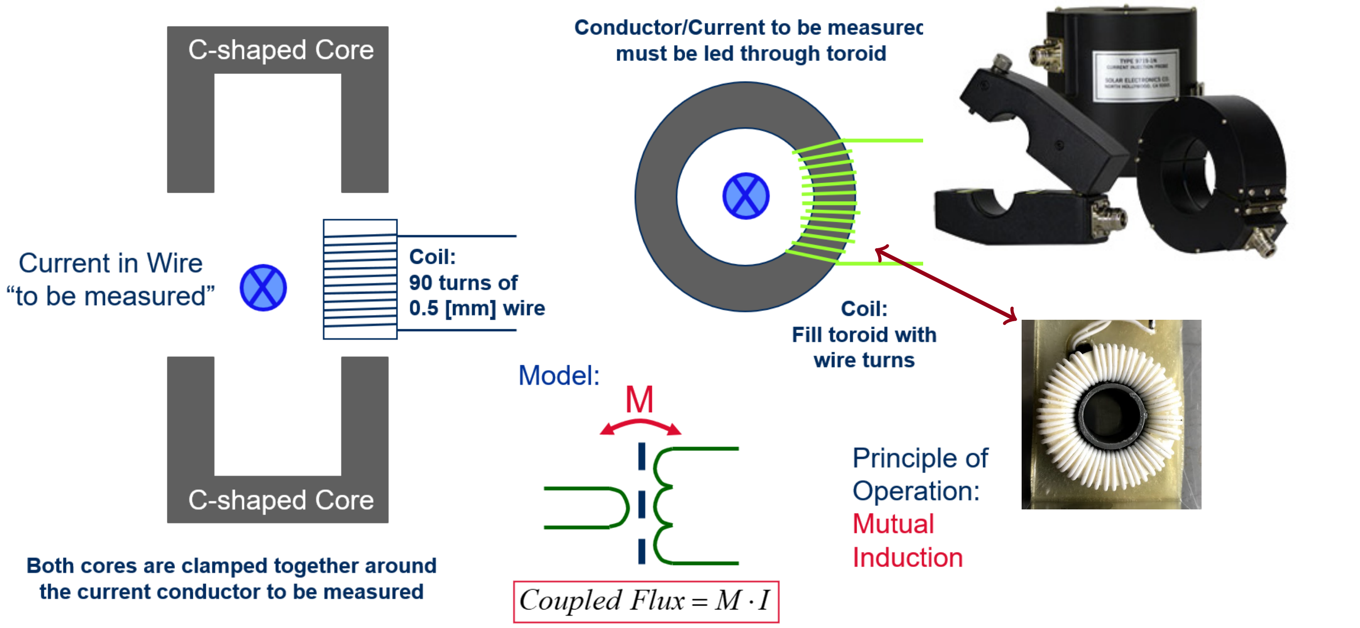
Faraday's law defines the voltage, $V_{noise}(t)$ here, developed in $loop_2$ as $V_{noise}(t) = -\frac{\partial \Phi_{loop_2}}{\partial t}$.
This noise voltage can be rewritten as: $V_{noise}(t) = -\frac{\partial M_{12} I_{loop_1}(t)}{\partial t}$.
The $V_{noise}(t)$ describes the effects[3] seen in $loop_2$ due to the interference caused by the current $I_{loop_1}(t)$ in $loop_1$.
Therefore it is called a *substitute* –voltage– source, characteristic for inductive –magnetic– interference (Aspect 60).
In the frequency domain, this substitute source can be written as:

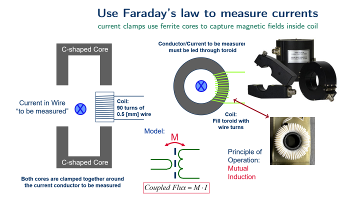
$V_{noise}(j\omega) = -j\omega M_{12} I_{loop_1}(j\omega)$
Note:
1. Lenz's law states that the direction of the electric current induced in a conductor by a changing magnetic field is such that the magnetic field created by the induced current opposes changes in the initial magnetic field. This is the **minus sign** in Faraday's law (Aspect 61).
2. Inductive cross-talk has a substitute *voltage* source (see Aspect 60).
3. Capacitive cross-talk has a substitute *current* source (see Aspect 55).

[3] If $loop_2$ is loaded, the self-induction, L_2 will have to be considered to calculate I_2.

Use Faraday's law to measure currents

current clamps use ferrite cores to capture magnetic fields inside coil





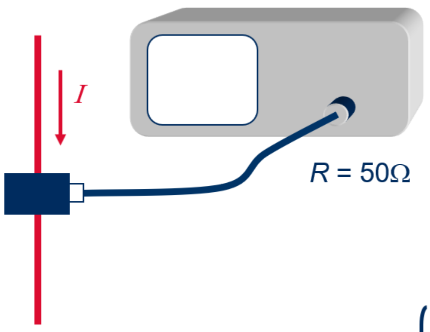
The cross-talk mechanism through mutual induction can be used to measure currents. This is usually performed by actually building a transformer. It has a ferrite –or metal– core on which the measuring loop is wound as a coil[4]. On the left in the slide a picture of a coil using two ferrite C-cores, as used in the old TV-receivers high-voltage transformer. At the right hand side, a toroid core with wire windings[5]. In both cases the current-to-be-measured flows in a wire or cable that is led through the center of the magnetically closed ferrite loop: a single turn. The device is called a current-transformer, current-probe or current-clamp. The disadvantage of the former, separable core is that this specific ferrite shape is not manufactured any more. The disadvantage of the latter ring model is that it cannot be "opened". Hence it can only be used if the cable or wire connector fits through the center hole. Note: these are EMC-experiment devices. Use a ready made commercial current clamp for formal tests –see e.g. inset top-right–.

[4]Separable core: the older ferrite 3C8 (These current-clamps were designed and built in 1991).
[5]Ferrite core 3E27 size TX40/24/16, 53 turns of insulated AWG 22 wire, $F_{CO} = 300$ Hz on 50Ω.

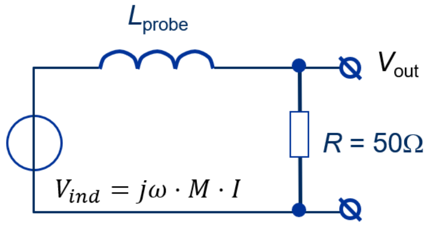
Use Faraday's law to measure currents

current probe response using the substitute voltage source

Measurement setup:



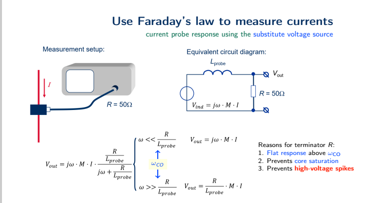
Equivalent circuit diagram:



$$V_{out} = j\omega \cdot M \cdot I \cdot \frac{\frac{R}{L_{probe}}}{j\omega + \frac{R}{L_{probe}}} \begin{cases} \omega << \frac{R}{L_{probe}} & V_{out} = j\omega \cdot M \cdot I \\ \updownarrow \; \omega_{CO} & \\ \omega >> \frac{R}{L_{probe}} & V_{out} = \frac{R}{L_{probe}} \cdot M \cdot I \end{cases}$$

Reasons for terminator R:

1. Flat response above ω_{CO}
2. Prevents core saturation
3. Prevents **high-voltage spikes**

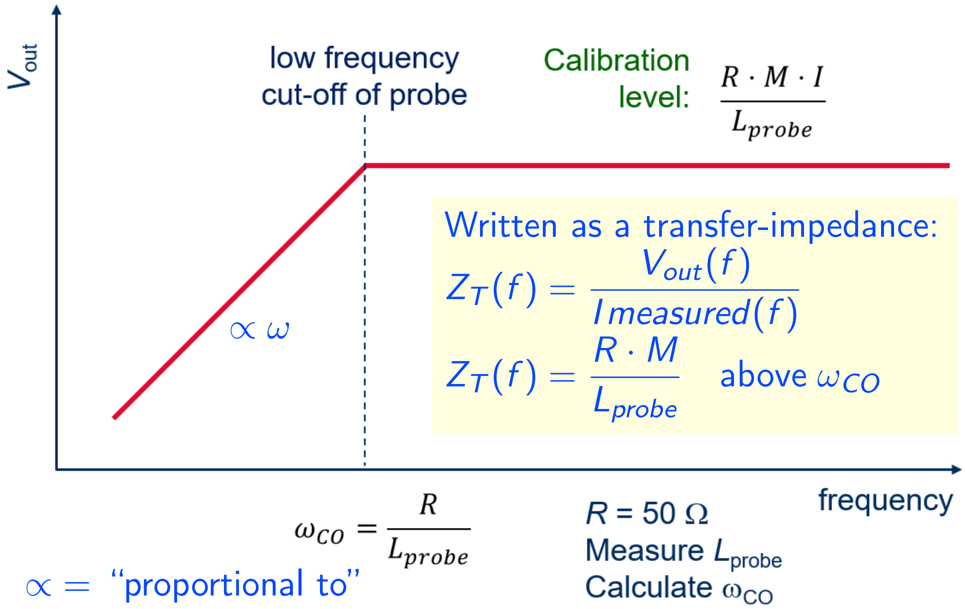


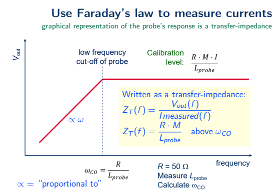
As for the case of the capacitive probe on slide 18, an equivalent circuit diagram can be drawn for the current-probe.
It shows the substitute voltage source, $V_{ind}(j\omega) = j\omega MI(j\omega)$ in series with a coil, L_{probe}.
This current-probe shall be terminated in a resistor –at the end of the coax cable on the measurement instrument–, usually 50ω, the coax characteristic impedance (Aspect 62).
This inductive-probe has a high-pass behavior as the capacitive-probe and a clear cut-off frequency, $f_{cut-off} = \frac{R}{2\pi L_{probe}}$.

The coil, L_{probe}, of a home-made –or commercial– version can be measured with an LCR-meter, to use in the $f_{cut-off}$-equation (if applicable: probe core should be closed).
The current-transformer transfers a current into a voltage. Their ratio, $Z_T = \frac{V_{out}}{I}$ is an impedance. For that reason manufacturers provide a transfer-impedance (Z_T) in their specification.
A probe that yields 1 Volt –over 50Ω– for 1 Ampère of current has a transfer impedance of 1Ω. This, too, could be specified in decibels: $Z_T = 1\Omega$ equals $Z_T = 0$ dBΩ –use $20 \log Z_T$–
.

Use Faraday's law to measure currents

graphical representation of the probe's response is a transfer-impedance





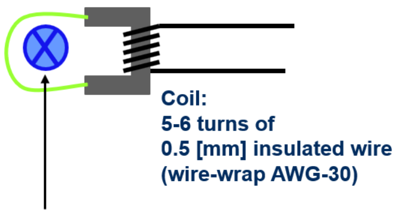
The tranfer function of the current probe is drawn in this slide.
The high-frewequency response is flat –frequency independent–.
Below the cut-off frequency, the response increases *proportional to*, $\propto$, the frequency and the measured results need to be corrected.
For that reason, manufacturers often provide a graph of the correction-factor[6] –or antenna factor (AF)–, comparable to the graph shown on slide 19.
Note: at some high frequency the flat response ends and resonances will be seen –usually during a calibration–. This marks the end of the usable frequency range.
The "home-made" current transformers shown perform well up to at least 10 MHz.

[6]Usually in deciBells, dB, to be **added** to the output voltage, in dBμV to obtain the current in dBμA. (read the manual for details)

Use Faraday's law to detect magnetic fields

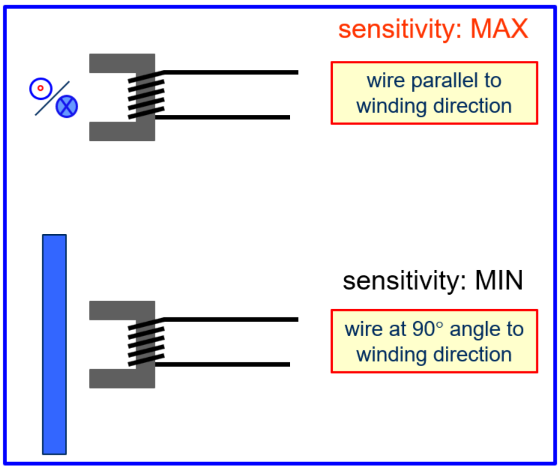
5-turn 'sniffer' probe with ferrite core to concentrate flux lines

Small C-shaped core



Conductor under test
(or external field)

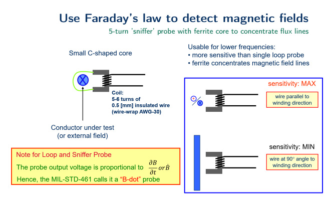
Usable for lower frequencies:
- more sensitive than single loop probe
- ferrite concentrates magnetic field lines



Note for Loop and Sniffer Probe

The probe output voltage is proportional to $\frac{\partial B}{\partial t}$ or $\dot{B}$

Hence, the MIL-STD-461 calls it a "B-dot" probe



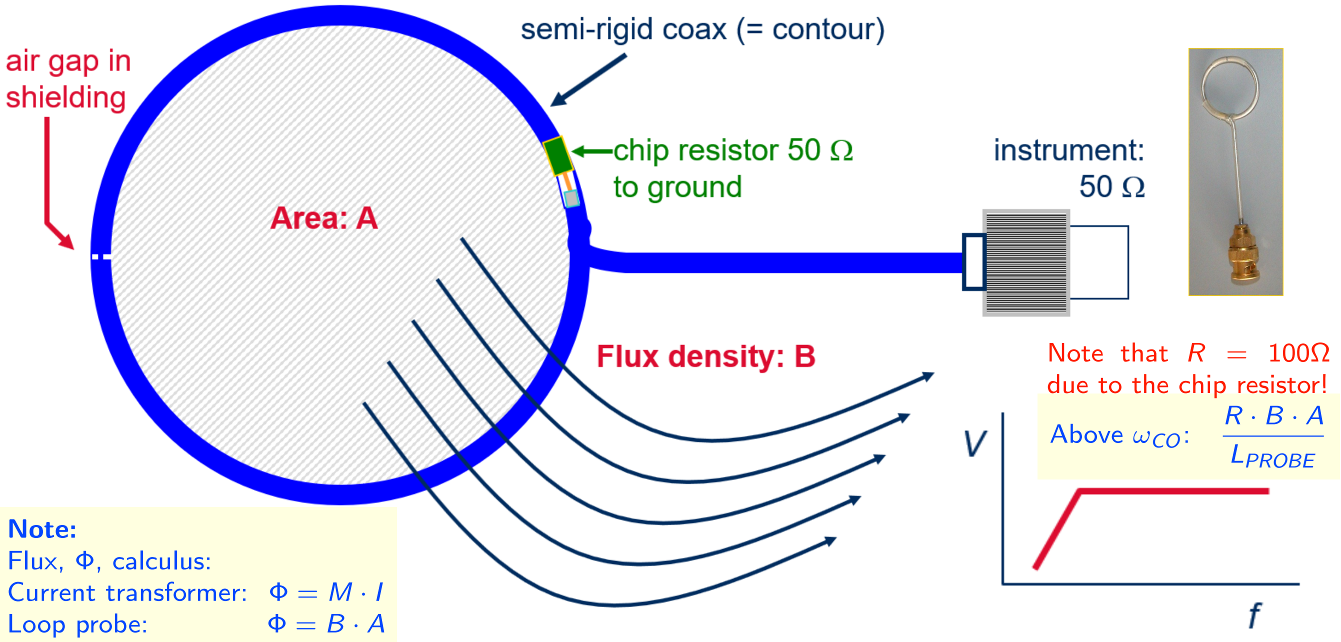
A few turns of wire-wrap wire on a small ferrite C-core can make a handy magnetic field "sniffer" probe to look for magnetic fields around wires or cabinets.
The ferrite core concentrates magnetic field lines inside the turns of the coil.
Even a single turn of wire, without ferrite, can be used for very high frequencies –see picture top left–.
The circuit diagram of the current-probe applies.

These devices cannot be calibrated as the amount of flux in the coil or loop depends on the distance between the field-source and the coil[7].
But the devices rely on Faraday's law i.e. their output is proportional to the *derivative* of the magnetic flux, B, enclosed by the loop or turns: $V_{out} \propto \frac{\partial B}{\partial t}$ or $\dot{B}$, hence these devices are labeled B-dot probes in e.g. MIL-STD-461 (Aspect 63).

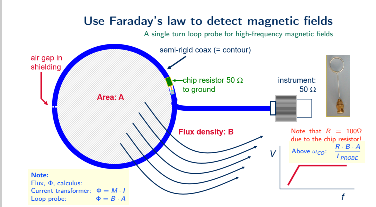
[7] If you have another ferrite C-core, you can close the magnetic loop and obtain a current-transformer.

Use Faraday's law to detect magnetic fields

A single turn loop probe for high-frequency magnetic fields



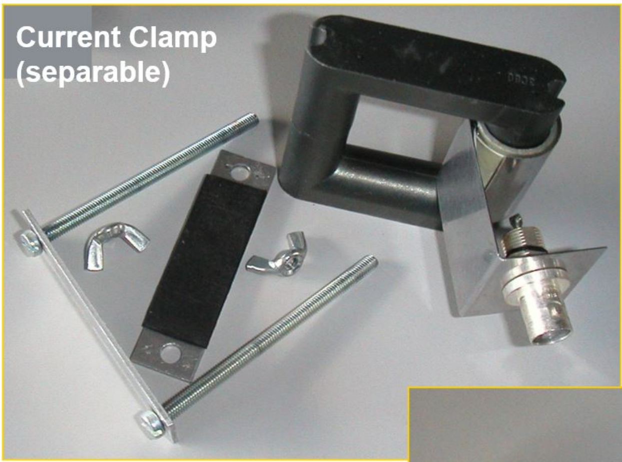
Note:
Flux, Φ, calculus:
Current transformer: $\Phi = M \cdot I$
Loop probe: $\Phi = B \cdot A$

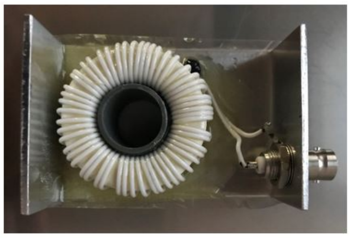


For high frequency work, this single turn $\dot{B}$ or B-dot probe can be used. A single turn, and no ferrite, means very low inductance, L, i.e. a high $F_{CO} = \frac{\omega_{CO}}{2\pi}$. But for the rest, the model on slide 27 applies. This is actually the way loops in equipment pick up magnetic fields –part of the complete EM-field– in a hazards of electromagnetic radiation to ordnance (HERO) situation. The device, shown at the top right hand side in the slide, is built using semi-rigid coax. This is a coax that has a massive –galvanized– copper shield. This shield is used for electric field shielding of the probe so it will only react to magnetic fields. To make that possible, the shield must have an "air-gap", shown at left. The flux, Φ, in the equations is now calculated as the magnetic flux-density, B, times the area, A inside the loop: $\Phi = B \cdot A$. This replaces the original $\Phi = M \cdot I$ in the current transformer model on slide 27. Like the 5-turn "sniffer" probe on slide 29, it is used to check for the presence of magnetic fields, close to cabinets or printed circuit boards (PCBs) rather than *measure* them. The loop-probe diameter shall be much smaller than a quarter wavelength of the highest frequency. In case of doubt, build a smaller loop! The 50Ω chip resistor terminates the center wire of the coax to the shield. Use $R = 100\Omega$ in the equations!

