Electromagnetic Shielding Concepts and Applications

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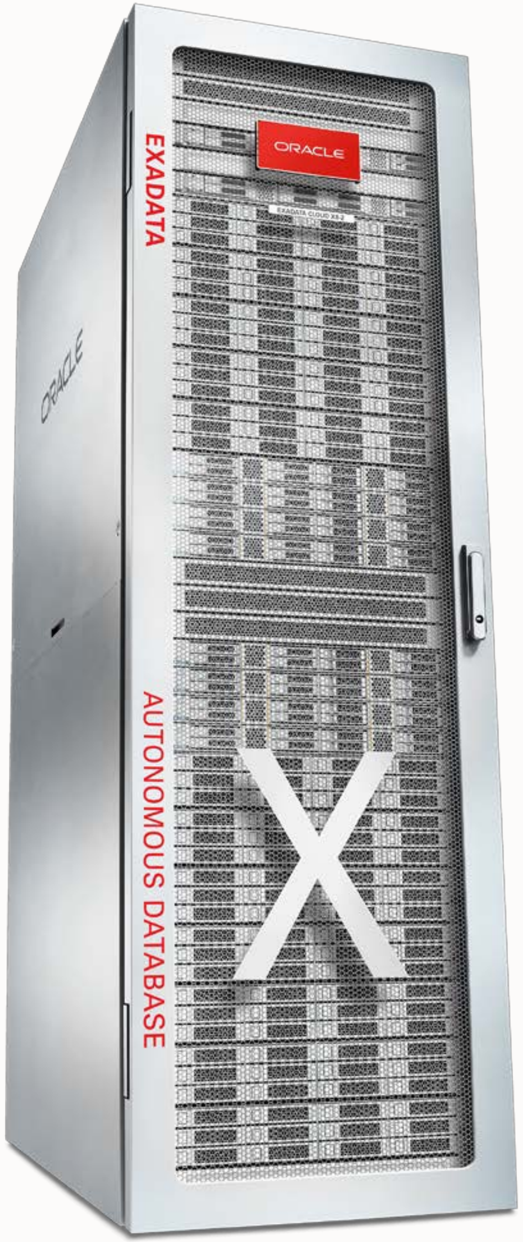

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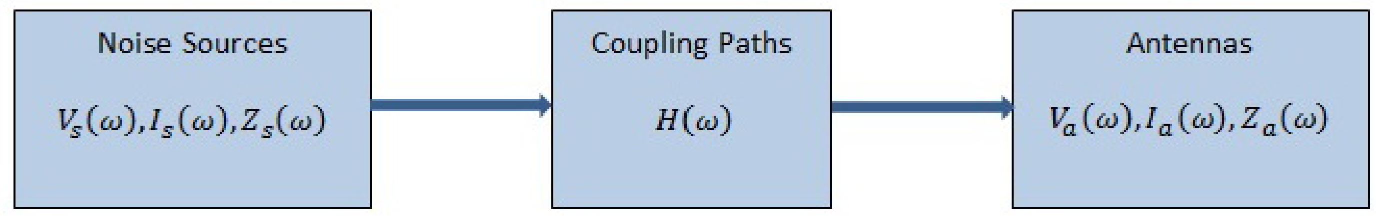
Abstract

All electronic equipment uses some type of shielding. From a theoretical point of view, electromagnetic shielding is among the most difficult areas of EMC. Basic shielding concepts, their practical limitations, and some typical engineering problems associated with shielding are presented. Among the topics discussed herein, there are theoretical approaches for shielding, chassis resonances, shielding integrity problems (seams, joints, apertures, perf patterns), aperture coupling and internal sources, coatings, corrosion, evaluation of shielding effectiveness and materials used for shielding, and grounding of the shield. The examples used are from the IT industry, and even if the underlying physics is the same, there are many specific shielding aspects in automotive, aerospace, medical, and mobile or portable industries.

Outline

1. Introduction, near field, far field, electric, magnetic and electromagnetic shielding
2. Analytical approaches to shielding. Field theory method (Kaden) and TL impedance method (Schelkunoff)
3. Limits of the theoretical approaches, numerical simulations, reciprocity
4. Chassis resonances, internal compartmentalization, absorbers
5. Practical aspects of shielding, aperture coupling, slots, seams, rivets
6. Shielding and thermal issues, holes, perf patterns, honeycomb
7. Shielding and internal sources, heatsink grounding, local shielding
8. Shielding and coatings, corrosion
9. Evaluation of shielding effectiveness, materials, chassis
10. Shielding and safety certification
11. References
12. Acknowledgments

Introduction



Noise Sources

$V_s(\omega), I_s(\omega), Z_s(\omega)$

Coupling Paths

$H(\omega)$

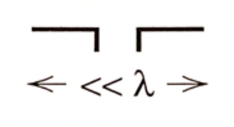
Antennas

$V_a(\omega), I_a(\omega), Z_a(\omega)$

1. Any EMC problem is comprised of one or multiple noise sources (typically the CPUs, ASICs, DC/DC, other ICs, etc.), one or multiple coupling paths (capacitive, inductive, galvanic, field coupling), one or multiple antennas (cables, apertures, holes, seams, etc.).
2. The best way to mitigate an EMC problem is to control EMC at the source whenever possible and to add containment measures at every other level, reducing coupling paths, eliminating unwanted antennas.
3. The noise sources, the coupling paths, and the antennas are all frequency dependent, which increases the complexity of the problem.
4. EMC related parasitics for components, sources and coupling paths are not typically available in the data sheets of the components or on the schematics.
5. The characteristics of some coupling paths can be extracted from the geometry on the PCB, using empirical and simplified models, but usually is based on experience and simplified models for system level coupling paths.
6. EMC deals with unintentional antennas, and their shapes, material composition, and the corresponding boundary conditions are not the same as typically described in Antenna textbooks.
7. All these factors result in substantial complexity of EMC problems, with numerous unknowns and ambiguities, and the prediction models lack accuracy.

Introduction – Near Field vs. Far Field for Point Sources

Small electric dipole



Small dipole

E_θ, E_r, H_φ

$$(Z_W)_e = \frac{dE_\theta}{dH_\varphi} = \frac{1}{\varepsilon}\,\frac{\frac{1}{j\omega r^3} + \frac{1}{vr^2} + \frac{j\omega}{v^2 r}}{\frac{1}{r^2} + \frac{j\omega}{vr}}$$

$$v = c = 3 \cdot 10^8 \ \text{m/s} \ \text{(in vacuum)}$$

Small current loop

Small loop

E_φ, H_θ, H_r

$$(Z_W)_m = \frac{dE_\varphi}{dH_\theta} = \mu\,\frac{\frac{j\omega}{r^2} - \frac{\omega^2}{vr}}{\frac{1}{r^3} + \frac{j\omega}{vr^2} - \frac{\omega^2}{v^2 r}}$$

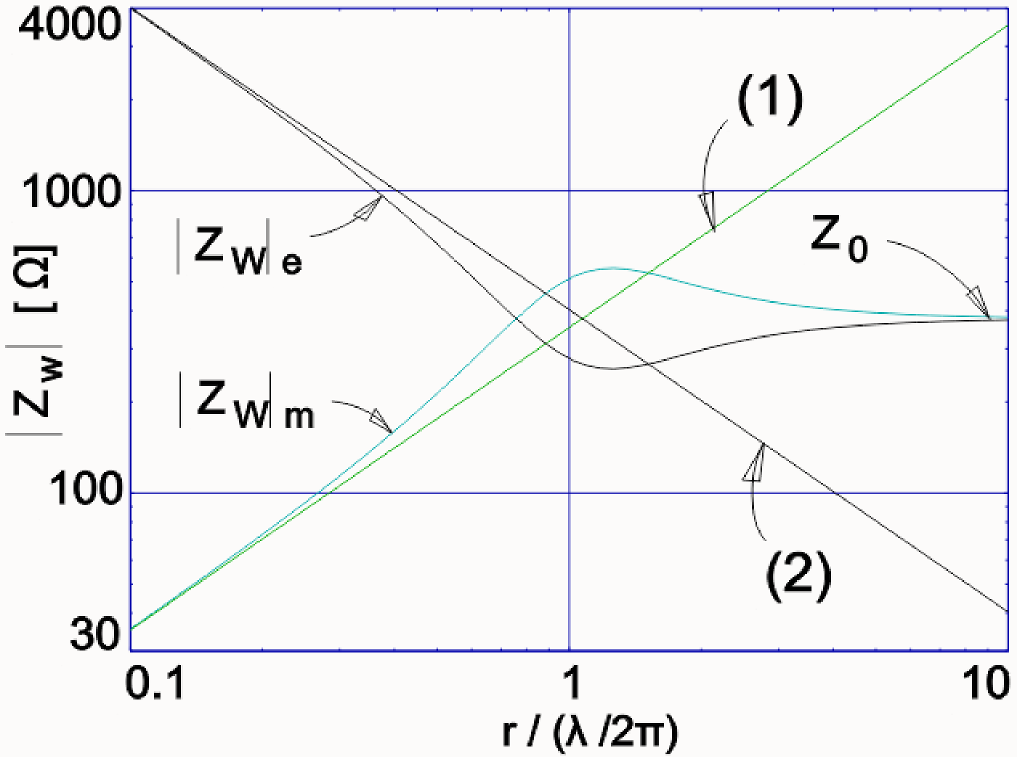
Near Field (Reactive Field, Fresnel Zone)

$$(Z_W)_e = \frac{dE_\theta}{dH_\varphi} \xrightarrow{r \to 0} \left|\frac{1}{j\omega\varepsilon r}\right| \geq 377\Omega \qquad (Z_W)_m = \frac{dE_\varphi}{dH_\theta} \xrightarrow{r \to 0} \left|j\omega\mu r\right| \leq 377\Omega$$

Far Field (Radiation Zone, Fraunhofer Zone)

$$Z_W = \frac{dE_\theta}{dH_\varphi} \xrightarrow{r \to \infty} \frac{1}{\varepsilon v} = \mu v = \frac{1}{\varepsilon \frac{1}{\sqrt{\varepsilon\mu}}} = \sqrt{\frac{\mu}{\varepsilon}} \simeq \sqrt{\frac{\mu_0}{\varepsilon_0}} = 120\pi = 377\Omega$$

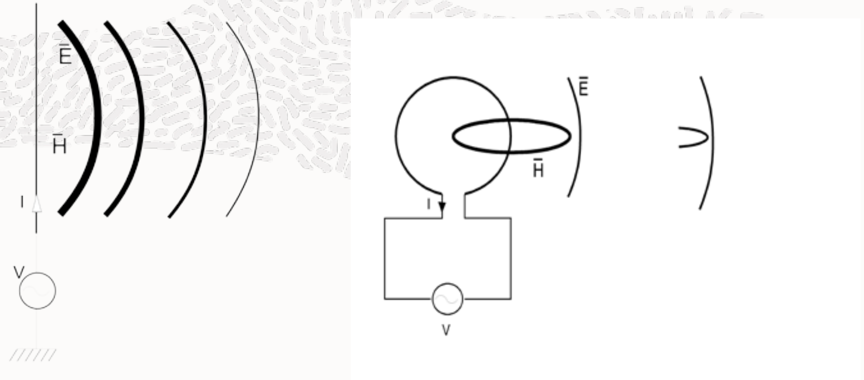
$$\varepsilon_0 = \frac{1}{36\pi}10^{-9}\,F/m \qquad \mu_0 = 4\pi 10^{-7}\,H/m \qquad v = \frac{1}{\sqrt{\varepsilon\mu}} = \frac{c}{\sqrt{\varepsilon_r \mu_r}} \qquad Z_0 = 377\Omega$$

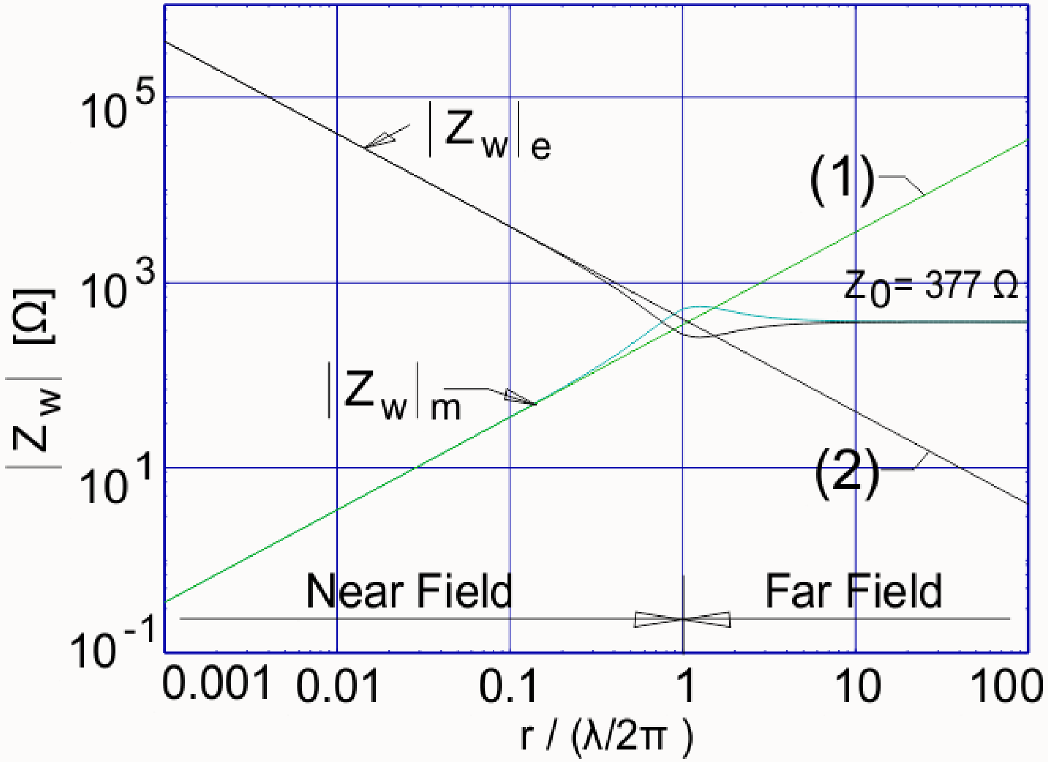


Introduction – Near Field vs. Far Field

1. For point sources, below λ/2π the E and H fields are in quadrature, and the power density is imaginary, indicating reactive power, hence no time average radial power flow, only stored energy in the vicinity of the source. The field structure is strongly dependent on source type and distance r, with $E, H \sim \frac{1}{r^2}, \frac{1}{r^3}$.
2. Above λ/2π the E and H fields are in phase, all power is radiated power, and the radiation pattern is independent of distance r, with $E, H \sim 1/r$. The calculations for point sources are in spherical coordinates. At distance, on a small wavefront area, the field is well approximated by a plane wave.
3. The power density associated with radiation exists everywhere and passes through the near field region. At λ/2π, the maximum (θ=90°) radiated power and reactive power are equal.
4. For antennas of maximum dimension D, the regions at distance r:

Reactive near field:	$0 \le r \le 0.62\sqrt{D^3/\lambda}$	
Radiating near field:	$0.62\sqrt{D^3/\lambda} \le r \le 2D^2/\lambda$	(Fresnel)
Far field:	$2D^2/\lambda \le r$	(Fraunhoffer)



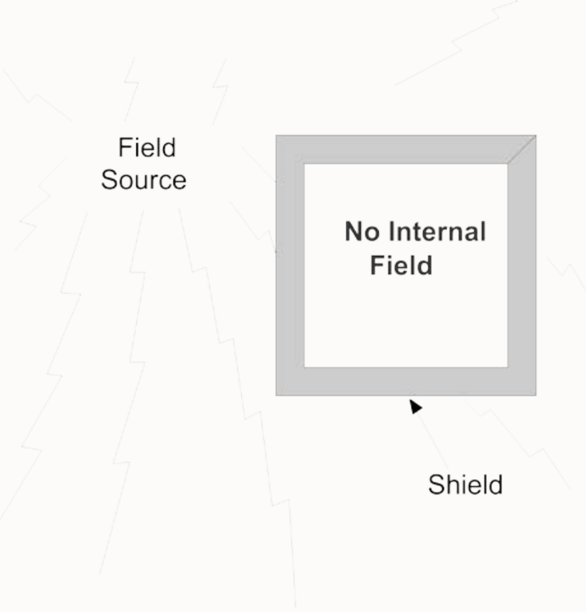


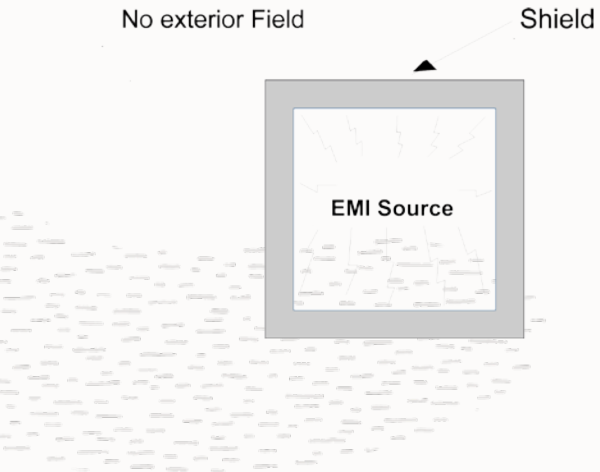
Introduction – Shielding Effectiveness

1. Radiated immunity problem: uniform external field of 3V/m applied to EUT in the frequency ranges 80 MHz - 1000 MHz, 800 MHz, 960 MHz, 1.4 GHz – 6 GHz. The system must work properly (Criterion A). ESD (Criterion B), transients of ±4kV on metal, ± 8kV on plastic, is also strongly related to chassis shielding performance.
2. Radiation emissions problem: for 30 MHZ – 1 GHz the external field is measured at 10m, and for 1 GHz – 40 GHz is measured at 3 m, with antenna from 1 up to 4 meters high. The electric field limits do not change much in the whole frequency range because a 10 m limit is extrapolated to 3m by adding 10.46 dB.
3. Electric field is measured, and it needs to be below FCC class A limit levels, which are 40 dBμV/m (30 MHz – 230 MHz), 47 dBμV/m (230 MHz – 1 GHz), 56 dBμV/m (1 GHz – 3 GHz), 60 dBμV/m (3 GHz – 40 GHz), plus the typical 6 dB margin.
4. The typical figure of merit is Shielding Effectiveness (SE), for external source:

$$S.E. = 20\log\left|\frac{E_{exterior}}{E_{interior}}\right| \quad [dB] \qquad S.E. = 20\log\left|\frac{H_{exterior}}{H_{interior}}\right| \quad [dB] \qquad S.E. = 10\log\left|\frac{P_{exterior}}{P_{interior}}\right| \quad [dB]$$

5. Similar definitions can be used for internal source, practically is the ratio of the field magnitudes in a point away from the source without the shield and with the shield.
6. The antennas used for testing are for electric field, but there are near field probes for both electric and magnetic field.
7. The difficult problem is the radiated emissions, the radiated immunity is not usually a problem at the commercial levels listed above.





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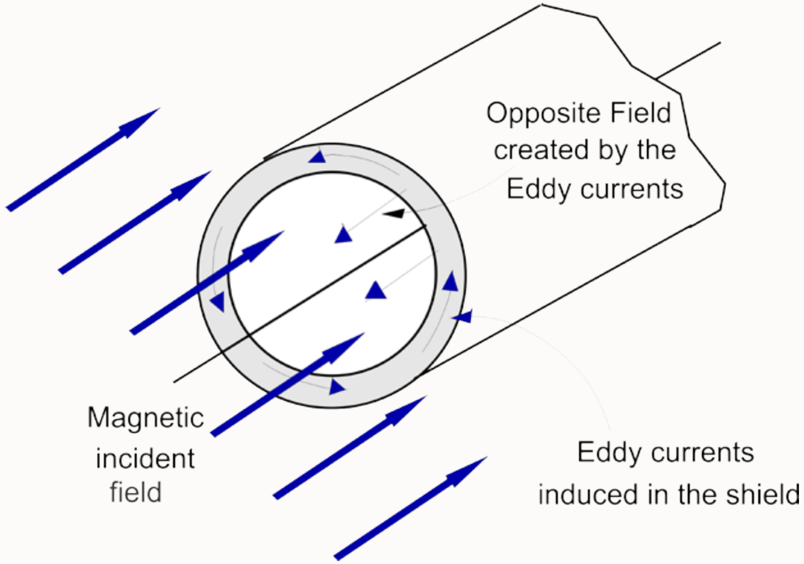
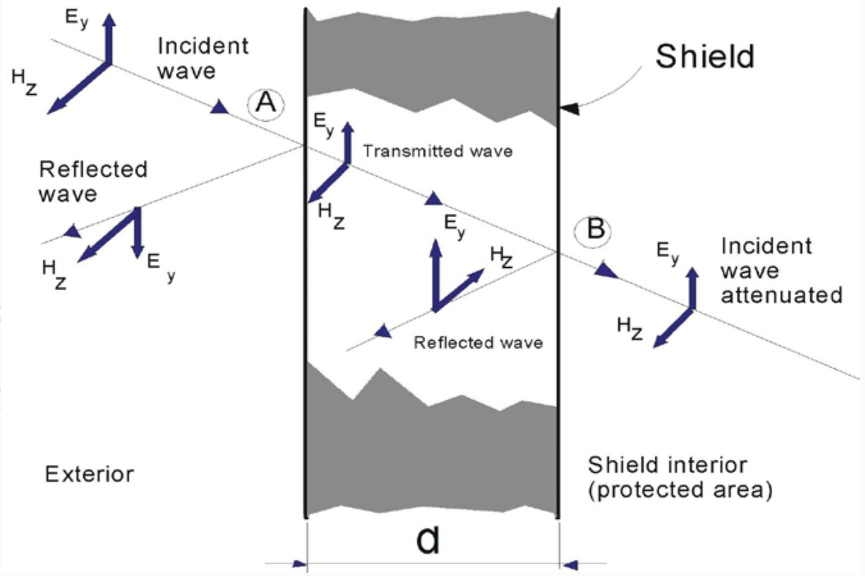
Analytical Approaches to Shielding

1. Shielding Effectiveness (SE) is rarely used as a single number characterizing a chassis in the whole frequency range. In the general case of a real-life enclosures, SE will depend on many parameters like source polarization, source impedance, source-aperture distance, internal chassis resonances, radiated beam pattern vs. receiving antenna position, receiving antenna – enclosure (antenna) coupling strength, etc.
2. One may find the SE as a function of frequency, but that graph will depend on how much electronics was inside, how was excited by internal sources, and will be difficult to extrapolate that graph for other electronics in the same chassis.
3. The study of SE in simplified cases may give a general understanding of the parameters which play a role in shielding and a better physical understanding useful in design and troubleshooting.
4. There are two traditional analytical approaches for shielding: the transmission lines (TL) impedance method ([9] Sergei A. Schelkunoff, 1933) and the field theory method ([10] Heinrich Kaden, 1950).





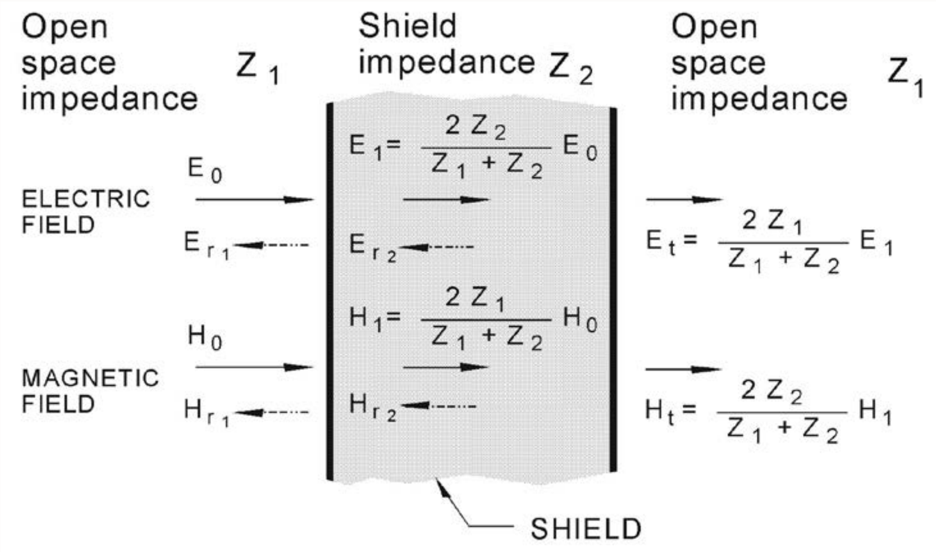
Analytical Approaches to Shielding – TL Impedance Method

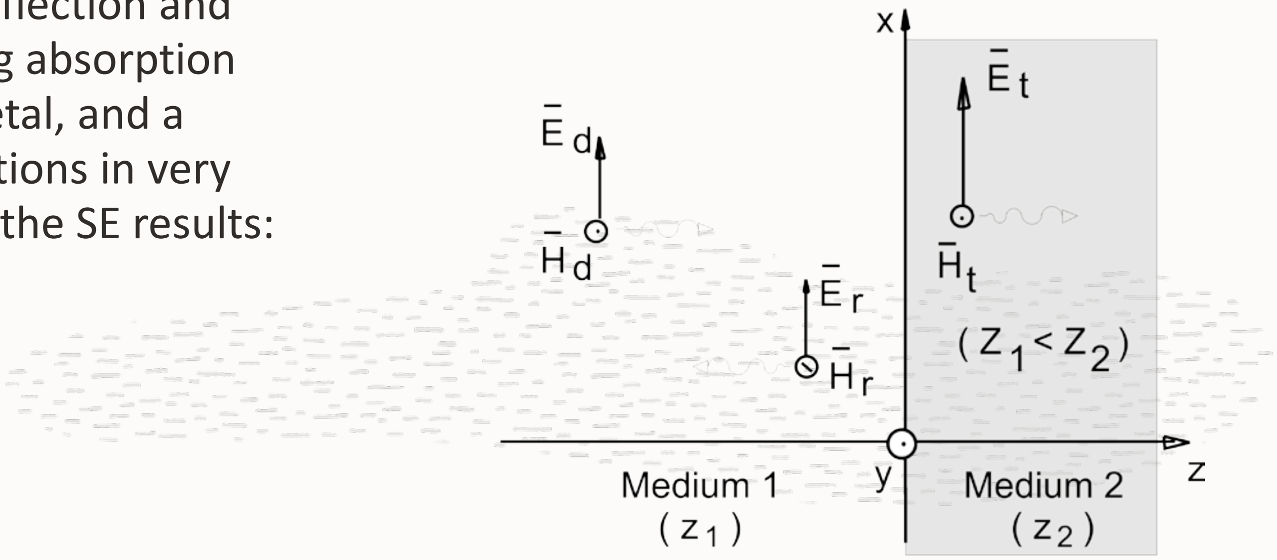
1. Schelkunoff observed that equations describing the propagation of a plane wave (TEM) are like the equations describing the propagation of current and voltage waves in a transmission line, with the fundamental mode TEM.
2. Analysis of shielding is reduced to study of cascaded transmission lines (TL) with different characteristics (impedance, propagation constant) dictated by the medium (air, metal). The method can be extended to multilayer shields.
3. For an **infinite** in 2D shield (much larger than the wavelength λ) shielding is reduced to reflection losses R, using reflection and transmission coefficients for each interface, adding absorption losses (A) due to attenuation of the field in the metal, and a correction factor (B) to account for multiple reflections in very thin shields for which the absorption is small, and the SE results:

S.E. [dB]=R [dB] + A [dB] + B [dB]

$$\gamma = \sqrt{(R + j\omega L)(G + j\omega C)} \qquad \Gamma = \sqrt{j\omega\mu(\sigma + j\omega\varepsilon)}$$

$$Z_0 = \sqrt{\frac{R + j\omega L}{G + j\omega C}} \qquad Z_{medium} = \sqrt{\frac{j\omega\mu}{\sigma + j\omega\varepsilon}}$$





Analytical Approaches to Shielding – TL Impedance Method

1. Reflection loss R calculated using TL theory:

$$R[dB] = 20\log\left|\frac{(Z_s+Z_0)^2}{4Z_sZ_0}\right| \qquad |Z_s| = \sqrt{\frac{\omega\mu}{\sigma}} \qquad (Intrinsic\ \ impedance\ \ metal)$$

$$Z_0 = 377\Omega \qquad (Z_0)_E = \left|\frac{1}{j\omega\varepsilon r}\right| \geq 377\Omega \qquad (Z_0)_H = |j\omega\mu r| \leq 377\Omega$$

2. Absorption losses (δ_s is skin depth is few μm at 30 MHz):

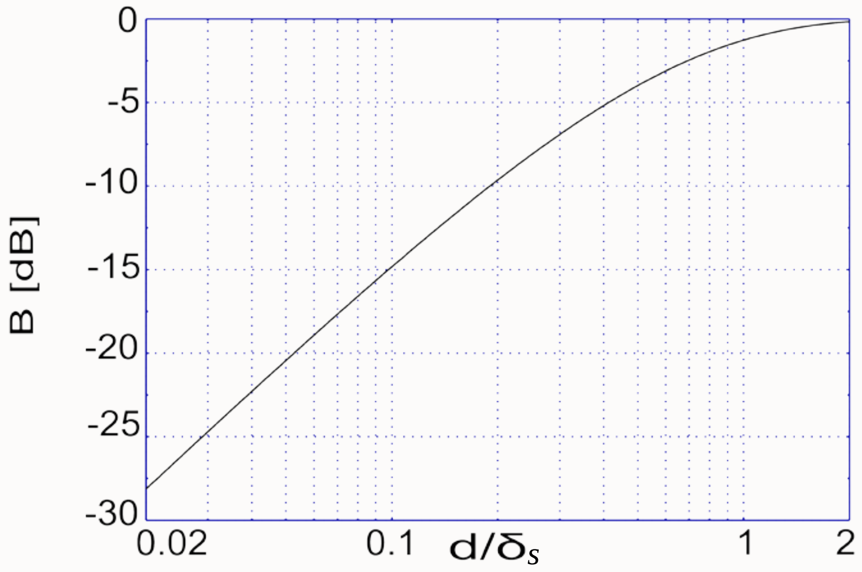
$$A[dB] = 20\log\left|\frac{H_{ext}}{H_{metal}}\right| = 8.69\frac{d}{\delta_s} \qquad \delta_s = \sqrt{\frac{2}{\omega\sigma\mu}} \qquad |H_{metal}(\delta_s)| = \frac{H_{metal}(0)}{e}$$

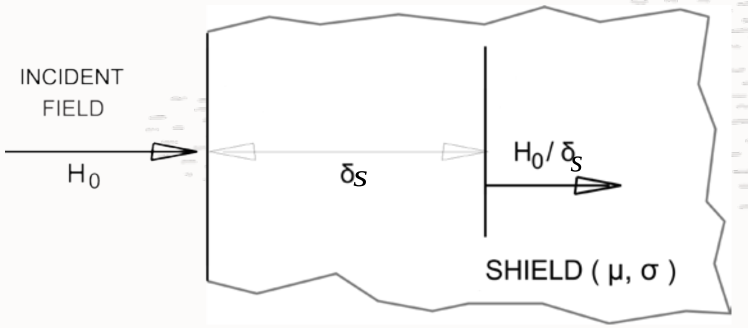
3. Correction factor for multiple reflections in thin shields:

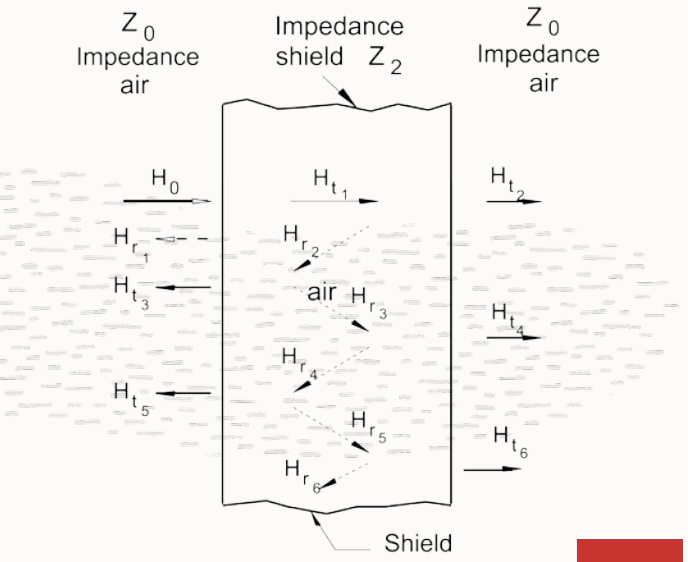
$$B[dB] = 20\log\left(1-e^{-\frac{2d}{\delta_s}}\right)$$

4. Shielding effectiveness:

$$S.E.[dB] = R + A + B$$







Analytical Approaches to Shielding – TL Impedance Method

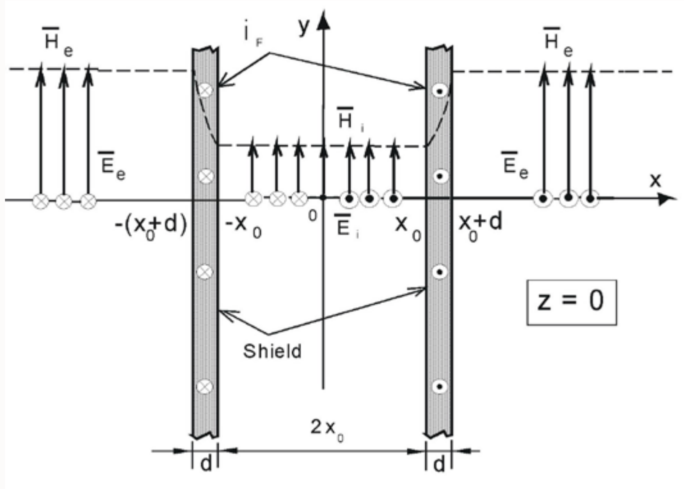
1. Schelkunoff’s method considers the shield material properties (μ, ε, σ) through the intrinsic impedance of the shield Z_s and skin depth δ_s, and their impact on SE.
2. In principle, is applicable at LF and HF, for plane wave (TEM) field and can also be used for multilayer shields (metal, air, metal, metal/coating, etc.). Some books are also using near field expressions of Z_w to estimate near field R losses.
3. This method does not account for any specific geometry of the shield rather than an “infinite” (>>λ) flat 2D wall or a layer. Internal resonances of enclosures, slots, gaps, and apertures cannot be directly accounted for, and currents induced in the shield must close at infinity.
4. Can be useful to rank materials like metal coated plastic, conductive plastic, composite materials with difficult to predict properties. Test methods for shielding materials in IEEE 299 were based on this approach.
5. The SE value calculated is for the material and can’t be extrapolated for any realistic chassis (the results are >100 dB!).
6. The other analytical method, the circuit method (Kaden) is based on solving the Helmholtz equation outside the chassis, in the wall of the chassis, and inside the chassis, with proper boundary conditions. Using the continuity of the field components at each air/metal interface the integration constants can be calculated and the unique solution for the field is obtained. The SE is calculated directly using the resulting fields inside and outside the shield.
7. Helmholtz equation $\left(\Delta\bar{E} = \Gamma^2\bar{E},\ \Gamma^2 = j\omega\mu\left(\sigma + j\omega\varepsilon\right)\right)$ is solved using separation of variables, possible only for canonical geometries, with orthogonal coordinates and where the surface of the shield coincide with a constant coordinate.
8. There are only 11 such geometries (cartesian, cylindrical, spherical, elliptical, parabolic, toroidal, etc.), and typical shields analyzed with this method will be plan-parallel, cylindrical (Bessel), spherical (Legendre), elliptical (Mathieu).
9. For such cases, this method can account for geometry and can give some physical insight about the effect of slots and apertures.

Analytical Approaches to Shielding – Field Theory Method

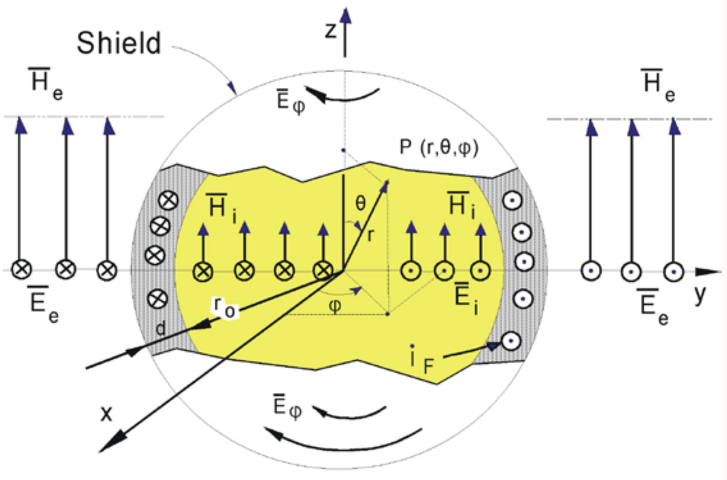
1. Plan-parallel shield

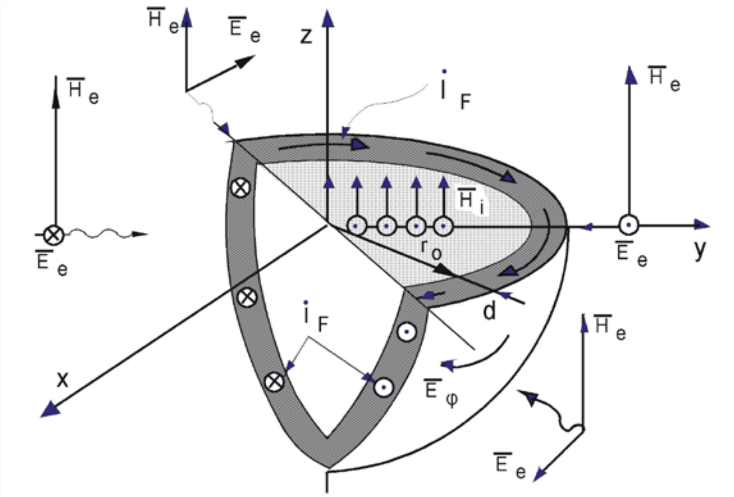
$$S.E.=10\log\left[\frac{1}{2}\left(\cosh\frac{2d}{\delta}+\cos\frac{2d}{\delta}\right)+\frac{\mu_0 x_0}{\mu\delta}\left(\sinh\frac{2d}{\delta}-\sin\frac{2d}{\delta}\right)+\left(\frac{\mu_0 x_0}{\mu\delta}\right)^2\left(\cosh\frac{2d}{\delta}-\cos\frac{2d}{\delta}\right)\right]$$

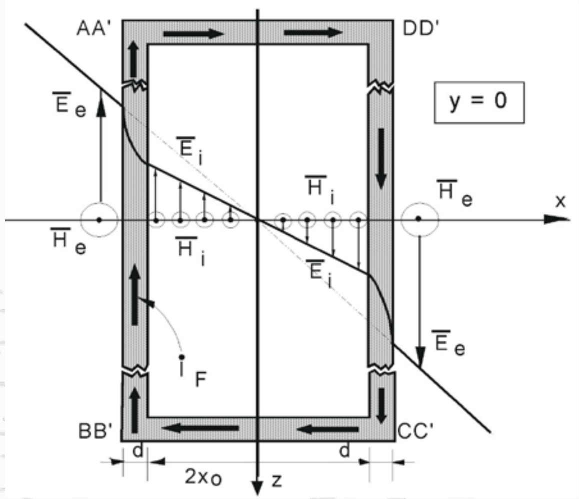
$$\delta=\delta_s=\sqrt{\frac{2}{\omega\sigma\mu}}$$



2. The incident field is inducing current in the shield, which creates a field opposing the incident field, and therefore a reduced field inside.
3. Spherical shield







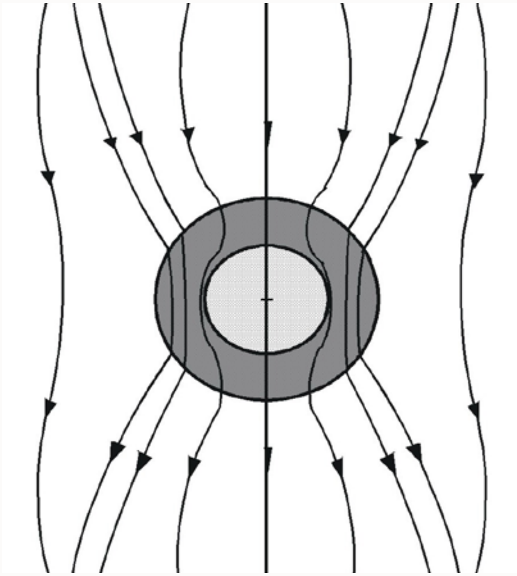
$$S.E.=10\log\left[\frac{1}{2}\left(\cosh\frac{2d}{\delta}+\cos\frac{2d}{\delta}\right)+\left(5A^2+2B^2-2AB\right)\frac{1}{2}\left(\cosh\frac{2d}{\delta}-\cos\frac{2d}{\delta}\right)+A\left(\sinh\frac{2d}{\delta}+2\sin\frac{2d}{\delta}\right)+B\left(\sinh\frac{2d}{\delta}-\sin\frac{2d}{\delta}\right)\right]$$

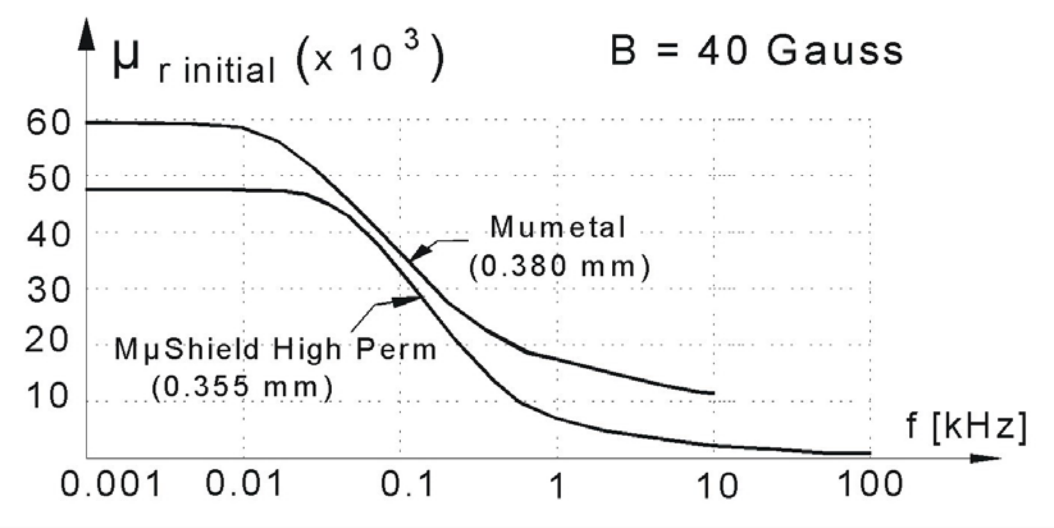
$$A=\frac{\mu\delta}{3\mu_0 r_0}\qquad B=\frac{\mu_0 r_0}{3\mu\delta}$$

Analytical Approaches to Shielding

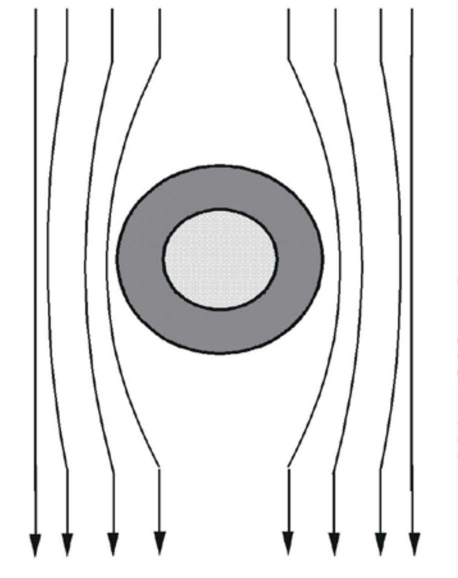
1. While they are of little practical use, these methods are intuitive and provide some **physical** insight about which parameters may be important for shielding, how multilayer shields may behave, how the flow of current may be affected by a slot or an array of perf. Calculations using these analytical methods will be way off in a practical case.
2. For example, for H field, from the analysis of the spherical shield, if the metal of the shield is non-magnetic, at LF will not modify the structure of the exterior field and will not provide shielding. However, for a magnetic material at LF the tangential component of the field is neglectable while there is a strong radial component, and the field lines are practically perpendicular on the surface of the shield, providing shielding. At HF (above few MHz) magnetic materials behave as non-magnetic materials and for both magnetic or non-magnetic the radial component disappears and only the tangential component is present, like the field is going around the shield.

LF magnetic



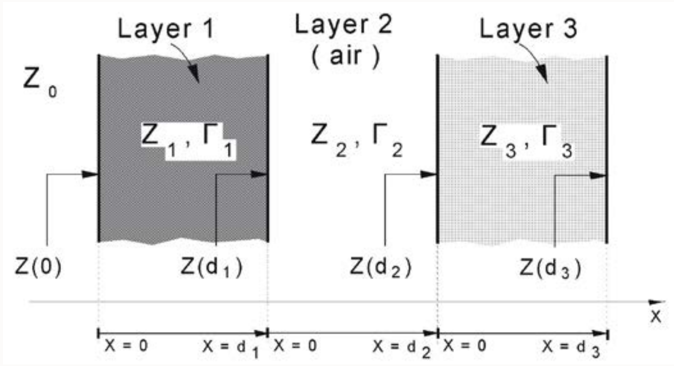


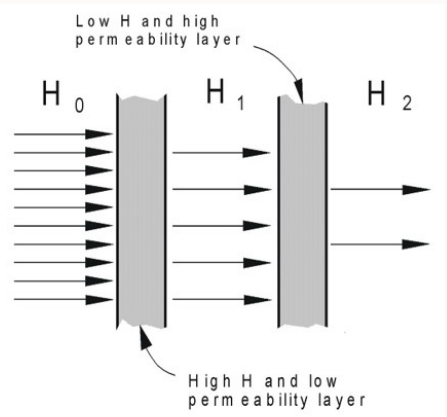
HF magnetic or non-magnetic



Analytical Approaches to Shielding – Multilayer Shields

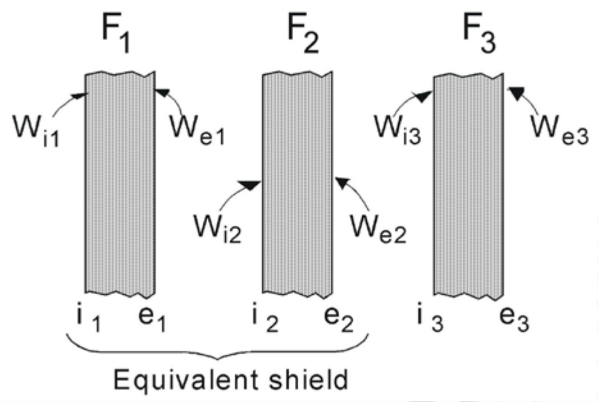
1. The TL impedance method provides ways to estimate the effect of a multi-layer shield, like a base metal with a coating, by simply cascading TLs with different intrinsic impedances.
2. For a magnetic shield, the layer toward the excitation is typically with lower μ_r and higher H field saturation, while the other layer is with higher μ_r . Dielectric layers in between help.
3. Field theory method also helps to build models for multilayer shields. For example, in LF a double shield from materials with the same magnetic behavior (Cu + Al) is worse than a single shield with equivalent thickness. If the materials have different magnetic behavior the double shield is 6dB better (for example Cu+ Ni). In general, equal thickness for the two layers is best.
4. The triple shield works best with materials of equal thickness and alternating magnetic behavior, with the most conductive layer toward the noise source.
5. A triple layer shield is evaluated by using repeatedly the two layers shield formalism.
6. There is a difference between laminated shields and double shields (air or dielectric separation).

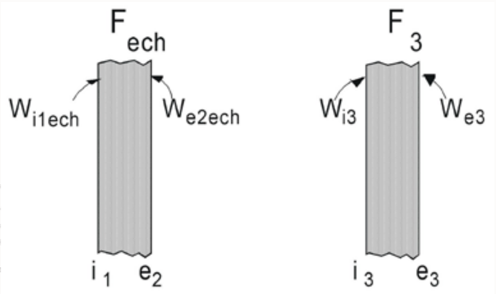




$$SE = 20\log\frac{1}{|F|}$$

W = reaction factor





(W - how much the exterior field is distorted by the currents in shield wall)

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Limits of the Analytical Methods and Far Field Estimations

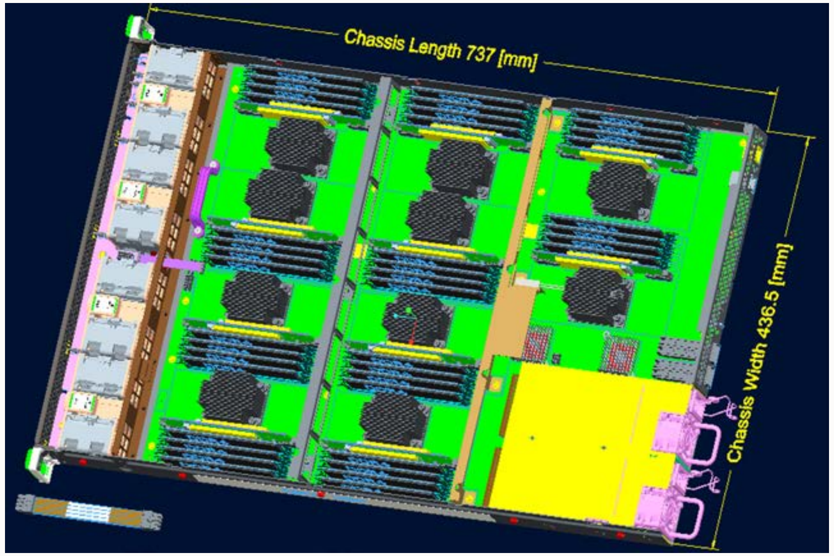
1. Schelkunoff’s method applies for an infinite sheet of metal, Kaden’s method is limited to only 11 canonical geometries.
2. They provide some physical insight and precision, good benchmark to evaluate a numerical method or software.
3. Numerical simulations based on FDTD, FEM, MoM, EFIE, etc. can be used for simplified situations, to check a trend, a principle, an A to B comparison, but not for real life prediction, because of complexity and the poor characterization of the sources and coupling paths.
4. Experimental methods are the only real validation of the analytical or numerical models and methods.
5. Eventually, all these methods are about material type and thickness, and to little extent about geometry.
6. For most real-life chassis in IT, a 0.8 mm steel will be enough, the real performance is mostly related to geometry, slots, apertures, perf areas, gasketing, etc. A typical chassis rarely has SE>25 dB in the whole frequency range.
7. If the field in an aperture is estimated correctly, the far field can be calculated using the field equivalence principle (Huygens’ principle), some software packages attempt to do this. For experimental validation, it is very difficult to measure the amplitude and phase of the field in an aperture across a wide frequency bandwidth.
8. Other methods try to measure the Q of the enclosure and use the data to estimate the field in the apertures and the far field but is quite inaccurate across the frequency range of interest.
9. Analytical formulations for slots, or radiation from holes (H.A. Bethe, 1944, theory of diffraction through small holes) are combined with the 3D methods in order to reduce the meshing and increase the precision.
10. Numerical methods are excellent to check the relative effect of a change, but don’t accurately predict the far field for a real complex product over the frequencies of interest. Commercial EMC in IT is very much a wide band problem as the system is designed and tested for radiated emissions in the 30 MHz – 40 GHz range.

Difficulties with Numerical Methods and Far Field Estimations

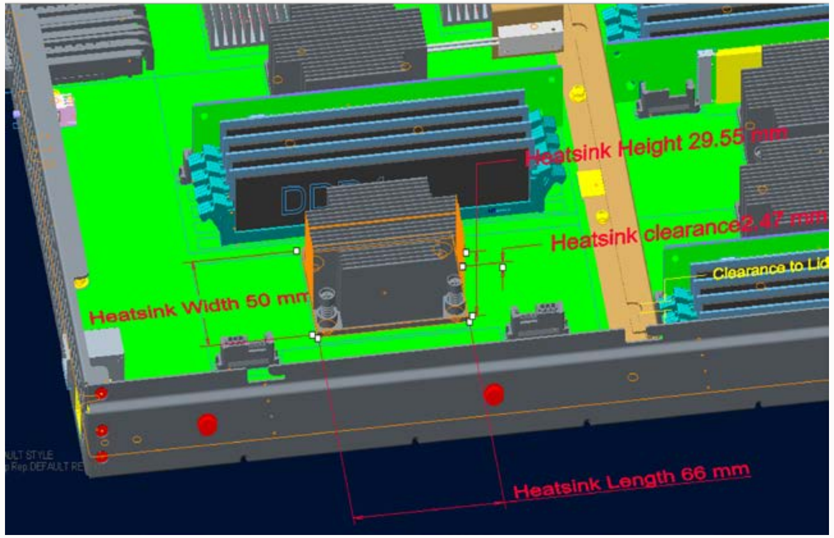
1. Using numerical methods for the entire complex system is very difficult, even for relatively simple systems like a two-socket server. The thermal analysis model doesn't include the correct materials or all the relevant details for EMC, and the mechanical model is too complex to use directly even on a powerful workstation with hardware acceleration.
2. There are HPC installations which can handle a complete system simulation, but the extreme complexity and poor characterization of internal noise sources from an EMI signature point of view limits their use. The results can't be an accurate prediction, but only A/B comparisons based on various internal changes or assumptions about sources frequency, strength, polarization and various coupling paths. Including all system level parasitics is practically impossible.
3. If the entire mechanical assembly file (like *.asm in PTC Creo) is imported directly, the model will have difficulties healing, and the simulation will take a very long time because irrelevant mechanical details will force very fine meshing and will need a lot of memory even if adaptive meshing is used.
4. Simplification of models to capture only electromagnetically meaningful features is needed. One possibility is to simplify the complete assembly file first in its native software (PTC Creo, Solid Works, etc.), removing details irrelevant for EMC, and export the simplified model as a .sat file in the 3D full wave simulation software (CST, Ansys, Remcom, etc.). One difficulty is to have an EMC engineer fluent in both mechanical and EMC software tools.
5. The alternative approach is to build the whole model in the 3D full wave tool, geometry, sources, etc. This is possible, but extremely time consuming, especially for projects in development, which change constantly.
6. The typical approach is to build models for limited problems of interest in the system and do what-if scenarios with the variables available from a design point of view.
7. Validating the model experimentally is another difficult step, especially in pre-P0 phase, before the electronics is available and when feedback is needed.

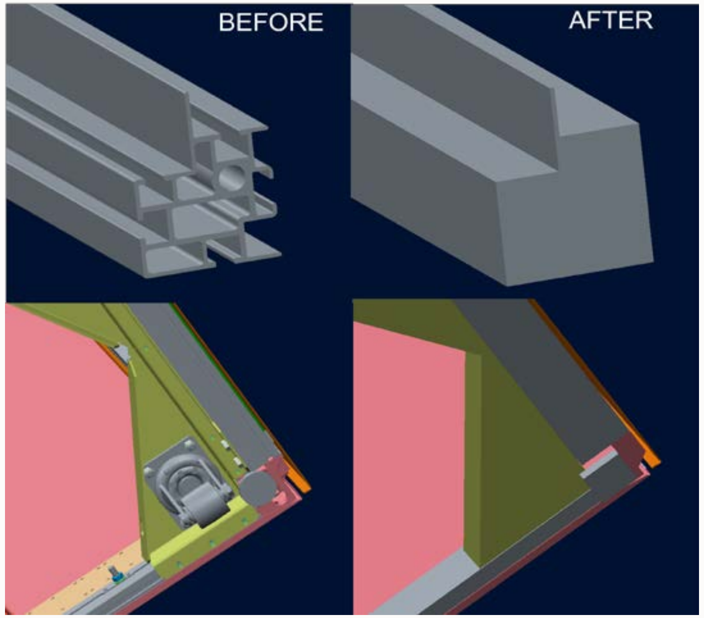
Difficulties with Numerical Methods and Far Field Estimations

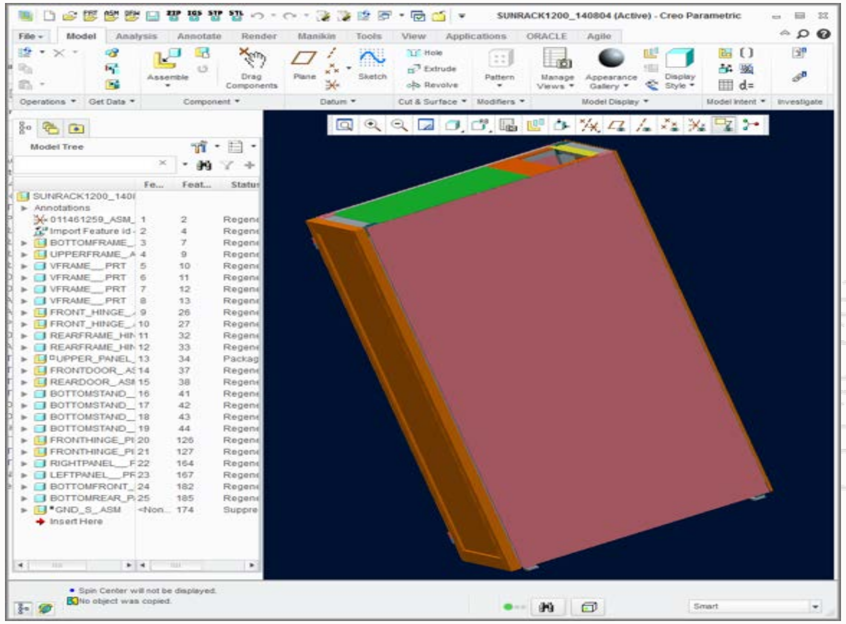
1. A model built directly in CST (right) and looking at heatsink coupling to an aperture nearby.
2. A model simplified in PTC Creo, exported as .sat file and imported in HFSS (below). Filets, small holes, screws, wheels, etc., that don't affect EM analysis in the frequency range of interest are removed.