Example magnetic field based transducers

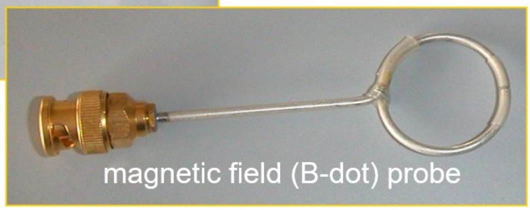
'home made' current and magnetic field probes

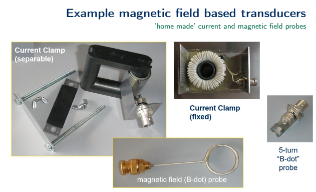




Current Clamp (fixed)

5-turn "B-dot" probe

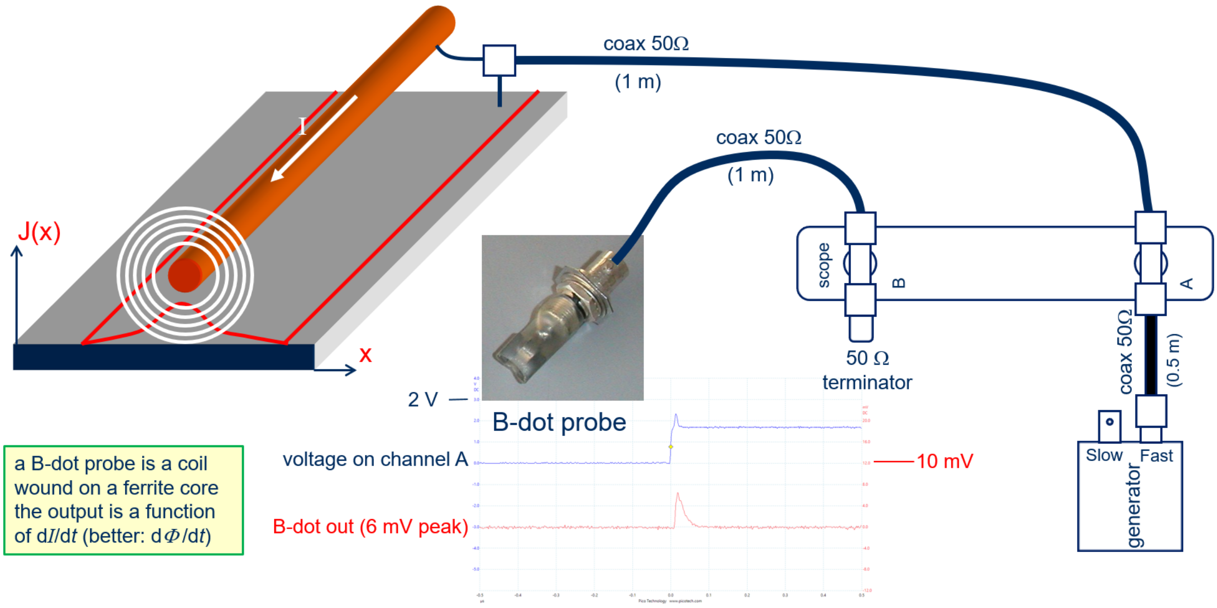


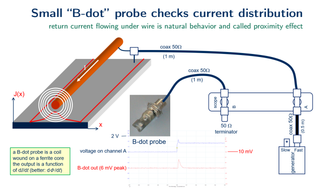


Photographs of the 4 described magnetic induction based transducers are shown in this slide.

Small "B-dot" probe checks current distribution

return current flowing under wire is natural behavior and called proximity effect

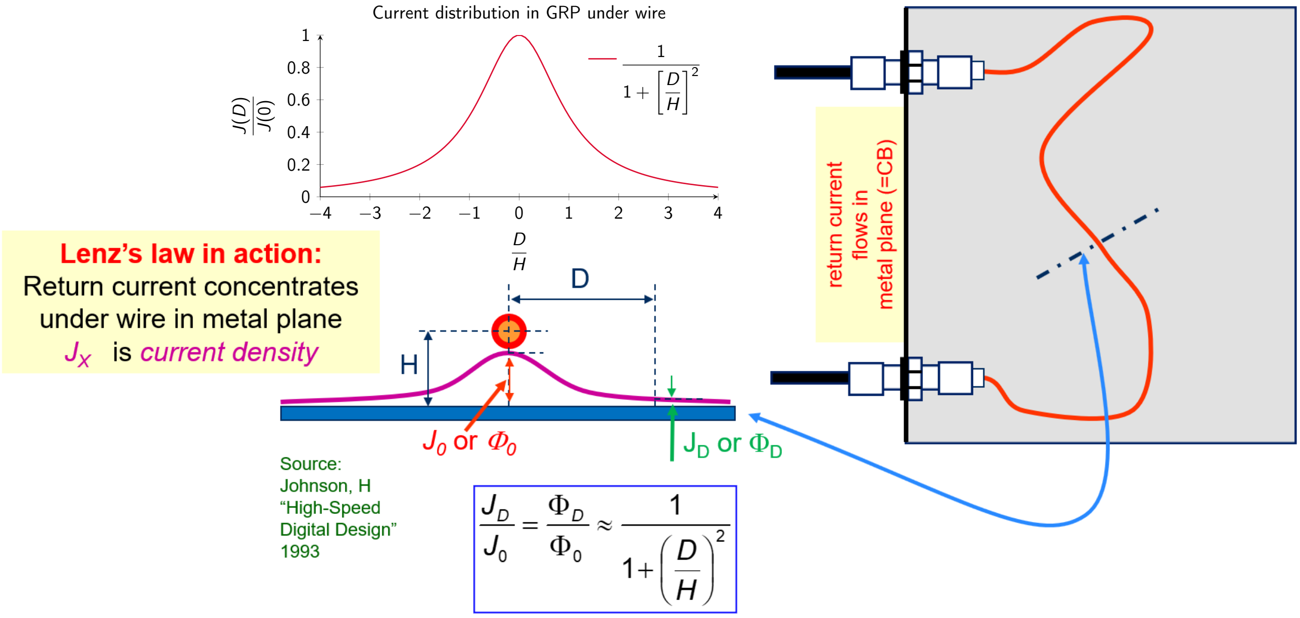




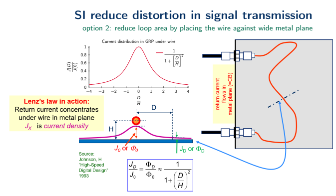
The five turn B-dot probe an be used to investigate the magnetic field distribution around a wire –with current– over a ground reference plane (GRP).

SI reduce distortion in signal transmission

option 2: reduce loop area by placing the wire against wide metal plane



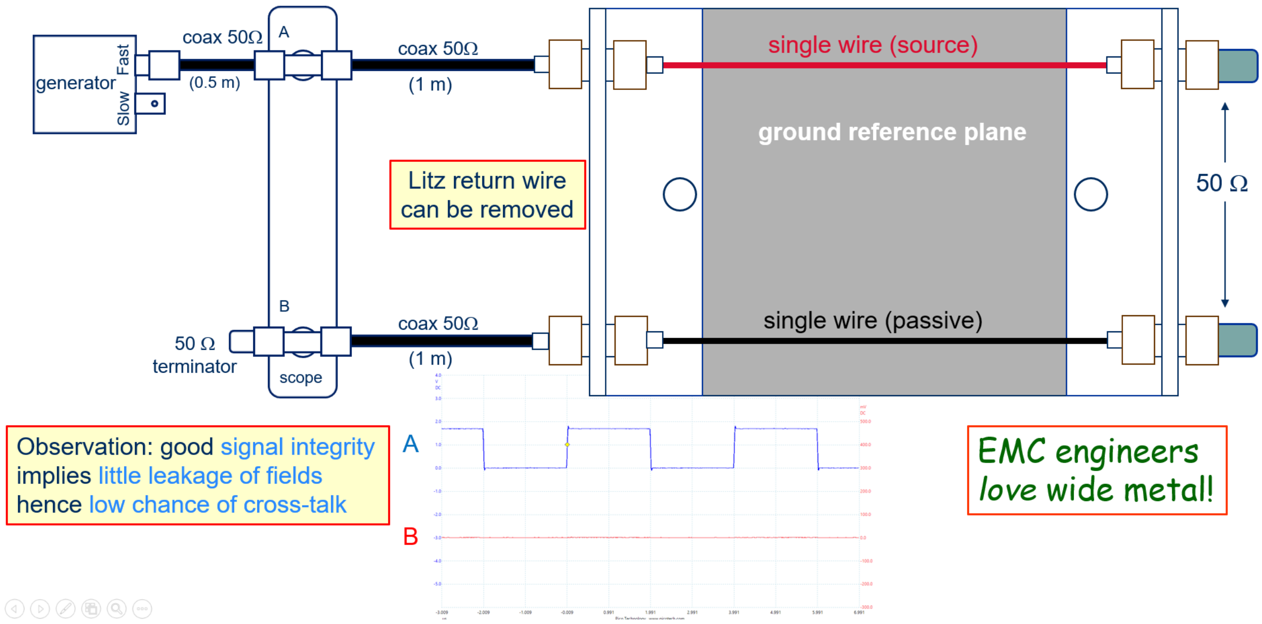
$$\frac{J_D}{J_0} = \frac{\Phi_D}{\Phi_0} \approx \frac{1}{1+\left(\frac{D}{H}\right)^2}$$



By moving the B-dot probe over the wire, the current distribution in the GRP can be inspected.
Directly over the wire, the current in the wire is seen as a spike on the oscilloscope.
Immediately next to the wire, the polarity of the spike reverses, showing that the current in the GRP is flowing in the opposite direction.
The return current in the GRP distributes in a bell shape under the wire.
If the wire is lifted, this bell becomes wider.
This is another manifestation of **Lenz's law** that states that the direction of the electric current induced in a conductor by a changing magnetic field is such that the magnetic field created by the induced current opposes changes in the initial magnetic field (Aspect 64).
Another way to say this: the return current's magnetic field in combination with the field of the line above it will occupy the minimum necessary volume between line and GRP i.e. it is a natural conservation mechanism.
Please use it to reduce electromagnetic interference (EMI)!

Minimize mutual induction to reduce cross-talk

best solution: wide return conductor –ground reference plane (GRP)–



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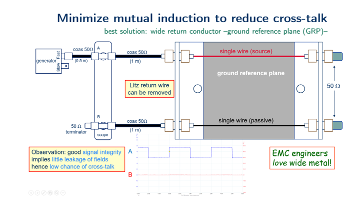


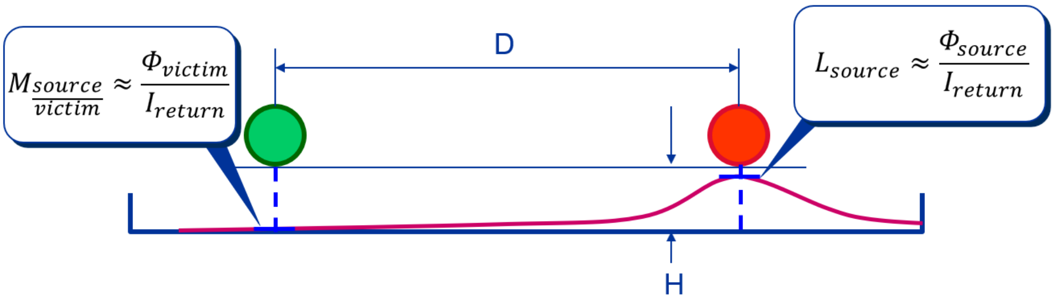
This bell shape of the return current under the source wire in our cross-talk experiment has an immediate impact on the amplitude of the cross-talk observed in the victim wire.
In fact, we can do the same measurements with the victim wire in place of the B-dot probe in slide 33.
Move the victim wire closer to or away from the source wire while keeping both at the same height over the GRP –use a book under the wires–.

Ratio height to distance determines XTK

wire distance is important once the return path, GRP, is in place

Current distribution of I_{return} in metal plane is measure of flux density, Φ, coupling into wire: at source $\sim L$, at victim $\sim M$!

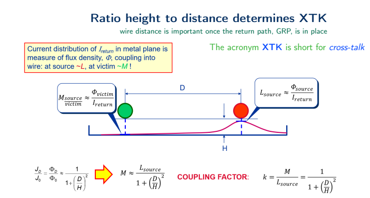
The acronym **XTK** is short for *cross-talk*



$$\frac{J_D}{J_0} = \frac{\Phi_D}{\Phi_0} \approx \frac{1}{1+\left(\frac{D}{H}\right)^2} \Rightarrow M \approx \frac{L_{source}}{1+\left(\frac{D}{H}\right)^2}$$

COUPLING FACTOR:

$$k = \frac{M}{L_{source}} = \frac{1}{1+\left(\frac{D}{H}\right)^2}$$



The current-density, J_0, in the GRP directly under the source wire –D=0– corresponds to the local flux-density between the source wire and the GRP.
In the same way, the current-density under the victim wire –D=D– corresponds to the flux-density under the victim wire.
The –total– flux captured between the source wire and the GRP corresponds to the (self-)inductance, L_{source}, under this wire.
In the same way, the flux captured between the victim wire and the GRP corresponds to the mutual-inductance, M, between the source and the victim wire loops formed with the GRP.
EMC literature in the USA predominantly uses L and M, in Europe the *coupling-factor* is preferred, $k = \frac{M}{L}$.

Part 4: The meaning of long and short in electromagnetics

Transmission-lines

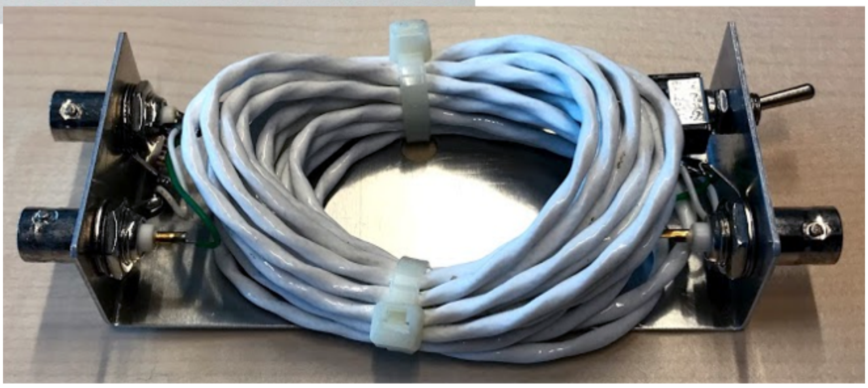
Part 4: The meaning of long and short in electromagnetics

Long and short in electro-magnetics relates to the wavelength of the transmitted signals. To really demonstrate long and short on a 200 MHz oscilloscope and a 4 ns rise time generator, we need to make our experiments longer, say, 2 meters or more.

Cross-talk experiment capacitive and inductive

monitor both near (source) and far (load) end of passive line

Shown in the Basics module...



Seven times longer...
To expose more detail

The concept of reflections, demonstrated in slide 9, has implications for another experiment, performed in the basic EMC module: Capacitive and inductive cross-talk and their combination, shown using the aluminum frame with two brass bars, visible at the top of this slide.
It is possible to show the effect of reflections on transmission lines in two ways:

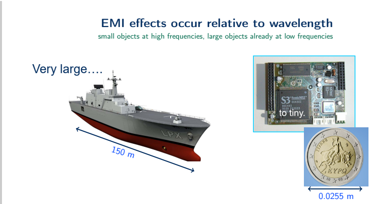
- Using faster –smaller risetime– signals in the same experiment;
- Using the same signals and increasing the length of the experiment.

Given the available oscilloscope and generator, a faster signal requires a faster generator and oscilloscope which is quite an investment.
Making the experiment larger is easy and cheap: in the bottom of the slide the same experiment is shown built from shielded twisted pairs (STP)-transmision line of 2.5 m, about seven times the original length (Aspect 65).

EMI effects occur relative to wavelength

small objects at high frequencies, large objects already at low frequencies

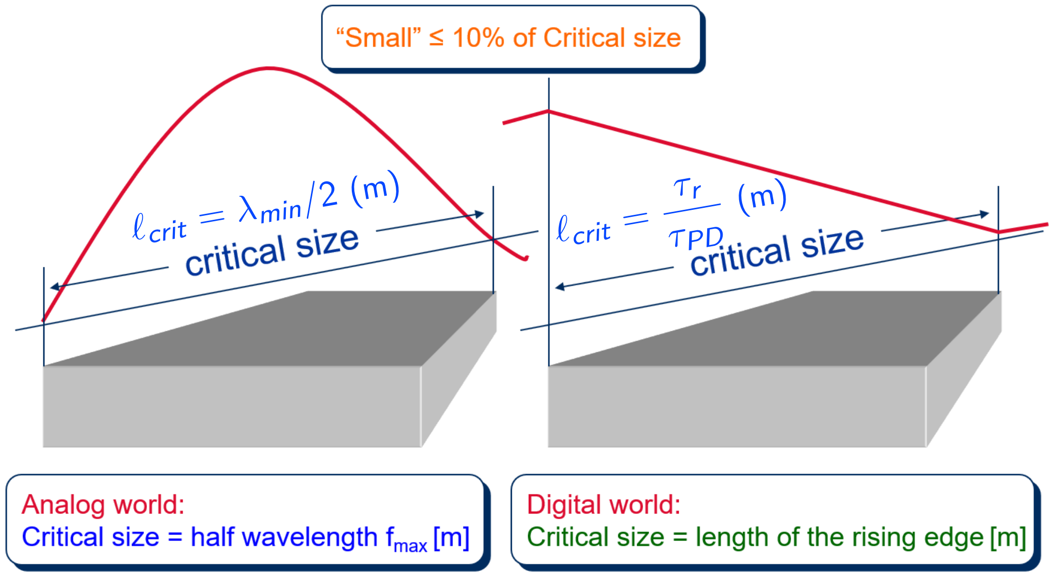
Very large....



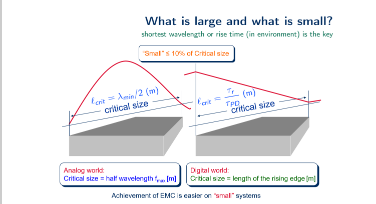
The interference phenomena occur in systems of any size.
Key is the wavelength of the interference in relation to the size of the source or victim...
When scaled to wavelength, behavior is similar.

What is large and what is small?

shortest wavelength or rise time (in environment) is the key



Achievement of EMC is easier on "small" systems



In the analog world the shortest wavelength, λ_{min}, of the highest frequency in the environment, f_{max}, is calculated using the parameters below. Divide that wavelength by 2 to find ℓ_{crit}:

$$\nu = \frac{1}{\tau_{PD}} \leq 3 \times 10^8 \text{ m/s}$$

Thi is the propagation speed of the EM-waves (Use "=" in free space)

$$\lambda_{min} = \frac{c}{f_{max}} \text{ [m]}$$

$$\ell_{crit} = \frac{\lambda_{min}}{2} \text{ [m]}$$

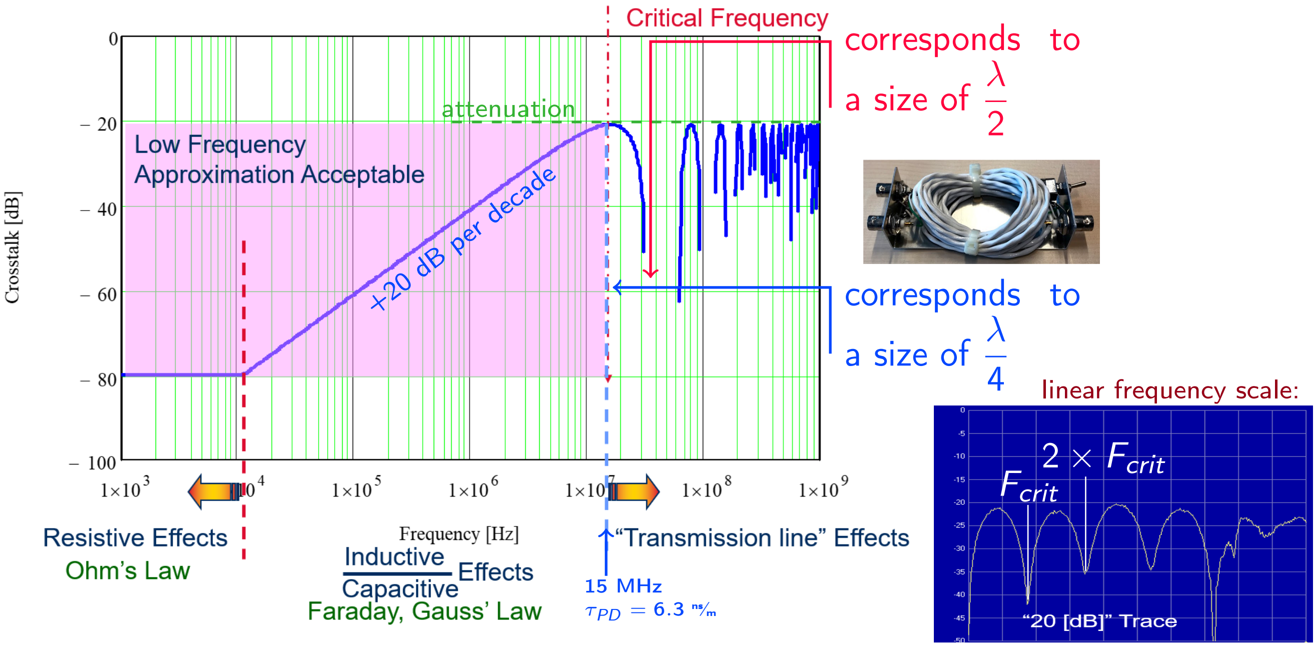
To calculate the critical size, ℓ_{crit}, in the digital world the key parameters are: $\tau_{PD} = \ldots$ s/m propagation delay (about 6.3s/m in wires) $\tau_r = \ldots$ [s] rise time of logic signals

$$\ell_{crit} = \frac{\tau_r}{\tau_{PD}} \text{ [m]}$$

One of the basic rules to achieve EMC is *do not use high frequencies*. What is a high frequency depends on the size of the system or rather, the length of the longest interconnection used. As we will see in the upcoming slides, the EM-behavior of a wire or cable changes drastically as soon as its size approaches one of the criteria in this slide.