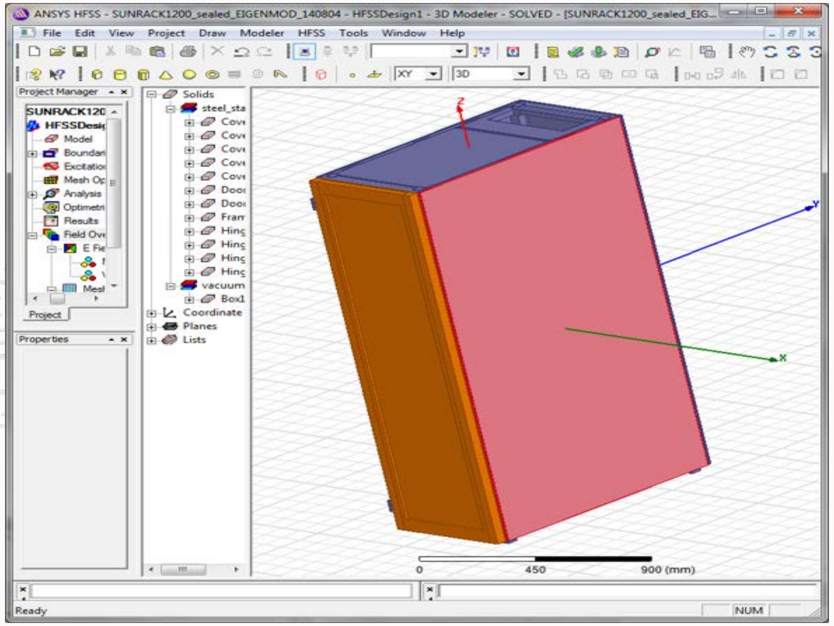


CPU heatsink cavity dimensions



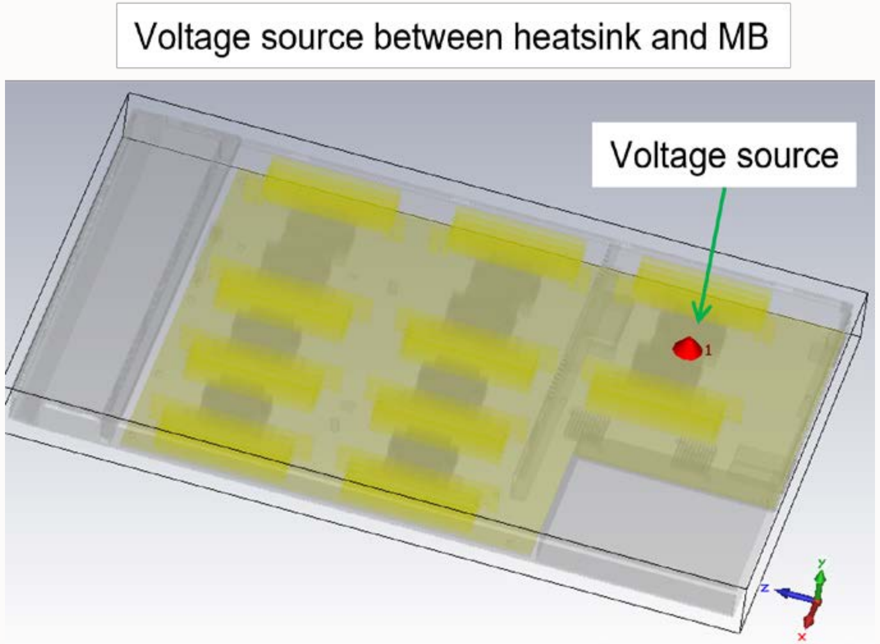


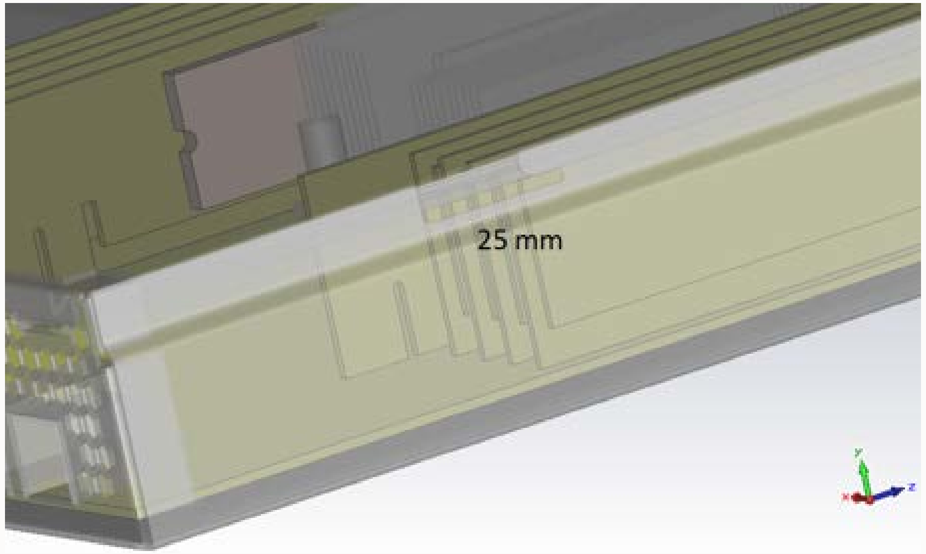


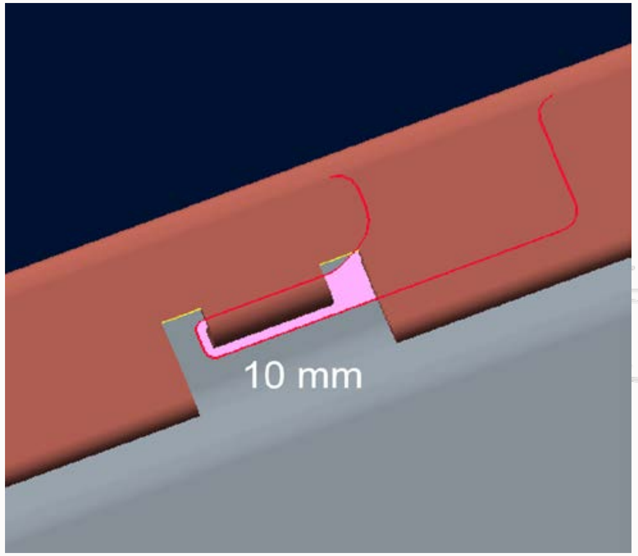


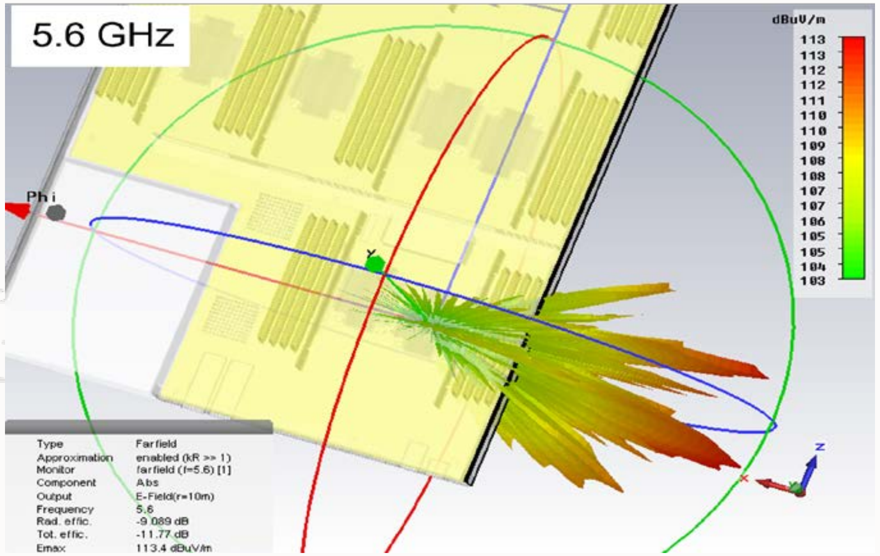
Difficulties with Numerical Methods and Far Field Estimations

1. A CPU heatsink close to a 25 mm side wall slot (lid) is excited with a voltage source toward PCB, and E-field probes are placed at the slots on the chassis walls. Heatsink (cavity) resonances at 5.6 GHz and 6.5 GHz couple to the slot, leading to radiated emissions.
2. The 25mm long slot on the side wall is λ/2 resonant at 6 GHz, close to the 5.6 GHz and 6.5 GHz internal resonances.
3. Based on the results, it was suggested to reduce the length of the slot to 10 mm, which is λ/2 resonant at 15 GHz, far away from the frequencies detected with this simplified model and the CST simulation.


Voltage source between heatsink and MB







```
Type            Farfield
Approximation   enabled (kR >> 1)
Monitor         farfield (f=5.6) [1]
Component       Abs
Output          E-Field(r=10m)
Frequency       5.6
Rad. effic.     -9.089 dB
Tot. effic.     -11.77 dB
Emax            113.4 dBuV/m
```

Shielding and Reciprocity

1. Reciprocity principle is applicable only to **passive** and **linear** circuits (CPUs, ASICs, IC are active and non-linear).
2. Reciprocity requires to consider only those parts of the system that meet the criteria of "passivity" and "linearity". Within an electronic system, the properties of radiators/sensors can be often assumed to be reciprocal, just as is the case with antennas.
3. In shielding practice, we can find non-linear effects, like the "rusty bolt" effect, resulting in nonlinear and non-reversible conductivity (diode). Another typical LF example is a multilayer shield, with one of the layers having nonlinear magnetic permeability.
4. If the shield material properties are linear, the reciprocity principle requires that $Z_{21}=Z_{12}$. The same requirements hold for passive, linear, circuits. If the shield is a passive network and its material exhibits linear electric and magnetic properties, the reciprocity theorem holds, and the same shielding parameters are true for emissions and susceptibility. However, this is true only at the levels of the **shield transfer function** as a frequency dependent graph, but not at the **system** level frequency dependent shielding effectiveness.
5. For example, a dominant source of emissions in the system may have a high level of signal and is not necessarily the dominant victim in the same system (immunity case). A Tx (transmitting) output signal is 20 dB stronger than the Rx (receiving) input signal, and the Tx will be more immune to EMI.

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Shielding and Chassis Resonances

1. For illustration, consider an empty chassis which behaves as a rectangular cavity and the equations for the field in the cavity are derived from the equations for a rectangular waveguide with specific boundary conditions.
2. RF cavities are linear and time-invariant (LTI) systems and allow to separate the time dependence out of Maxwell's equations (time-invariant) and to use superposition (linear).
3. To simplify the solution, consider that the cavity is entirely closed by a perfect electric conductor (PEC) and the cavity volume does not contain any lossy material. There would exist solutions to Maxwell's equations with non-vanishing fields even without any excitation.
4. These eigen solutions are the cavity resonant modes, and each mode is characterized by its (eigen-)frequency and its characteristic field distribution inside the cavity. The mode with the lowest frequency is the **fundamental mode**.
5. For a cavity with W, H cross-section and length L the formula for eigenfrequencies, where at most only one of the indices i, j or k is zero, and approximated for vacuum is included below.

$$f_{i,j,k} = \frac{c}{2\sqrt{\mu_r \varepsilon_r}} \sqrt{\left(\frac{i}{W}\right)^2 + \left(\frac{j}{H}\right)^2 + \left(\frac{k}{L}\right)^2} \qquad f_{i,j,k} = 150\sqrt{\left(\frac{i}{W}\right)^2 + \left(\frac{j}{H}\right)^2 + \left(\frac{k}{L}\right)^2} \quad \text{(m, MHz, vacuum)}$$

6. The energy stored in the lossless cavity is the sum of the electric field energy and magnetic field energy, constantly swapping back and forth between them at twice the frequency of the mode; with E and H in quadrature the sum is constant in time but will depend on the air volume (V) of the empty cavity. At resonance W_E=W_M in the cavity.

$$W_E = \iiint_{cavity} \frac{\varepsilon}{2}\left|\vec{E}\right|^2 dV \qquad W_E = \frac{1}{8}\varepsilon E_0^2 V \cos^2(\omega t) \qquad W_M = \iiint_{cavity} \frac{\mu}{2}\left|\vec{H}\right|^2 dV \qquad W_M = \frac{1}{8}\varepsilon E_0^2 V \sin^2(\omega t) \qquad W_{cavity} = W_E + W_M = \frac{1}{8}\varepsilon E_0^2 V$$

Shielding and Chassis Resonances

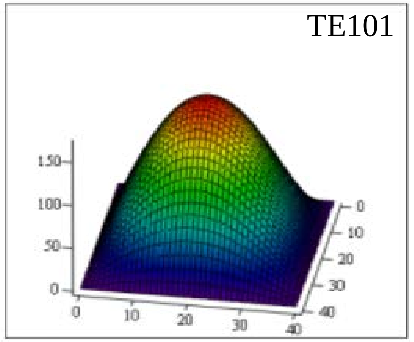
1. If the cavity walls are made of a good rather than a perfect conductor, or if the cavity is open (apertures), modes still exist and are useful to characterize the cavity, but their eigenfrequencies will become complex, describing damped oscillations, so each mode will be characterized by its frequency and its decay rate. If the field amplitudes of a mode decay as $\sim e^{-\alpha t}$, the stored energy decays as $\sim e^{-2\alpha t}$. If P_{loss} is the power lost into the cavity walls (or any other loss mechanism), then the same power must be supplied to the cavity to keep the stored energy at a constant value W_{cavity}. The larger the Q, the smaller will become the power necessary to compensate for cavity losses and excite the cavity at a given mode. The quality factor Q:

$$Q = \frac{\omega_0 W_{cavity}}{P_{loss}} = \frac{\omega_0 W_{cavity}}{-\frac{dW_{cavity}}{dt}} = \frac{\omega_0}{2\alpha}$$

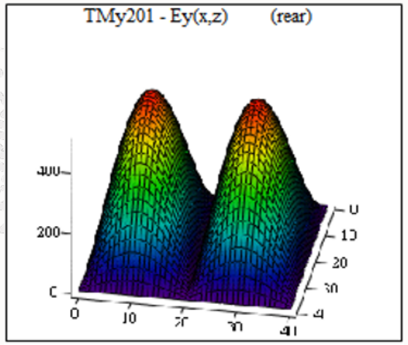
2. An empty cavity may have exceptionally high Q, over 10,000, a problem in simulations with PEC (convergence), and some losses must be artificially inserted, like a 50 Ω serial with a small monopole antenna excitation.
3. Real life chassis have much lower Q, typically in the 5-25 range, because of the loss in the conductive walls and in all the components inside the chassis, that also reduce the available volume for the cavity stored energy.
4. The Q factor tends to increase with the order of the resonance mode, because there are more internal peaks and nodes of the field, but the losses are still only at the walls.
5. Consider an empty typical server chassis L=800 mm, W=533.4 mm (21”) and H=88.9 mm (3.5”=2 RU), without any internal compartmentalization. The resonances will be well separated in frequency below 1-1.5 GHz, but will become increasingly close to each other for higher order modes: 337.75 MHz (101), 468.40 MHz (102), 592.45 MHz (201), 628.44 MHz (103), 675.51 MHz (202), 794.89 MHz (203), 863.63 MHz (301), 922.59 MHz (302), 1013.27 MHz (303), etc. At higher frequency, the cavity is over-moded and the effect of a specific mode is difficult to study.

Shielding and Chassis Resonances

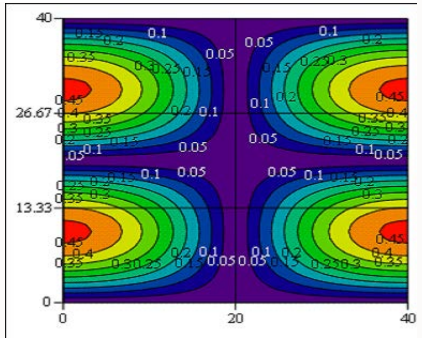
1. The field distribution for a mode is complex and not uniform inside the chassis, therefore any shielding effectiveness number will be frequency dependent and highly dependent on the position inside the cavity.
2. Different sources can excite different modes inside a resonant chassis, and they will radiate through slots, apertures, and perf areas.
3. If there is a high Q chassis resonance, the far field radiation will be dominated by the frequency of this resonance, even if the slots are not resonant at that frequency, if the mode can couple to the slots.
4. An internal source of the right type and polarization may easily excite a resonant mode if placed in a max field position for that mode.
5. If a slot is not excited and is not radiating for a certain resonant mode, it may be excited and radiate for another resonant mode.
6. Using internal compartmentalization, the eigenfrequencies of the chassis will be much higher, maybe outside the bandwidth of a certain internal source which may excite a resonance.
7. A typical server chassis has significant internal compartmentalization (PSUs, Fans, I/O areas, etc. with various walls breaking potential resonances. There is very little empty space in a modern system, having large internal loss from all the components, therefore the Q is very low, in the 5-10 range or lower.



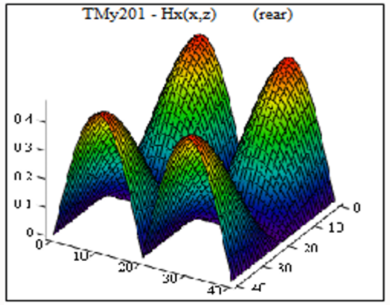
EyXZ



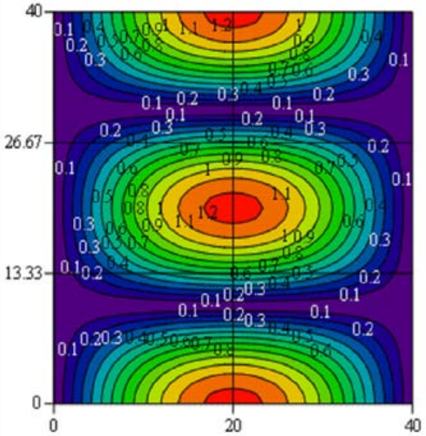
EyXZ



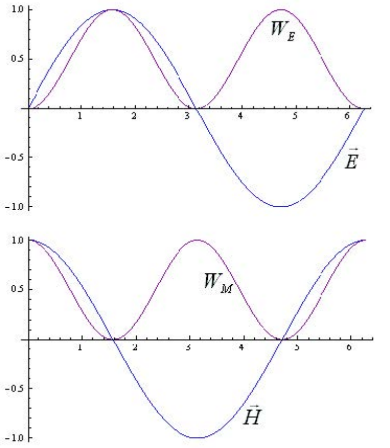
HxXZ



HxXZ



HzXZ

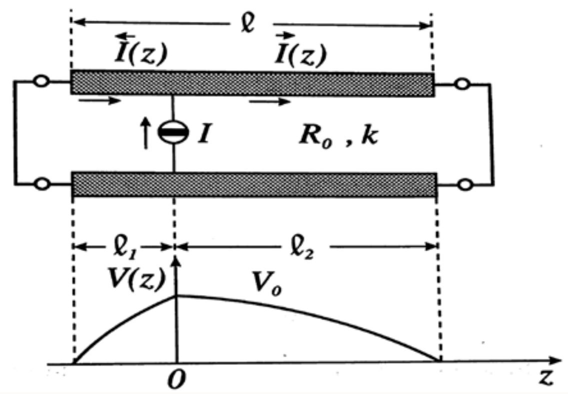


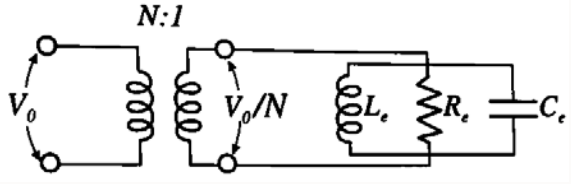
Shielding and Cavity Losses, Absorbers

1. The cavity at or near the resonance can be modeled as a TL short-circuited at its ends and driven by a current generator I. An equivalent resonant parallel circuit can be derived, and the parameters to model the cavity around the resonance can be obtained. The circuit includes an ideal V transformer N:1 to account for the source-cavity coupling coefficient.
2. Slightly off-resonance the stored electric and magnetic energy are no longer equal, and the noise source provides the reactive power to compensate for the stored energy imbalance.
3. The field distribution at resonance for the lossless cavity is typically used to compute the conductive, dielectric or magnetic loss in its lossy elements.
4. The dielectric loss in the cavity is ~f (for 30 MHz – 40 GHz) and ~ε'', the imaginary part of the permittivity associated with the dissipation of energy within the dielectric medium, while the real part ε' is associated with the stored energy within the medium.
5. Similarly, the magnetic loss in the cavity is ~f (for 30 MHz – 40 GHz) and ~μ'', while the real part of the permeability μ' is associated with the stored energy within the medium.
6. The conductive loss are ~σ and ~√f considering the skin effect. Total power loss per unit volume for a harmonic field is:

$$P = P_\sigma + P_\varepsilon + P_\mu = \frac{1}{2}\sigma(\omega)|E|^2 + \frac{1}{2}\omega\varepsilon_0\varepsilon_r''(\omega)|E|^2 + \frac{1}{2}\omega\mu_0\mu_r''(\omega)|E|^2 \quad [\mathrm{W/m^3}]$$

7. Absorbers can be used to detune a cavity or detune a noise source inside the chassis. Light carbon loaded foam absorbers are easier to use and available for wide frequency ranges. Ferrite loaded and magnetic absorbers are also available but can be brittle. There are thermal interface materials (TIM) which can be a chip-heatsink absorber in some cases.





$$k_n = n\frac{\pi}{l} \qquad \sin kl = 0 \ \ \text{(resonance)}$$

$$N = \sin k_n l_2$$

$$\varepsilon^* = \varepsilon' - j\varepsilon'' = \varepsilon_0\left(\varepsilon_r' - j\varepsilon_r''\right)$$

$$\tan\delta = \frac{\varepsilon''}{\varepsilon'} = \frac{\varepsilon_r''}{\varepsilon_r'}$$

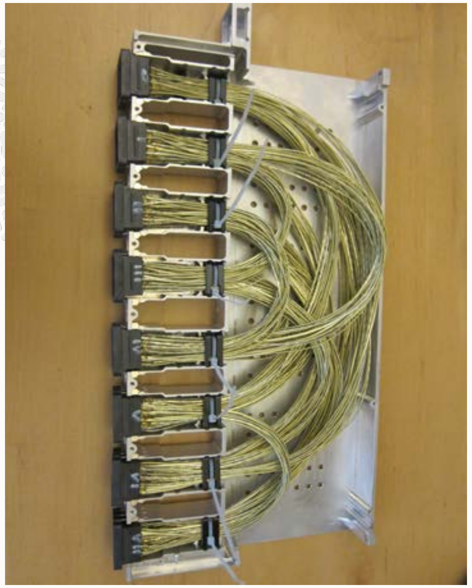
$$J_{tot} = j\omega\varepsilon(\omega)E = j\omega\varepsilon'(\omega)E + \omega\varepsilon''(\omega)E$$

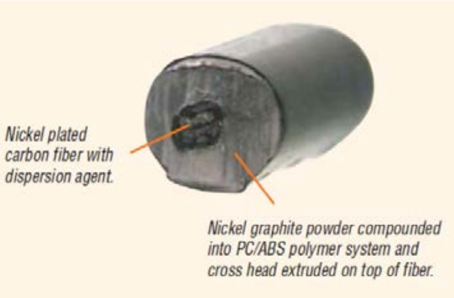
$$\frac{dP_{loss}}{dV} = \frac{1}{2}\mathrm{Re}\left[J_{tot}E\cdot E^*\right] = \frac{1}{2}\omega\varepsilon''(\omega)|E|^2$$

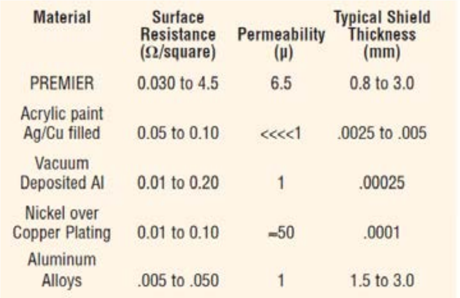
$$\delta_s = \sqrt{\frac{2}{\omega\sigma\mu}} \qquad R_s = \frac{1}{\delta_s\sigma} \ \ \text{(surface resistivity)}$$

Shielding and Absorbers

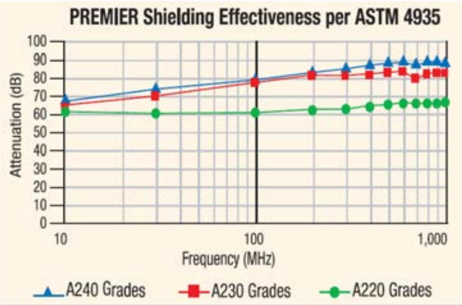
1. An M8 system uses twinax connections for connectivity at 19.2 Gbps. Relatively large DM/CM conversion in the connectors resulted in 19.2 GHz (λ=1.56 cm) radiation through the plastic handle, even if shielded inside. Using Chomerics Premier absorber material solved the issue. It is a blend of PC/ABS thermoplastic polymer alloys and conductive fillers with stable electrical, mechanical, and physical performance. The conductive filler utilizes nickel plated carbon (Ni-C) fibers as base.
2. In a storage system a layout mistake from the HDD vendor created a strong 12 GHz (SAS3) emission, still OK for a single system, but too much for a full storage rack sold as an appliance. The solution was to use a small absorber (ARC-Tech) on the SAS connector of each drive. This flexible, loaded, silicone, rubber microwave absorber entirely solved the 12 GHz issue, with some extra cost.

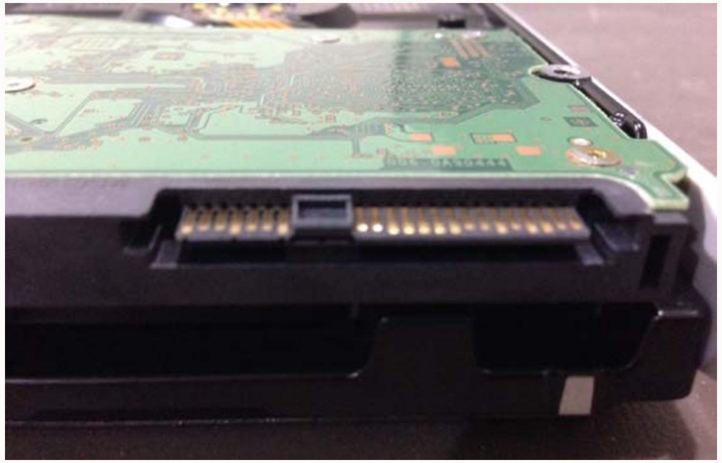






Material	Surface Resistance (Ω/square)	Permeability (μ)	Typical Shield Thickness (mm)
PREMIER	0.030 to 4.5	6.5	0.8 to 3.0
Acrylic paint Ag/Cu filled	0.05 to 0.10	<<<<1	.0025 to .005
Vacuum Deposited Al	0.01 to 0.20	1	.00025
Nickel over Copper Plating	0.01 to 0.10	≈50	.0001
Aluminum Alloys	.005 to .050	1	1.5 to 3.0





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Shielding and Aperture Coupling Rules

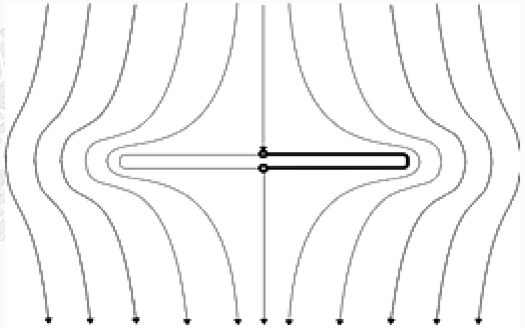
1. Solving practical shielding problems means understanding the potential internal sources (frequency, strength, potential field distribution) and their relation to the chassis apertures (characteristics, resonance, polarization, etc.).
2. A realistic chassis may radiate power through slots, apertures and perforated panels necessary for functional and thermal reasons. However, the power loss through radiation is too small to affect the Q of an internal resonance. For FCC class A radiated emissions limits, the Effective Radiated Power (ERP, relative to an ideal dipole) from the EUT is between -46.92 dBm to -37.38 dBm, or 0.03 µW - 0.30 µW (!). About 0.5 µA on a cable will fail FCC class A.
3. The two main aspects to look for are: the effect of chassis resonances, and the effect of direct coupling from a potential internal noise source to a potential antenna (slots, apertures, perf, cables, etc.).
4. For each chassis resonance, the maximum energy is at the fundamental (lowest) frequency of resonance. For the higher modes, the energy decreases with the order of the mode. From a practical point of view, for a typical 19” or 21” server chassis, the most critical internal resonances will be below 1 GHz, and to be on the safe side we may extend this to 3 GHz. The exact frequency will depend on the box dimensions. The energy at the very high modes, above 10 GHz to be small enough that the chassis resonances will not be a problem (for normal sized).
5. One may use 3 GHz (λ=10cm=4”) as upper limit for calculating the riveting (tux) pitch for the areas which are **not in the path of a direct coupling** to an internal source, in order to account for potential chassis resonances. Therefore, 1” between rivets will be safe, and the pitch can be increased if the risk of resonance is low (and material flatness good).
6. For near-field direct coupling between an internal source and an aperture over an unobstructed direct path, the most important parameter is the **source-aperture distance**.
7. For any significant internal source, the distance to the closest aperture should be **at least 3-5 times the longest dimension of the aperture**, or to close the aperture with a gasket or the path by internal compartmentalization.

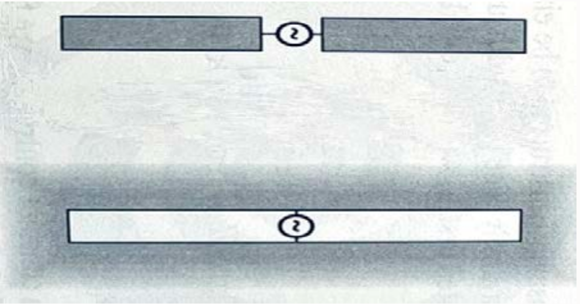
Shielding and Slot Radiation

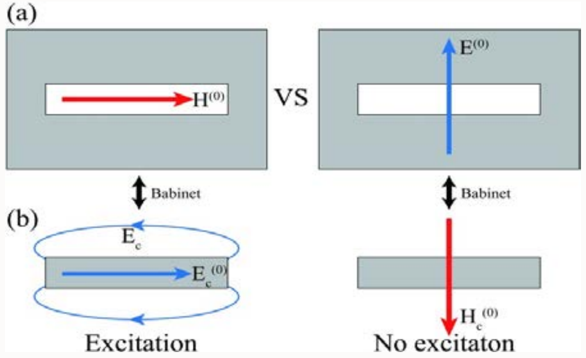
1. The field inside the chassis from the internal noise sources will induce currents on the wall of the chassis which will return to the source following the least impedance path.
2. A slot is forcing a different current flow and a voltage drop V=ZI across the slot impedance Z , hence a field E=V/t across it. The slot will radiate, behaving like its complementary dipole rotated 90 degrees, based on Babinet's principle (E, H directions interchanged) .
3. A dipole is resonant when its length l= λ/2. A ribbon dipole, with width "t", is resonated by reducing its length slightly below a half-wavelength, and its impedance is Z_{dipole}=70 Ω [6].
4. The impedance of a slot at resonance can be calculated from the general formula, in this case resulting 506.8 Ω. If the slot and the equivalent dipole are self complementary and one can exactly overlay its complement through translation and/or rotation, their impedance will be equal, and the slot impedance in this case will be only 188.35 Ω.

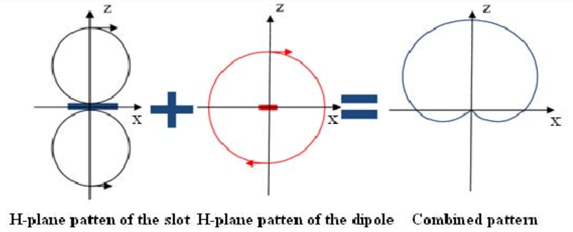
$$Z_{metal}Z_{air} = \frac{Z_0^2}{4} = \frac{(376.7)^2}{4} = 35{,}475.7 \;\; \Omega \qquad Z_{slot} = \frac{Z_0^2}{4Z_{dipole}} = \frac{35{,}475.7}{70} = 506.8 \;\; \Omega \qquad Z_{metal} = Z_{air} = \frac{Z_0}{2} = 188.35 \;\; \Omega$$

5. The combined field from the slot and its complementary dipole will create in far field the same field as the original source, without slot or dipole (Babinet).
6. The practical implication, a horizontal slot will radiate in V-polarization, and a vertical slot will radiate in H-polarization, useful to remember while troubleshooting an emission.
7. The various types of fabric over foam gaskets used to close a slot have typically a surface resistivity of 50-100 mΩ/□, therefore some voltage drop may still exist across a gasket.

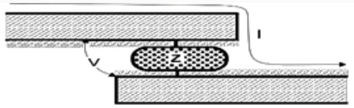


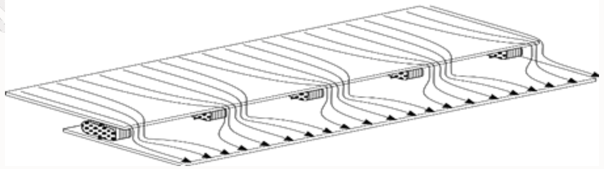


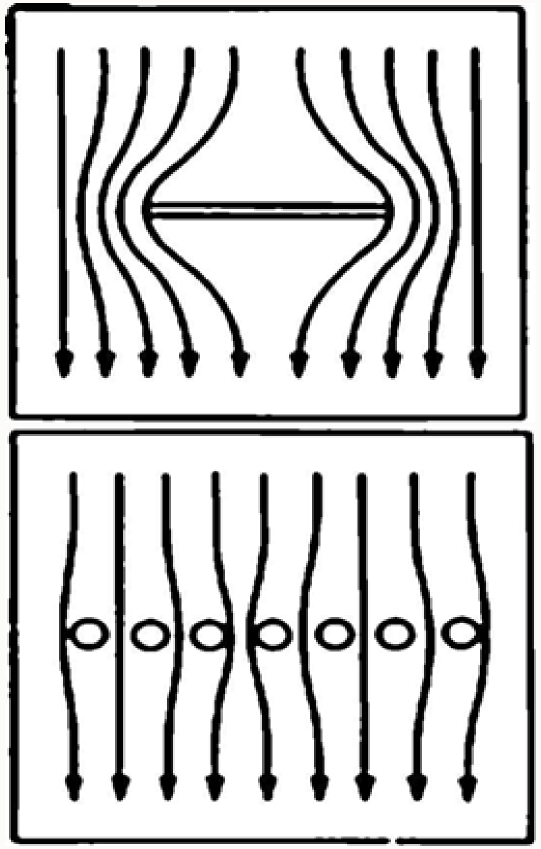




Shielding and Slot Radiation







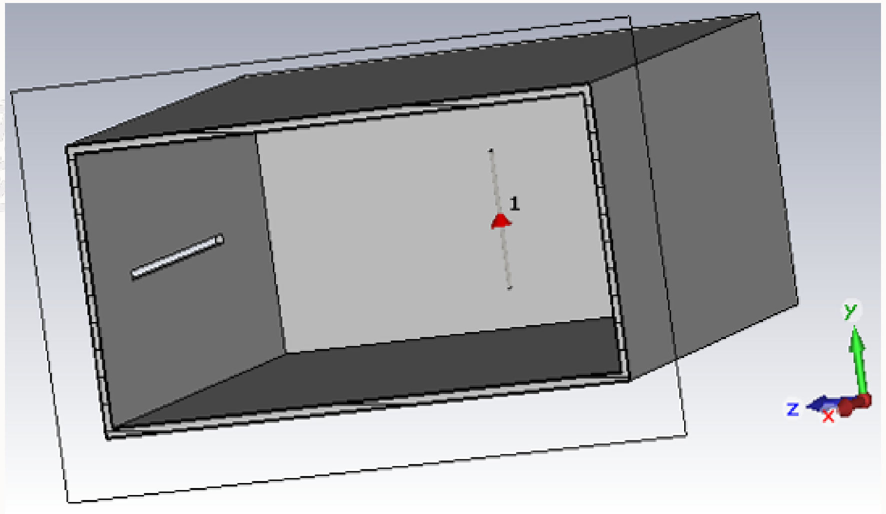
1. For a slot excited by a specific internal frequency close to the λ/2 slot resonance, the simplest solution is to cut the length of the slot with a gasket placed in the middle, where the max E field is. Usually, the entire slot is gasketed because there are other internal frequencies driving shorter slots, and because even below the slot resonance will be some slot radiation.
2. The currents induced in the chassis wall will return to the source on the path of least impedance, and this will be the path least disturbed by the presence of long slots.
3. Multiple holes of equal size, closely spaced, at less than λ/2 of each other are a much better solution than a long slot with the same open area.
4. If a λ/2 resonant slot provides zero SE, a shorter slot will still radiate but will provide some SE [4]:

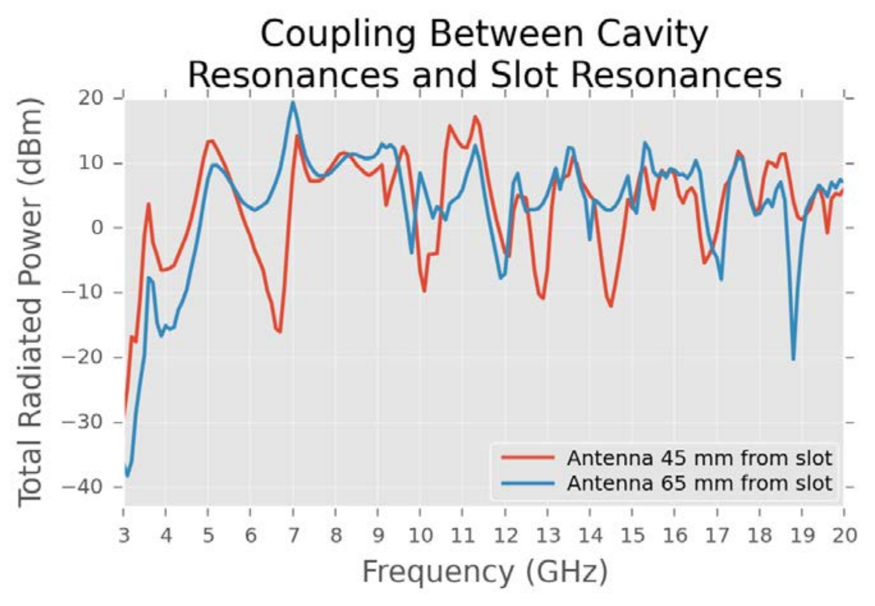
$$S.E._{slot} = 20\log\left(\frac{\lambda}{2l_{slot}}\right) = 20\log\left[\frac{150}{f_{MHz}l_{slot\ \mathrm{m}}}\right]\ \mathrm{dB}$$

5. For example, a 15 cm slot is resonant at 1 GHz, but it provides an attenuation of 1.93 dB at 800 MHz, 4.43 dB at 600 MHz, 6 dB at 500 MHz, therefore not much.
6. A relatively long slot may be fine in a certain design if is not driven by the internal source because is not close enough to couple, doesn't have the necessary polarization, frequency or strength. Leveraging the same experimentally proven "good" chassis to a different design with different internal sources can be a problem.
7. The safe approach is to treat every slot as potentially driven by an internal source.

Shielding and Slot Radiation

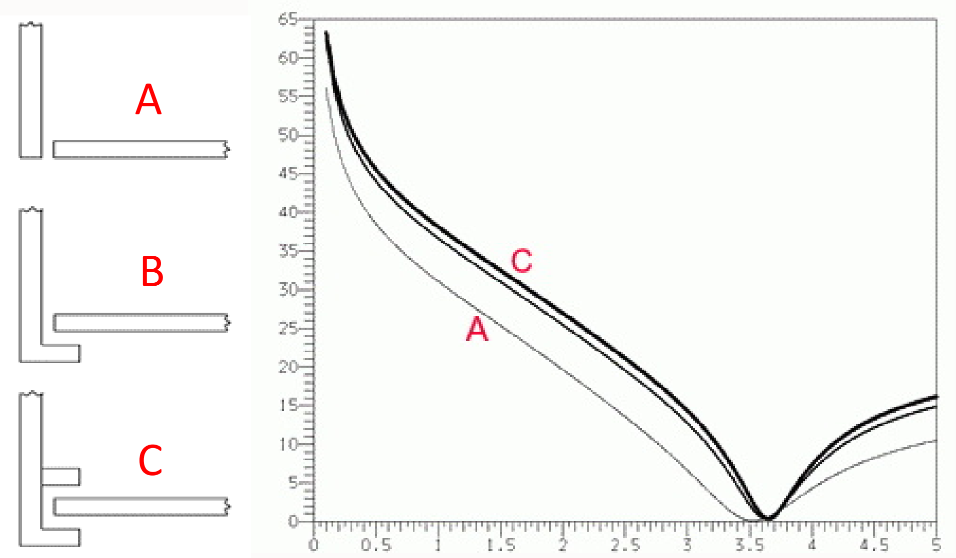
1. The presence of a slot in a chassis greatly affects the fields inside very close to the slot (<10 mm), but the fields inside the box further away from the slot are less affected except at some of the cavity resonances. The general modal field patterns inside the chassis are relatively unaffected by the presence of the slot (little energy radiated).
2. The air gap thickness of slots and overlapping seams are important in determining the shielding effectiveness. Doubling the air gap width leads generally to about 3 to 6 dB increase in the radiation through the slot. Less increase may occur around the slot resonance.
3. In this model (right) cavity mode-based coupling appears to dominate direct coupling between the slot and the internal dipole. The radiated power is maximized when the dipole is positioned in such a way that the modes strongly excited lead to voltage drop across the slot.
4. For example, let’s consider a 50×30×70 mm empty box with 1 mm thick walls (σ=1000 S/m). The box has a 25 mm long and 1 mm wide slot, resonant at 6 GHz. A 15 mm long internal dipole, resonant at 10 GHz, is excited by a 50 Ω discrete port. The results in CST Studio indicate cavity mode-based coupling dominating direct coupling between the slot and the dipole for this model. The modes that lead to voltage drop across the slot are strongly-excited. All the max values in the graph are related to cavity resonances and not the slot resonance.

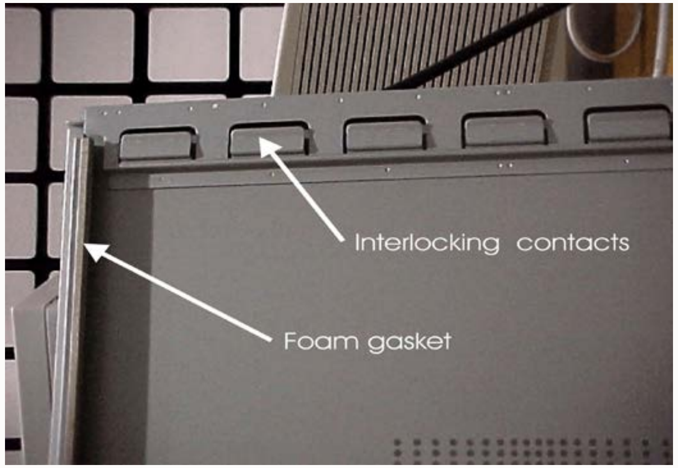


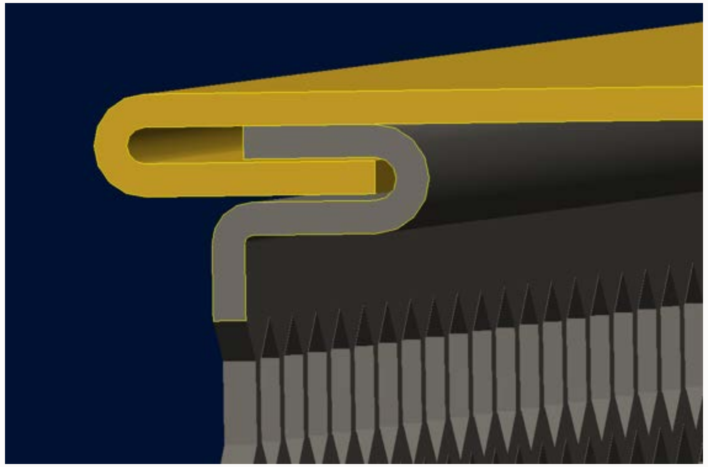


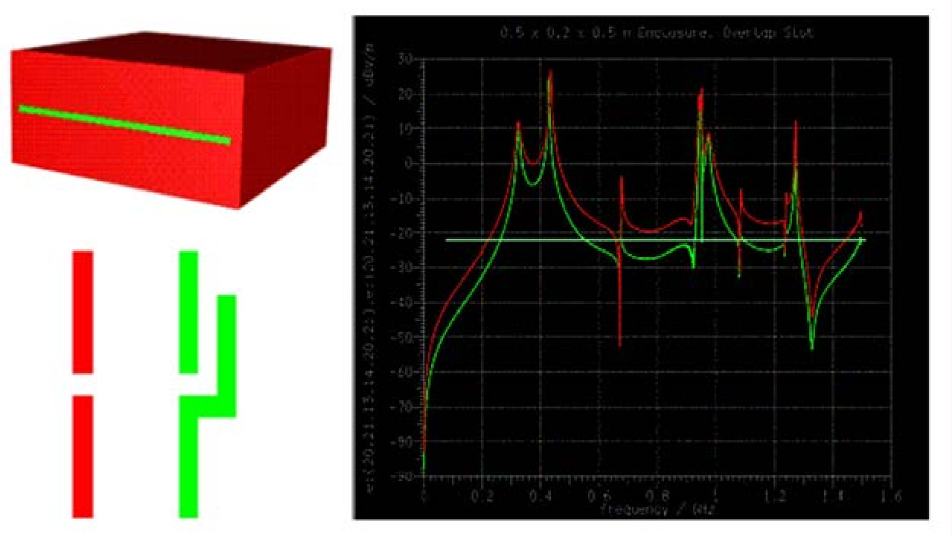
Shielding and Seams

1. For seams, the resonance frequency is still λ/2 length and will not change much with the type of the seam.
2. The attenuation below the seam resonance can be 10 dB better for overlapping seam (C vs. A). Even partial overlap helps a lot (B).
3. Overlapping seams are effective at shielding mainly due to incidental mechanical contact and/or acting as waveguides below cutoff rather than due to waveguide bends.
4. The direct source – aperture coupling is very difficult to study even in an empty box because of the cavity resonances. To simulate this in an empty box, one needs to use absorbers on all internal walls to eliminate the effect of the internal cavity resonances and multiple reflections.







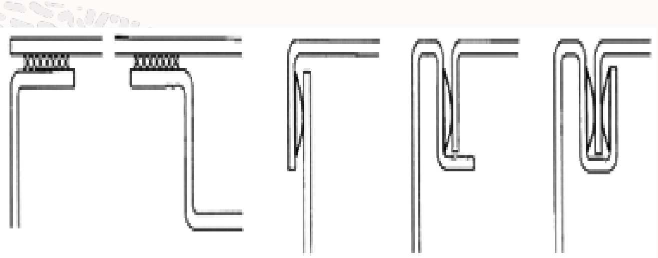


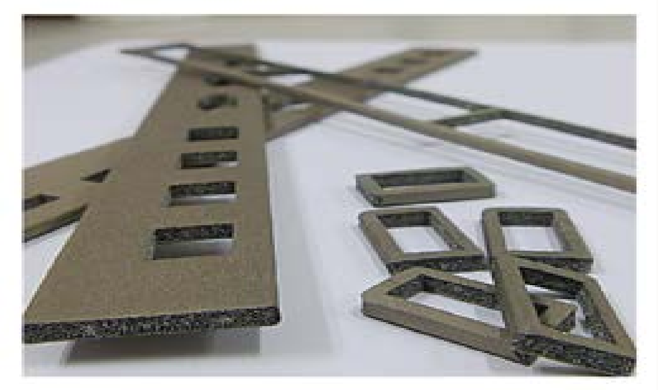
Shielding and Gaskets

1. Gaskets are critical for the integrity of the chassis, to reduce the impact of slots, apertures, seams and joints, therefore, to help pass legal EMC tests for radiated emissions, radiated immunity and ESD.
2. There is a wide variety of profiles and materials for gaskets: spring fingers, fabric over foam, conductive elastomer, conductive foam, knitted wire, woven fabric. Beryllium-Copper spring fingers are among the best (conductivity, resilience, corrosion), and still RoHS compliant. Typical surface resistivity of a gasket is 50 mΩ/□ or less.
3. Each type of gasket material requires a different minimal **compression** for max performance (lowest impedance): BeCu 4-6%, knitted wire 10-15%, conductive elastomer 25-35%, however the safest and optimal compression is 30-70%, because the correct compression must also break the 1-2 µm oxide or coating of the mating surfaces.
4. To avoid under-compression a tolerance analysis needs to be done before choosing the gasket type, profile and material. Chassis sag creates a point of minimal compression. Avoid over-compression, use compression stops.
5. It is recommended to use a groove channel for the gasket, especially in shear applications (removable modules).
6. The gasket adhesive can be conductive or nonconductive. If nonconductive adhesive is used, make sure that the compression will mushroom the gasket around the adhesive strip for continuity of the contact.
7. Spring fingers are better at breaking through the oxide or coating of the metal contact areas. For example, independent of the type of gasket, a chromate conversion coating (CrAl) requires about 20 psi for low readings of about 170 mΩ, while bright aluminum has about 10 mΩ at 8 psi. The effect is important mostly above few GHz.
8. The contact resistivity will typically increase with time (oxidation, corrosion), sometimes from 10 mΩ to 100 mΩ.
9. Nickel coating on aluminum may require up to 200-250 psi for a reading below 30 mΩ.
10. Most die-cut I/O gaskets do not have fabric continuity between the two faces, only higher conductive elastomer in between, across the Z axis.

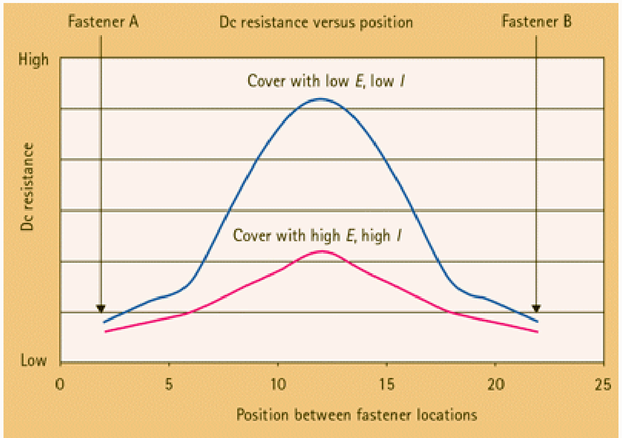
Shielding and Gaskets

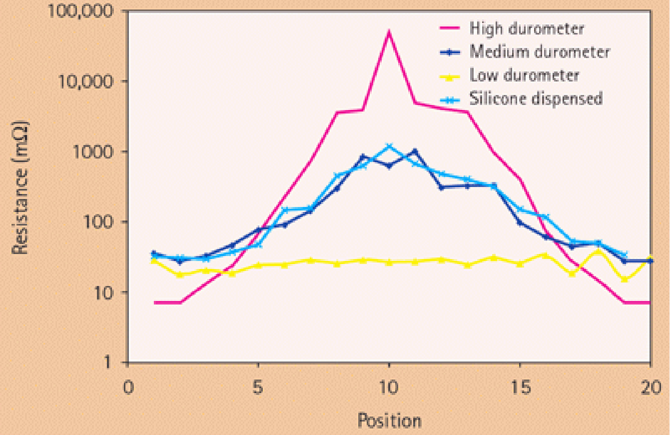
1. Constant compression on the length of the gasket is critical, and because of the dimensions of the chassis, its lid flatness and rigidity are very important. Adding stiffening cross-ribs that cover long spans is a simple way to improve the lid. Side hooks are also necessary to reduce the distance between two compression points.
2. To illustrate the importance of constant compression, consider two fasteners 30 mm apart, a PCB with 25 gold plated contacts between the fasteners, a gasket above and a piece between fasteners to simulate the cover. The actual DC resistance at each contact was measured for soft gaskets (low durometer) and harder gaskets (high durometer). The theoretical and measured values show high variation even if 30 mm between the compression points (fasteners) is actually very short.
3. If F is force (N), X is distance between fasteners (mm), E is elasticity module (steel 200 kN/mm^2, plastic 3 kN/mm^2), I is the geometry specific moment of inertia (kg m^3), then the displacement δ_{max} of a loaded beam (mm) is:





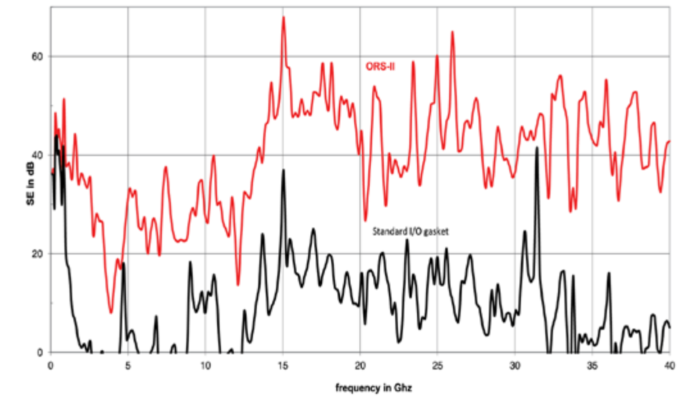
ORS-II Die-cut with fabric continuity





$$\delta_{max} = \frac{FX^3}{384EI}$$

$$\delta_{max} \sim X^3$$



Shielding and Rivets

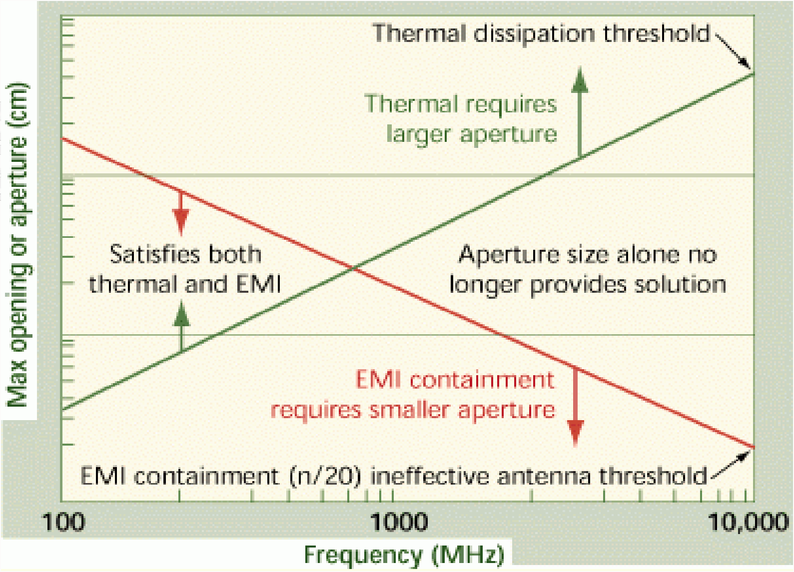
1. The pitch of the rivets will depend on their position with respect to the known noise sources. Some rivets may have a direct path to a noise source, other rivets may be out-of-sight from a noise source.
2. If in direct sight, source – rivet slot coupling is possible, and to reduce the coupling through physical separation, the pitch must be less than λ/2 and 1/5 the source to slot distance.
3. At the lower frequencies, chassis resonances are still possible, and a potential slot between rivets can be excited by the internal field.
4. The maximum internal fields are associated with the fundamental resonance and lower modes, and such a coupling from resonances to rivets slots is likely to occur only below 2-3 GHz for a 19” server.
5. Below 2-3 GHz the wavelength is λ/2≥2”- 3”, and for a pitch smaller than λ/4, the rivets can be spaced at 1”-1.5” for a 19” or 21” server.
6. Whether a resonant mode is a problem or not will depend on the field distribution for that mode with respect to the rivets.
7. The above statements assume the worst-case scenario when there is a slot between two rivets, because of oxidation, flatness issues, etc.
8. The lid of servers are sometimes a problem and need to use gaskets and overlapping seams for reliable contact. The locking mechanism has plastic parts which need internal shielding.

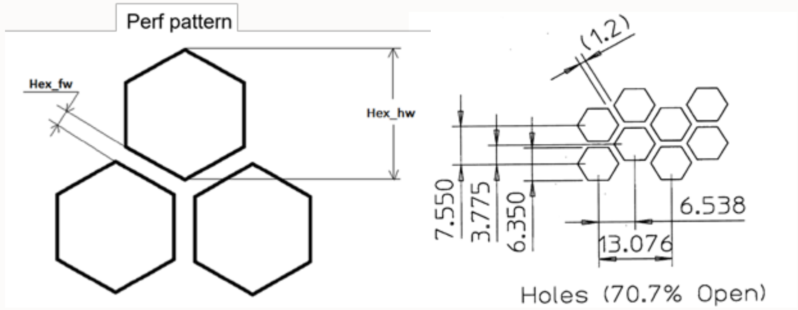
Outline

1. Introduction, near field, far field, electric, magnetic and electromagnetic shielding
2. Analytical approaches to shielding. Field theory method (Kaden) and TL impedance method (Schelkunoff)
3. Limits of the theoretical approaches, numerical simulations, reciprocity
4. Chassis resonances, internal compartmentalization, absorbers
5. Practical aspects of shielding, aperture coupling, slots, seams, rivets
6. Shielding and thermal issues, holes, perf patterns, honeycomb
7. Shielding and internal sources, heatsink grounding, local shielding
8. Shielding and coatings, corrosion
9. Evaluation of shielding effectiveness, materials, chassis
10. Shielding and safety certification
11. References
12. Acknowledgments

Shielding and Perf Panels

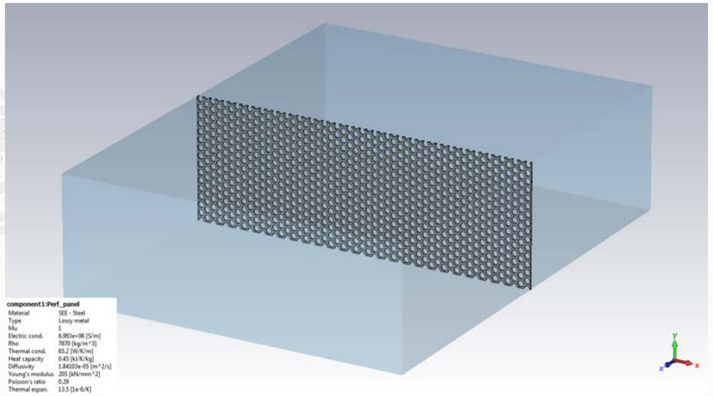
1. The perf panels are used in all large equipment, and they are always a tradeoff with the thermal requirements. The internal frequencies can be as high as 25.78 GHz for 100 G Ethernet, or 32 GHz for PCIe Gen5 switches, while CPUs and GPUs are over 350 W. Open area is 70% - 90 %, depending on pattern.
2. The size of the perf vertex to vertex is typically 7.5 mm for a rack door, 6.5 mm for a power supply, 5 or 4.57 mm for regular cooling, 3 or 2.85 mm for areas illuminated by over 20 GHz field, or even 2 mm in some 100 G cards.
3. The size of the perf cannot be chosen based on the highest frequency inside the chassis but based on the highest frequency of the **internal source likely to couple to the perf panel**, its strength, and distance from the perf panel.
4. The radiated emission from a perf panel doesn't have a dominant H or V-pol, like is the case for slots and cables radiation, useful info in troubleshooting.
5. Understanding the internal sources and their characteristics (frequencies, heatsink or not, its size, power, etc.) is very important for practical shielding.
6. Equations provided in books assume constant illumination of the entire perf area by a plane wave, and this is quite different from the actual illumination by a specific internal field distribution, mostly in the near field region.
7. The simplest way to get more realistic values for the perf attenuation is by using numerical simulations, which even if not perfect in all situations, can provide relatively accurate A/B comparison and how many dB better a perf is.

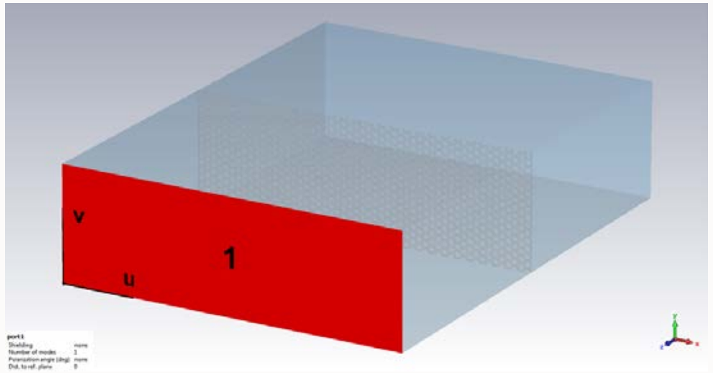


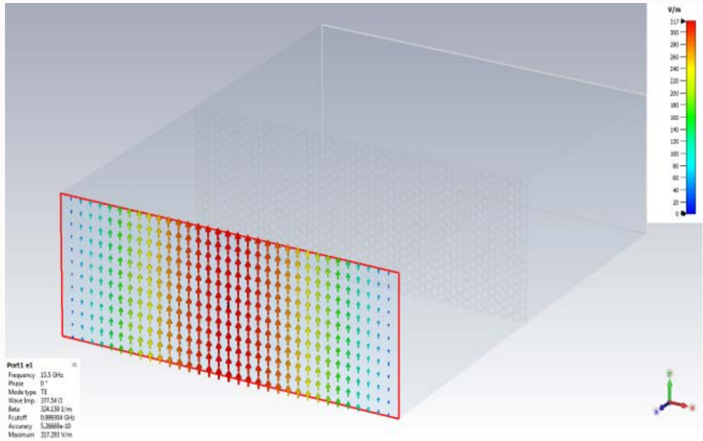


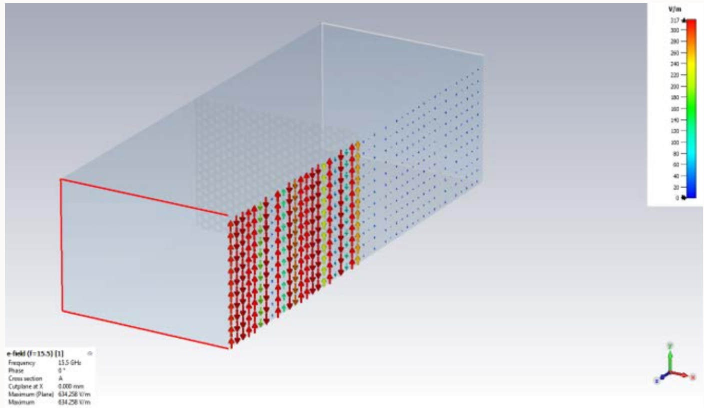
Shielding and Perf Panels

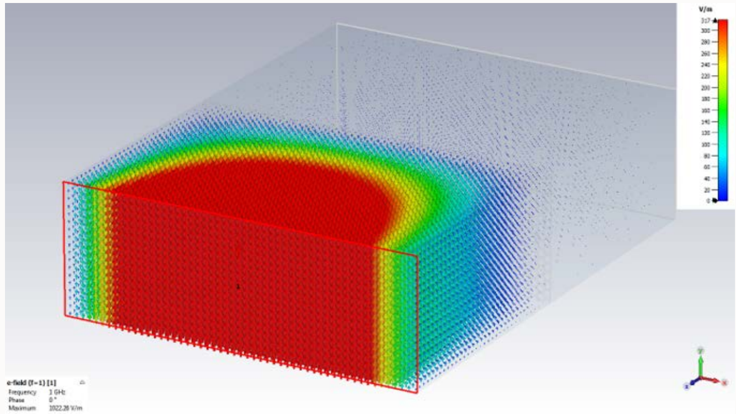
1. For example, you may find in literature [4] that if one aperture has a SE, an array of n such apertures will have the **SE reduced by -10log (n),** therefore an 10x10 array of perf will be 20 dB worse than a single hole, and in this case using perf will not be possible in practical cases.
2. We simulate in CST, using a x=150, y=50, z=200 mm waveguide, two waveguide ports, and the perf in the middle; for these dimensions, the cutoff for the TE10 mode (fundamental) is 1 GHZ. Other type of models are possible.
3. CST-T time domain solver is used as a better choice for broadband simulations.
4. EM energy conversion between different propagation modes is negligible.
5. The antenna is measuring E-field, and this model is a good first approximation for the E-field attenuation.
6. Because it is a waveguide, is not supporting TEM mode, which is equivalent to a plane wave. The E-field field is not uniform across the perf pattern for TE10.