The critical size implies a critical frequency

example: crosstalk between two 2.5 m long parallel lines (experiment slide 44)



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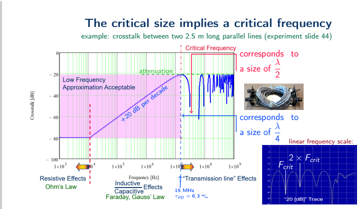
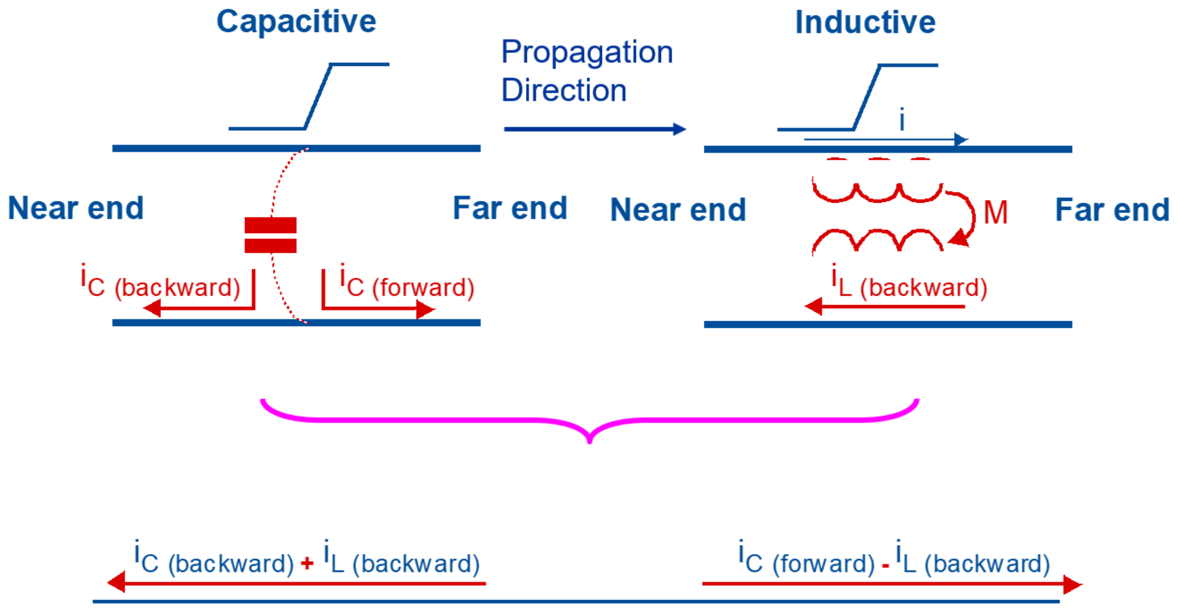
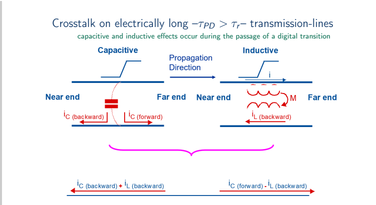


The analog world model for the critical size in EM-terms, shows a *critical frequency* related to the *physical dimensions* of an electronic system or module. This slide shows the frequency *spectrum* of the cross-talk measured at one end of a (victim) line/wire running in parallel to a (source) line/wire inside a common shielding: shielded twisted pairs (STP). The horizontal scale shows the *logarithm* of the frequency of excitation on the source line. The vertical scale shows the cross-talk voltage measured on the victim line relative to the voltage of the source line excitation. It is a logarithmic scale and the units are decibels, $20 \log \frac{V_{victim}}{V_{source}}$. At low frequencies, the ratio is determined by resistive effects in the common return metal shield. For higher frequencies, the cross-talk increases proportional to frequency, shown by the inclined straight line. This ends at *half* the critical frequency, determined by the physical length of both (source and victim) conductors. Here the cross-talk value reaches a *maximum* value. At the critical frequency, the cross-talk has a *minimum* and a low value at regular intervals as shown in the blue inset picture at the right hand bottom with a *linear* frequency scale (Aspect 66). This shows that at frequencies from $\frac{F_{crit}}{2}$ the cross-talk reaches an asymptote. Its height is determined by the *geometry of the cable* and is labeled ***attenuation***.

Crosstalk on electrically long $-\tau_{PD} > \tau_r-$ transmission-lines

capacitive and inductive effects occur during the passage of a digital transition



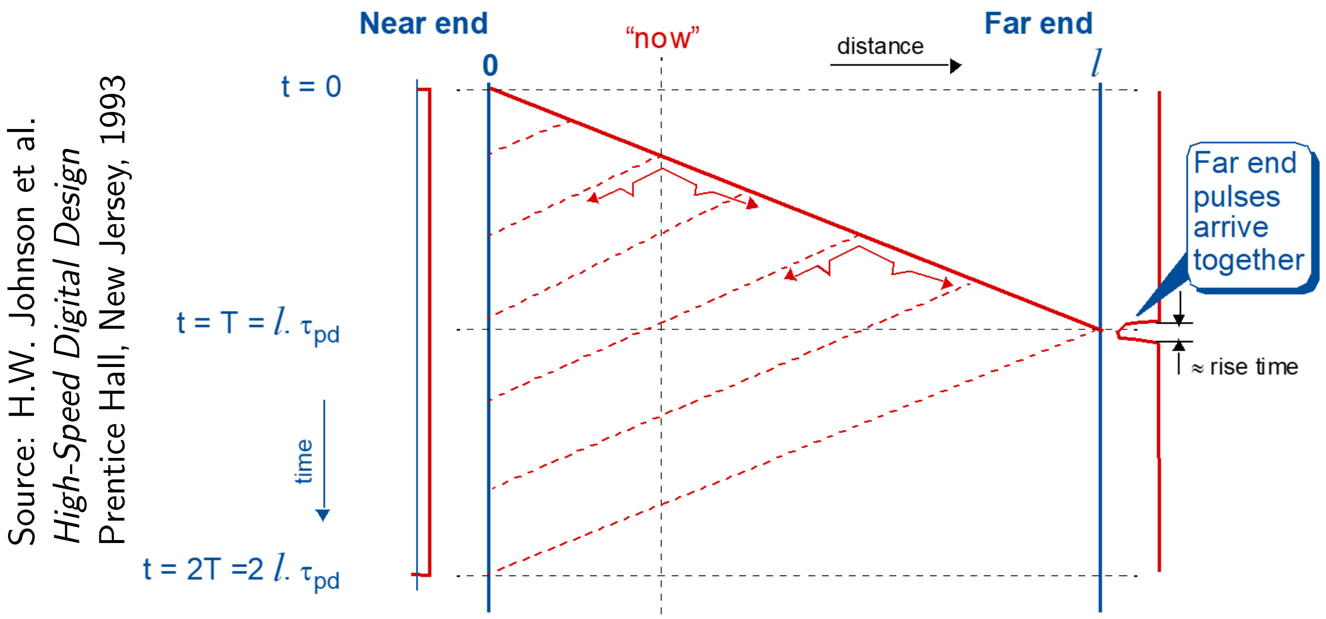


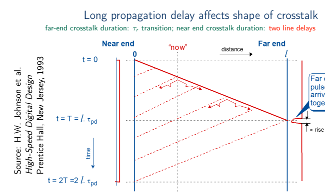
If the length of the experiment is much larger than the critical length defined on slide 39, Cross-talk occurs only *locally* at the location where a signal transition is passing at the moment (Aspect 67). The source signal propagation direction is from left to right in the drawing, the source resides at the *near* end of the line.
Both the Gauss and Faraday laws show that energy is transferred based on the derivative of electric or magnetic flux, a change in voltage and/or current in the source line.
The capacitive cross-talk generates a current flowing between the source line and the victim line. This capacitive current, based on voltage changes, I_C, flows in both directions on the victim wire.
A current change in the source wire induces a current, I_L, in the victim wire that flow in the *opposite* direction,[8] towards the source or near end of the wire.
The combined effect, generated *at the location of the propagating transition* on the source line is a current *backwards* towards the near-end of the victim line with amplitude $I_C + I_L$ and a current $I_C - I_L$ *forward* towards the far end of the victim line. These are called backward- and forward-cross-talk respectively.

[8] Remember the minus sign in the Faraday law –i.e. Lenz's law–

Long propagation delay affects shape of crosstalk

far-end crosstalk duration: τ_r transition; near end crosstalk duration: two line delays

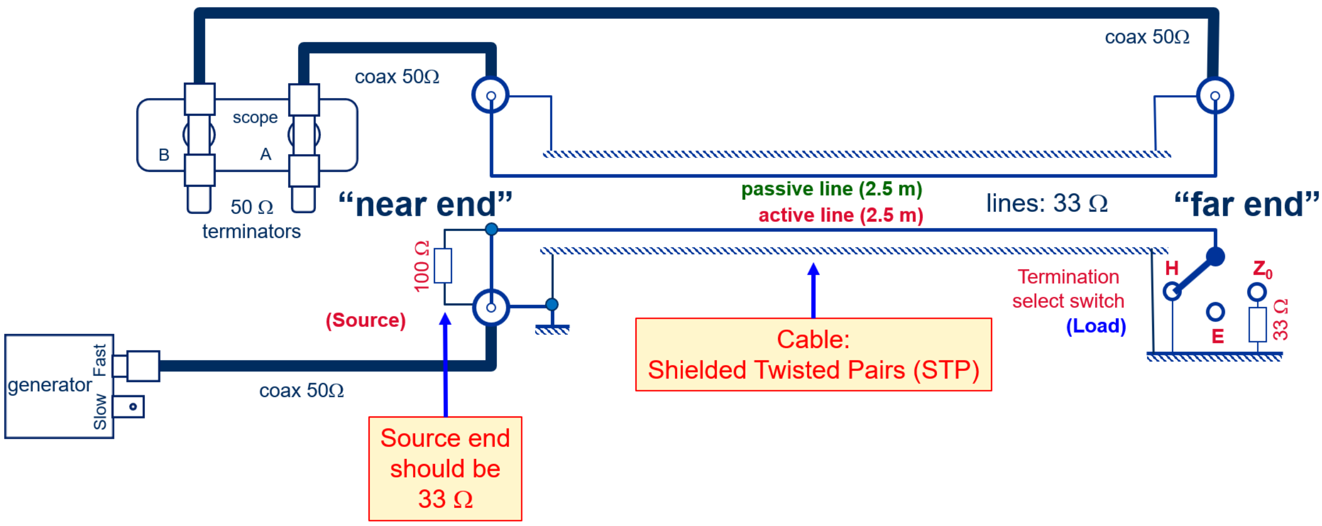




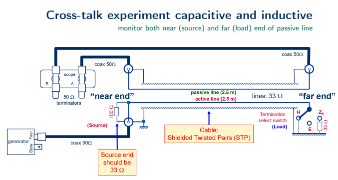
This graph has a horizontal distance scale and a vertical time scale.
The slanted red line is the path along the victim line of cross-talk caused by a logic transition on the source line, moving from the source –the near end– starting at time 0 at distance 0 –top left– to the end of the victim line –far end–.
After $T(ime) = \ell\tau_{PD}$, where ℓ m is the total line length and τ_{PD} s/m the propagation delay, the cross-talk arrives at the end of the line at distance ℓ.
During the passage of the logic transition on the source line, the *forward* cross-talk builds up and moves at the same speed in the direction of the *far* end on the victim line where it arrives at the same time as the logic transition arrives at the far end of the source line.
The duration of this cross-talk spike is τ_r, the risetime of the logic transition.
All this time, the *backward* cross-talk is generated and moves from the location of the logic transition towards the near end.
After the arrival of the transition at the end of the line, the last amount of *backward* cross-talk starts on its way towards the near end of the line and arrives at $2T(ime) = 2\ell\tau_{PD}$.
The duration of the backward cross-talk at the near end is therefore $2T$, Much longer than the duration of the *forward* cross-talk, τ_r.

Cross-talk experiment capacitive and inductive

monitor both near (source) and far (load) end of passive line



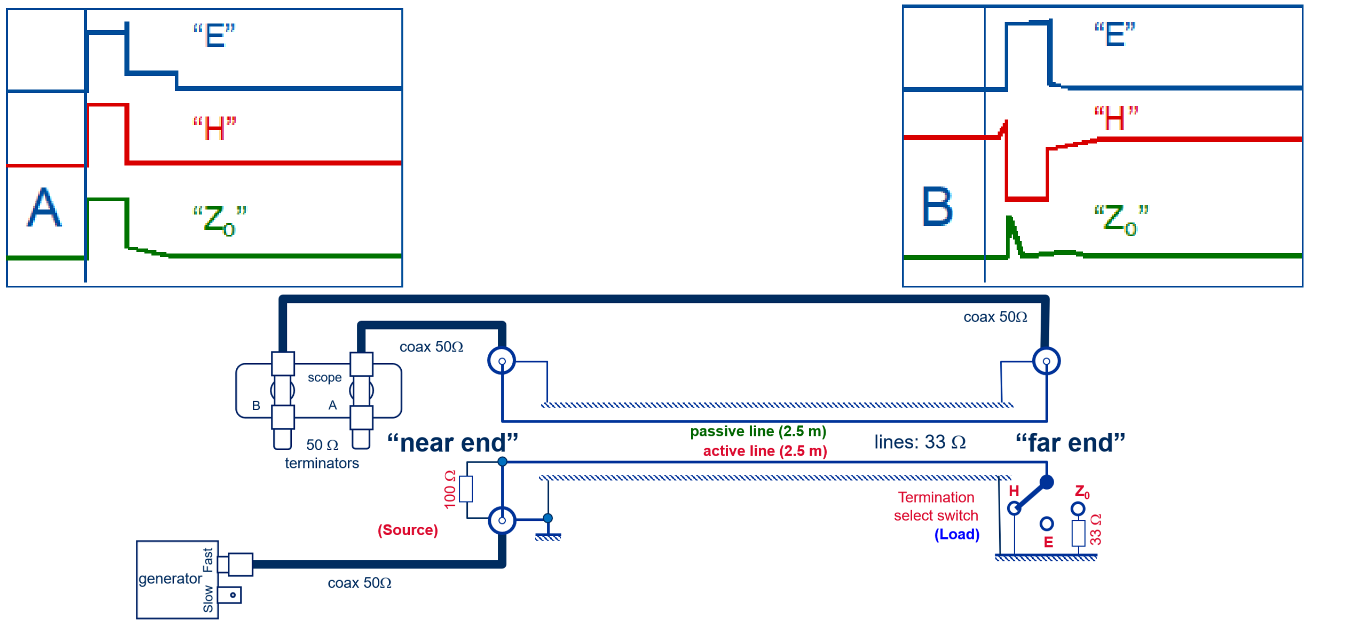
└Cross-talk experiment capacitive and inductive

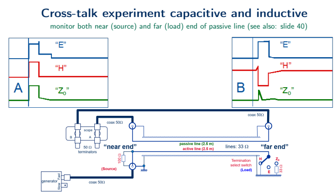


The setup of the experiment is identical to that in the basic EMC-module and shown in this slide.
The impedance of the STP lines with respect to the shield conductor is 33Ω.
At the end of the source line there is the familiar switch with three positions:
H: short circuit to the shield
E: open circuit
Z_0: a 33Ω resistor to the shield.
The impedance at the *near* end of the *source* line needs to be 33Ω to make it a characteristic termination which will not reflect whatever comes back from the *far* end. That is accomplished with a 100Ω resistor.
This will result in some reflection on the cable back towards the generator. At the generator that reflection is absorbed there and we will not see it.
The "weak spot" is the termination of the victim line: just 50Ω at either end of this 33Ω line. There will be some secondary reflection effects (Aspect 68).

Cross-talk experiment capacitive and inductive

monitor both near (source) and far (load) end of passive line (see also: slide 40)





The oscilloscope displays seen at the three switch positions show, that the two spikes in the same direction seen in the E and the spikes in opposite direction are actually *near* end cross-talk effects with a duration of two line delays.

Notice that all *far* end spikes are shifted one line delay, T with respect to the spikes at the *near* end. Interesting are the $2T$ crosstalk pulses at the far end in the E and H switch positions. This is also *near* end cross-talk, resulting from the transition on the *source* line which is ***reflected*** at the switch at the end of the *source* line (Aspect 69).

Then, finally, the switch position Z_0. At the *near* end of the *victim* line, we again see *near* end cross-talk. But since the source line is now characteristically terminated, there is only *far* end cross-talk at the *far* end.

This *far* end cross-talk is not zero but very short, τ_r.

The *far* end cross-talk is the difference $I_{C_{forward}} - I_{L_{backward}}$ shown in slide 41.

This could be zero if the shape in time and amplitude of the two components is identical.

The ratio of these two components is determined by the *geometry* of the two lines and the shielding over the length of the line and not by whatever impedance is placed at the *end* of the *source* line (Aspect 70).

Part 5: Crosstalk between cables

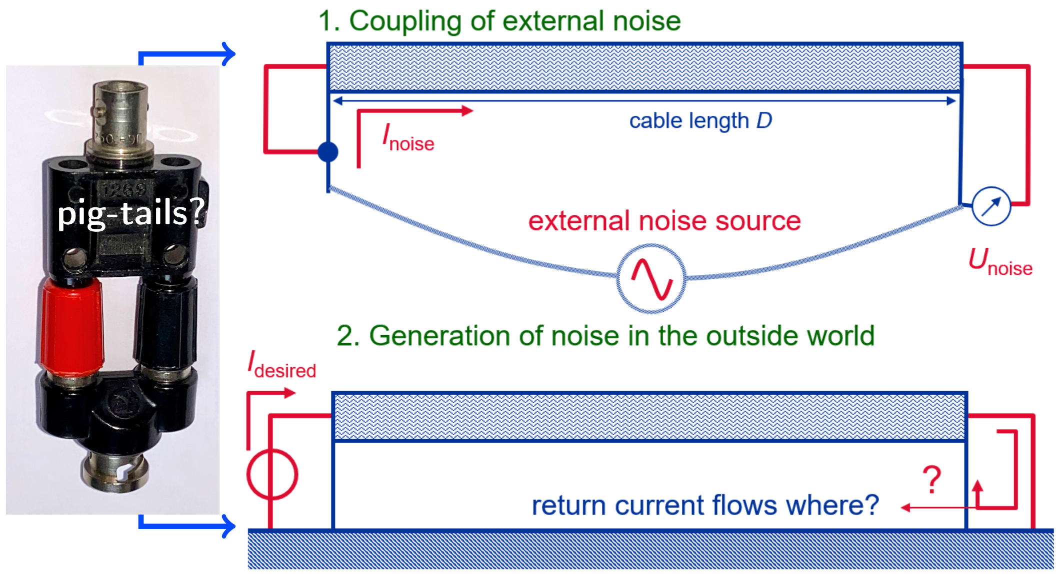
Transfer-impedance

Part 5: Crosstalk between cables

We have already seen cross-talk between two cables. The cause is transfer-impedance. We will analyze this phenomenon some more to get a better grip on what we can do to avoid this cross-talk.