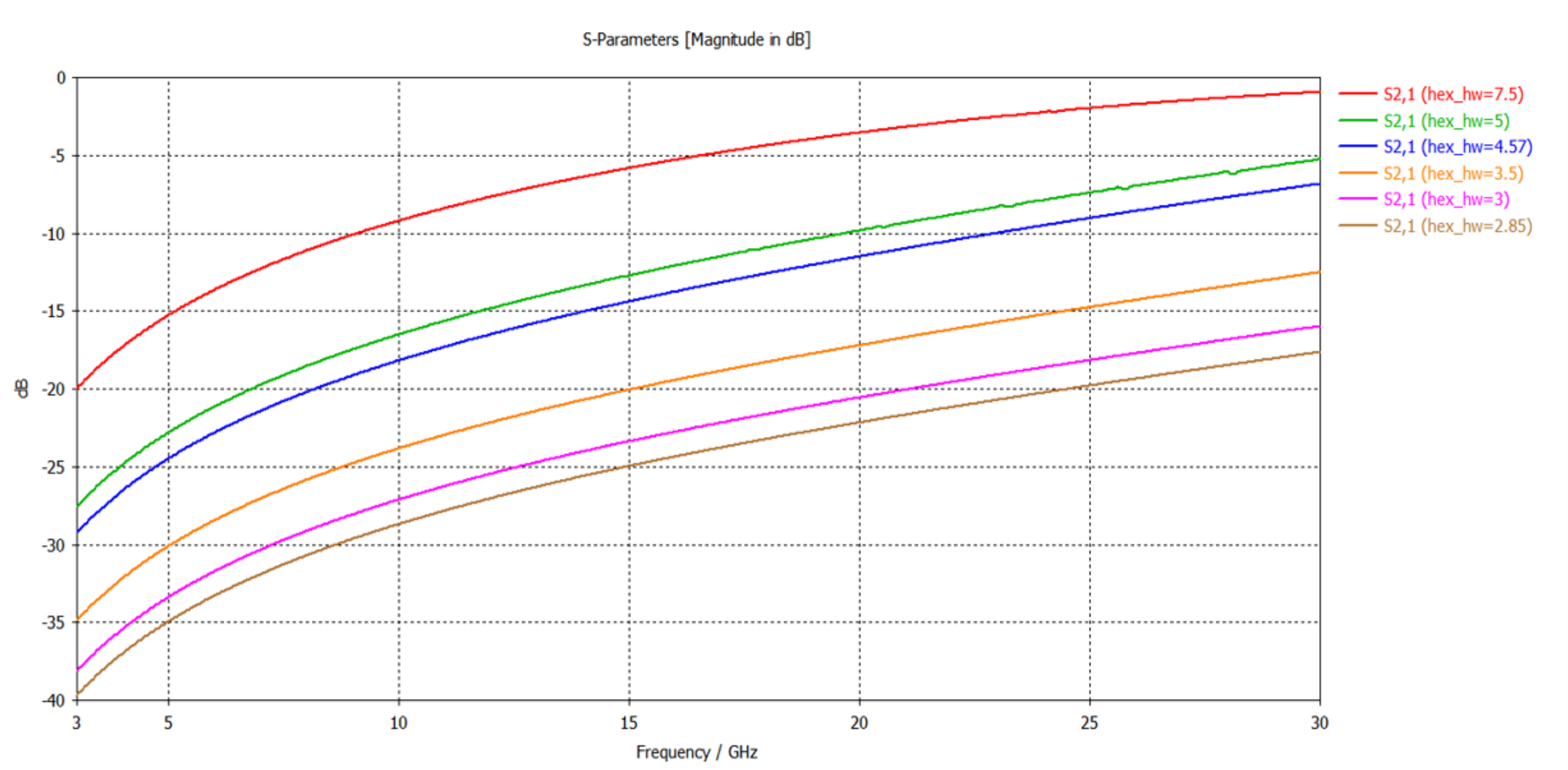






Shielding and Perf Panels

1. For example, at 25.78125 GHz, the half-wavelength is 5.82 mm, changing the perf from 5 mm to 3 mm the shielding is improved by 10 dB.
2. This type of curves can help in finding the right balance between thermal and shielding requirements. The perf doesn’t exhibit resonance in the frequency range of interest.
3. Compare 2mm holes with 5 mm holes, later on reduced to 3 mm hex pattern (CX5 card in I/O module -2.8 dB, Xyratex disk bracket, Oracle disk bracket).



Shielding and Perf Panels

1. One approach for apertures uses H.A. Bethe (1944) theory of diffraction by small holes (r, d<<λ), observing that a small aperture in a conducting wall is equivalent to an electric dipole (**P**) normal to the aperture and having the strength proportional to the normal component of the exciting electric field, plus a magnetic dipole (**M**) in the plane of the aperture and having the strength proportional to the exciting tangential magnetic field.
2. The electric and magnetic polarizabilities of an aperture (α_e, α_m) depend on the aperture size and shape and characterize the coupling or radiating properties of the aperture. For circular aperture of radius r, or square aperture with side d, notice the dependence with the **cube power** of the radius ($\sim r^3$) or side of the square ($\sim d^3$) .

Circular (r): $\alpha_e = -\frac{2}{3}r^3 \quad \alpha_m = \frac{4}{3}r^3$ Square (d): $\alpha_e = -0.114d^3$, $\alpha_m = 0.258d^3$

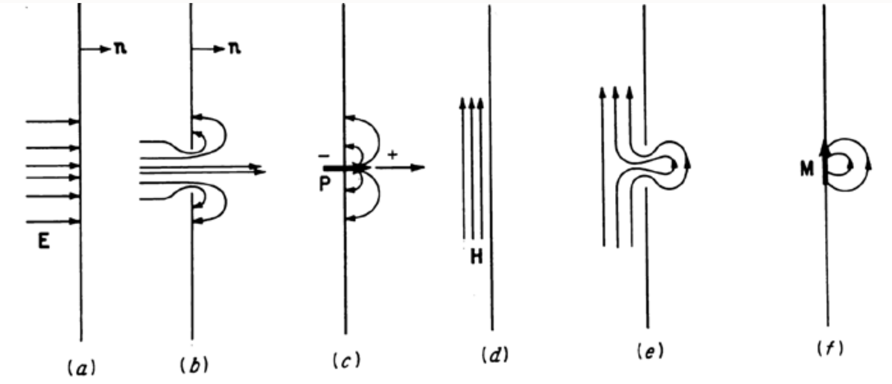
$$\vec{P} = -\varepsilon_0 \alpha_e \left(\hat{n}\cdot\vec{E}\right)\hat{n}$$

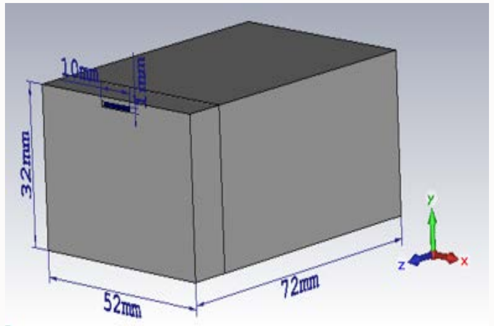
$$\vec{M} = -\alpha_m \vec{H}_t$$

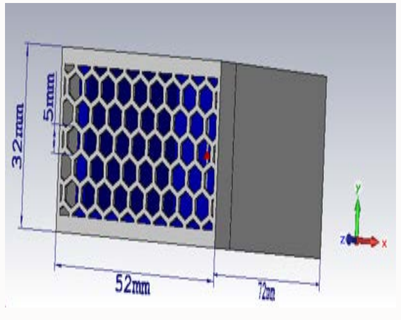
3. A comparison between two apertures is close to the numerical results:

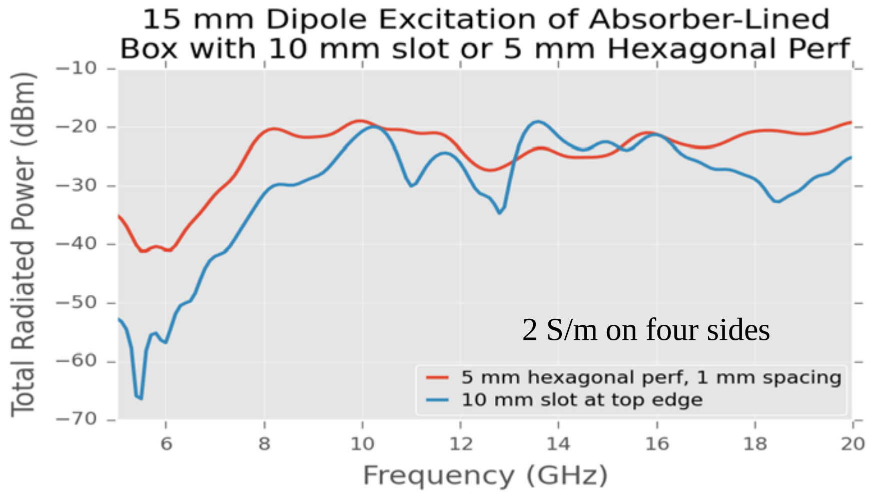
$$SE_{\text{difference}} = 20\log\left(\frac{r_1^3}{r_2^3}\right) = 60\log\left(\frac{r_1}{r_2}\right) \qquad SE_{\text{difference}} = 60\log\left(\frac{5\text{ mm}}{3\text{ mm}}\right) = 13.31\text{ dB}$$

4. The radiation from a single 10 mm slot, 1 mm width, in an absorber-lined empty box excited by a 15 mm dipole driven by 1V source along y-axis at (0,0,-10), exceeds or is comparable to the radiation from a full 30 x 50 mm hex perf panel using 5 mm hex holes at 1 mm spacing, at frequencies close and below the slot resonance (15 GHz) , when the slot length is close to λ/2.



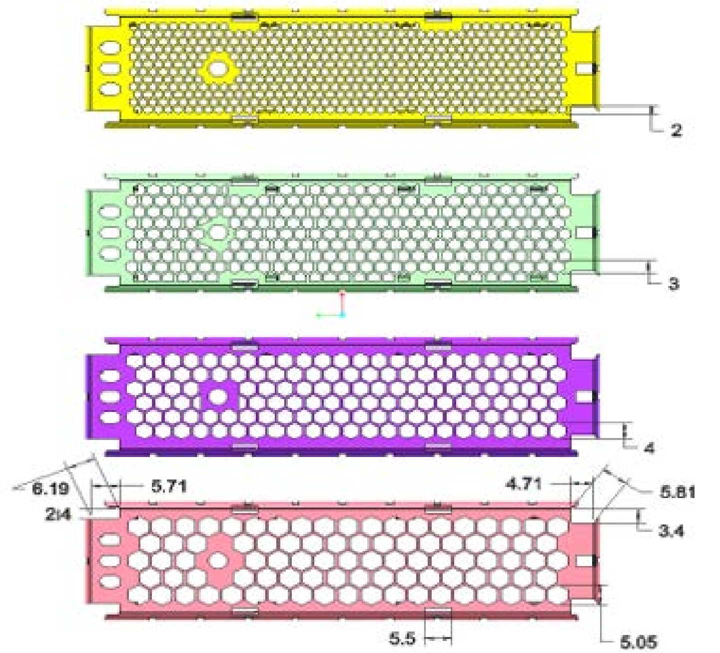




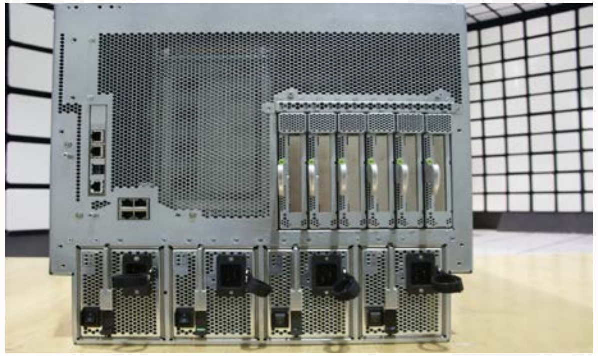


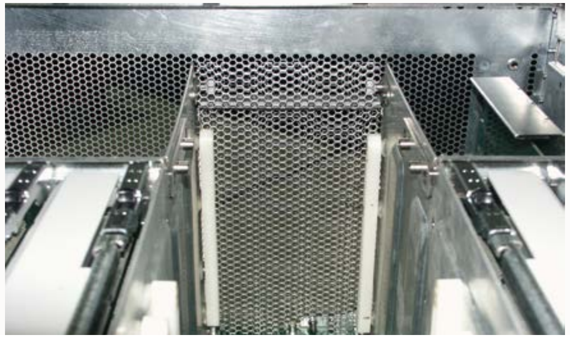
Shielding and Perf Panels

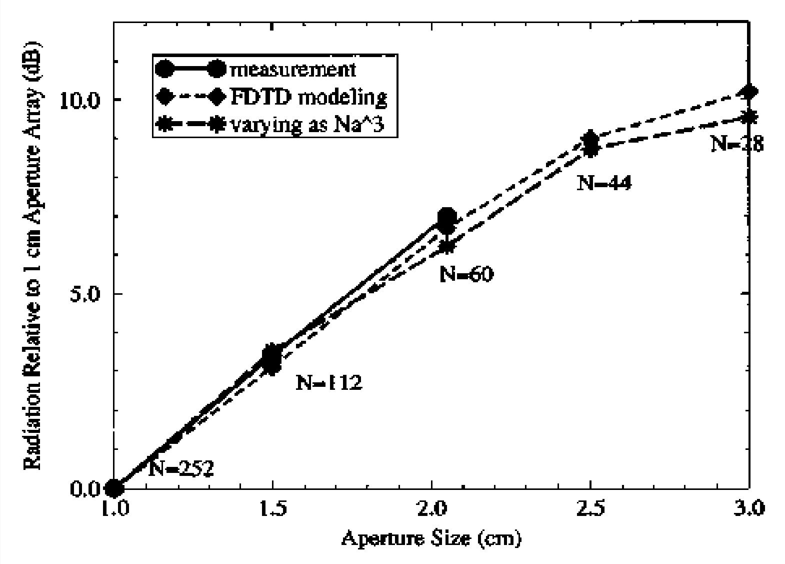
1. One concern might be that reducing perf size open area for thermal purposes is reduced. The impact may be small because there are more small holes (see table).
2. In some cases, when a very strong radiator must be near perf for thermal reasons, and honeycomb might be a solution, however two perf panels with some distance between them can be an effective and cheaper solution, without changing the front.

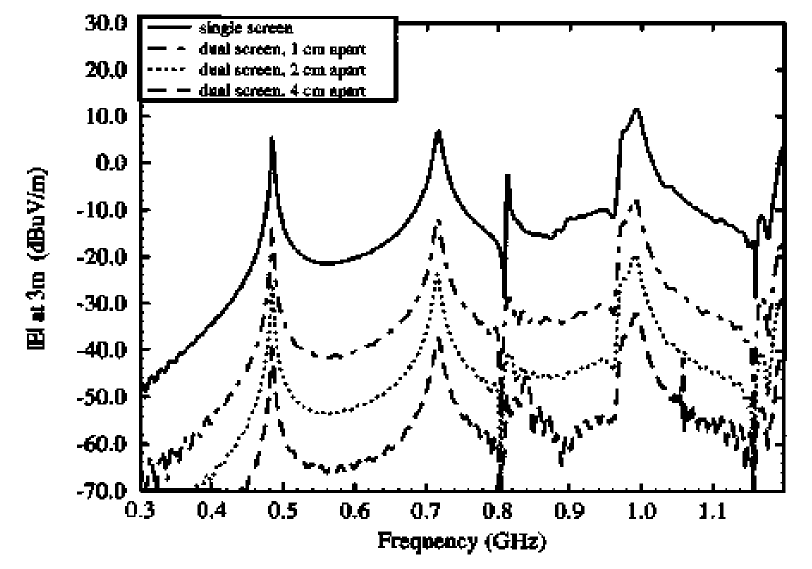


Perforation Size	Total Perforation Area	Area Reduction (Compare to 5mm)
2mm	1291.24mm	-12.41%
3mm	1450.7mm	-1.60%
4mm	1392.57mm	-5.54%
5mm	1474.22mm	--



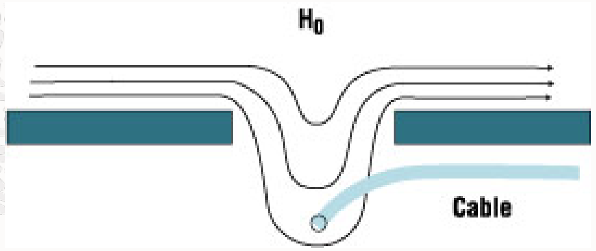


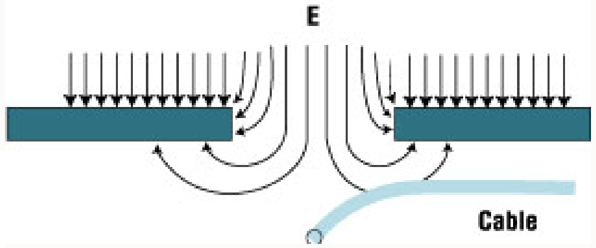


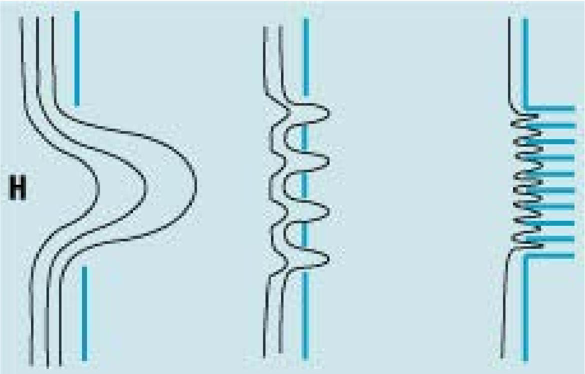


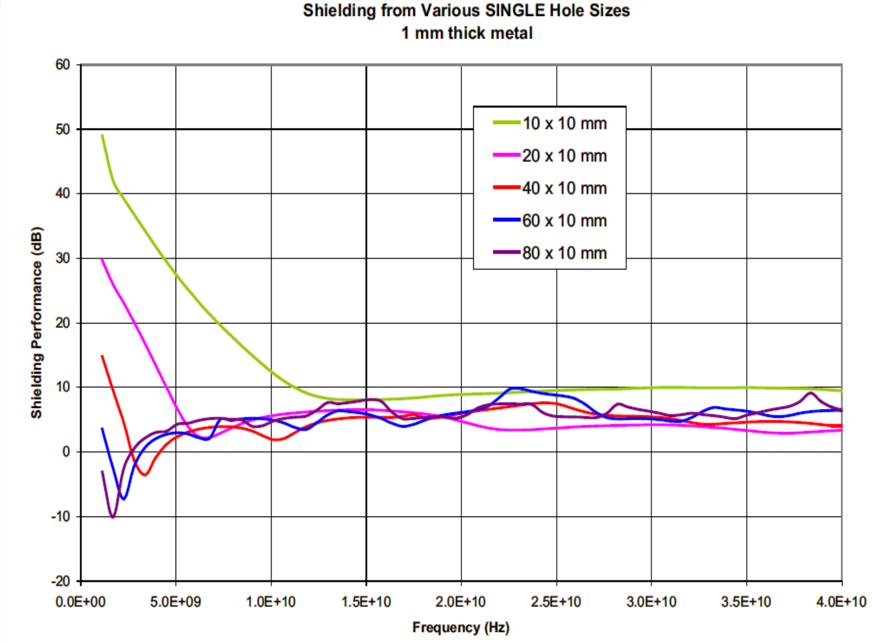
Shielding – Perf and External Cables

1. Typically, radiated emissions below 1 GHz are from cables, and radiated emissions above 1 GHz are from slots and apertures in the chassis. About 0.5 µA on a cable will fail class A.
2. A (shielded) cable can radiate if it has some parasitic CM current on it (or its shield).
3. The cables routed in front of a perf pattern may couple to the field from the perf and re-radiate. The EMC standard asks to find the cable position for which radiated emissions are maximized. This is not necessary for well controlled cable management in racks.
4. Sometimes, multiple shielded cables may block radiation from perf without re-radiating.

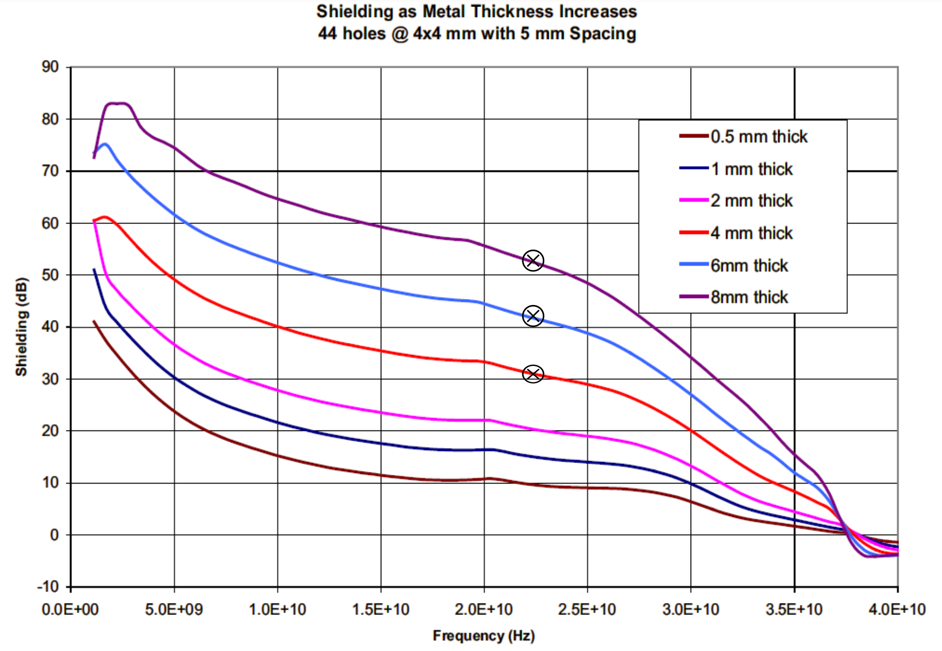







Shielding from Various SINGLE Hole Sizes
1 mm thick metal

Wire at 20 mm from the hole


Shielding as Metal Thickness Increases
44 holes @ 4x4 mm with 5 mm Spacing

Wire at 20 mm from the hole pattern

⊗ Waveguides below cutoff $f_{cr} = 37.5\ \text{GHz}$

Shielding and Waveguides Below Cutoff

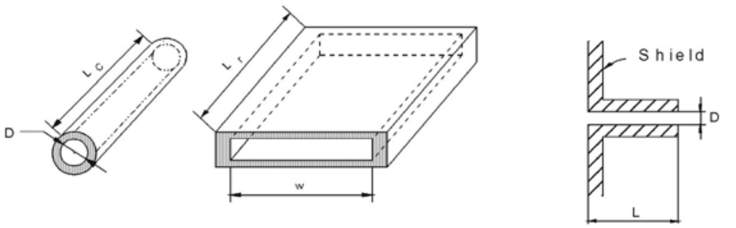
1. A single conductor waveguide doesn't support the TEM mode, therefore can't carry DC and low frequency, but support electromagnetic waves propagation above a critical cutoff frequency f_c. The lowest f_c is for the fundamental mode, TE_{10} for the rectangular waveguide, or T_{11} for a circular waveguide.

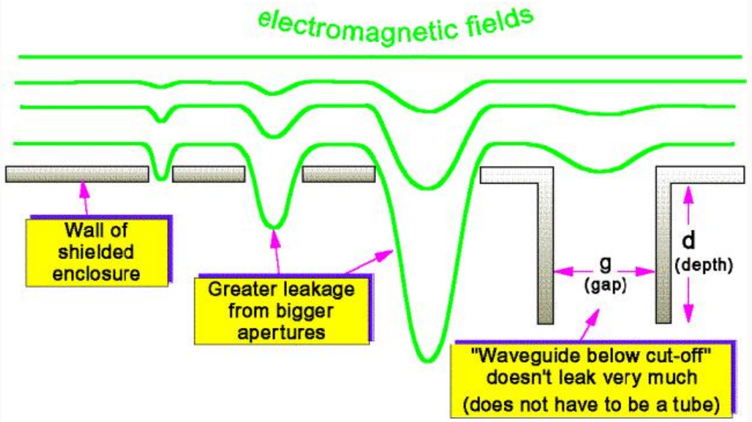
$$f_{cc} = \frac{175.768}{D\ [\mathrm{m}]}\ [\mathrm{MHz}] \qquad f_{cr} = \frac{150.114}{W\ [\mathrm{m}]}\ [\mathrm{MHz}]$$

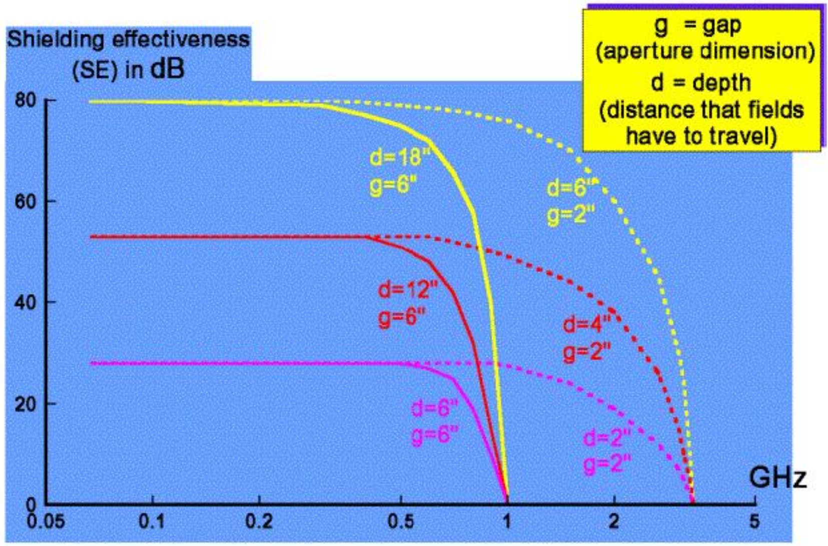
2. Below this cutoff frequency electromagnetic waves will not propagate and will be attenuated by the waveguide below cutoff.

$$SE_c = 32\frac{L_c}{D}\ [\mathrm{dB}] \qquad L_c \gg D \qquad SE_c = 27.2\frac{L_r}{W}\ [\mathrm{dB}] \qquad L_r \gg W$$

3. These equations are not valid if the length of the waveguide is not larger than the cross section of the waveguide.
4. Never pass a wire through a waveguide, it will support the TEM mode below the cutoff frequency and will defeat the purpose (if a fan is on the exterior side).
5. Increasing the thickness of a shield wall from 1 mm to 2mm, for example, will not create any waveguide below cutoff, but may slightly affect the field outside.
6. The typical application for waveguides below cutoff are the honeycomb panels, used mostly for thermal reasons (96% open area, laminar flow).

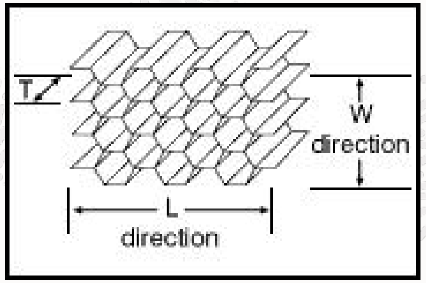


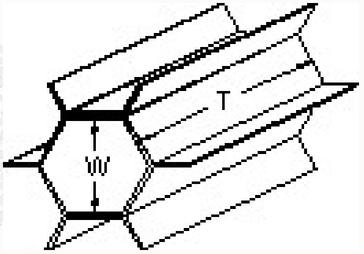


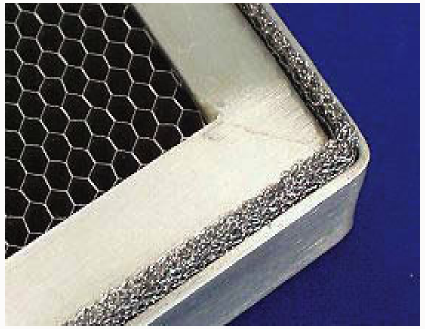


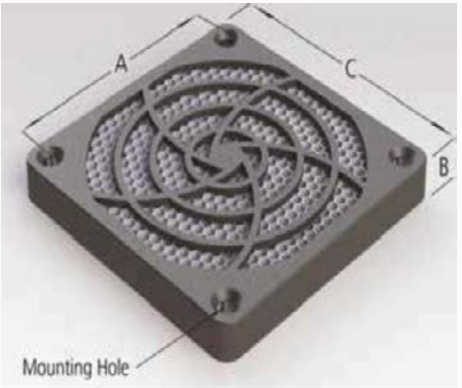
Shielding and Honeycomb Panels

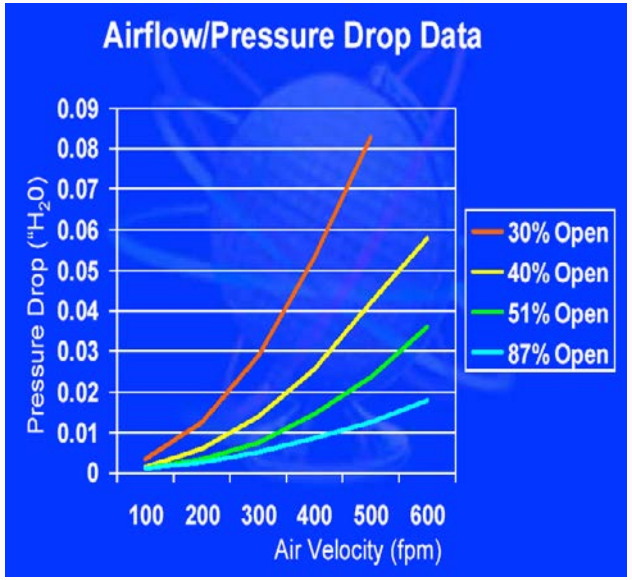
1. Honeycomb panels use the principle of below-cutoff waveguide and provide excellent shielding, better than 40 dB, the cross-section of the cell controls the cutoff frequency (W<<T), the panel thickness controls the attenuation.
2. Usually, using Aluminum or metalized plastic (whole panel metalized after assembly), typical ¼” or ½” thick, custom made, with or without metal frame.
3. Excellent from a thermal point of view, 96% open area and laminar flow, however, expensive and bulky. The air pressure drop will depend on air flow volume, velocity and cell cross section but is lower than for perf panels.
4. From an EMC point of view, the main difficulty with honeycomb is gasketing to the chassis, along the entire periphery, with reliable constant compression.