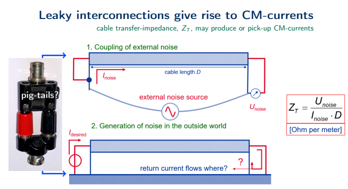
Leaky interconnections give rise to CM-currents

cable transfer-impedance, Z_T, may produce or pick-up CM-currents



$$Z_T = \frac{U_{noise}}{I_{noise} \cdot D}$$

[Ohm per meter]



In the basic EMC-module we have seen that cables can leak.
The process is called transfer-impedance, Z_T.
In the situation at the top of this slide, an external current flows over a cable shield.
If the cable leaks, this results in a voltage, essentially, over the shield:

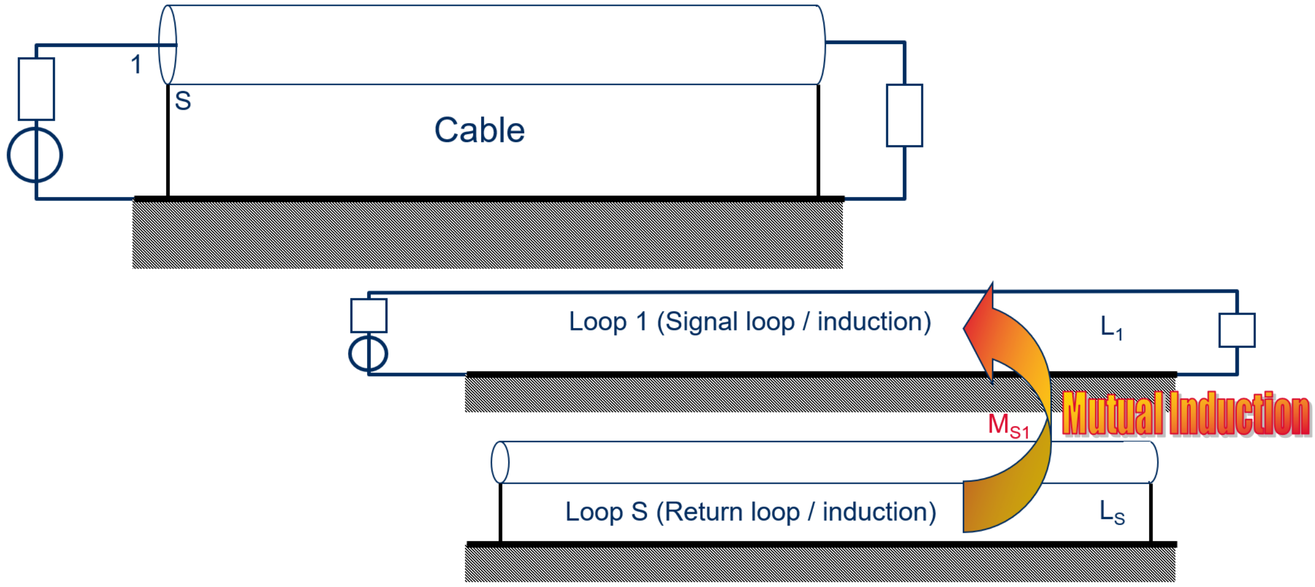
- caused by the external, common-mode (CM)-current flowing over it
- caused by the internal, differential mode (DM)-current flowing over it

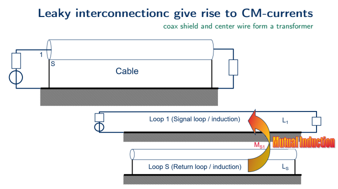
In addition to the leaking shield, there could be pig-tails involved, usually at one of the ends of the cable.

Note that "shield" could be "return conductor" e.g. in the unshielded twisted pairs (UTP)-case. The effect is called transfer-impedance because the noise voltage –U_{in} in the top-drawing– is caused by a current, I_{noise}, so the ratio manifests as an impedance.
The leaking shield –or other return– leaks *per meter*. For that reason, the dimension is Ω/m.
The cable at the bottom on the slide is a –possibly nasty– DM-current which is sent through the cable. Transfer-impedance here means that essentially a voltage is created by the DM-current between the ends of the shield –or other return– (Aspect 71). This in turn causes CM-currents in the outside world.

Leaky interconnectionc give rise to CM-currents

coax shield and center wire form a transformer

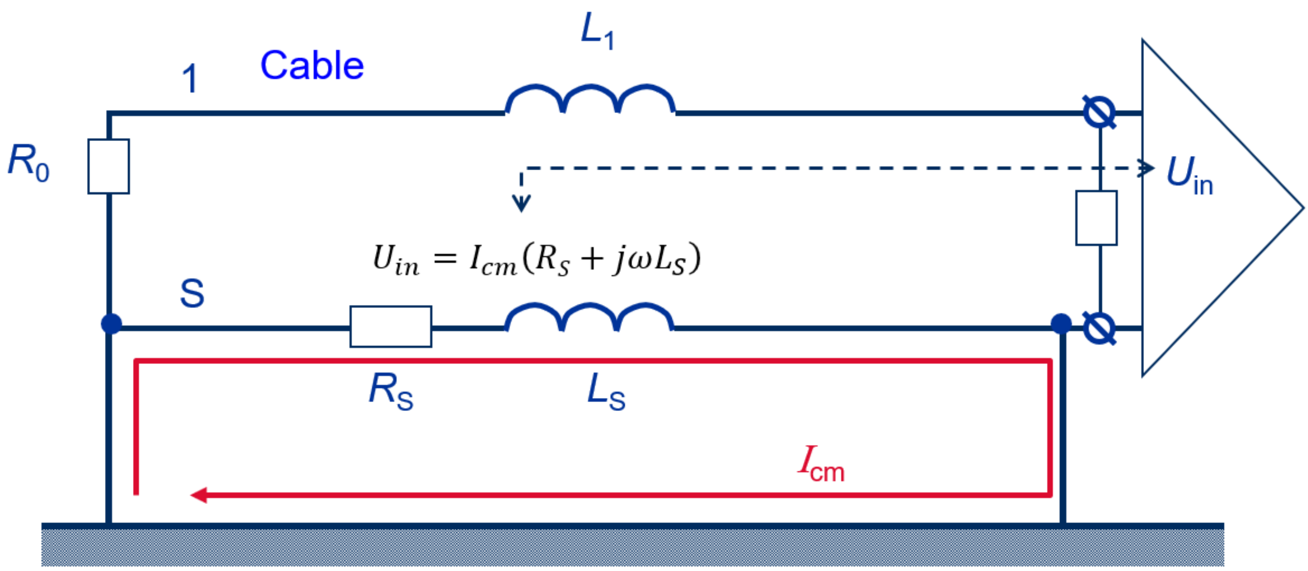




CM-current flowing over cable is *transferred into* a DM-“substitute” source in the center conductor in the first example on the previous slide.
To see how this works, the shielded cable –in this example– can be seen as two separate loops in combination with an external return path, loop 1 and loop S in the drawing in this slide.
These two loops are intimately coupled, –this is a good thing, they should be–, the shield loop current magnetic field is picked up by the signal loop through mutual induction M_{S1}.

Leaky interconnectionc give rise to CM-currents

mechanism of transfer-impedance; CM-current flows over shield impedance



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EMC for Engineers

2023-11-29

└Leaky interconnectionc give rise to CM-currents

If we, for the moment, disregard the mutual inductance between the shield and the center wire loops, we observe that the shield current, I_{cm}, flows over the shield impedance consisting of a resistor, R_S, with a series inductance, L_S. This means that, even if there is no mutual inductance, there is transfer-impedance as the current, I_{cm}, will induce a voltage which is added to the DM-voltage transported by the cable:

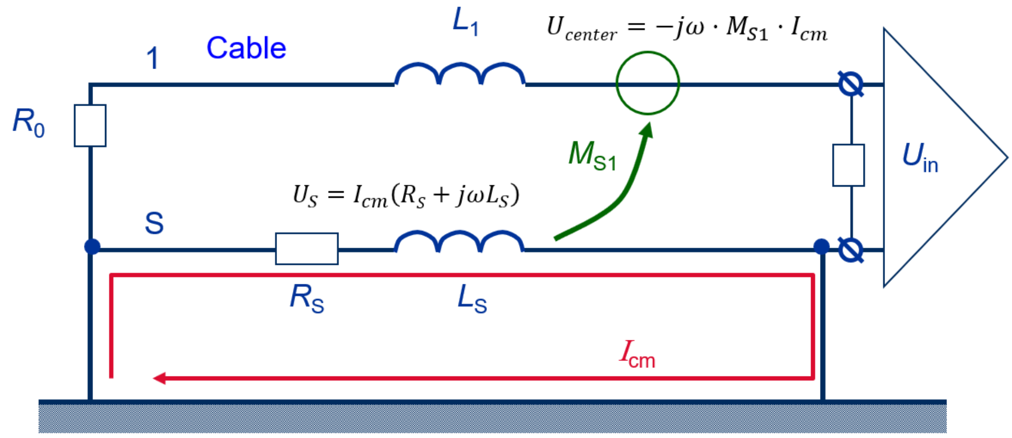
- Via Ohm's law: $I_{cm} \cdot R_S$
- Via the shield inductance: $I_{cm} \cdot j\omega L_S$

So, $U_{in} = I_{cm}(R_S + j\omega L_S)$
Both the R_S and L_S are *per meter* or per unit length (PUL)-parameters.

In this first approach we obtain, $Z_T = \dfrac{U_{in}}{I_{cm}} = R_S + j\omega L_S \; \dfrac{\Omega}{\mathrm{m}}$

Leaky interconnectionc give rise to CM-currents

mechanism of transfer-impedance; CM-current flows in external circuit



$$U_{in} = U_S + U_{center} = I_{cm}\big(R_S + j\omega(L_S - M_{S1})\big)$$

$Z_T = R_S + j\omega(L_S - M_{S1})\ \Omega/\mathrm{m}$
R_S, L_S and M_{S1} are
per unit length (PUL)-parameters.

Ohm's Law

Transfer Inductance (L_T) (Don White)

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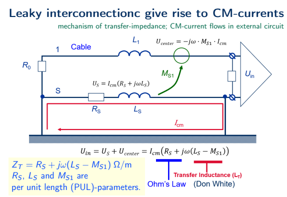
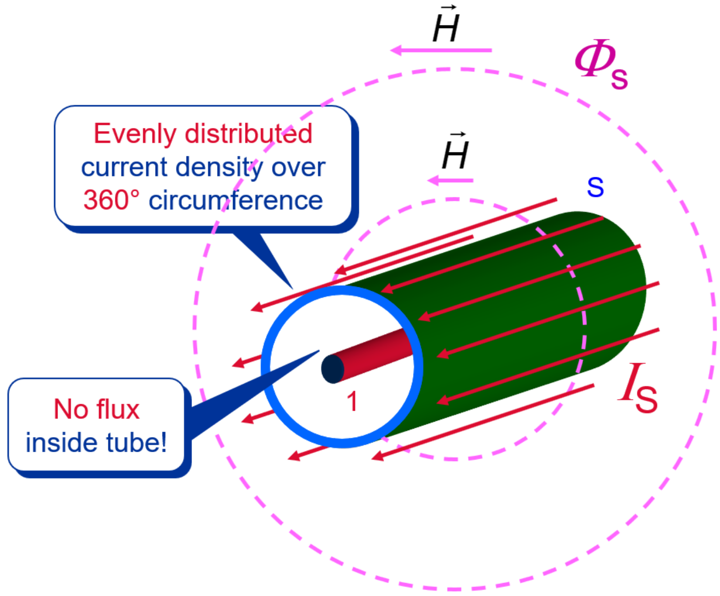


But, obviously, mutual inductance, M_{S1}, between the shield and the center conductor should be included in the analysis. The shield-loop wit I_{cm} induces a voltage $U_{center} = -j\omega M_{S1} I_{cm}$ V through Faraday's law. Notice this voltage component has a negative sign! So it must be subtracted from the voltage component, U_S, already found in the shield.
The noise voltage, added to the desired DM-voltage at the receiving amplifier at the end of the line now becomes: $U_{in}(j\omega) = U_S(j\omega) + U_{center}(j\omega) = I_{cm}(j\omega)\left(R_S + j\omega\left(L_S - M_{S1}\right)\right)$ V
Therefore the transfer-impedance becomes: $Z_T(j\omega) = \frac{U_{in}(j\omega)}{I_{cm}(j\omega)} = \left(R_S + j\omega\left(L_S - M_{S1}\right)\right)\ \Omega/\mathrm{m}$. The inductive element in the last equation, $L_S - M_{S1}$ is called the transfer-inductance, L_T, by the late Don White (1927-2017), a well known EMC-expert. Imagine the ideal case, where all magnetic fields produced by the shield-loop would couple into the center conductor-loop. Then the transfer-inductance, L_T, would become zero (0) –but remember possible pig-tails–. The only remaining element in the transfer-impedance would be the per unit length (PUL) shield-resistance, R_S. R_S can be reduced by the skin-effect, slide 51, so it becomes important at very high currents at low-frequencies as seen e.g. in lightning (Aspect 72).

Inside a metal tube with current there is no field

if this current is evenly distributed over the 360° circumference of this tube



$L_S = \dfrac{\Phi_S}{I_S}$

if all flux, Φ_S, is **outside**

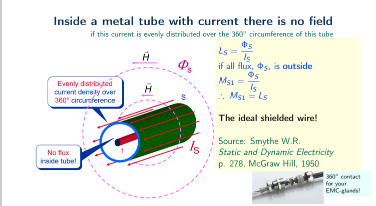
$M_{S1} = \dfrac{\Phi_S}{I_S}$

$\therefore\ M_{S1} = L_S$

The ideal shielded wire!

Source: Smythe W.R.
Static and Dynamic Electricity
p. 278, McGraw Hill, 1950

360° contact for your EMC-glands!

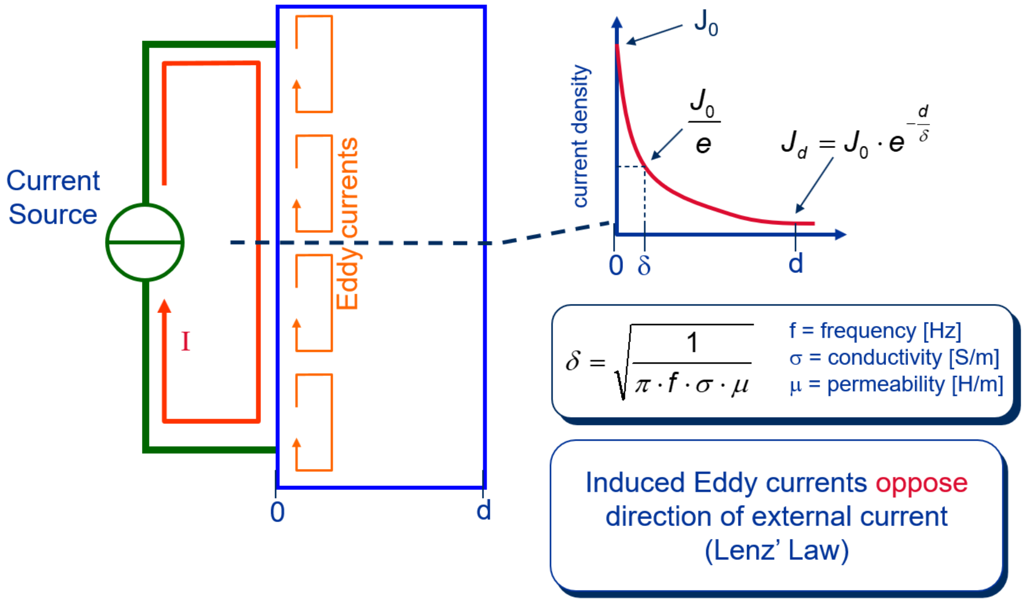


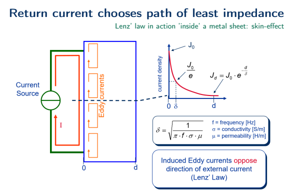
As shown in slide 49, it is theoretically possible that Dow White's transfer-inductance is zero. This would occur as soon as $M_{S1} = L_S$. In that case $Z_T(j\omega) = R_S$ i.e. frequency independent. Can it be achieved? In the book *Static and Dynamic Electricity* by W.R. Smythe, the author proves that if a current on a tubular conductor flows evenly distributed over its circumference –same current density over 360°– there is *no magnetic field –from the shield current– inside the tube*: Smythe's theorem. Adding a center conductor does not change that situation. In slide 47 the coax cable is shown as a transformer having two single turns: Loop 1 is the center conductor and external return path, shown as a return through "ground". Loop 2 is the tubular shield with this same external return path through "ground". Loop2 is the current-loop, I_S. This loop generates the flux Φ_S which –according to Smythe's theorem– exists only on the *outside* of the tubular shield.

This loop has an inductance $L_S = \frac{\Phi_S}{I_S}$. As all flux is outside the tube, the mutual inductance, the flux contained in the center-conductor loop divided by the current in the tube is identical: $L_S = \frac{\Phi_S}{I_S} = M_{S1}$. To achieve this uniform current distribution **connector/gland shells** shall **contact cable shields** over 360° (Aspect 73).

Return current chooses path of least impedance

Lenz' law in action 'inside' a metal sheet: skin-effect



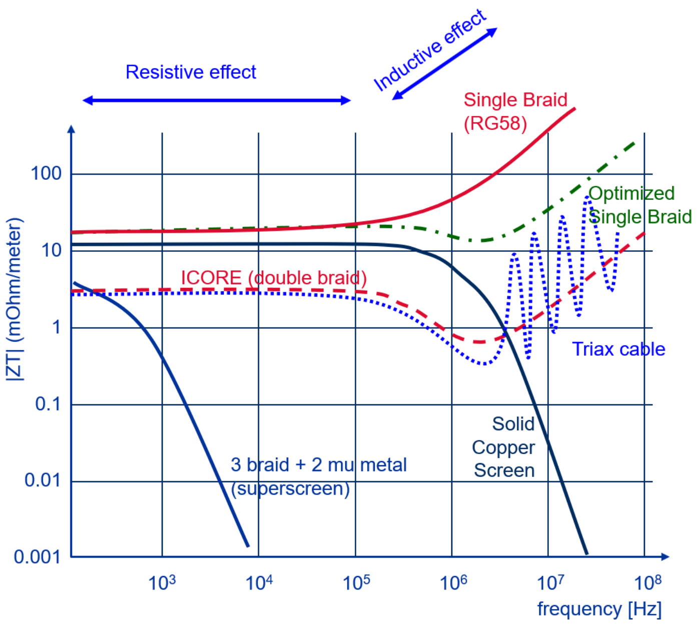


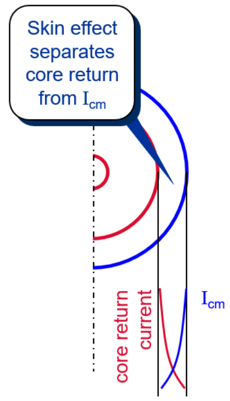
Slides 46 and 47 show the weak spot in the efforts to create ideal interconnections: their interface with the real world which can easily resemble *pig-tails*.
In addition we can ask ourselves what happens if an –intentional, DM– current flows on the center conductor and back through the shield? Well, there still is another mechanism that can help: the **skin effect**.
If an alternating current flows in a metal its field will induce, so called, Eddy currents in the metal which, due to the minus sign in Faraday's law, flow in the opposite direction of the current that generated the magnetic field.