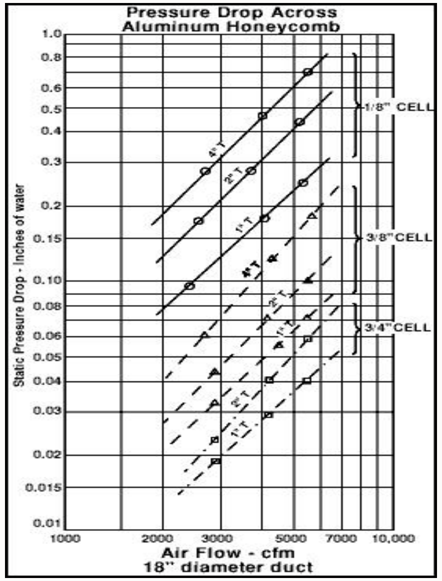


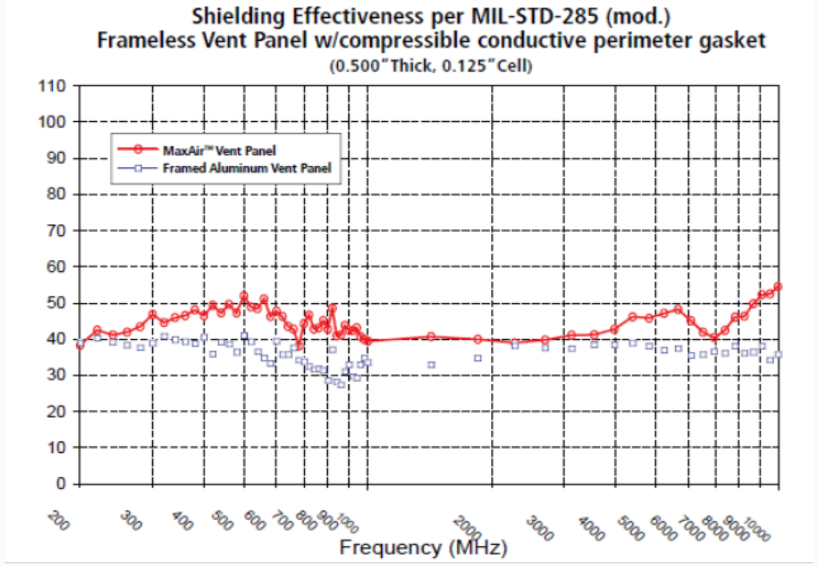






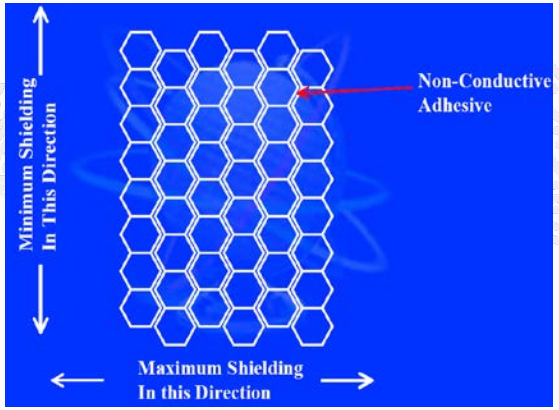


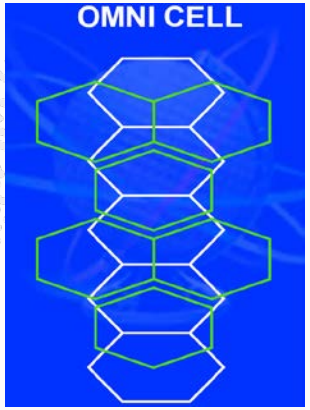


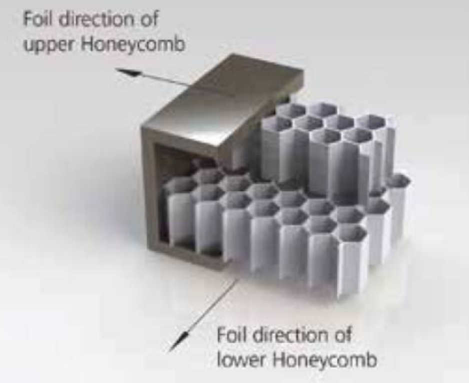


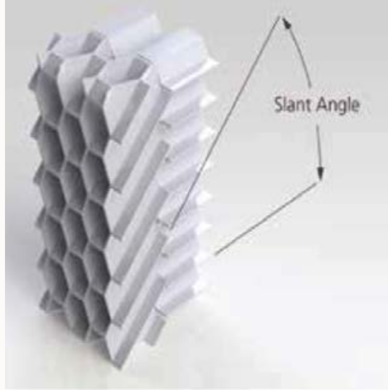
Shielding and Honeycomb Panels

1. The honeycomb panel is assembled from metal strips, with non-conductive adhesive to the adjacent metal strip, therefore excellent path for current flow in the direction of the metal strip, and less in perpendicular direction.
2. The solution is using two ½" panels with the metal strips in perpendicular direction, therefore excellent current flow in both directions.
3. The high conductivity plating of the panel will greatly improve the shielding effectiveness at high frequency, above 1 GHz.
4. If two layers are used, one layer can have a slant angle if this is better for thermal performance but will slightly impact the air pressure drop.
5. Some air inlets for rack level systems will need air filters, and in order to compartmentalize the large air inlet from electronics and avoid resonances, it is better to use conductive (carbon loaded) foam as air filter.

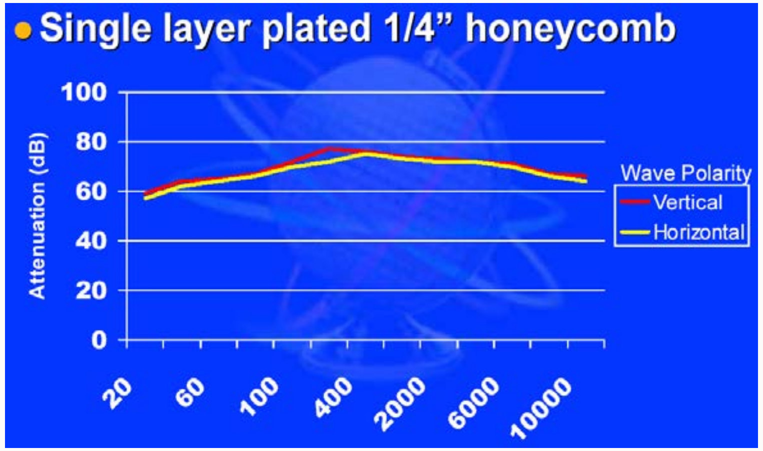


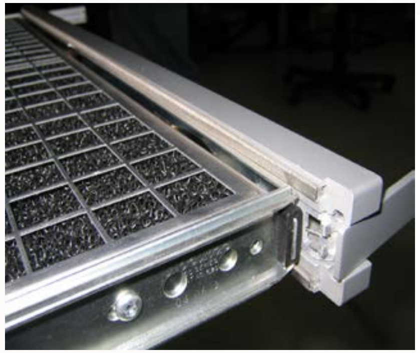


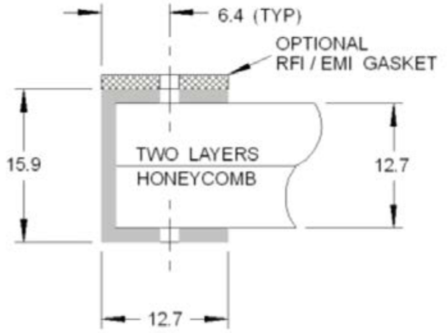





- Single layer Un-plated 1/4"honeycomb


- Single layer plated 1/4" honeycomb


Conductive foam (air filter)



Outline

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3. Limits of the theoretical approaches, numerical simulations, reciprocity
4. Chassis resonances, internal compartmentalization, absorbers
5. Practical aspects of shielding, aperture coupling, slots, seams, rivets
6. Shielding and thermal issues, holes, perf panels, honeycomb
7. Shielding and internal sources, heatsink grounding, local shielding
8. Shielding and coatings, corrosion
9. Evaluation of shielding effectiveness, materials, chassis
10. Shielding and safety certification
11. References
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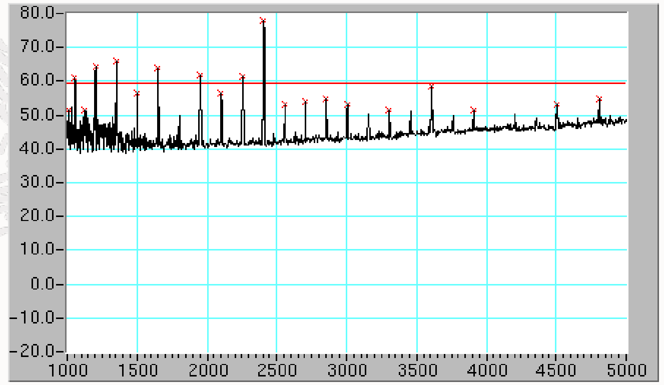
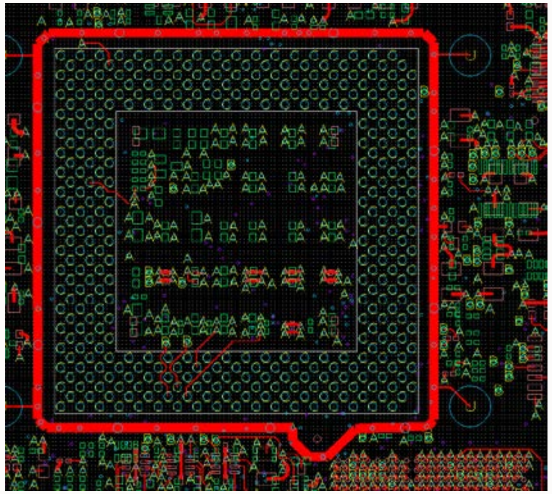
Shielding - Internal Sources

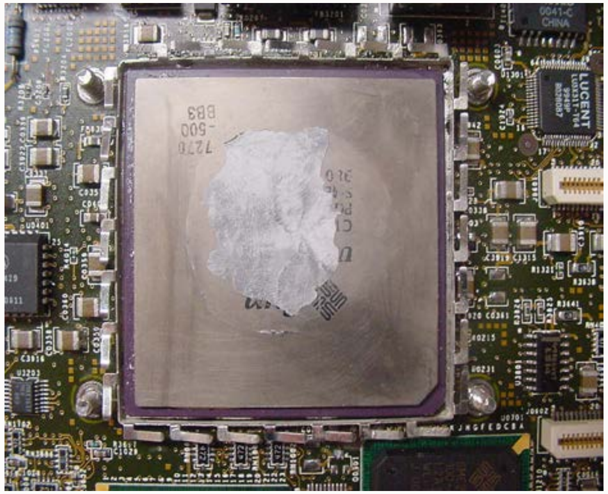
1. In order to understand the radiated emissions from a chassis, try to decompose the problem into three simpler problems: (1) radiation from slots, holes and apertures when excited; (2) characterization of unintentional antennas inside the chassis; (3) coupling mechanisms between sources and apertures in the presence of the chassis.
2. In most cases, the large VLSI chips, with large power dissipation and large heatsinks are among the biggest noise sources, especially if they operate on buses which do not support spread spectrum clock generation (Ethernet, IB).
3. Even if CPUs and GPUs can be over 350 W, they are much less of a problem than they used to be. With up to 80 cores, divided in multiple chiplets, these cores work much more asynchronous than an old CPU with only 1-2 cores.
4. The core clock of a VLSI chip at 1-2 GHz can be a bigger problem than the bus speed used by the SerDes, for example 25.78125 GHz for 100 G Ethernet, as the data stream is typically randomized, and these chips also use data scrambling to reduce the amplitudes in the frequency spectrum.
5. Buses which typically support spread spectrum and are not an EMI problem, can become a problem when shared between two pieces of equipment, for example a server and a GPU box connected by PCIe. Separate Reference Clock with Independent Spread (SRIS) can have a lot of SI difficulties, because not all PHYs are supporting SRIS, therefore the solution might be a common PCIe bus without spread spectrum in both equipment in order to communicate properly.
6. Heatsinks can be a problem, and two situations are typical, tall heatsink and flat heatsink: (1) the tall heatsink not extending outside the perimeter of the chip may act as a **monopole** antenna against the first solid plane of the PCB. The excitation is the voltage drop on the power distribution of the chip to the PCB power buses; (2) the flat heatsink is low profile, extending beyond the perimeter of the chip, creating a **patch antenna** with possible resonant frequencies. The excitation is the voltage drop on the power distribution of the chip to the PCB power buses.

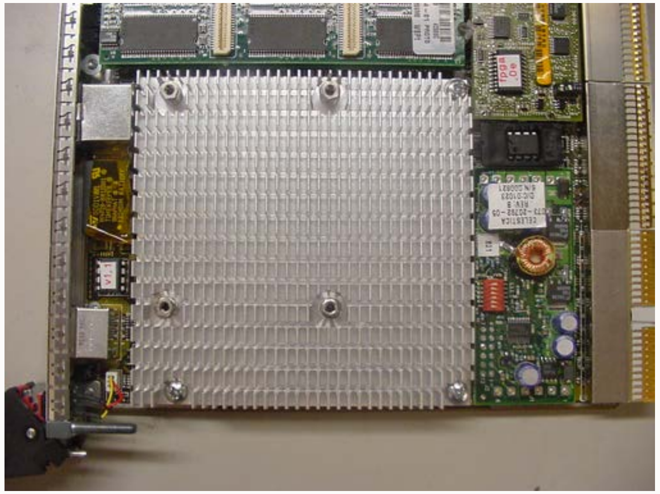
Shielding - Internal Sources

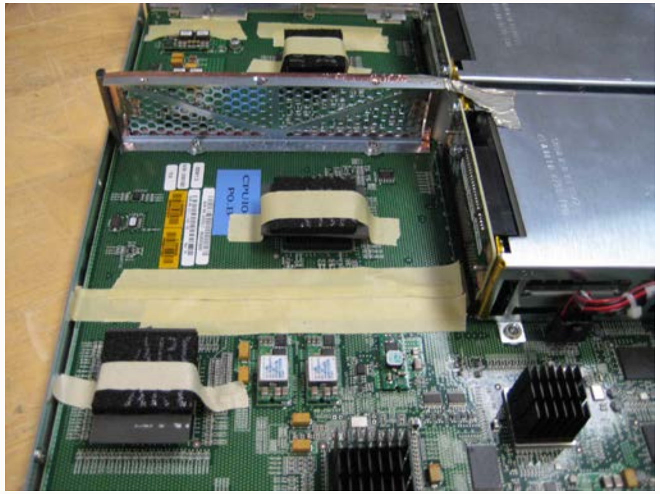
1. Both situations illustrated here: strong emissions from the chip, the heatsink height close to λ/4. When removing the heatsink and using suitable carbon loaded foam absorber, for diagnostic, the emission was mitigated. Solution: ground the heatsink, change height, ground the lid of the chip.
2. The second situation, the heatsink – GND patch antenna is worse when close to an I/O port, as in the figure, putting a lot of pressure on the gasketing of the connectors, even with the noise propagating beneath the connector with an unshielded bottom, and re-radiating on unshielded cables, like Serial or 1 GbE. The solution can be again to ground the heatsink if the placement cannot be changed.
3. Grounding the heatsink is a tradeoff, is not so effective above 1 GHz (inductance), blocks pin escapes on top, affects the routing on some internal layers (GND vias), pushes critical SI components on top further away (GND ring on top).





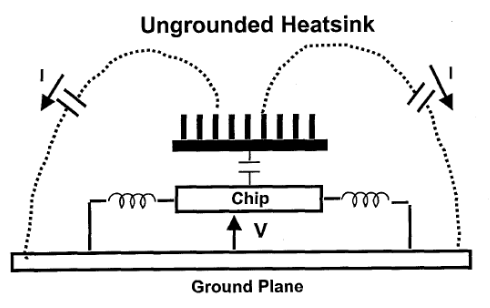


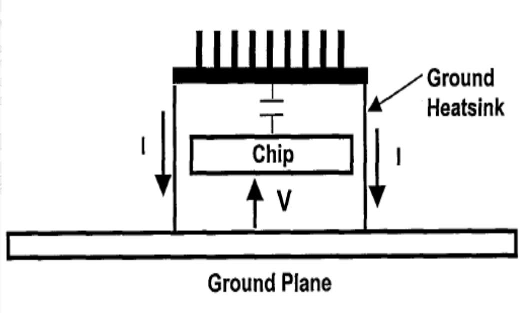


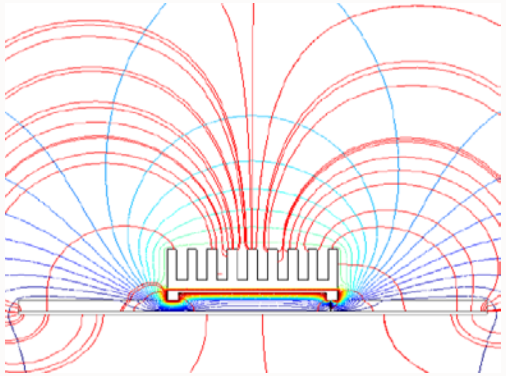


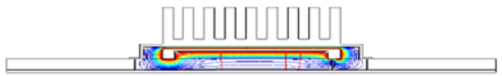
Shielding - Internal Sources

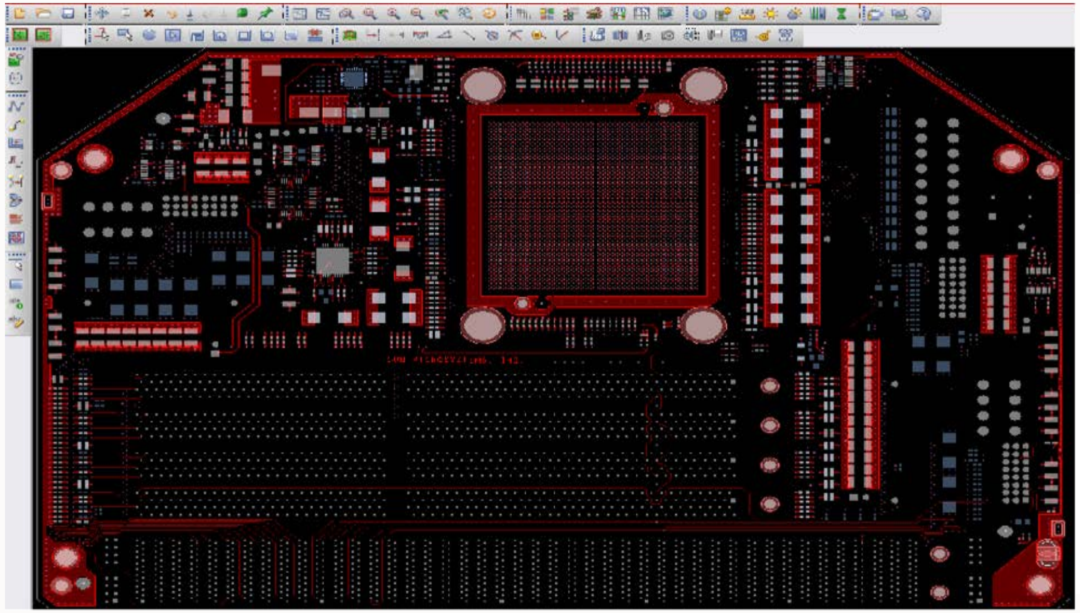
1. Grounding the heatsink is a form of local shielding and needs 360 degrees low inductance grounding. Grounding the chip lid is better.
2. Local shielding is used extensively in laptops, phones, and equipment with plastic enclosures.
3. The main difficulties with one-piece or two-pieces local shielding are the effective grounding, too inductive at high frequency, the layout impact of grounding, and thermal issues for chips without heatsink.
4. Available in cold rolled steel, brass, stainless steel and nickel silver, and sometimes combined with local absorbers.

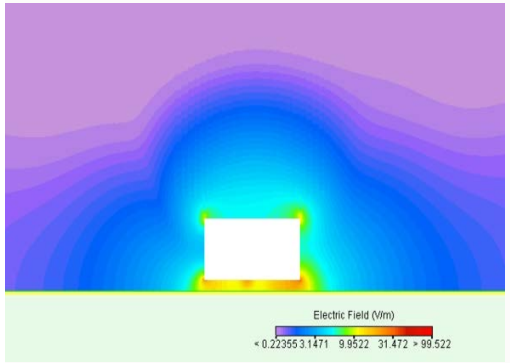


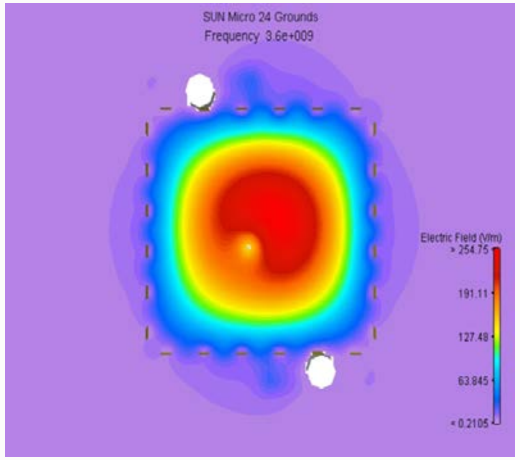






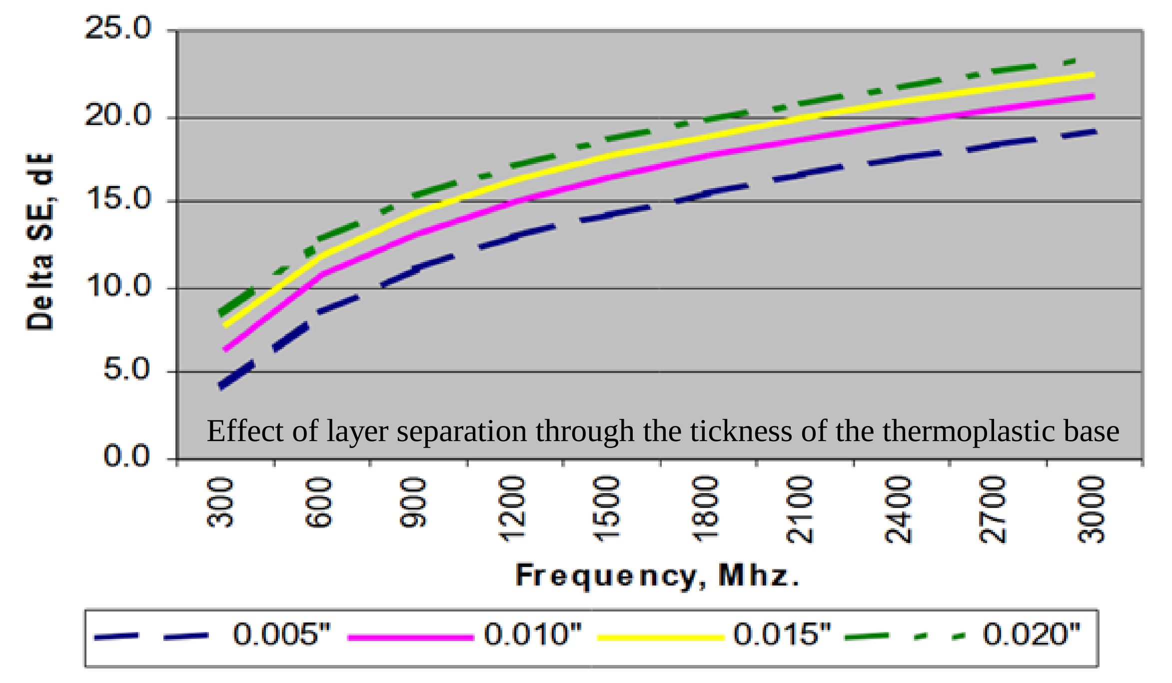


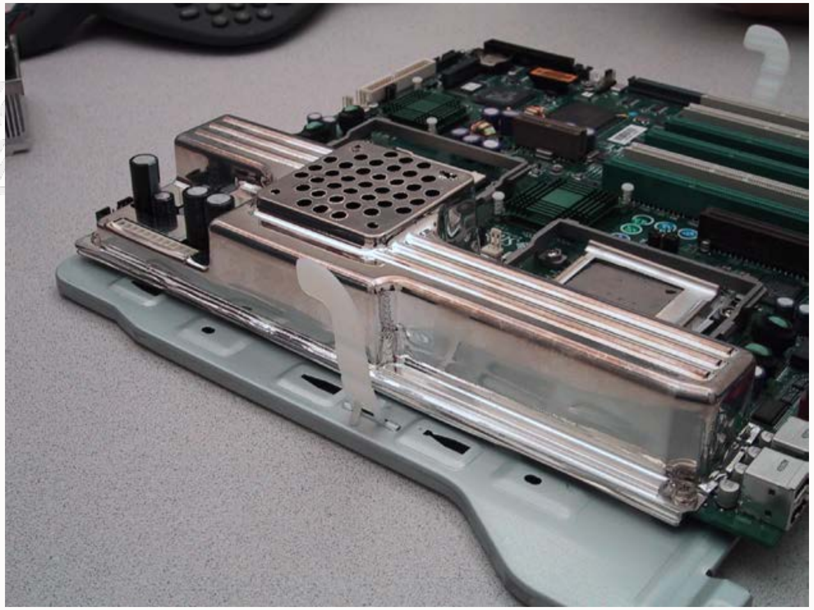


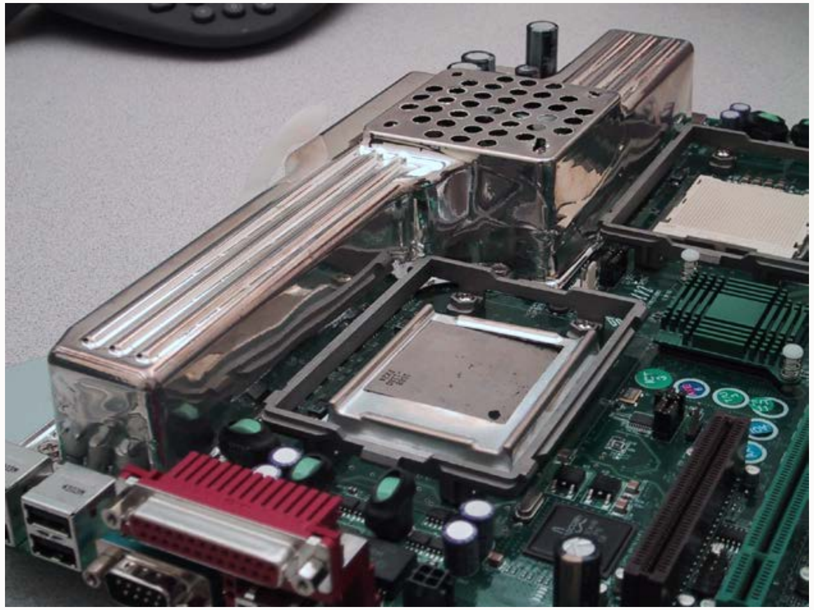


Shielding - Internal Sources

1. A particular form of local shielding used thermoplastic materials, easy to mold in any necessary shape. For example, a 25 mils molded plastic support with multilayer metallization (stainless steel 2 µm, copper 4 µm, stainless steel 2 µm), providing excellent attenuation below 1 GHz.
2. Single layer metallization (Al 1 µm) was not efficient for this memory shielding experimental setup.
3. Same problems with grounding and thermal issues. Good example for multilayer shielding. Larger separation between layers was better.



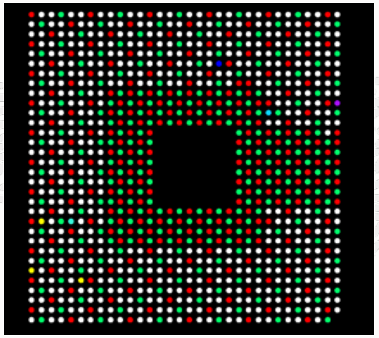




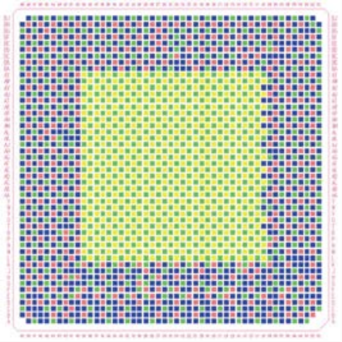
Shielding - Internal Sources

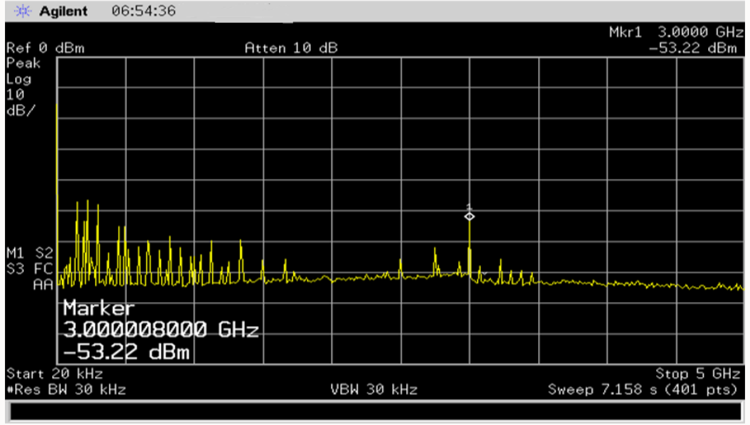
1. A typical source of potential problems up to about 300 MHz are the DC/DC converters. These are not harmonics of the switching frequency, but resonances of the parasitics and layout around the serial power MOSFET, triggered by sharp switching t_r, t_f. A resistor in the bootstrap loop will reduce the t_r, t_f.
2. The main problem is magnetic field coupling from output inductors on each phase to internal cables and poorly grounded internal metallic structures.
3. Voltage drops across VLSI power distribution pins and internal connectors pins are potential problems, and a checkerboard structure of the Vdd/GND pins provides the lowest inductance.
4. On the other hand, a very low inductance of the power distribution of VLSI, may allow noise injection from chip die to PCB, which may re-radiate. An example is measuring 3 GHz from a SAS controller on the PCB power distribution, even if this frequency was entirely local to the die.
5. Unified GND, most mounting screws toward the I/O are, shorting potential V_{CM}.