This has the effect that the original current is pushed to the surface. The graph in the top-right hand corner shows the resulting exponential decay of the current density, J_0, with depth into the metal. The depth, δ, at which the current density has decayed to $J_\delta = e^{\frac{J_0}{e}}$ is called the skin depth. The ultimate effect is that the original current density is reduced to $J_d = J_0 e^{-\frac{d}{\delta}}$ at the opposite side of the metal –depth d–.
This skin-effect is the main component of the shielding effect of all-metal shielding walls including cable braids and tubes (Aspect 74).

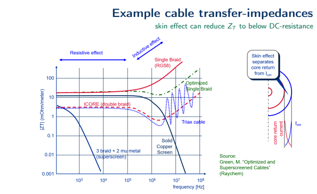
Example cable transfer-impedances

skin effect can reduce Z_T to below DC-resistance





Source:
Green, M. "Optimized and Superscreened Cables" (Raychem)

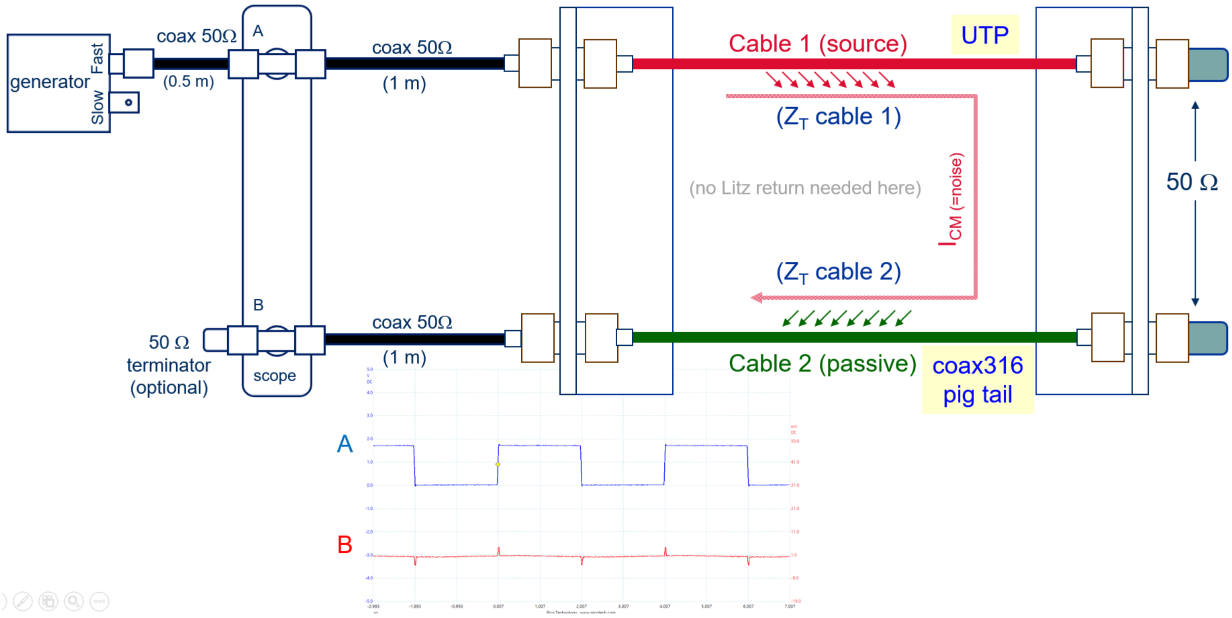


This slide shows examples of cable transfer-impedances.
At the top a single braid coax, like the RG-58 type used for experiments in this workshop.
At the left side of the graphs the shield resistance, R_S is shown –the horizontal part– e.g. $R_S = 22$ mΩ/m for the RG-58.
For most cables the inductive effects set in around 1 MHz. In the RG58 case, the curve slants up proportional to frequency –note the double logarithmic axes–. At the same level as the RG58 we see the optimized single braid that starts out at the same 22 mΩ/m but already shows some skin-effect –the downward dip before the slanting up proportional to frequency–.

More pronounced we see skin-effect in a double braid example, e.g. the dashed red line with the name ICORE. One of the effects of the double braid is the much lower shield resistance, $R_S = 3$ mΩ/m here. Notice that the skin effect is the only mechanism that allows the transfer-impedance (Z_T) to dive *below* the shield resistance! The skin effect separates the shield internal from the shield external currents as shown in the diagram at the right hand side.
The blue dotted line with resonances at high frequencies is a triax cable. This appears to be a measurement error as the insulated braids were shorted at the cable ends. The resulting shorted transmission-line resonates.

CM-currents run in loops too –see slide 12–

cable cross-talk similar to wire cross-talk due to transfer-impedance



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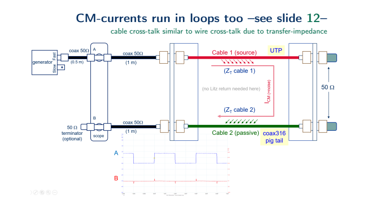


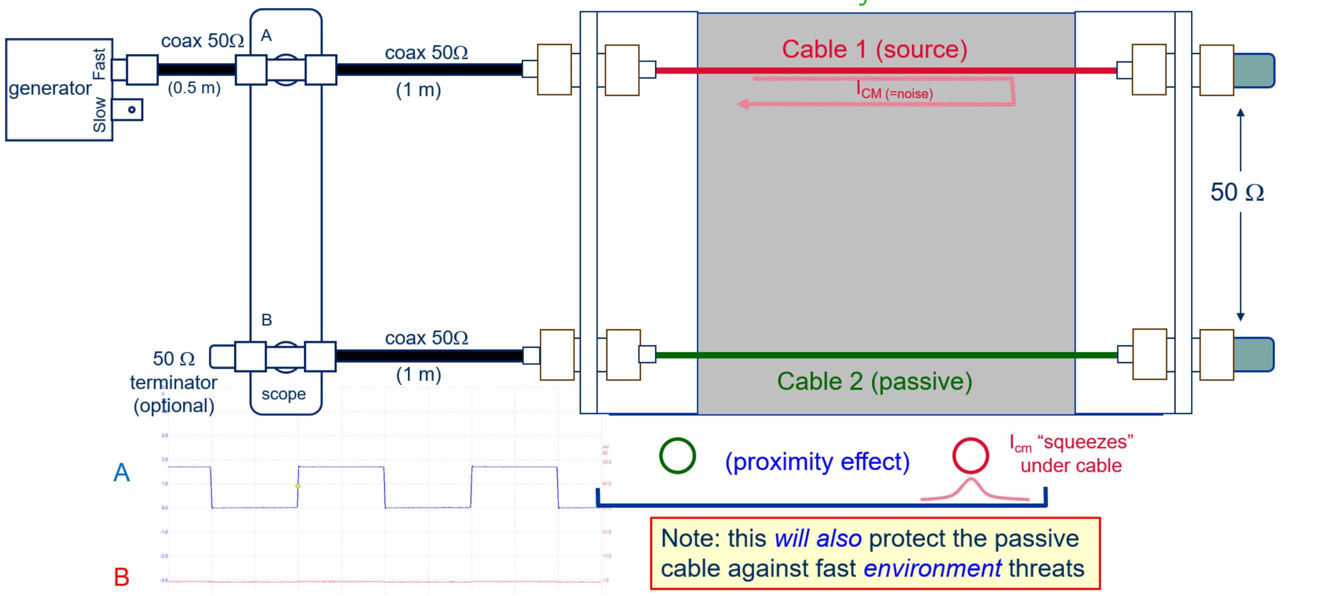
As we have seen in the basic module, cable cross-talk via transfer-impedance looks just like the cross-talk seen for single wires.
The only difference in the set up is the missing separate return conductor of the wire case as this return is built into the cables.
The explanation in the slide uses the CM-current produced by the source wire which flows over both cables and their connecting current boundaries (CBs).

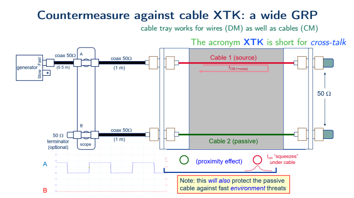
The victim cable will pick-up this CM-current through its own transfer-impedance.
Another metaphore would be that *magnetic fields* leak from the source-cable and *induce* a CM-current in the loop formed by the source and victim return conductors.
Separating the cables does not help to diminish the cross-talk as the I_{CM} will follow the path of least impedance which is through the cable-returns (Aspect 75).

Countermeasure against cable XTK: a wide GRP

cable tray works for wires (DM) as well as cables (CM)

The acronym **XTK** is short for *cross-talk*





Using a wide ground reference plane (GRP) does help.
The CM-current now has a choice: the other cable or the GRP.
Whichever conductor is closer will have –most of the– the CM-return current.

By placing the cables
1. against the GRP
2. separated by ample[9] distance

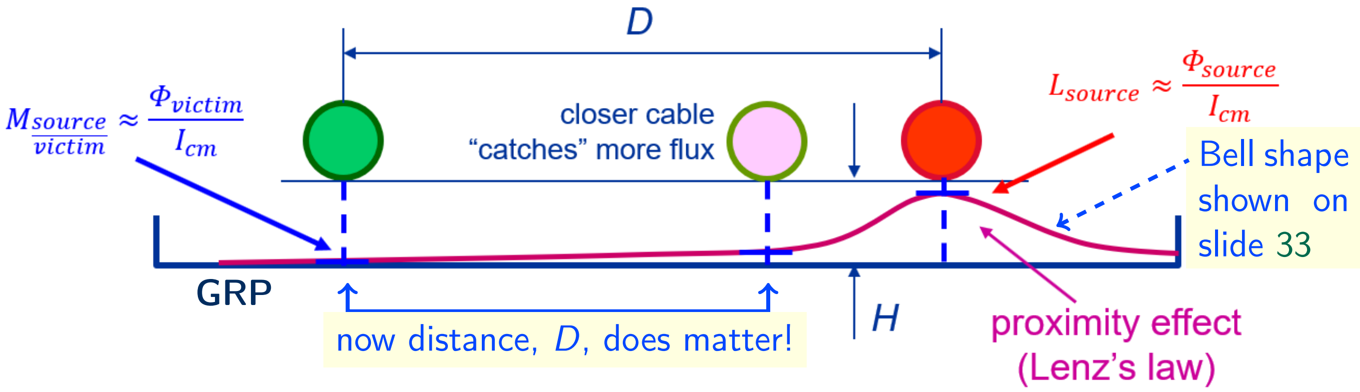
we force the return current to select the GRP.
This is an example of "Design for EMC".

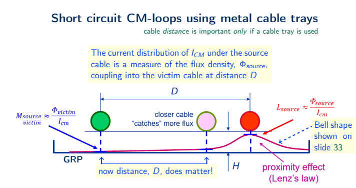
[9] If no other information is available, the maximum possible distance over the width of the CB could be chosen. Otherwise estimates based on DM-current in the noisy cable, its Z_T, the required reduction or permissible mutual induction, see slide 55. You will also need the Z_T of the sensitive cable and the acceptable noise signal component there.

Short circuit CM-loops using metal cable trays

cable *distance* is important *only* if a cable tray is used

The current distribution of I_{CM} under the source cable is a measure of the flux density, Φ_{source}, coupling into the victim cable at distance D

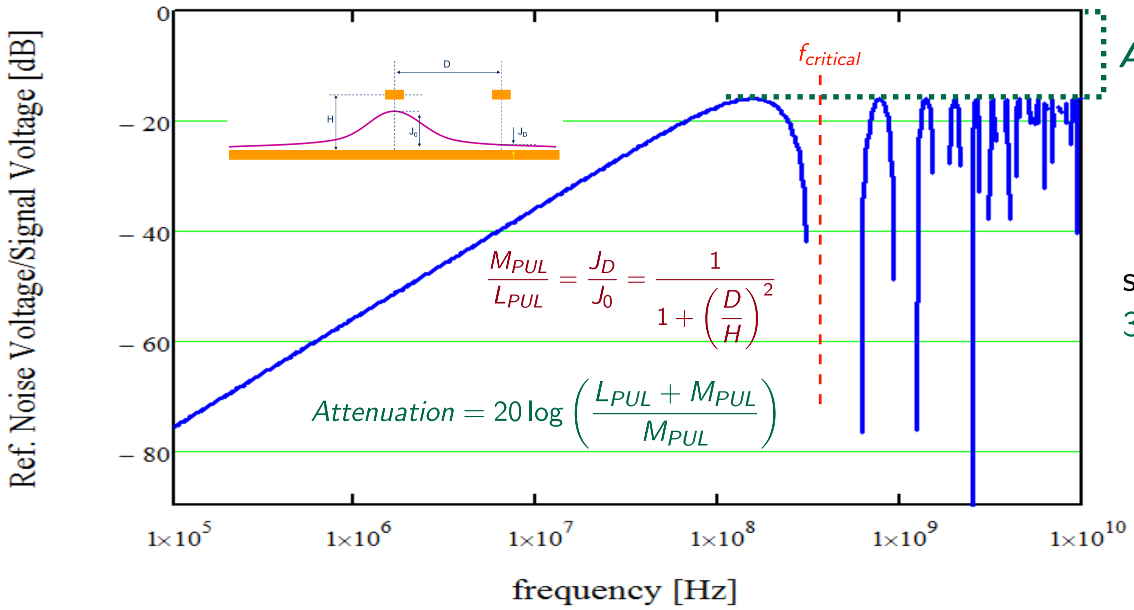




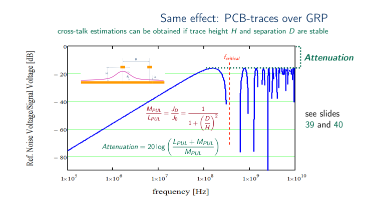
The wide metal cable-tray, or GRP, is actually a short-circuit of the source-cable shield *substitute* voltage source that generates the CM-current in this cable shield and in the connected wide metal strip.
Obviously, this short-circuit only works if the tray is connected –low-impedance– to the CB-strips with the connectors feeding the cables.
As soon as the cable-tray is connected and the cables are laid on it over their full length, the tray will serve as a return path for the generated CM-current.
The CM return-current will concentrate under the source-cable and the current density over a cross-section perpendicular to the cable will look like the graph shown on slide 33.
The current density in the GRP directly under the source cable, $D = 0$, corresponds to the magnetic flux, Φ_{source}.
Directly under the victim cable, $D = D$, this flux has come down to Φ_{victim}.
Dividing both fluxes by I_{CM}, we obtain L_{source} and $M_{\frac{source}{victim}}$.
Note that both L_{source} and $M_{\frac{source}{victim}}$ are per unit length (PUL) parameters i.e. (H/m)!
Bottom line: a cable tray reduces the cross-talk between cables by short-circuiting CM-currents from noisy cables and reducing the loop with sensitive cables that pick-up noise through mutual induction (Aspect 76).

Same effect: PCB-traces over GRP

cross-talk estimations can be obtained if trace height H and separation D are stable



see slides 39 and 40



The CM-cross-talk behavior of cables in a metal tray corresponds to the cross-talk behavior between two parallel traces on a printed circuit board (PCB). As discussed on slide 40, cross-talk will increase proportional to frequency until a ceiling level –asymptote–. The frequency at which this ceiling is reached is half the critical frequency, pointed out on slide 39. The ceiling level is *Attenuation* dB below the signal level on the source trace. This level is determined by the per unit length (PUL) induction, L_{PUL}, of the PCB-traces and the PUL mutual induction between the PCB-traces, M_{PUL}. The ceiling –*Attenuation*– is only determined by the geometry of the traces and the GRP over which they run. Note that if the *trace length is increased*, the *same* ceiling will be reached at a *lower frequency*. So, PCBs-trace-GRP combinations can be designed to have a minimum *Attenuation* , independent of their length. For digital logic -20 dB may be acceptable. Analog and mixed systems may need 30 dB or more (Aspect 77). Note that in the cables-in-the-tray situation both the parameters height-over-metal, H, and separation, D, are usually ill defined while cables will have varying diameters. Cables in trays are laid against the metal –ty-raps could be used[10]– and established source lines kept as far as possible removed from their victims.

[10] Modern cable trays are perforated.

Part 6: Current boundaries

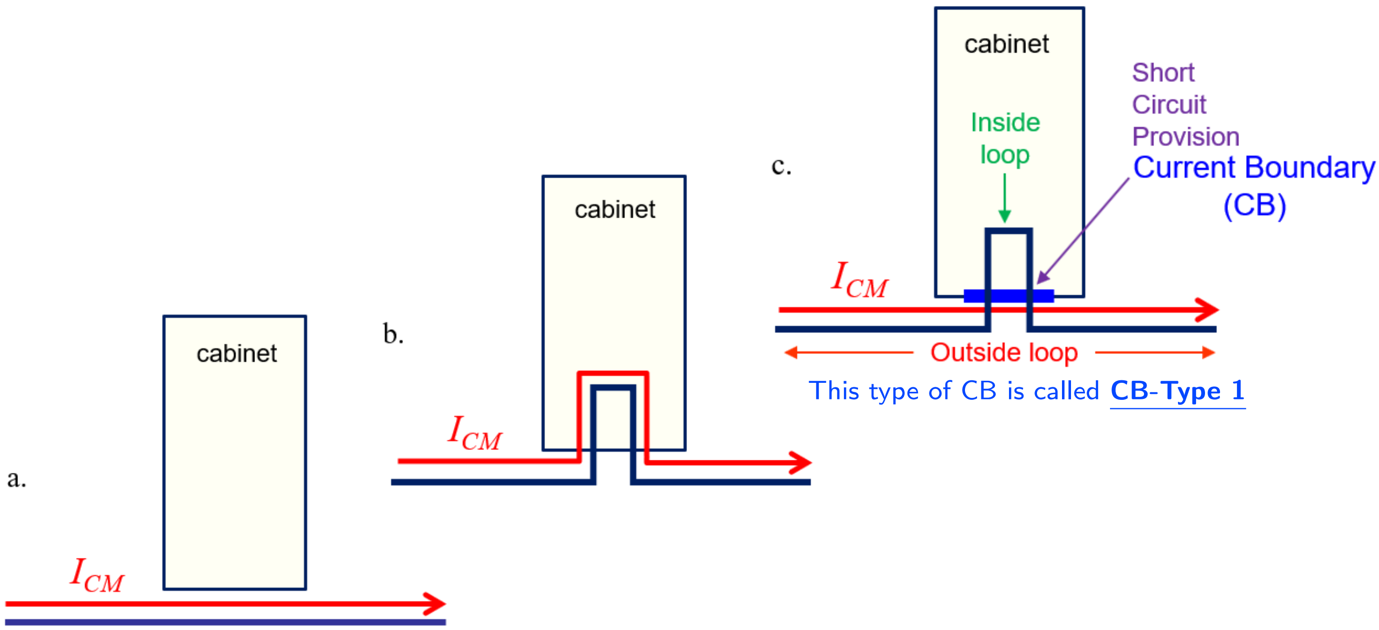
Short circuit CM-currents

Part 6: Current boundaries

In the part on signal integrity we have seen that if clean signals arrive at the end of an interconnection this means that little energy is lost "en route".
If signal energy is lost this usually happens via transfer-impedance and CM-currents occur on the cable shield-returns. These can then cross-talk to other cable shield-returns which transfer the CM-currents to DM-voltages that add to the signal transported by that cable.
The trick to avoid these effects is to short-circuit the CM-currents as soon as they are generated.