μPGA (more inductive)



LGA (less inductive)





Agilent 06:54:36

Ref 0 dBm Atten 10 dB

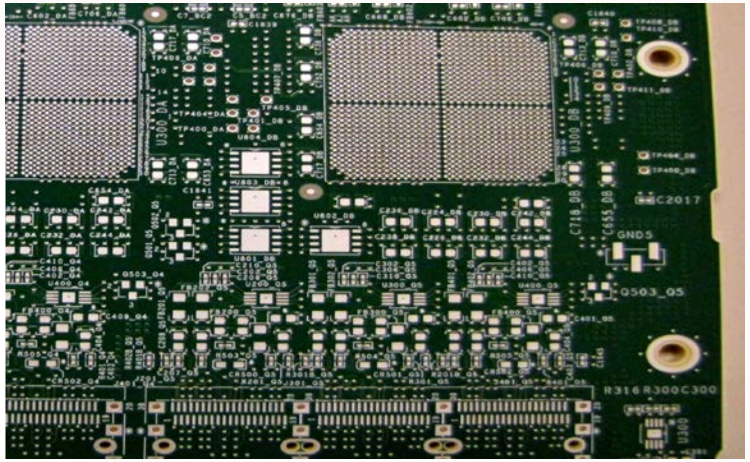
Mkr1 3.0000 GHz
-53.22 dBm

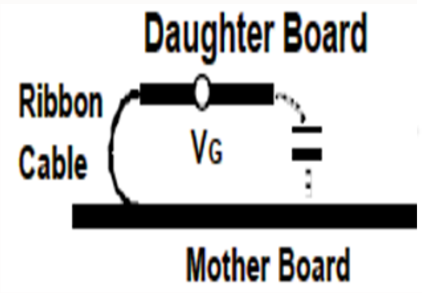
Peak
Log
10
dB/

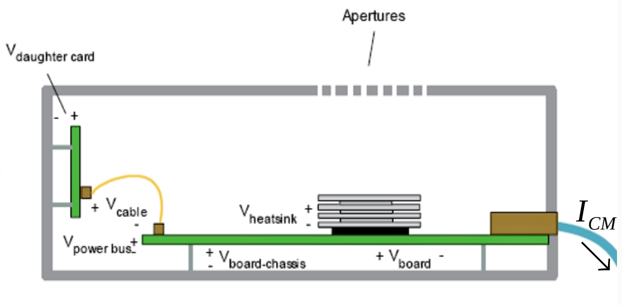
M1 S2
S3 FC
AA

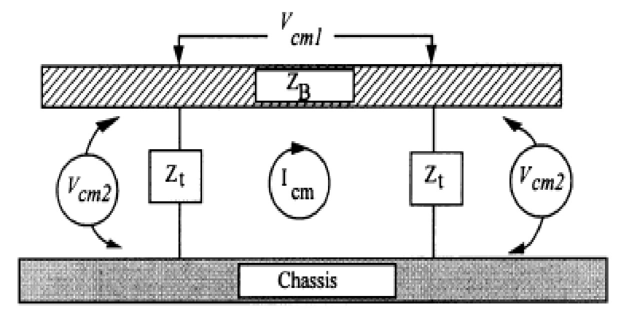
Marker
3.000008000 GHz
-53.22 dBm

Start 20 kHz Stop 5 GHz
#Res BW 30 kHz VBW 30 kHz Sweep 7.158 s (401 pts)



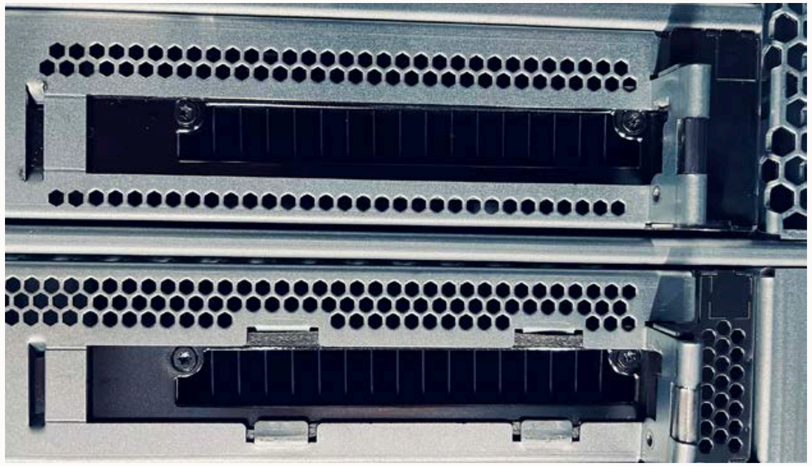






Shielding - Internal Sources

1. Shielding is a **system** problem, the system must pass RE, RI, ESD, and without taking into consideration the whole context of the design some choices can be very puzzling.
2. An example of double shielding (right) for a Telco (NEBS) system which needed air filter and 8 kV contact ESD. The detachable bezel provides the second perf panel, both with 7.5 mm hex holes.
3. Below, the Nvidia A10, a 150 W card, with a wide opening in front exposing the heatsink. The structure creates 20 cm long waveguides with around 7 GHz cutoff, and lossy anodized walls. Just behind these “waveguides” the CPU, memory and all the electronics. The internal source will need H polarization and a frequency above cutoff to propagate, but the system has very good margins. A40 (double slot) will have taller heatsink and about 3.5 GHz cutoff.



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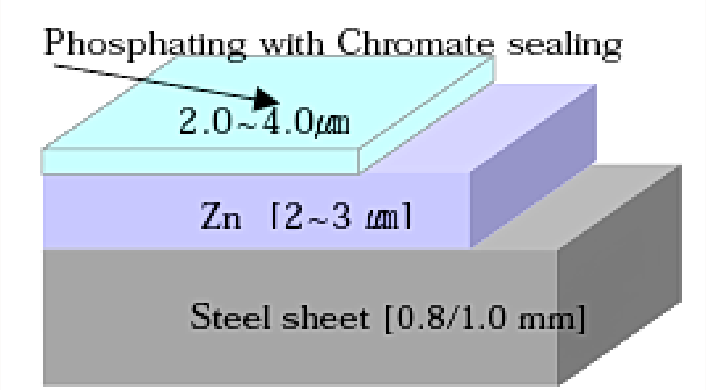
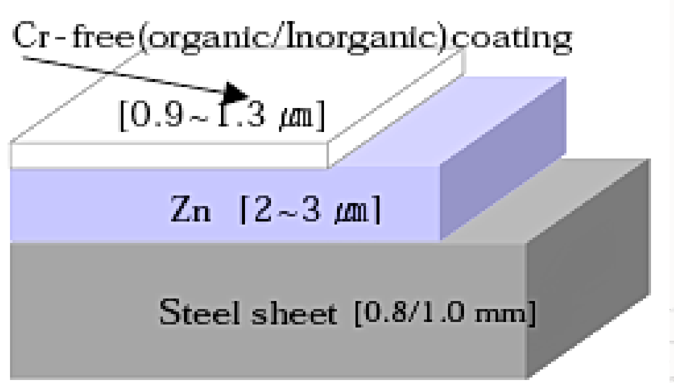
Shielding and Coatings

1. The steel chassis is typically coated by electroplating a thin layer of the chromium onto metal, to provide protection against wear and corrosion. However, **RoHS** requires less than 0.1% Hexavalent Chromium (Cr^{6+}, toxic) present as a "homogeneous material", not intentionally added.
2. Forms of chromium that <u>are</u> **RoHS** compliant: pure chrome plating (shiny, metallic), trivalent chromium (Cr^{3+}) conversion coatings ("Clear" or "Clear blue bright").
3. "Safe" substitutions for hexavalent chromium conversion coatings for steel: use aluminized steel (T2-65 or T2-LC), not galvanized (electro-galvanized steel is linked to zinc whiskers, but also NEBS GR78 CORE compliance prohibits the use of ANY zinc in systems, even as an alloying element if over 15%). NEBS stands for Network Equipment Building System.
4. It is critical that areas using anodization, alodine, phosphatizing or painting, are not overlapping with areas for gaskets contact, as all these treatments create a dielectric barrier and compromise the gasketing.
5. If Telco (NEBS) compliance is not an issue, then use "Clear blue bright" trivalent chromium with acid bath electrolytic zinc underplate. Zinc plating with clear trivalent chromate conversion coating must withstand min. 96 hours of Salt Spray (ASTM B-166) . No additional dips or coatings are necessary for this surface treatment.
6. For aluminum determine if a coating is really needed, and if needed, preferred use is trivalent chromium coating.
7. There should be no additional topcoats over the clear blue bright trivalent chromium finish. Any additional topcoat will most likely be organic and will interfere with the EMI attenuating properties. A surface resistivity check can likely determine the presence of a topcoat.

Shielding and Coatings

1. Some less-than-reputable coaters will try to get away with a hexavalent (or trivalent) chromium conversion coating to which they add a blue dye to make it look like the clear blue bright trivalent. The blue color comes from added cobalt to the trivalent coating bath, which also enhances the corrosion resistance. We may need to send samples for an external hex chrome "spot check" at a lab to be sure.
2. The coating must pass multiple type of tests: Conductivity, Anti-Finger printing test, Adhesive tape testing without peel off, Good corrosion resistance, Stamp/Paintability Test(scratch and peel off), Humidity Test (60°C / 95%H / 192 hours) without blister or rust, Anti-Chemical Resistance.
3. The coatings can have a big impact on the chassis-gasket contact, increasing the resistivity of this contact with shielding degradation, especially at HF. This contact will depend on the surface resistivity of the chassis and gasket compression.
4. Measuring the surface resistivity of the chassis is difficult, and there are not well standardized methods to do this measurement in a rigorous and repeatable way.





Shielding and Surface Resistivity

1. Chromate type conversion coatings have been observed to require around 20 psi to obtain low readings. Even then, in comparison to bright tin, the resistivity is high. In SI, 1 psi=6894,76 N/m^2, or Pa (Pascal).
2. Chromate on aluminum has been measured to be around 170 milliohms at 20 psi, while bright tin on aluminum has been observed to be less than 10 milliohms at only 8 psi. Very important in the GHz range.
3. There is data that shows the surface contact resistance to be several hundreds milliohms and does not decrease below 30 milliohms until a pressure of around 200-250 psi for a nickel coating on aluminum (because of oxide).
4. Use mechanical stops either stamped into or bends added to prevent exceeding the maximum compression value, typically 90-95% (recommended 30-70%) of the gasket height.

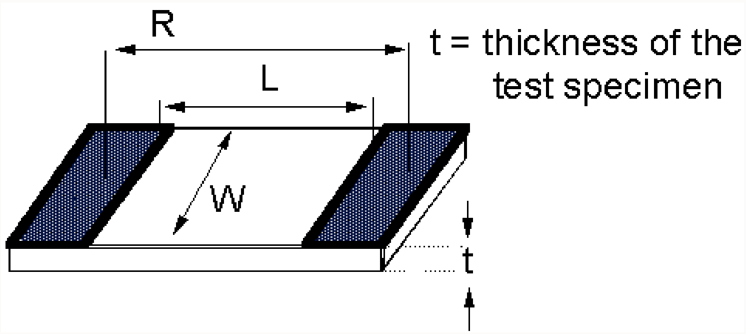
A. For coated plastic, a shielding increase of >10 dB can be observed on a frequency of interest (radiated emissions through the DVD bezel slot).

B. Vapor deposition is the most environmentally sound process for high quality conductive coatings on plastic. Aluminum is not considered a hazardous material and is thus completely compliant with the EU RoHS Directive. The coating is approximately 0.5 µm to 3 µm thick.

C. Coated plastic is difficult to recycle, usually not compliant with European Union (EU) Waste Electrical and Electronic Equipment Directive (WEEE).

Shielding and Surface Resistivity

1. Surface resistivity is measured with a µ-ohmmeter for a square sample and is typically reported in Ω/□. The method was initially developed for highly conductive coatings used in microwaves (silver plating, gold plating), much more conductive than the base material.
2. For thin, homogeneous and volume conductive materials, the **apparent** surface resistivity ρ_s is equal to the **true** volume resistivity of the coating ρ_v divided by the thickness of the coating t.
3. If the coating is more resistive than the base metal (organic coating), and L dimension is much greater than t, the circuit can be approximately as is shown below. It is the volume resistivity which is being measured. Surface resistivity, in fact, a non-existent material property, acquires its apparent value depending on **volume resistivity of the coating** and specimen geometry (two serial resistors, connected by the conductive substrate). Herein, the area involved in calculation is the area of the probe. The pressure applied by the two large, square, and highly conductive µ-ohmmeter probes, will also play a role, and most companies apply 1-2 pounds on the probes. The measurement result is mostly good for A/B comparison and material ranking, but not as an absolute value. For a coating more conductive than the base material, surface conductivity σ_s makes sense.
4. For the chassis, look for surface resistivity of 100 mΩ/□ or less, as fabric over foam gaskets have typically 50 mΩ/□.

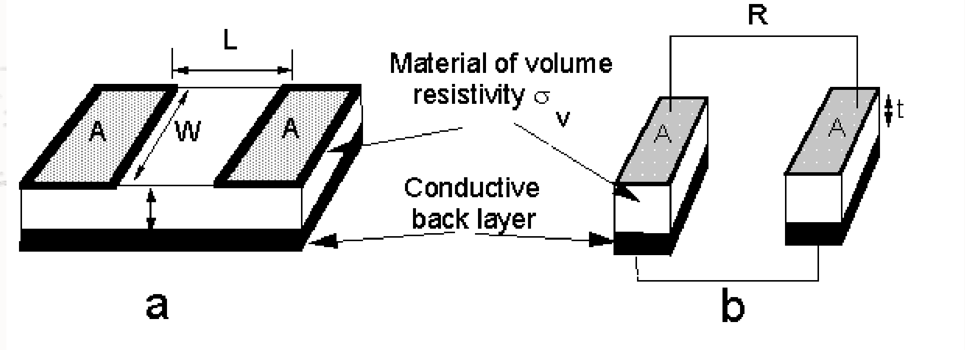


$$R = \rho_v \frac{L}{Area} = \rho_v \frac{L}{tW} = \frac{\rho_v}{t}\frac{L}{W} = \rho_s \frac{L}{W}$$

$$\rho_s = R \ \text{if} \ L = W \ [\Omega/\square]$$

$$R = \rho_v \frac{t}{A} + \rho_v \frac{t}{A} = \rho_v \frac{2t}{A}$$

where A is area of the probe



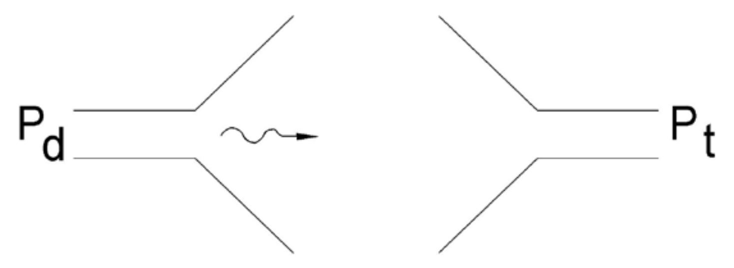
Shielding and Corrosion

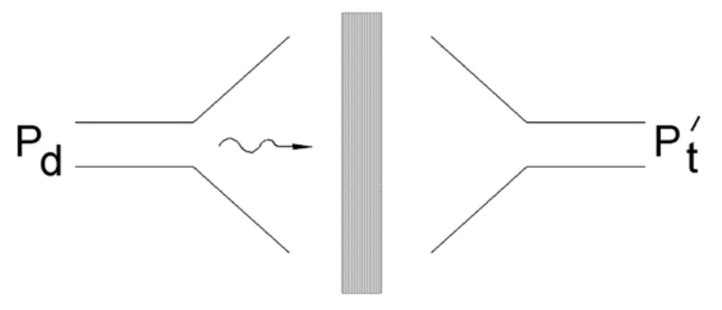
1. Galvanic or electrolytic corrosion may reduce shielding effectiveness by affecting the gasket and the chassis material. Their contact may become an insulator. Rust may even act as a diode, creating new frequencies by the nonlinear mixing. It can be a serious problem in humid environment or marine environment (saline fog). No significant corrosion due to electrochemical action in any working, storage, or transport environment is allowed per **UL 1950 standard**.
2. **Galvanic corrosion** is due to contact between two dissimilar metals in the presence of moisture (electrolyte), creating practically a wet-cell battery. The larger the potential developed between the two metals, based upon their relative position in the electro-chemical or galvanic series, the faster the corrosion. Area of the cathode and anode and the pH of the electrolyte play also a role.
3. Galvanically compatible materials are those whose electrochemical potentials differ by less than 0.25 V. For commercial applications where the environment is controlled, the range can be increased up to 0.5 to 0.6 V. If large contact voltages occur, the more anodic material will eventually be destroyed. To prevent this problem, either the gasket material or mating surface, or both, will need to be plated with a material or finish that is compatible with the base material. The closer the materials that are in contact with each other are in the galvanic series, then the lower the risk of corrosion.
4. A galvanic series with standard electrode potential, respective metal ion, in sea water: **Cathodic** →Gold^{+} (+1.69 V), Platinum^{2+} (+1.2 V), Mercury^{2+} (+0.85 V), Silver^{+} (0.8 V), Copper^{2+} (0.34 V), Hydrogen^{+} (0 V), Lead^{2+} (-0.13 V), Tin^{2+} (-0.137 V), Nickel (-0.257 V), Iron^{2+} (-0.44 V), Chromium^{3+} (-0.74 V), Zinc^{2+} (-0.76 V), Aluminum^{3+} (-1.67 V), Magnesium^{2+} (-2.37 V) ← **Anodic (destroyed)**. UL recognized protective coating must be adequately bonded to the substrate material.
5. **Electrolytic corrosion** is due to (stray) current flow between two metals in the presence of an electrolyte (which could be just slightly acidic ambient moisture). The major contributors to this problem are surface contact area, material dissimilarity, and the presence of an electromotive force (V), while the electrolyte is usually moisture.

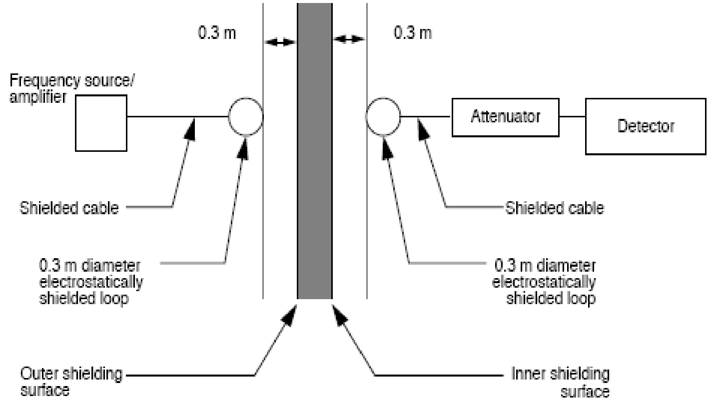
Outline

Shielding Evaluation - Materials

1. In most design cycles the chassis is designed in parallel with the electronics, usually including new chips with unknown EMI signature. The EMC engineers need to provide feedback for the chassis design based on understanding of the noise sources from schematics, component placement, and experience. Some numerical simulations where possible. Preliminary inspection and experimental investigation on the soft tooled (MX0) chassis.
2. Before the functional system is available for a full EMC design verification and troubleshooting , an experimental evaluation of a shielding enclosure can look at three aspects: (1) the properties of the shielding material itself; (2) the internal resonances and field coupling between internal compartments; (3) the integrity of the chassis using wideband artificial noise sources.
3. Methods of evaluation of shielding materials and shielding enclosures were proposed in MIL-STD-285, ASTM D4935, IEEE 299-2006, IEC 61000-5-7 standards. The basic idea is to use two ports, measure S21 with and without a material sample illuminated by a known field. Some methods are useful only for ranking materials with complex, difficult to predict properties (metalized plastics, conductive plastics, conductive elastomers, composites, gaskets, etc. They will not provide new info about well known materials typically used for enclosures, like Steel, Al or Cu.
4. The sample must be large in 2D compared with the Tx, Rx antenna to be able to neglect impact of fringing fields and reflections from the edges of the test sample. Magnetic field or electric field can be tested using suitable antennas. Especially used below 1 GHz, no path for the currents induced in the shield.



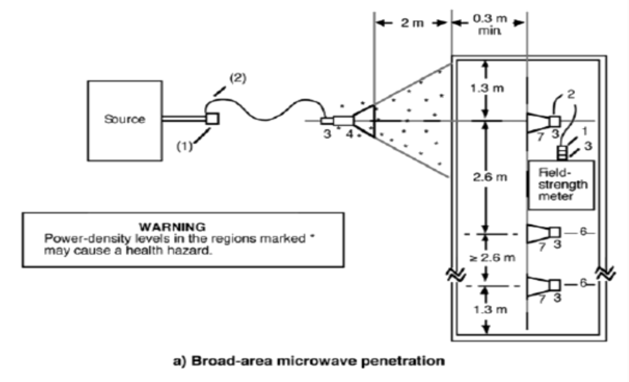




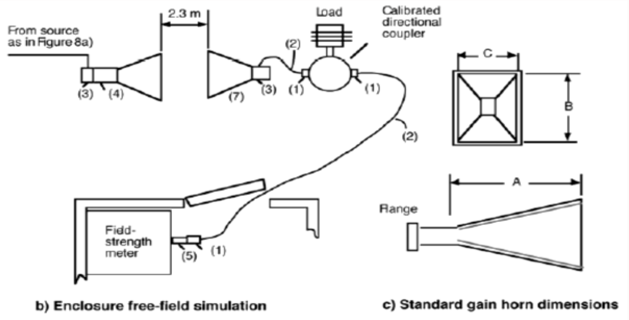
Magnetic tests diagram 9 - 20 MHz (IEEE 299)

Shielding Evaluation - Materials

1. Above 1 GHz the setup proposed in MIL 285 and the old IEEE 299 uses horn antennas, and the method was used for evaluation of (semi) anechoic rooms.
2. The coaxial TEM cell proposed in ASTM D4935 uses a coaxial structure with constant 50 Ω impedance by keeping the D_{ext}/D_{int} ratio constant. The sample is illuminated by a plane wave, the TEM mode in the coaxial cell, but the upper frequency is limited to 1.5 GHz by higher modes which start to affect the plane wave field.
3. The dual TEM cell is simply a pair of TEM cells with the added feature of an aperture in shared wall. The apertures served to drive power from the driving cell to the receiving cell. The insertion loss provided by the sample is compared with the power coupled through the empty aperture. The advantage of the method is the use of isolated fields, low power, broadband, well analyzed structure, but has upper-frequency limitations.
4. All these are qualitative methods. Controlling the contact resistance can be a big issue.

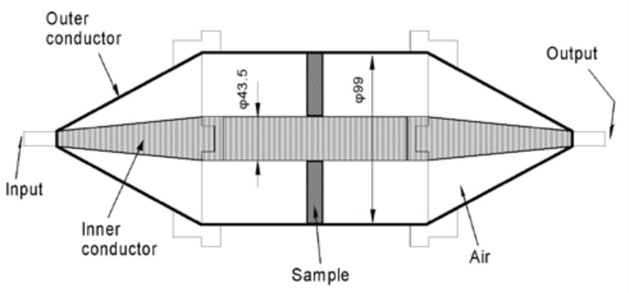


Test setup over 1 GHz (IEEE 299)

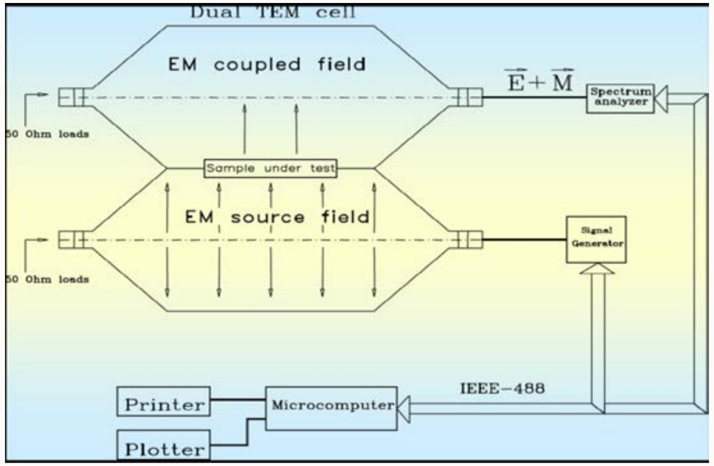


NOTES
1—Type N adapter coax to waveguide (if needed).
2—Coaxial cable or waveguide.
3—Adapter (if needed).
4—Transmitter antenna, Table 4, or ridged horn.
5—Attenuator (if not within field-strength meter).
6—Additional centerlines so that all areas are illuminated.
7—Receiving horn antenna, Figure 8c) and Table 4; dimensions relate to standard EIA waveguides, flanges, and waveguide-to-coaxial transitions.

Coaxial TEM (ASTM D4935)

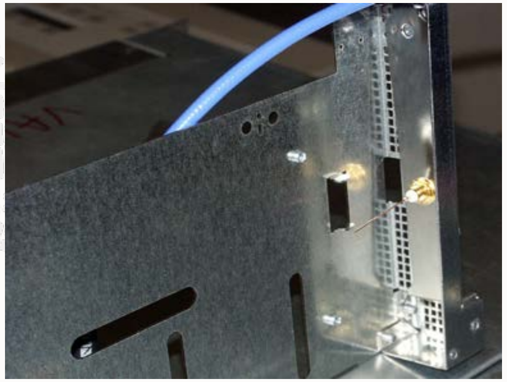


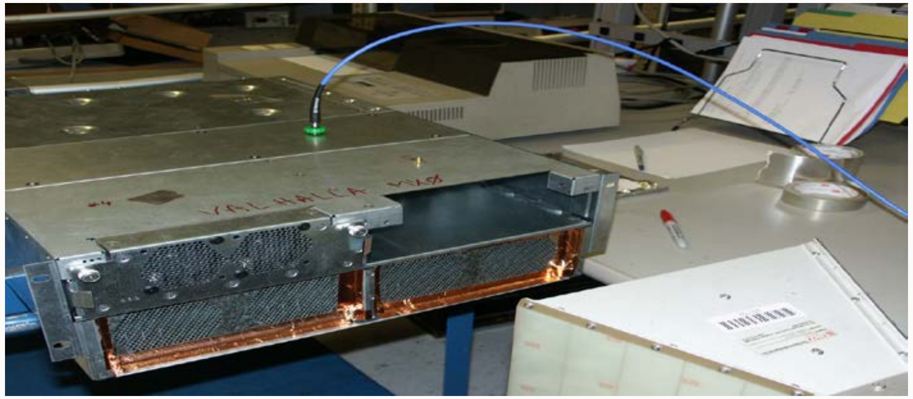
Dual TEM Cell (MIL 285)

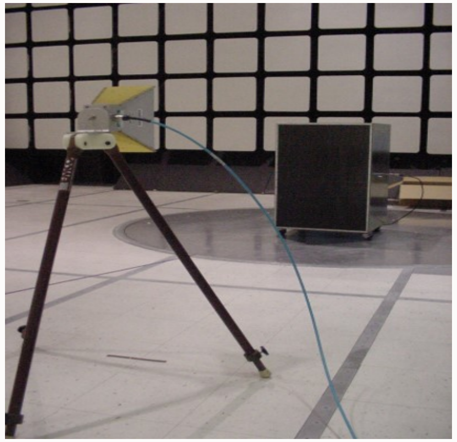


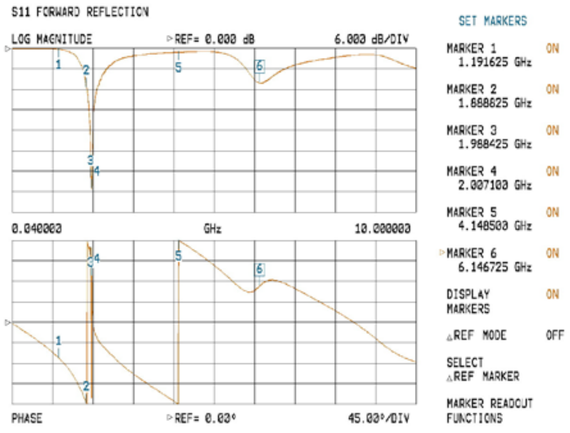
Shielding Evaluation - Resonances

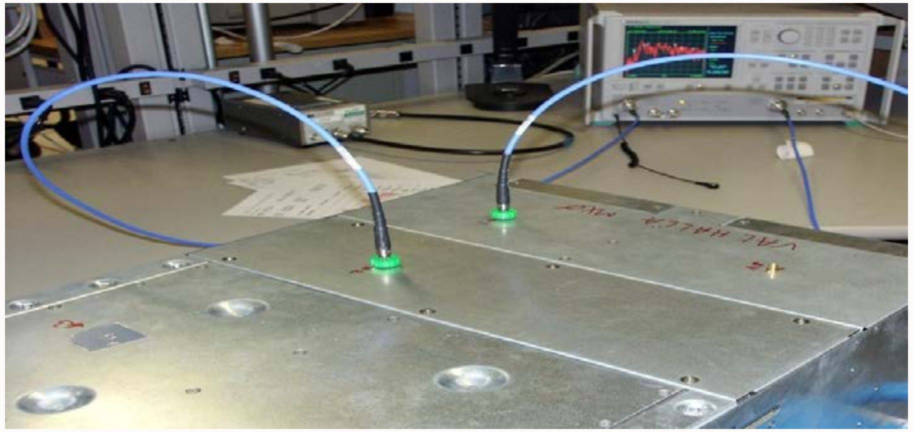
1. If a soft tooled chassis is available, preferable with the mock-up boards used for thermal evaluation, to be more realistic and include some losses, measuring S11 in different places will show potential internal resonances.
2. Measuring S21 between different internal compartments will measure potential internal coupling or isolation (CPU to I/O area for example).
3. A short monopole antenna can be used as a probe for both S11 and S21.
4. The short monopole must be characterized (2 GHz resonance) to be sure that the probe resonance is not confused with chassis resonances.
5. While exciting the chassis with the small monopole probe, one can scan around with a horn antenna or other probe for slots or apertures.
6. Wrong way to evaluate honeycomb (below); cavity modes distort the results; horn antenna is too large for the volume of the cavity.