Electronic assemblies need current boundaries

every electronic circuit –large or small– should have adequate protection



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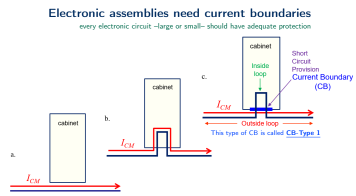
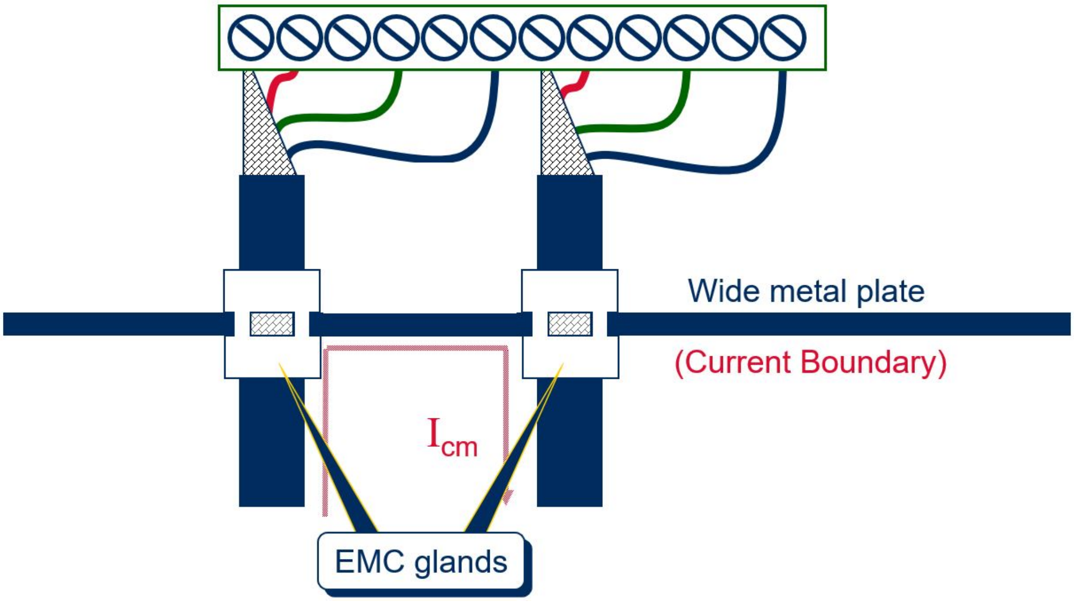
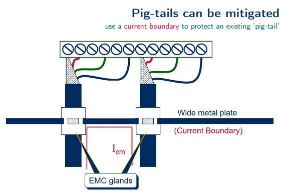


Cables may generate –or pick-up– EMI by CM-currents through transfer-impedance.
This makes it necessary to provide type-1 current boundaries (CBs) on equipment interconnected by those cables.
These CBs prevent these CM-currents from flowing into equipment cabinets and possibly creating EMI there.
The CBs short circuit CM-currents and do not affect the DM currents transported by the cables.
The principle of the CB was demonstrated in the basic EMC module.

Pig-tails can be mitigated

use a current boundary to protect an existing 'pig-tail'



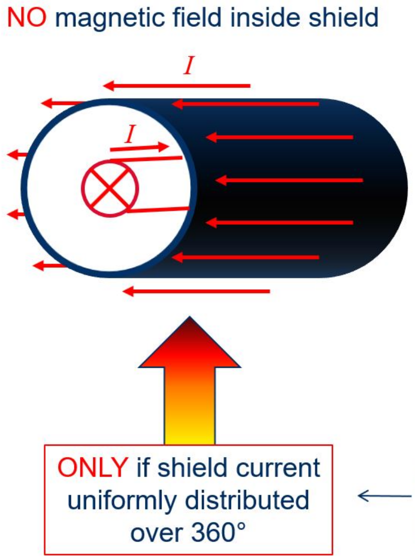


Also shown in the basic EMCmodule was the fact that CBs work both ways:
1. They keep external CM-currents out;
2. They keep internal CM-currents in.

The second, internal CM-currents, are often generated by equipment-internal *pig-tails* e.g. cable-core strips on which originally shielded cable cores are connected.

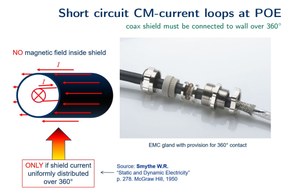
Short circuit CM-current loops at POE

coax shield must be connected to wall over 360°



Source: **Smythe W.R.**
"Static and Dynamic Electricity"
p. 278. McGraw Hill, 1950

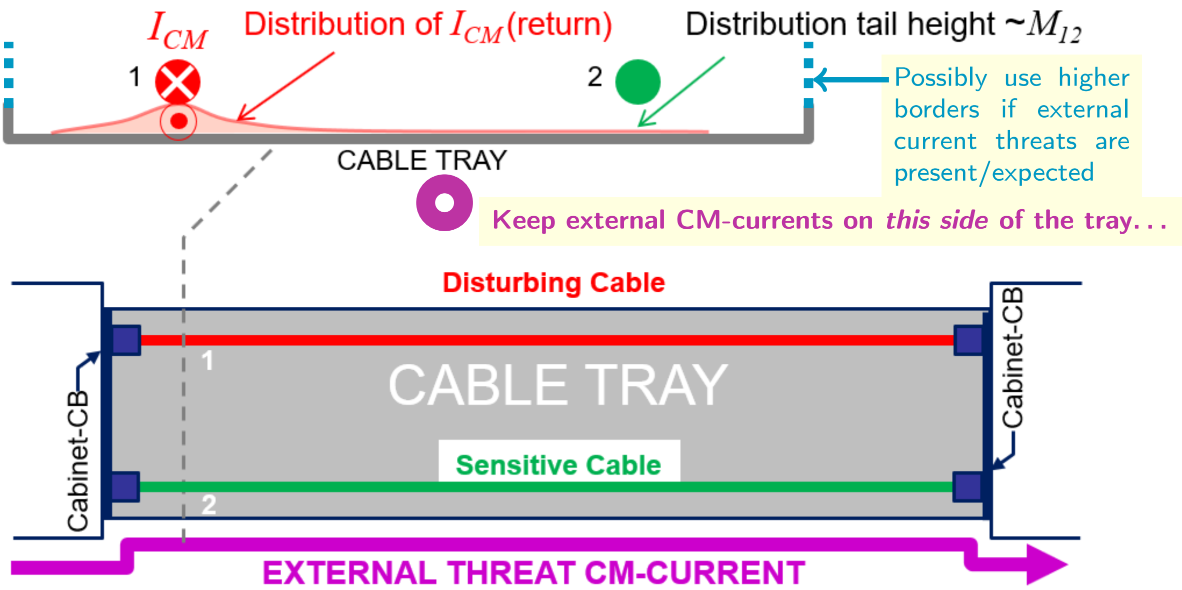
EMC gland with provision for 360° contact

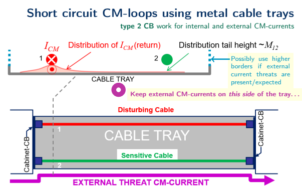


The EMC-glands are used to connect the cable shield to the metal gland-plate.
As we have seen in slide 50, it is important to connect the shields of cables to the current boundary (CB)-metal over the full 360° otherwise the cable transfer-impedance may seriously suffer.
EMC-glands have a provision to make this 360° connection.
But to make this work, the insulation around the cable shield shall be removed at the exact location where e.g. the metallic spring –see picture top-right– will contact.
The rest of the insulation stays intact as these EMC-glands also provide water-tightness.
Metal connector shells should be used to connect the cable-shields to the –in this case– connector plate that serves as CB.
As for the EMC-glands, EMC-connector shells also provide water-tightness.
Correct connection of the shield to the connector shells without compromising water-tightness should be easier as a connector is usually at the end of a cable.
But always remember **Smythe's theorem** and connect the cable shield directly to gland or connector over 360°.

Short circuit CM-loops using metal cable trays

type 2 CB work for internal and external CM-currents



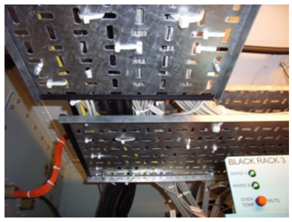


We have already shown the ability of a cable tray to protect a sensitive cable against CM-currents generated by a disturbing one.
These type-2 CB as these cable trays are labeled also protect all cables inside them against external CM-currents from the environment.
But try to keep the external CM-currents – lightning etcetera– on the opposite side of the cable-tray with respect to the system-=internal cables to benefit from the skin-effect (Aspect 78).
The connection with the CBs on either side should be very low impedance: think of lightning currents passing over the metal of the tray.

Equipment cabling: source or victim of CM-loops

use structural metal and/or walls as cable tray or GRP

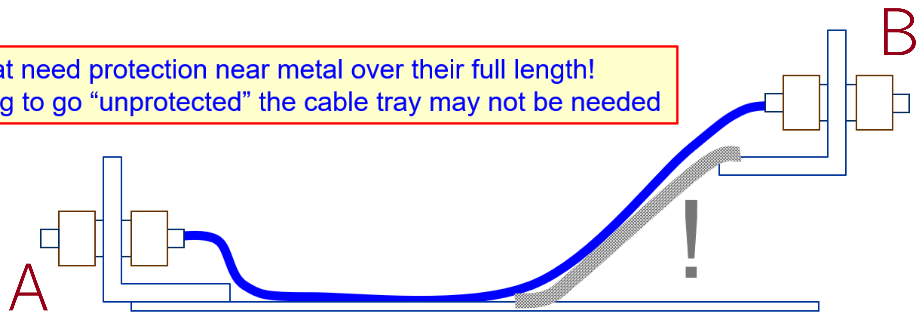
PROTECTED

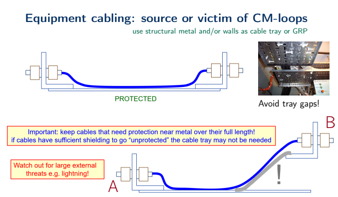


Avoid tray gaps!

Important: keep cables that need protection near metal over their full length!
if cables have sufficient shielding to go "unprotected" the cable tray may not be needed

Watch out for large external threats e.g. lightning!

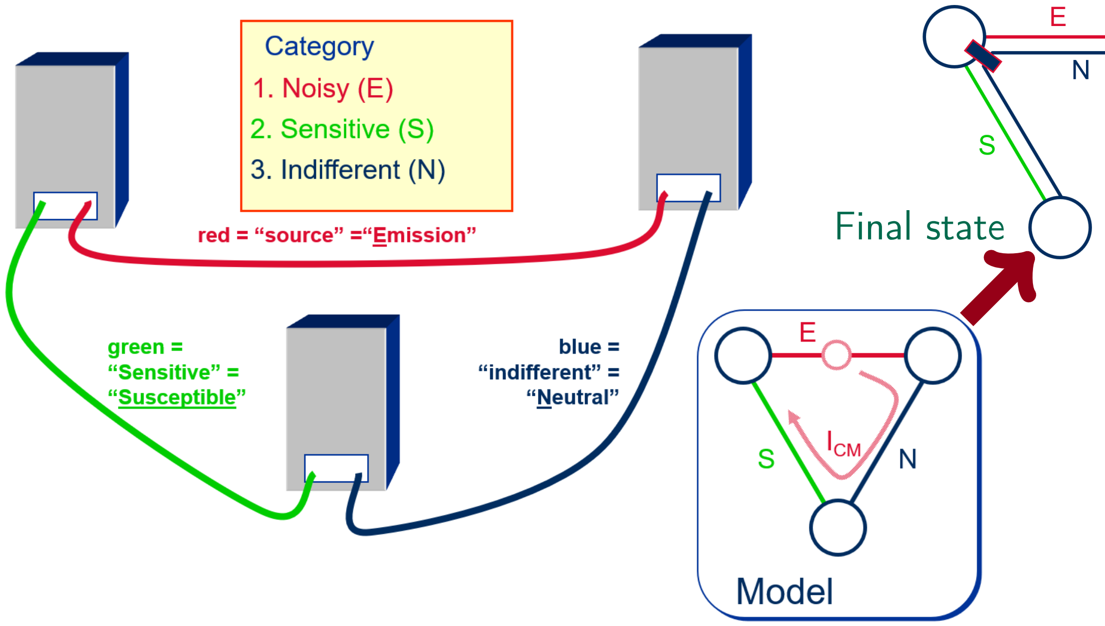




In the cable cross-talk experiment, we placed the source- and victim-cable against the metal of the cable tray. And noticed that, when we lifted a cable, cross-talk would increase.
The loops formed by cables and tray need to stay as small as possible for both source and victim cables.
In installations e.g. on ships, equipment is often placed on shock absorbers. Make sure that, if a cable tray in needed, it should follow the cables from equipment-A to equipment-B.
With shock absorbers, these cable-tray extensions should –of course– be flexible.
The cable-trays should be of one piece –at least electrically–. If separate tray sections need to be interconnected, make sure there is no paint or other insulator in the junction.
If trays need to be painted, first join them together and check the resistance with a milli-Ohm-meter, then paint.
Of course, gaps are disastrous: check the picture in the top right of the slide. Please note that a simple wire strap cannot repair this error (Aspect 79).

Categorize cables to detect sources and victims

use current boundaries to separate cables after classification into categories



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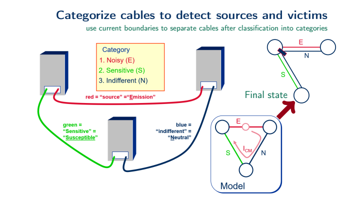


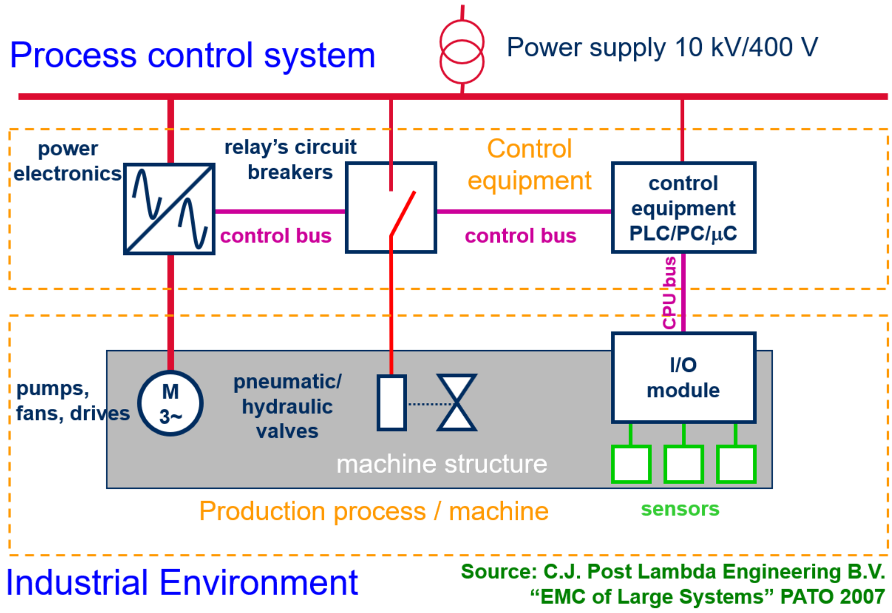
EMC is all about separating sources and victims. So it is important to find them in the first place.
For that reason we can start by categorizing our interconnections. Cables in the interconnected equipment situation. But you can do this on any scale e.g. by categorizing traces on PCBs or even integrated circuits (ICs).
The simplified three category version shown in this slide is used in the basic EMC-module to show the approach to find the *critical* CB, located here at the cabinet where both the Noisy (E) and Sensitive (S) cables are connected. The approach works for any 2 adjacent categories. But an Indifferent or Neutral (N) cable will never be used to separate (E) from (S) (Aspect 80).
Official standards are more carefull and call our (N) "potentially disturbing" i.e. not suitable as protective GRP (type-2 CB). In practice GRPs should be used –strip, tube– **OR** adequately shielded cables.

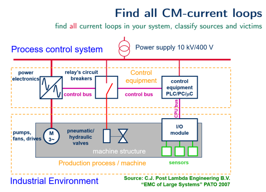
Example: Cable Categories in the IEC 60533:

3	Extremely sensitive	4	Extremely disturbing
2	Sensitive	5	Special
1	Potentially disturbing		

Find all CM-current loops

find all current loops in your system, classify sources and victims



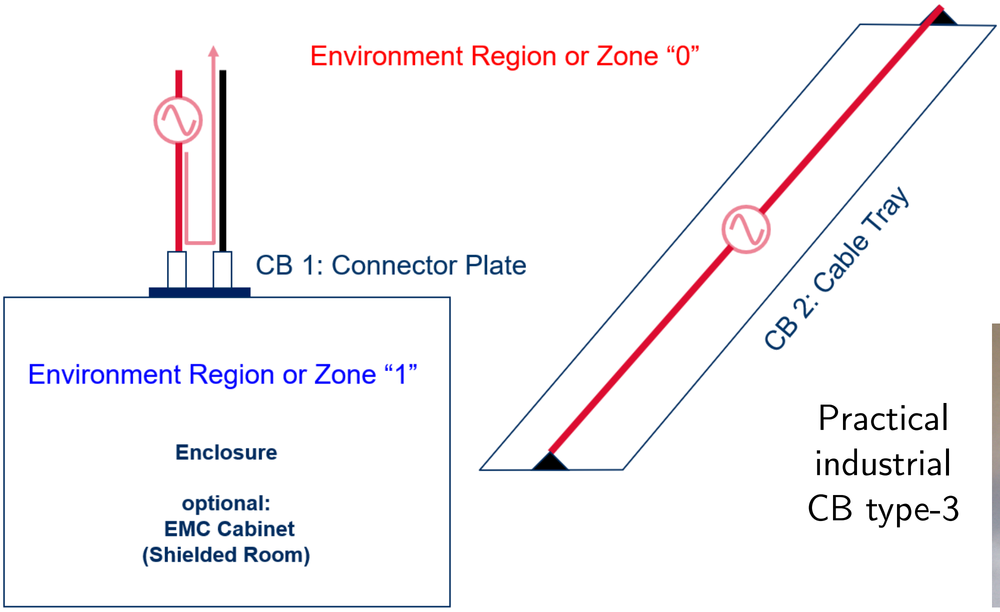


In a real system there are many potential CM-loops.
You will have to find them and categorize all cables.
Then you should either find adequately shielded cables or find metal structures to use as cable-trays.
Any metal can be used as long as the complete cable path is against a continuously galvanically connected metal structure. Be sure to also give proper attention to the connection of all type-1 CB to this type-2 CB structure –wires are inappropriate–.
A metal GRP under[11] the whole system is an option for smaller installations, a metal maze structure is the next best alternative.
Most technical installations have sufficient structural metal that can be used to protect disturbing or sensitive cables –but keep these two separated–.
One way to separate them is to use the skin-effect: use a metal strip *between* a sensitive and a disturbing cable.

[11] A GRP as a floor implies all cables run on the floor. A ceiling GRP or grid could solve this. In that perspective the type-1, *connector-plates* on cabinets are logically moved to the tops of the cabinets.

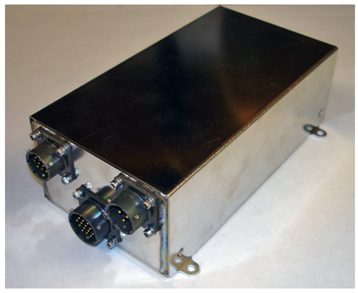
Dominant EM-fields: use type 3 CB (shielding)

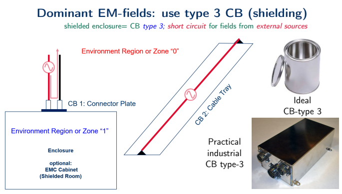
shielded enclosure= CB *type 3; short circuit* for fields from *external sources*



Ideal
CB-type 3

Practical
industrial
CB type-3





Current boundaries (CBs) as discussed so far are useful for interconnection/cable-related or *conducted* interference. CB type-1 –the connector plates– serves to short-circuit the CM-currents from all cables connected to cabinet or module in an installation. If all cables are connected to a single connector/gland plate, the cabinet itself does not have to be a shielded enclosure. CB type-2 –cable trays– serves to separate sensitive cables from noisy ones. Often, the installation metallic structure can be used as CB type-2. Then, finally, for environments polluted with EM-fields with half-wavelengths smaller or in the order of the size of the installation, we need to think of actually shielding equipment. This is CB type-3. Ideally a gas-tight metal container e.g. the paint can shown at the top right. But gas-tight metal is not practical for electronics as these need to communicate with devices in the outside world and electric power. The CB type-3 shown at the bottom right of the slide is more practical. It has a CB type 1 with connectors. This leads us to a new aspect: zoning or the definition of regions. The inside of a type-3 CB is an EM environment that is *different* from the environment around it. The outside, region 0, is always there. From there inward we find region-1, -2 and so on. This implies a hierarchy of CB type-3 (Aspect 81).

Part 7: Achieving EMC at the system level

Frame Title

Part 7: Achieving EMC at the system level

The next step is to apply the rules we found and obtain systems EMC.