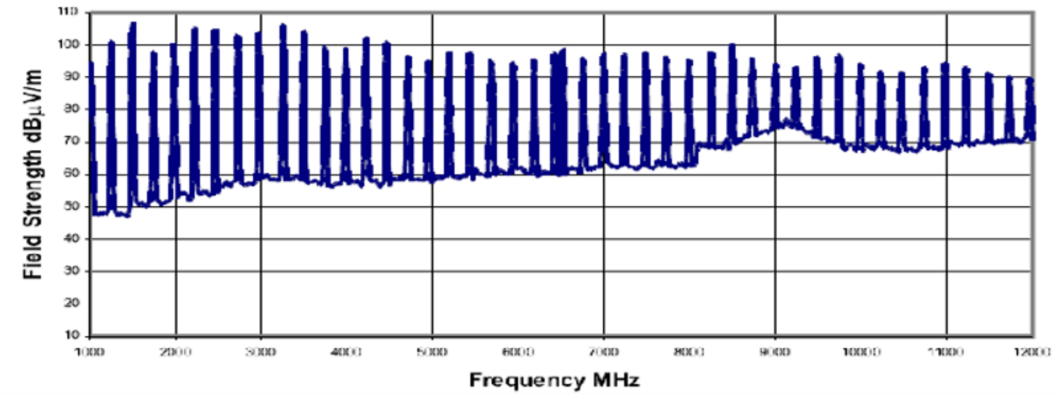


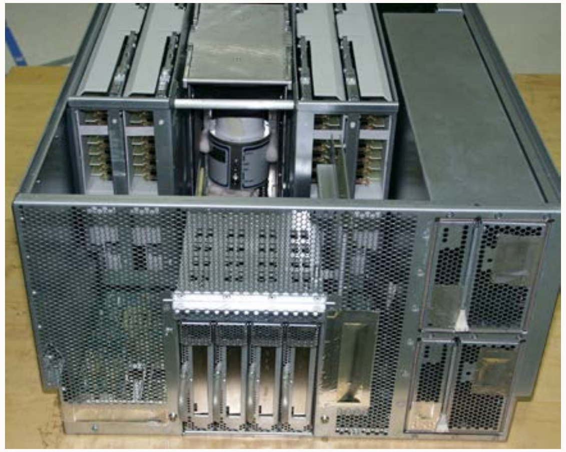
Shielding Evaluation – Comb Generators

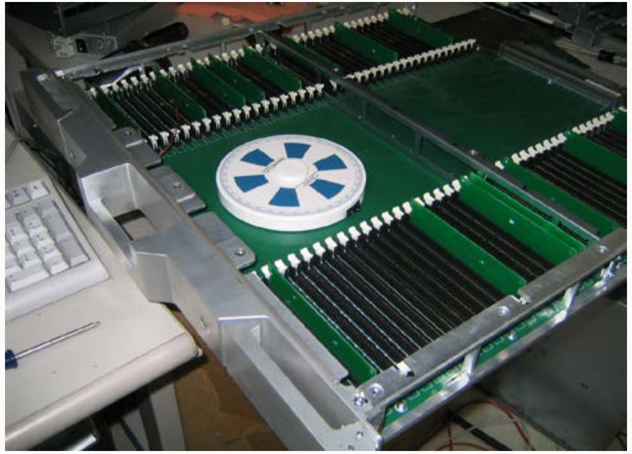
1. When the electronics is not yet available, we can use battery operated comb generators used also for Normalized Site Attenuation (NSA) as a broadband internal source.
2. The field generated is not uniform, covering up to 26 GHz, and there are comb generators with two selectable frequency steps.
3. The position and polarization of the internal field will matter, this method is useful especially for A/B comparison, measuring the field with and without the chassis at frequencies of interest and changing gasketing, perf, detecting apertures that may be a problem, etc.
4. Even if not perfect, this method still allows for a better evaluation and understanding of the chassis potential behavior.

CGO-5100 12 GHz (100 MHz steps)

CGE-02 26 GHz (250/256 MHz steps)

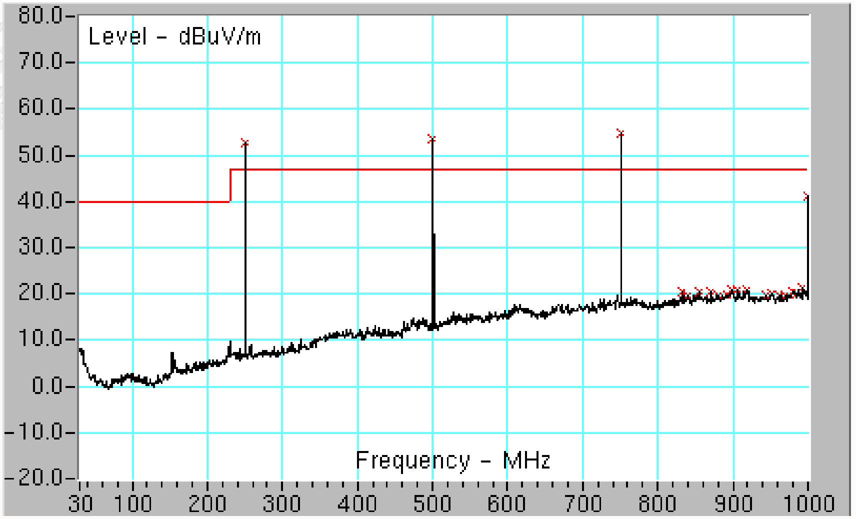


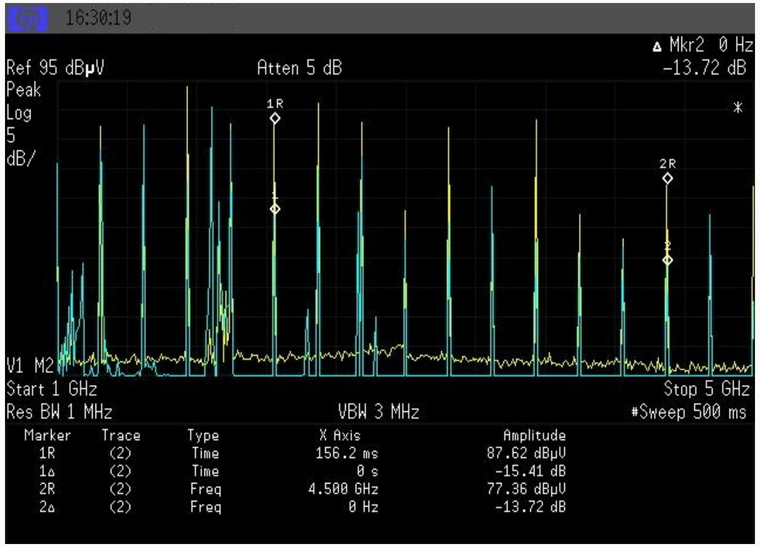




Shielding Evaluation – Comb Generators

1. An example of using a comb generator to check the difference of a lid gasketing for frequencies below 1 GHz – right, without and with gasket.
2. Another example shows the difference between the two types of gaskets for a slot radiating at 2.5 GHz, which was a known strong source.
3. These type of measurements are not providing actual SE values but may point out to design weakness and allow to compare specific solutions.
4. Identified areas of interest can be modeled numerically using simplified models, superposition, and looking only at the sources most likely to couple to a specific aperture or perf panel.





```
Marker   Trace   Type     X Axis      Amplitude
 1R       (2)    Time    156.2 ms    87.62 dBµV
 1Δ       (2)    Time        0 s     -15.41 dB
 2R       (2)    Freq   4.500 GHz    77.36 dBµV
 2Δ       (2)    Freq        0 Hz    -13.72 dB
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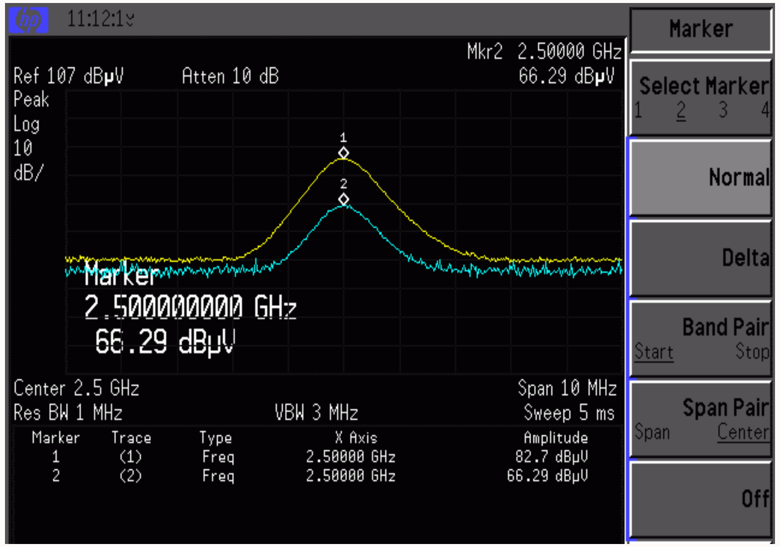

```
Marker   Trace   Type     X Axis        Amplitude
  1       (1)    Freq   2.50000 GHz    82.7 dBµV
  2       (2)    Freq   2.50000 GHz    66.29 dBµV
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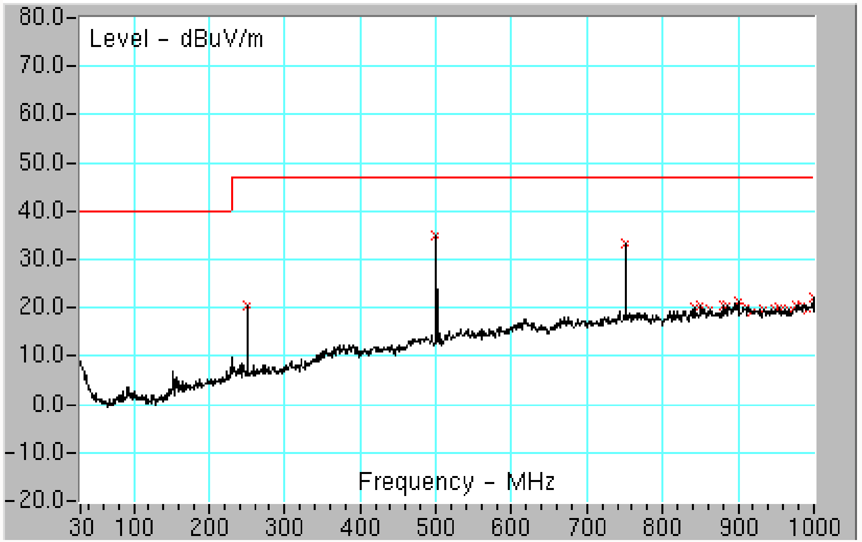
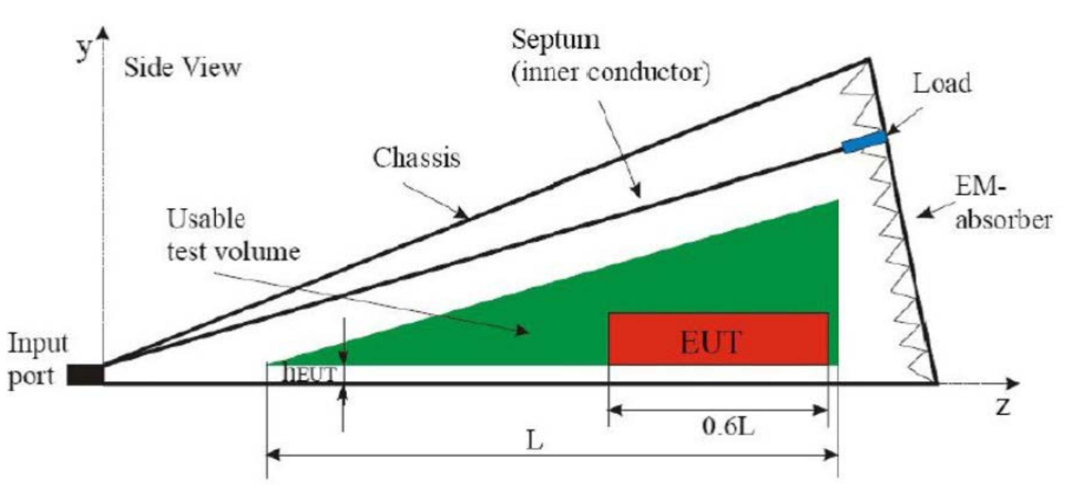


Shielding Evaluation – GTEM

1. The G-TEM (GHz - TEM) is a large version of the coax TEM idea and can be used for RE (9 kHz – 3 GHz), or radiated immunity (80 MHz – 6 GHz). The EUT must fit in a specified and relatively small volume, much smaller than the whole GTEM volume, therefore without significantly affecting the internal field.
2. Correlation GTEM to OATS is included in the software.
3. This TEM waveguide is based on IEC 61000-4-20 standard.
4. The frequency range can be extended up to 20 GHz and is very easy to switch from radiated emissions to radiated immunity using a receiver or a signal generator and amplifier as input.
5. Input impedance is 50 Ω and can achieve 10 V/m inside.

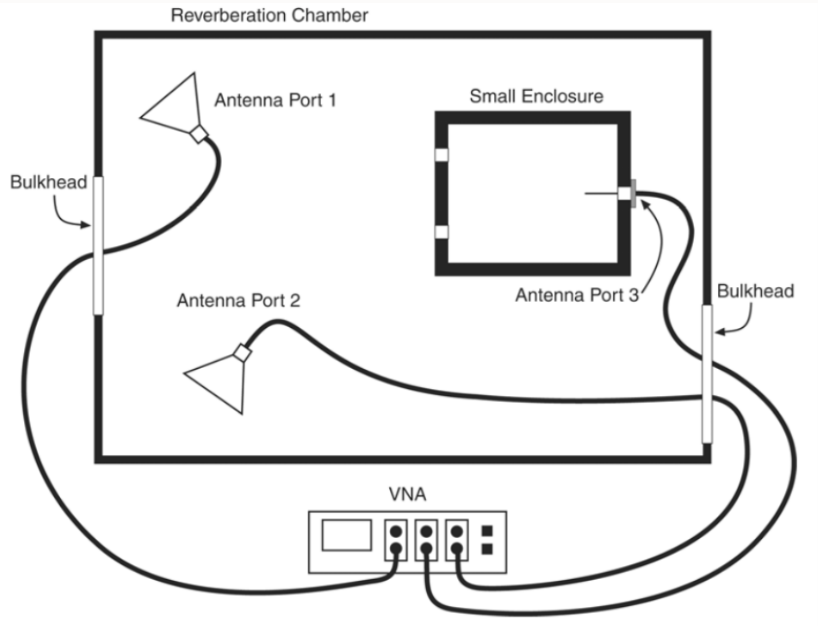


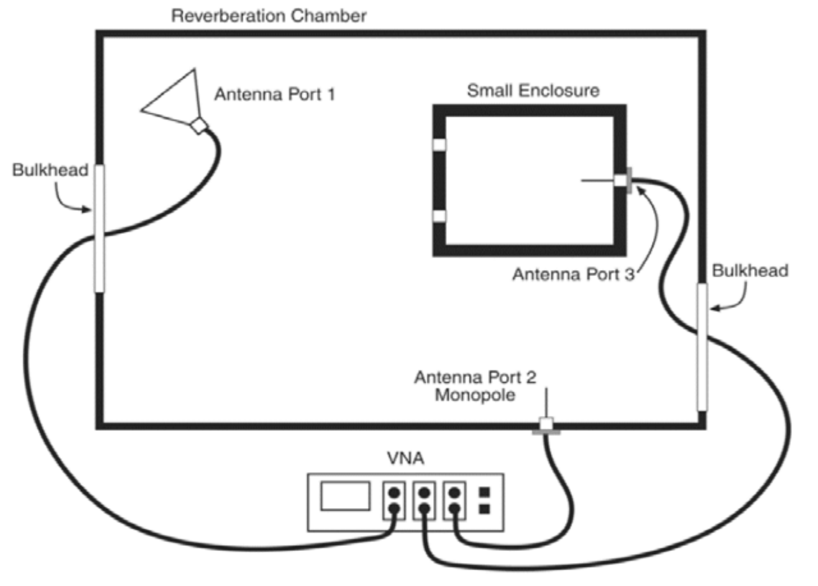
Shielding Evaluation – Reverberation Chambers

1. The reverberation chamber uses a slowly moving paddle to constantly change the boundary conditions and create a statistically uniform field.
2. For a physically small enclosure (<0.75 m max dimension) but electrically large, the IEEE 299-2006 proposes a nested reverberation chamber method.
3. Antenna Port 1 excite energy inside the outer reverberation chamber, and Antenna Port 2 monitors this field. Stirring in the outer chamber may be done with a mechanical stirring paddle.
4. Some portion of the energy in the outer chamber will couple into the small enclosure, and cause frequency stirring within the latter.
5. All points in the small enclosure statistically have the same field levels for the data averaged over some bandwidth of frequencies, no sampling location issue, the field inside the small enclosure can be measured anywhere.
6. The power levels in the small enclosure are monitored by a small monopole probe mounted on its wall. **Difficult to troubleshoot, unknown slot/seam.**
7. The $\langle\ \rangle$ represents the ensemble averaged over some frequency bandwidth (and paddle position if a combination of frequency and mechanical stirring in used in the outer reverberation chamber).

$$SE=-10\log\left(\frac{P_{in}}{P_{out}}\right)$$

$$SE=\frac{\langle|S_{31}|^2\rangle}{\langle|S_{21}|^2\rangle}\frac{1-|\langle S_{22}\rangle|^2}{1-|\langle S_{33}\rangle|^2}$$



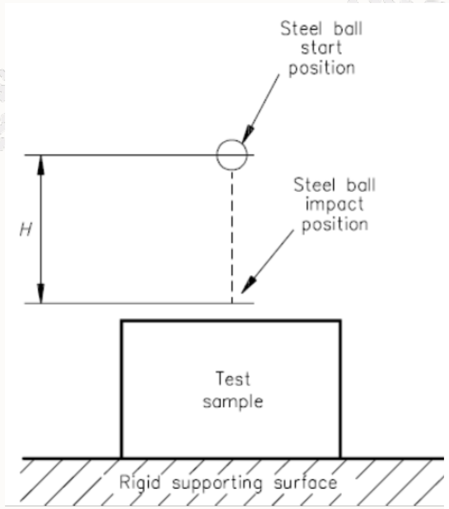


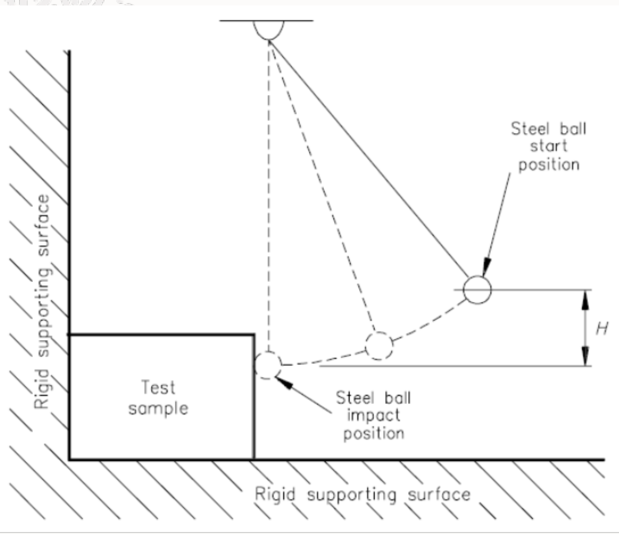
Outline

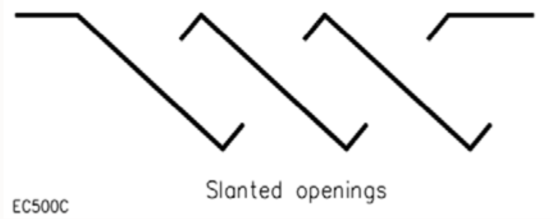
1. Introduction, near field, far field, electric, magnetic and electromagnetic shielding
2. Analytical approaches to shielding. Field theory method (Kaden) and TL impedance method (Schelkunoff)
3. Limits of the theoretical approaches, numerical simulations, reciprocity
4. Chassis resonances, internal compartmentalization, absorbers
5. Practical aspects of shielding, aperture coupling, slots, seams, rivets
6. Shielding and thermal issues, holes, perf patterns, honeycomb
7. Shielding and internal sources, heatsink grounding, local shielding
8. Shielding and coatings, corrosion
9. Evaluation of shielding effectiveness, materials, chassis
10. Shielding and safety certification
11. References
12. Acknowledgments

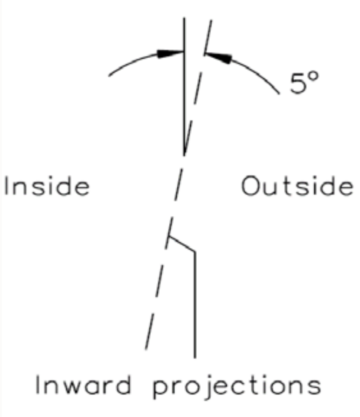
Shielding and Safety Certification – Mechanical Enclosure

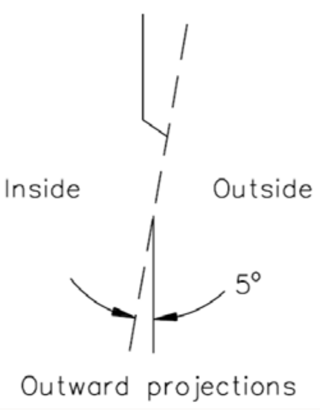
1. A chassis is a **mechanical enclosure**, providing structural strength, but also an **electrical enclosure** limiting access to parts at hazardous voltage (42.4 V peak, or 60 V d.c.) or hazardous energy (stored energy ≥ 20 J or available continuous power ≥240 VA at a potential of ≥2V) levels or are in TNV circuits. The chassis is also a **fire enclosure** limiting the spread of fire or flames from within. All these safety aspects regulated by **UL 1950** standard need to be considered. TNV stands for Telecommunication Network Voltage.
2. Mechanical strength is tested by applying a force of 250 N ± 10 N for 5s with a plane circular test tool with 30 mm diameter. There is also an impact test for external for external surfaces of enclosures. A steel ball, 50 mm in diameter and with a mass of 500 g ±25 g, is permitted to fall freely from rest through a vertical distance (H) of 1.3 m onto the sample. In addition, the steel ball is suspended by a cord and swung as a pendulum in order to apply a horizontal impact, dropping through a vertical distance (H) of 1.3 m. These tests will matter for plastic parts, for example a metalized plastic honeycomb panel.
3. Openings in the top and sides shall be so located or constructed that it is unlikely that objects will enter the openings and create hazards by contacting bare conductive parts.





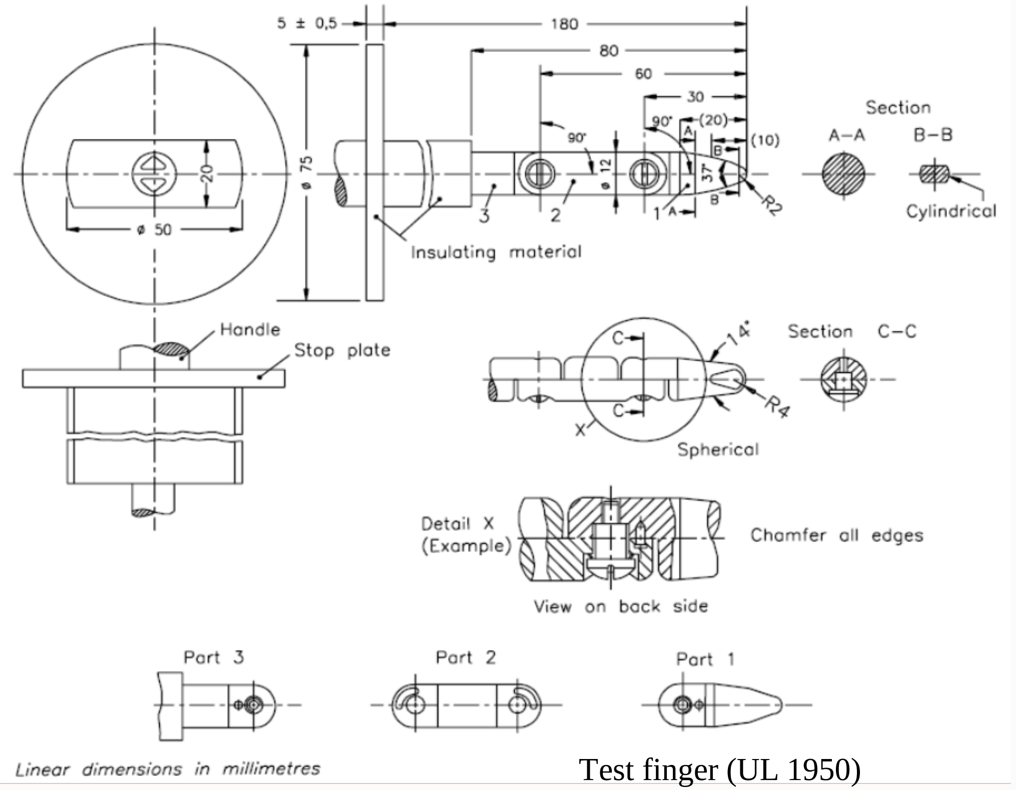




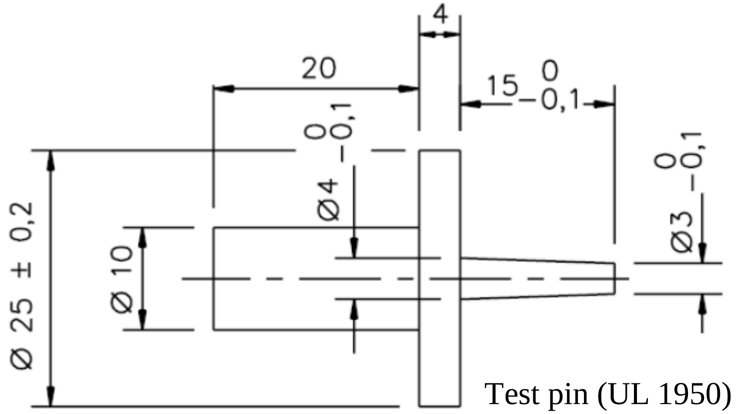


Shielding and Safety Certification – Electrical Enclosure

1. A chassis as an **electrical enclosure** limiting access to parts at hazardous voltage (42.4 V peak, or 60 V d.c.) or hazardous energy (stored energy ≥ 20 J or available continuous power ≥240 VA at a potential of ≥2V) or parts in TNV circuits is tested using a **test finger,** applied with a force of 30 N, which shall not contact parts described above when applied to openings in the enclosures after removal of parts that can be detached by an operator, including fuse holders, and with operator access doors and covers open. It is permitted to leave lamps in place for this test.
2. Alternatively, a test with the **test pin** which shall not contact bare parts at **hazardous voltages** when applied to openings in an external **electrical enclosure**. Parts that can be detached by an operator, including fuse holders and lamps, are left in place, and operator access doors and covers are closed during this test.
3. The lid lock must have **safety interlocks** so designed that the hazard will be removed before the covers, doors, etc. are in any position that will permit contact with hazardous parts by the test finger. If a failure of the interlock system during the normal life of the equipment is possible, the probable failure mode(s) will not create a hazard for which protection is required.



Test finger (UL 1950)

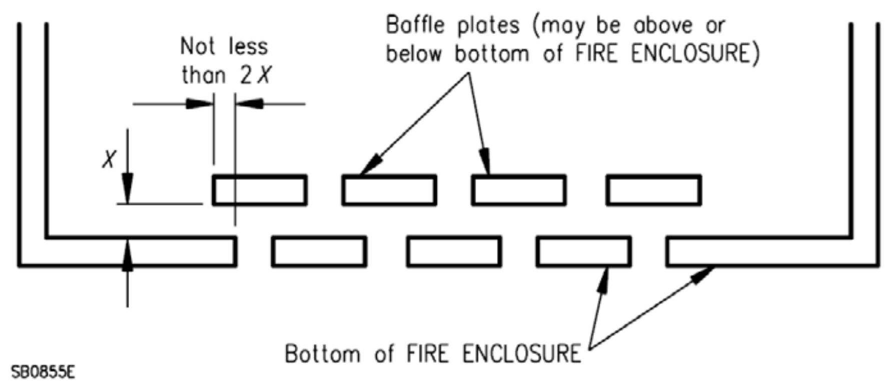


Test pin (UL 1950)

Shielding and Safety Certification – Fire Enclosure

1. A chassis as a **fire enclosure** must use components and other parts constructed so that the propagation of fire from within is limited. This typically means using materials in the V-0 or V-1 flammability class (UL1950), which may flame or glow but will extinguish in <1 min and glowing particles or flaming drops released do not ignite surgical cotton.
2. Inside **fire enclosures** materials for components and other parts can be of flammability class V-2, or HF-2 (UL1950).
3. Air filter assemblies shall be constructed of materials of flammability class V-2, or HF-2.
4. The bottom of a **fire enclosure**, or individual barriers, shall provide protection under all internal parts, including partially enclosed components or assemblies, which, under fault conditions, could emit material likely to ignite the supporting surface.
5. The bottom or barrier shall provide equivalent protection. An opening in the bottom shall be protected by a baffle, screen or other means so that molten metal and burning material are unlikely to fall outside the **fire enclosure**.
6. Size and spacing of openings in metal bottoms of fire enclosures are included in the table.

	Applicable to circular holes		Applicable to other shaped openings	
Metal bottom minimum thickness mm	Maximum diameter of holes mm	Minimum spacing of holes centre to centre mm	Maximum area mm2	Minimum spacing of openings border to border mm
0,66	1,1	1,7	1,1	0,55
0,66	1,2	2,3	1,2	1,1
0,76	1,1	1,7	1,1	0,55
0,76	1,2	2,3	1,2	1,1
0,81	1,9	3,1	2,9	1,1
0,89	1,9	3,1	2,9	1,2
0,91	1,6	2,7	2,1	1,1
0,91	2,0	3,1	3,1	1,2
1,0	1,6	2,7	2,1	1,1
1,0	2,0	3,0	3,2	1,0



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18. Standards – IEEE, MIL, IEC, etc.

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Electromagnetic Shielding Concepts and Applications

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