–> Use **no high frequencies**
if you do
–> **Do not transport** them
if you do
–> **Use adequate current boundaries (CBs)**

Fundamental rules of electromagnetic compatibility (EMC)
high frequency and transport to be interpreted as wavelength >> distance
–> Use no high frequencies
if you do
–> Do not transport them
if you do
–> Use adequate current boundaries (CBs)

Achieving electromagnetic compatibility (EMC) in a system, in principle is simple:
1. Use no high frequencies
But we want to build faster systems so we run into the next rule:
2. If you do use high frequencies: do not transport them
This dictates to keep high frequency interconnections extremely short e.g. place the frequency converter *on* the motor it is controlling. But there can be many reasons why this is not possible in a complex system.
3. Then, if you have to use longer interconnections, use adequate current boundaries (CBs)
There are numerous reasons to end up in rule 3: convenience, reliability, availability and maintainability (RAM), thermal, environmental...
So the bottom line is: try to stick to these rules by dividing your system into EMs-*independent* modules. Restrict your fast electronics to the smallest possible volume. Try to stick to the low-frequency approximation –keep interconnection lengths below the local shortest $\lambda/20$. Try to stick to this rule as the sizes get larger. This means the frequencies go down as the environment gets larger.
EM-independent modules further allow you to build better versions later that can simply replace them.

Use 'loosely (EM) coupled coherent modules†'

independent hierarchy of building blocks with 'abstract' behavior

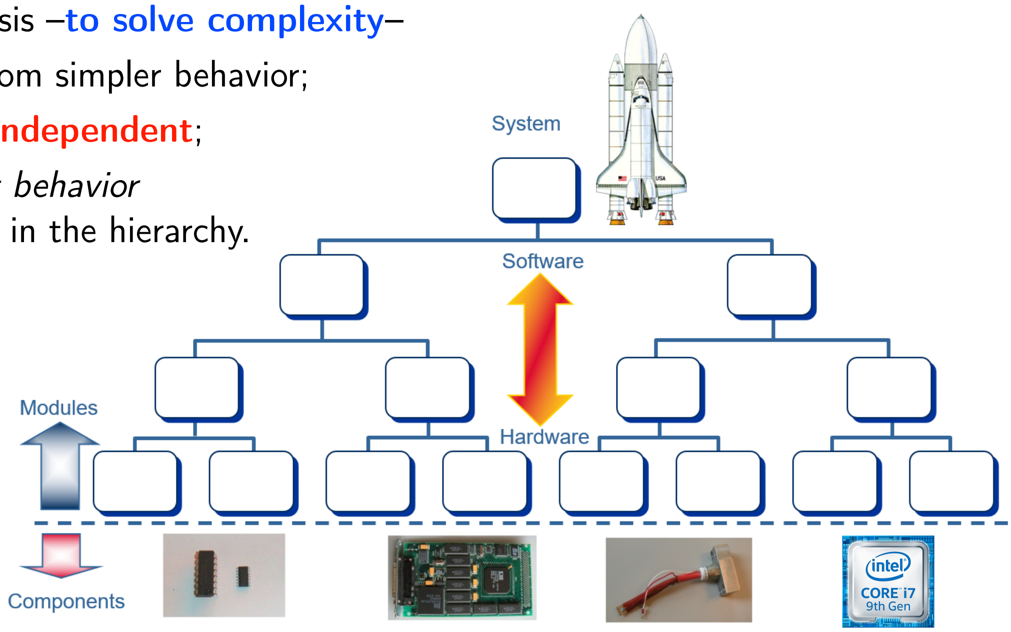
Object Oriented Analysis –**to solve complexity**–

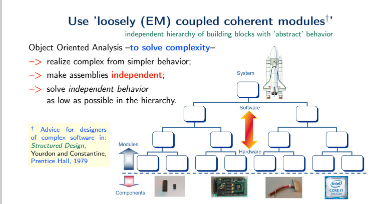
–> realize complex from simpler behavior;

–> make assemblies **independent**;

–> solve *independent behavior*
as low as possible in the hierarchy.

† Advice for designers of complex software in: *Structured Design*, Yourdon and Constantine, Prentice Hall, 1979

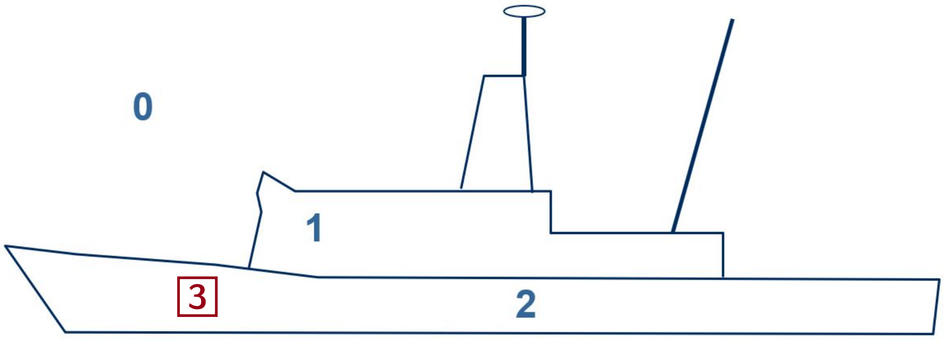




Electromagnetic interference (EMI) for larger installations becomes a problem of complexity. The idea is that if you work on a project with many people, the size of the problem is no doubt more complex than one engineer can completely cover. This is called the *semantic gap*. The solution is *hierarchy* and *abstraction*. Hierarchy means complex systems are decomposed into layers of less complex *independent* modules ultimately ending at components and assemblies available on the market. The other aspect is abstraction. Abstraction implies that the designer uses modules from a lower level based exclusively on that module's specifications. The modules are complex and the designer has no access to/cannot comprehend all internal details. In the light of our EMC dialogue, this abstraction implies that the hardware modules are electromagnetically compatible: EMC. This means the individual modules will not influence each others behavior i.e. they will behave according to their specifications at all times. The way to achieve this is the hierarchical application of current boundaries (CBs), starting at the IC level up to the level where software takes over the behavior. For more information: see *Reliable Systems Design using Current Boundaries*, IEEE-explore, DOI 10.1109/MEMC.2017.7931984. and a summary of Yourdon's book at https://www.win.tue.nl/~wstomv/quotes/structured-design.html

Regions are defined (specified) EM-environments

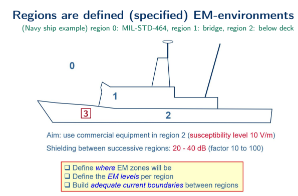
(Navy ship example) region 0: MIL-STD-464, region 1: bridge, region 2: below deck



Aim: use commercial equipment in region 2 (susceptibility level 10 V/m)

Shielding between successive regions: 20 - 40 dB (factor 10 to 100)

- ❑ Define *where* EM zones will be
- ❑ Define the *EM levels* per region
- ❑ Build *adequate current boundaries* between regions



One of the first steps in the hierarchical approach is the definition of electromagnetic (EM)-regions or zones.
EM-Regions are environments with specific levels for EM-phenomena.
The specification of these regions is the responsibility of the systems-engineer.
The outside world, i.e. the environment in which the system is supposed to operate according to its specifications but on which the designer has no influence whatsoever, is always Region 0. Region 0 will always be there regardless of further defined –system internal– regions.
The slide shows a navy ship as an example.

The ship's bridge area is closest to region 0 and is labeled region 1.
The ships construction may have some shielding effects making region 1 slightly less challenging than region 0. Deeper down in the belly of the ship is region 2. Inside region 2, there could be e.g. a radio hut, defined as region 3. If more regions 3 are defined, there needs to be specified whether they should be mutually shielded.
The slide specifies the situation where commercially available equipment (CAE) for the industrial environment can be used in region 2 without modifications.

MIL-STD-464 specifies military environments

it is up to the manufacturer to make equipment compliant in its environment (region 0)

DEPARTMENT OF DEFENSE
INTERFACE STANDARD

ELECTROMAGNETIC ENVIRONMENTAL EFFECTS (E^3)
REQUIREMENTS FOR SYSTEMS

Topics
- fieldstrengths
- lightning
- EMP
- ESD
- ordnance
- HERO
- electrical bonding

TABLE 4. Maximum external EME for ground systems.

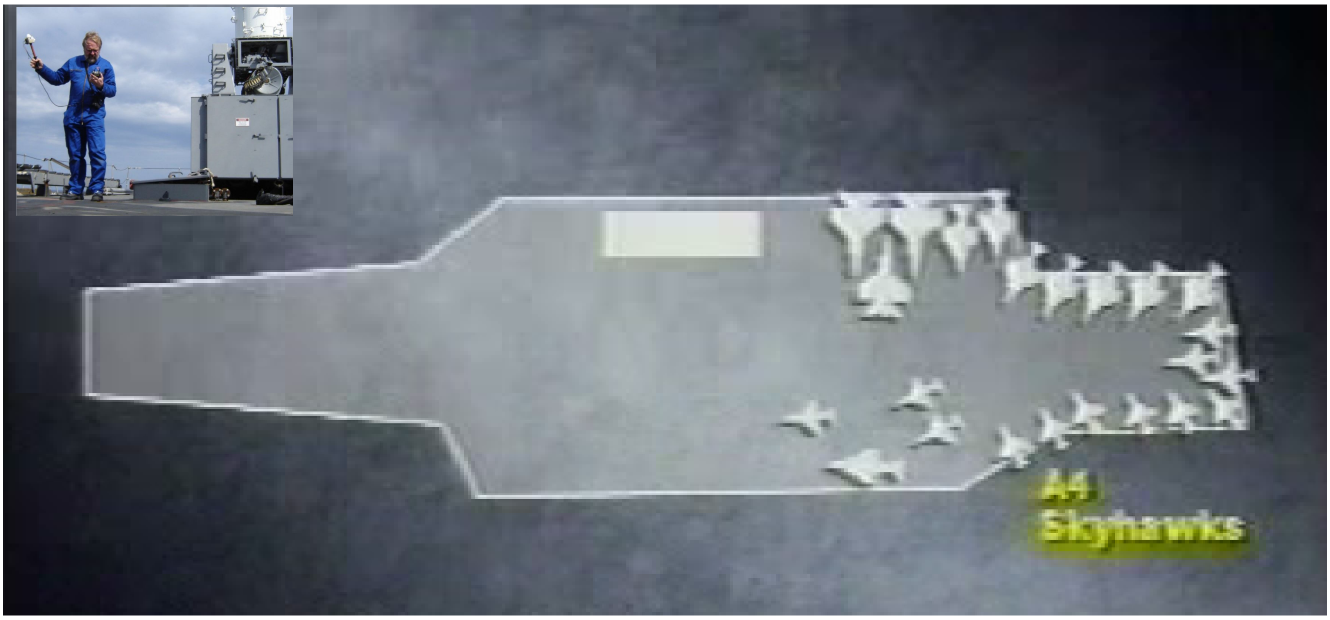
Frequency Range (MHz)		Electric Field (V/m – rms) Peak	Average
0.01	2	73	73
2	30	103	103
30	150	74	74
150	225	41	41
225	400	92	92
400	700	98	98
700	790	267	267
790	1000	284	267
1000	2000	2452	155
2000	2700	489	155
11000	14000	3454	110
14000	18000	8671	243
18000	50000	2793	76

Region 0 is the environment where your customer intends to operate your system. The systems-engineer needs to access the properties of this region 0 environment. As it is not under his/her control, the particular properties are usually found in a standard. The standard used in this slide in the MIL-STD-464 which specifies military environments for Navy, army and air-force. This standard example is used as it is available for free on the internet (Aspect 82).

Other standards exist.
e.g. the following, generic, standards:
for the domestic environment
– immunity: EN/IEC 61000-6-1
– emission: EN/IEC 61000-6-3
and for the industrial environment
– immunity: EN/IEC 61000-6-2
– emission: EN/IEC 61000-6-4

Electromagnetic Environmental Effects (E^3)

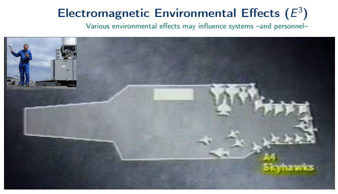
Various environmental effects may influence systems –and personnel–



EMC for Engineers

2023-11-29

└Electromagnetic Environmental Effects (E^3)



This is a 10 minute video showing various effects like:

- Lightning
- The Forrestal disaster (HERO accident)
- Animation "Uncommanded rocket launch"
- Handling of ordnance
- Radiation Hazards (RADHAZ) effect: induction on crane by short-wave transmitter
- Large EM-fields created by high-voltage switch at powerstation
- The Hindenburg disaster (an ESD accident)
- ESD accident at a fuel station
- EMI effects on disabled car

Stick to the ’low frequency approach’ if possible

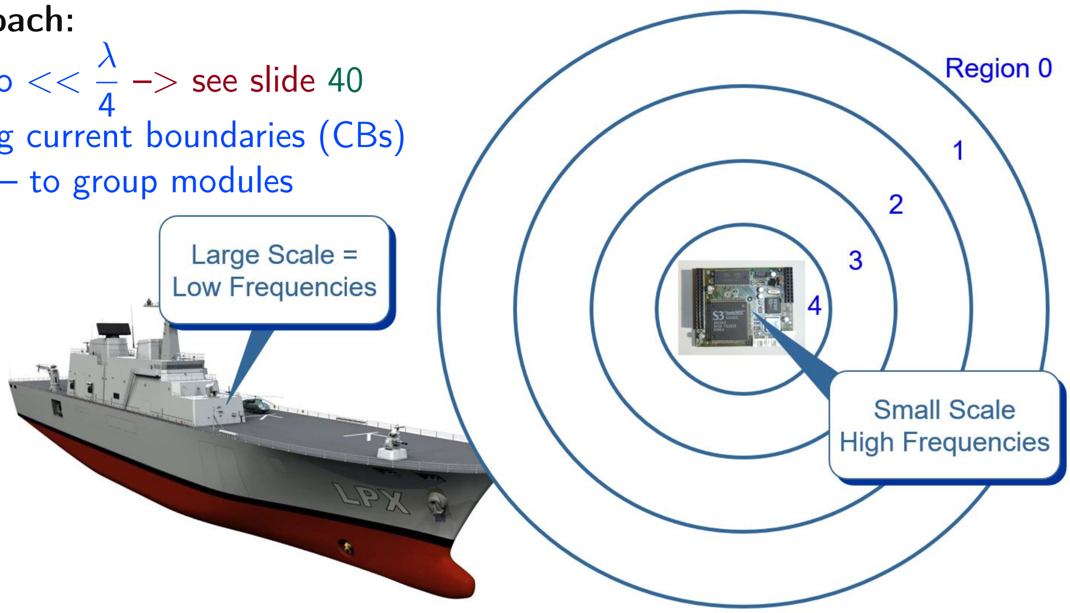
use current boundaries (CB) to ’engineer’ environment sizes to way below ⅟₄ (slide 81)

Low frequency approach:

Restrain module size to $<< \frac{\lambda}{4}$ –> see slide 40

Separate modules using current boundaries (CBs)

Use regions –with CBs– to group modules



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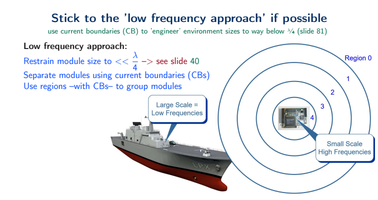


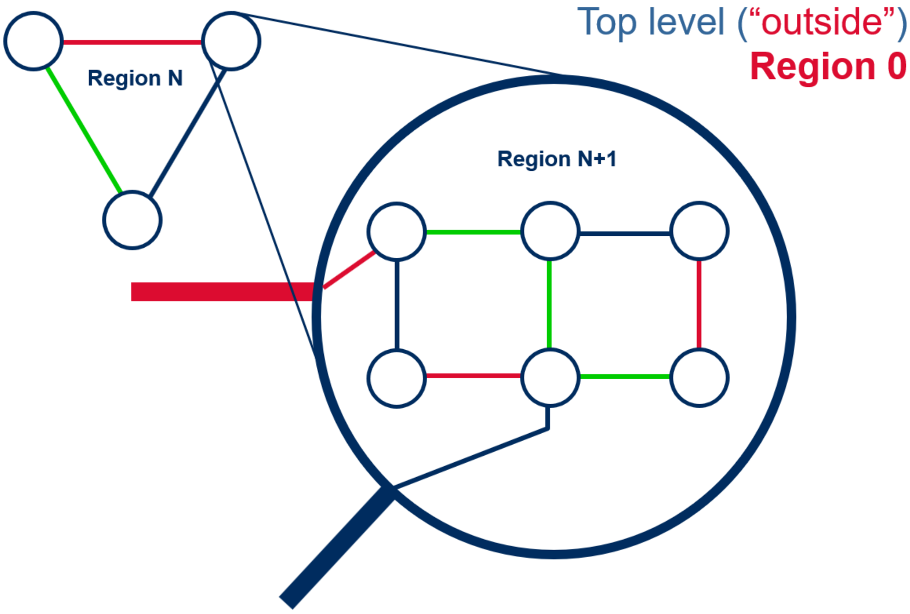
Even electronic modules could have regions: Current boundaries (CBs) should be used to bound areas/environments to much smaller than the critical size defined in slide 39 and 40. Building these CBs at strategic locations, the physical size of a module in terms of EM-effects can be restricted to be less than ℓ_{crit}. Trying to limit module sizes in this way allows the “Low frequency approximation” which simplifies the measures to achieve EMC (Aspect 83).

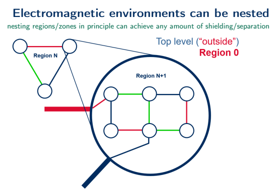
The effort implies the restriction of interface frequencies to values that allow the low frequency approximation in the next larger region/environment. See slide 40.

If this cannot be achieved, the interconnection needs special shielding care to obtain a *low* transfer-impedance (Z_T) , e.g. routing it within a steel pipe. The shielding/steel pipe could then be called a region in itself.

Electromagnetic environments can be nested

nesting regions/zones in principle can achieve any amount of shielding/separation





Defining regions or zones is actually the nesting of electromagnetic (EM)-environments. As you remember the three node –cabinet– model initially used to explain the approach to use CBs to separate noisy and sensitive cables, we can now continue this process. One of those nodes is expanded here to show that inside this particular cabinet there are six interconnected smaller nodes –modules or assemblies–, again with all three[12] cable types. At the module level, we can repeat the process and determine new CBs. By building systems as hierarchies of nested objects –loosely coupled and coherent– you place your CBs at natural boundaries and document your choices. This will greatly improve maintainability in the future (Aspect 84). Note that the analysis model shown here assumes the locations of cabinets and modules are fixed. Before doing the analyses, these locations should be verified as the noisiness or sensitivity increases with distance because transfer-impedance is distance dependent. In PCB-terminology: "placement is more important than routing" (Aspect 85).

The other important aspect to remember is that type-2 CBs like cable trays or machine-structures shall be connected to the cabinet and module type-1 CBs. And keep cables that need type-2 CBs close to these provisions.

[12] Or more if you use more categories.

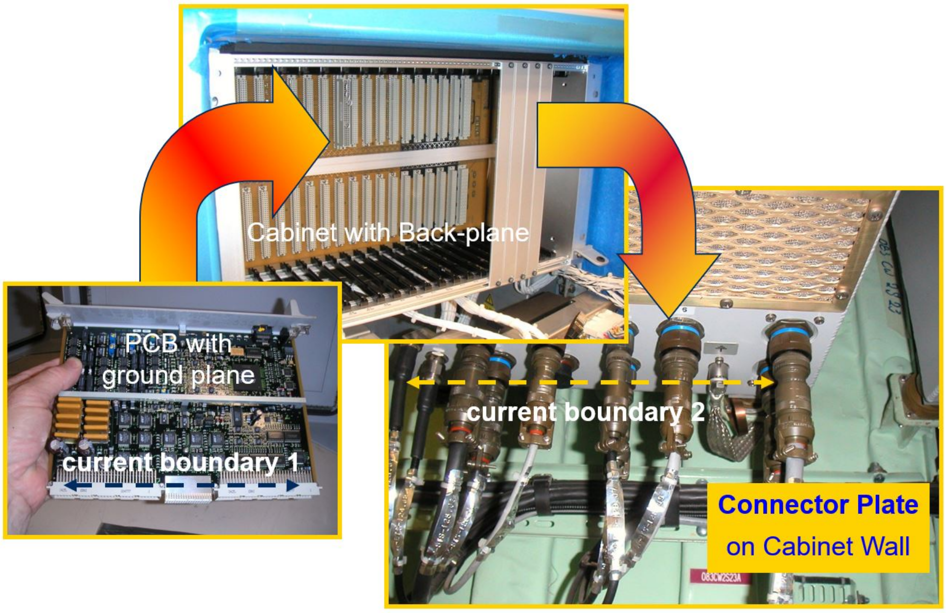
Hierarchically apply current boundaries

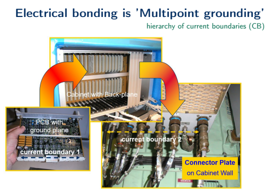
comparable to the well known Matryoshka dolls

An appropriate metaphor for the nesting of EM-environments using CBs are the well known Matryoshka dolls.

Electrical bonding is ’Multipoint grounding’

hierarchy of current boundaries (CB)





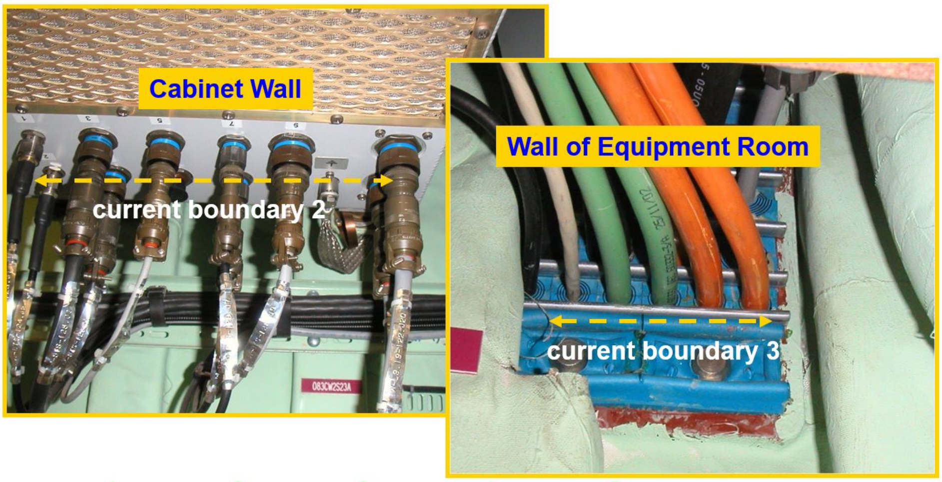
This and the next slide show examples of CBs at various levels in a system design. At the center –highest region number– we find PCBs. Modern PCBs have one or more ground reference planes (GRPs) over which traces are routed, a type-2 CB. At one edge[13] these PCBs have connectors, that serve as type-1 CB, to interface with the module back-plane to which many more PCBs are connected.

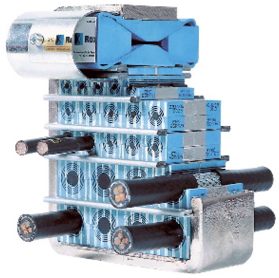
There are several provisions, explained in 10.1109/MEMC.2017.7931984, to make the PCB-edge connector a true CB. 40 years ago, in the transistor-transistor logic (TTL)-days, when an error was found on a digital board, a spare gate on an adjacent board would be used to correct it. That gate was wired with e.g. wire-warp wire from one board to another. Nowadays that would skip at least one CB, invalidating it. Drilling a hole in a shielding metal wall and feeding a insulated cable through it to service a newly installed device is a similar offense (Aspect 86)! The backplane interconnections in this example end up at the module or cabinet connector plate which is the next type-1 CB.

[13] If at two edges, this could cause the PCB to generate noise into or be sensitive to CM-currents from the next lower number region/environment. So this is not recommended.

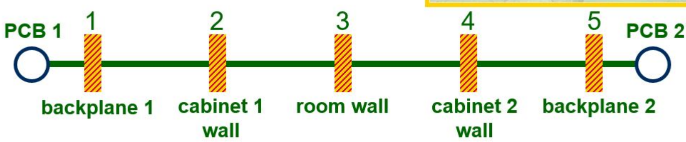
'Multipoint grounding'

EM-separating rooms in e.g. a ship is called 'zoning' (partition into EM-regions)

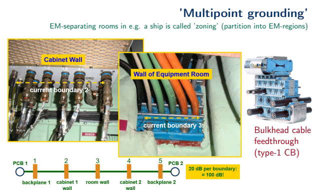




Bulkhead cable feedthrough (type-1 CB)



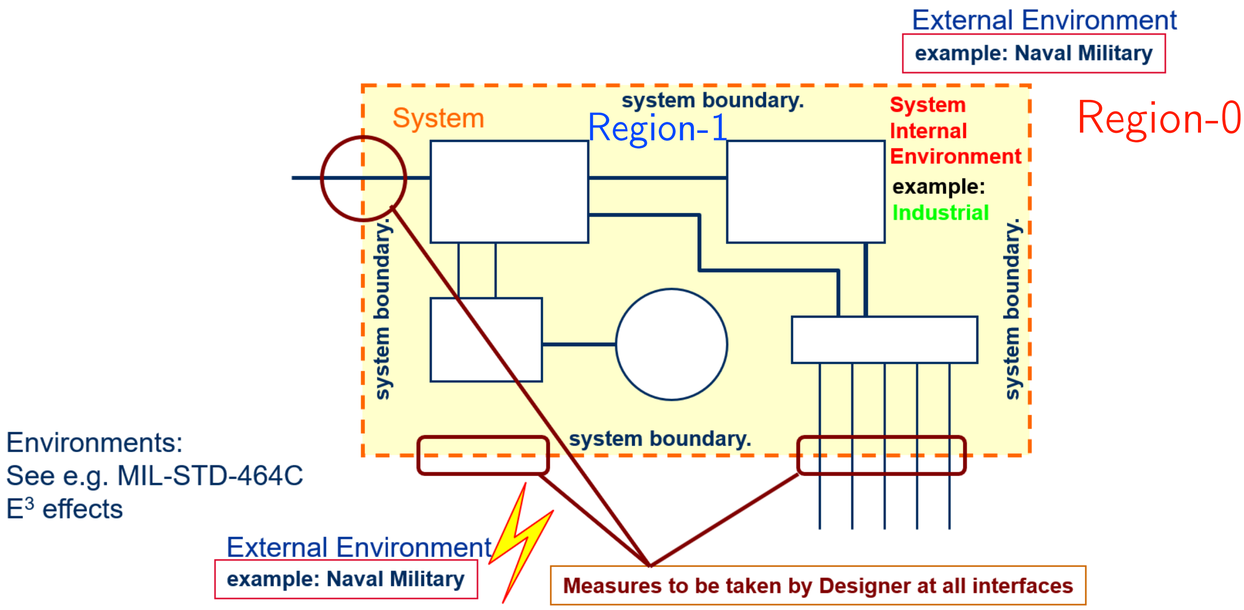
20 dB per boundary:
= 100 dB!

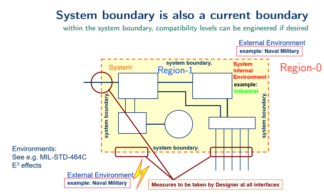


From the cabinet connector plate –type-1 CB–, cables are routed to other modules/cabinets in the same room or they run to the bulkhead/metal room wall to an adjacent room. Imagine a situation where the adjacent room has a similar installation.
In that case, an interconnection from a PCB in the first room encounters 5 type-1 CBs on its journey to a PCB in the second room.
If each CB provides only 20 dB CM-separation, the total separation is already 100 dB –a factor 100000–. This layered CB approach is the true meaning of *multipoint "grounding"* for EMC.
By the way, "*Earthing*" is a completely different subject and has nothing to do with reduction of interference or EMC.
Protective earthing (PE) is for personal safety. Everything that can be touched should be connected, ultimately to the "Earth" you walk on, so that no dangerous touch-voltages can develop. As all metal equipment has one or more PE-connections, CM-currents may flow through the PE-network. This was demonstrated with the long-wire experiment with the radio to explain the working of a CB in the basic EMC module. CBs for EMC use the same metal structure, hence the term "grounding" (Aspect 87).

System boundary is also a current boundary

within the system boundary, compatibility levels can be engineered if desired



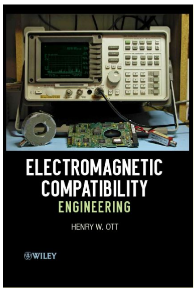


A top-down approach to EMC encompasses "thinking in environments":
1. What is the specified EMC-environment for which the devices you plan to use in your system have been tested or at least are guaranteed to work in properly –your region-1–?
2. In which environment will the new system have to operate –your region-0–?
3. Write/find a specification for your region-0
4. Build your region-1 and possibly extra regions for specific equipment.
This approach implies that you may have to build an environment –e.g. the industrial environment in this slide– as region-1 within your given region-0 –MIL-STD-464 in the example here. How difficult this is depends on the *EM-distance* between your region-0 and region-1. The least you need to do is look at all interconnection-crossings of the system-boundary. These should be type-1 CBs. This is easiest if you have a completely shielded system. Just create type-1 CBs at the locations where cables enter. As a compromise you could use the metal chassis of your machine to connect your type-1 CBs to. Then attempt to use *just one* location –CB type-1–. To make type-1 CBs possible at several *separated* entry points, you will need some amount of complete shielding as external –region-0– CM-currents will otherwise flow *through* your system (Aspect 88).

In the older MIL-HDBK-237B there is more information on development issues

MIL-HDBK-237B, DEPARTMENT OF DEFENSE HANDBOOK: GUIDANCE FOR CONTROLLING ELECTROMAGNETIC ENVIRONMENTAL EFFECTS ON PLATFORMS, SYSTEMS, AND EQUIPMENT

Read all details in Henry Ott's book. https://hott.shielddigitaldesign.com/EMCE_book_files/emce_book.html



http://everyspec.com/MIL-HDBK/MIL-HDBK-0200-0299/MIL-HDBK-237B_21779/
The later versions of the 237 Handbook focus on procurement rather than development.

Find Henry Ott's book EMC Engineering at:

https://onlinelibrary.wiley.com/doi/epdf/10.1002/9780470508510.fmatter

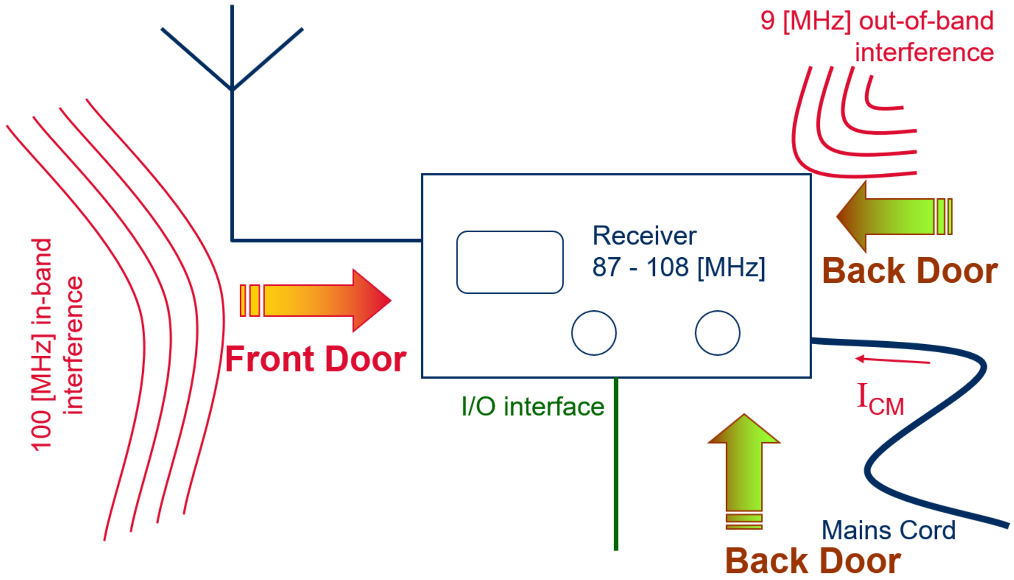
Part 8: Testing

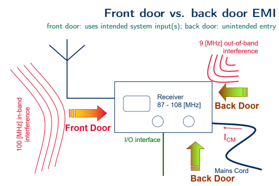
Specify, design and then test your system in its environment

Part 8: Testing

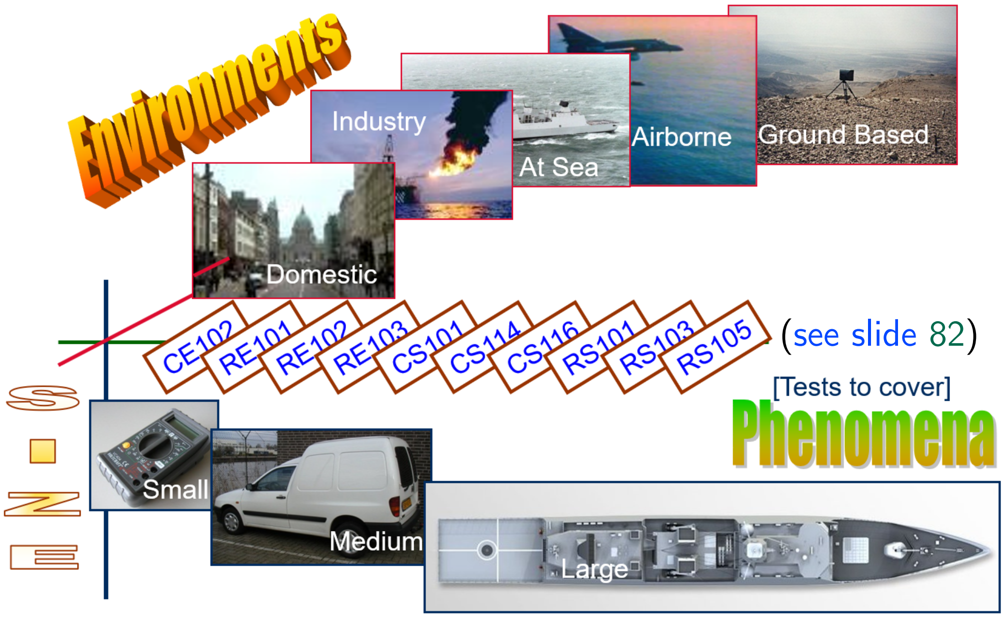
A brief introduction to where to start once you get to the part of finding out whether your system works in its intended environment.

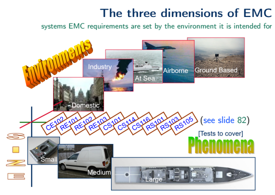
front door: uses intended system input(s); back door: unintended entry





The concepts *frontdoor* and *backdoor* are used used to describe different interference paths. If you have a radio receiver and receive two stations at the same frequency, this is called *frontdoor* interference i.e. you hear two radio stations at the same time. Interference through –more hidden– other paths is called *backdoor* interference. This could be mains-carried noise, a radio frequency (RF) signal coupling into your intermediate frequency (IF) circuits or disturbance on your wired I/O entries. In a *frontdoor* case, e.g. an illegal transmitter operating at the same frequency as the desired station, it may be difficult to solve this in your system. If it is a station at a nearby frequency, you may have insufficient selectivity. Then you could solve the problem by improving your input circuit. *Backdoor* interference usually implies you have to make your system stronger by shielding, filtering or even applying software protection (Aspect 89). An example is airplanes activating their brakes upon operation of a handheld transmitter by airport personnel. It appeared to be a spurious signal from some sensor. The software solution was to only react to that sensor after a certain number of pulses. In the early days of airbags, some would go off when a mobile telephone was used: typical backdoor. The latter is an example of a system safety related event. See e.g. MIL-STD-882.





The three dimensions to approach any EMC situation are shown here:
1. Environments
Shown are five environments or regions-0: three military: Ground based, Airborne and Navy. And two civil environments: Industrial and Domestic.
Each environment has its own requirements and interference levels.
Reason: in a EM-silent environment like the domestic, little measures are required which makes products less immune but cheap. It that means also that domestic products have stricter emission limits (Aspect 90).

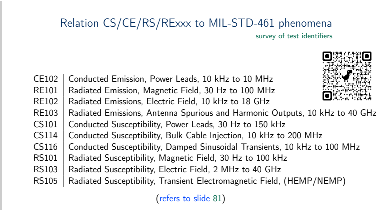
2. Phenomena.
For each environment there exist aspects to consider usually contained in EMC-standards. The sequence shown are tests intended for the Military Navy environment (Aspect 91). The description of the tests are summarized on slide 82.
3. Size
Depending on the size of equipment, different phenomena are dominant (Aspect 92).

The treatment of EMC requirements and the way to test them goes beyond the scope of this presentation.

CE102	Conducted Emission, Power Leads, 10 kHz to 10 MHz
RE101	Radiated Emission, Magnetic Field, 30 Hz to 100 MHz
RE102	Radiated Emissions, Electric Field, 10 kHz to 18 GHz
RE103	Radiated Emissions, Antenna Spurious and Harmonic Outputs, 10 kHz to 40 GHz
CS101	Conducted Susceptibility, Power Leads, 30 Hz to 150 kHz
CS114	Conducted Susceptibility, Bulk Cable Injection, 10 kHz to 200 MHz
CS116	Conducted Susceptibility, Damped Sinusoidal Transients, 10 kHz to 100 MHz
RS101	Radiated Susceptibility, Magnetic Field, 30 Hz to 100 kHz
RS103	Radiated Susceptibility, Electric Field, 2 MHz to 40 GHz
RS105	Radiated Susceptibility, Transient Electromagnetic Field, (HEMP/NEMP)

(refers to slide 81)

2023-11-29
EMC for Engineers
└─Relation CS/CE/RS/RExxx to MIL-STD-461 phenomena



For information on the details of the tests mentioned see the MIL-STD-461, available on the internet (follow the QR-code).

Relation CS/CE/RS/RExxx to MIL-STD-461 phenomena

survey of test identifiers

CE102	Conducted Emission, Power Leads, 10 kHz to 10 MHz
RE101	Radiated Emission, Magnetic Field, 30 Hz to 100 MHz
RE102	Radiated Emissions, Electric Field, 10 kHz to 18 GHz
RE103	Radiated Emissions, Antenna Spurious and Harmonic Outputs, 10 kHz to 40 GHz
CS101	Conducted Susceptibility, Power Leads, 30 Hz to 150 kHz
CS114	Conducted Susceptibility, Bulk Cable Injection, 10 kHz to 200 MHz
CS116	Conducted Susceptibility, Damped Sinusoidal Transients, 10 kHz to 100 MHz
RS101	Radiated Susceptibility, Magnetic Field, 30 Hz to 100 kHz
RS103	Radiated Susceptibility, Electric Field, 2 MHz to 40 GHz
RS105	Radiated Susceptibility, Transient Electromagnetic Field, (HEMP/NEMP)

(refers to slide 81)

List of abbreviations

E^3	Electromagnetic Environmental Effects
Z_T	transfer-impedance
AC	alternating current
AF	antenna factor
AWG	American wire gauge
BNC	Bayonet-Neill-Concelman
CAE	commercially available equipment
CB	current boundary
CM	common-mode
DC	direct current
DM	differential mode
EM	electromagnetic
EMC	electromagnetic compatibility
EMI	electromagnetic interference
EMP	electromagnetic pulse
ESD	electrostatic discharge
GRP	ground reference plane
HERO	hazards of electromagnetic radiation to ordnance
IC	integrated circuit
IF	intermediate frequency
PCB	printed circuit board
PE	protective earth
PUL	per unit length
RADHAZ	Radiation Hazard
RAM	reliability, availability and maintainability
RF	radio frequency
STP	shielded twisted pairs
TDR	time domain reflectometer
TTL	transistor-transistor logic
UTP	unshielded twisted pairs
VNA	vector network analyzer