Interaction Note

Note 617

April 2011

Allocating Indirect Lightning to Cables & Boxes at Program Inception

Application of Ohm's Law, Kirchhoff's Laws, Faraday's Law &
Scaling by Geometric, Electrical, & Spectral Parameters

Larry West
10215 Beechnut St, Suite 1003
Houston, TX 77072

abstract

Indirect lightning engineering needs early numbers for allocations of weight and space for protection in weight conscious composite systems' programs. This note attempts to do define the in-flight environments and provide the tools for scaling them to any size system. Waveform 5A (WF5A) in the standards has been changed to Waveform 4 (WF4) based on findings in IN615. All other allocations are affected as more high frequency penetrates onto interior conductors and less on the exterior. Lightning Components A, D, and H have about the same dI/dt and the same spectra above 2MHz therefore they will excite the same levels of Waveform 2 derivative excitation and Waveform 3 resonant excitation. Design guidelines help meet requirements.

1. Summary of Interactions that produce Indirect Lightning Transients

1.1. Introduction. Figure 1 is a simplified picture of aircraft cables subject to indirect lightning interactions.[3,13] Since the lightning phenomena are mostly low frequency, simple RLC models can adequately model most lightning interactions with cables. Zone 3 is assumed everywhere.

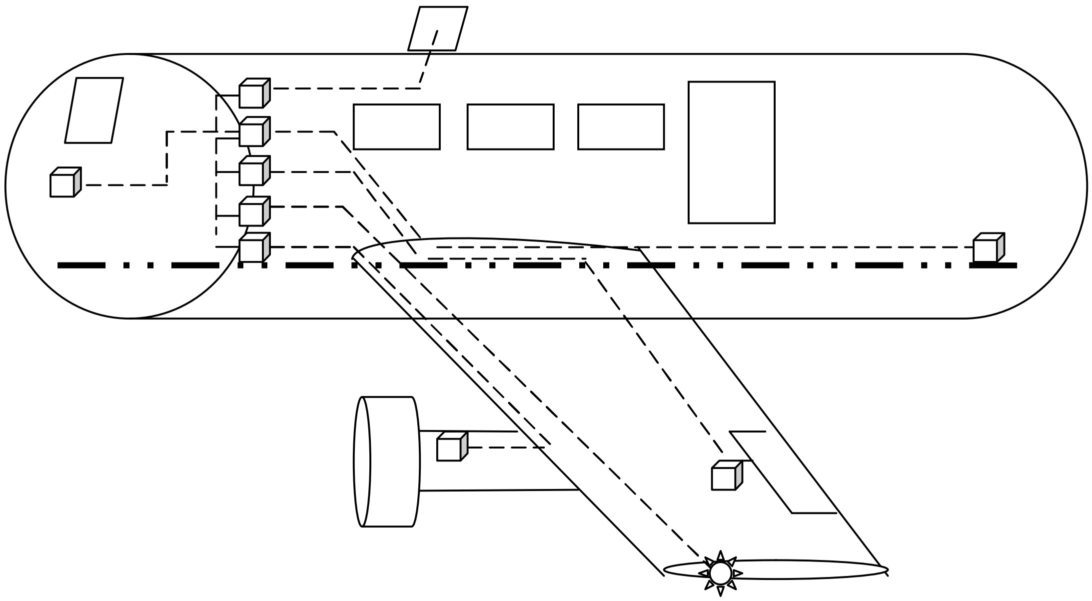


Figure 1. Notional Aircraft Cables that Couple to Lightning

Lightning interactions considered herein:

a. I·R-voltage drop across airframes and the corresponding induced cable current,
b. inductive coupling to cables along aircraft wing spars and rocket raceways,
c. inductive coupling through apertures – doors, windows, etc.,
d. cross coupling to less exposed cables from exposed cables such as in a, b, and c, above,
e. coupling to wiring connected to external loads, i.e. lights, pitot, temp, antennas, WOW, etc.,
f. inductive coupling of cables to fields from nearby strikes, and
g. surges on ground connections including lightning.

The last two are included because of the long periods of time all aerospace systems spend on the ground connected to ground power and ground support equipment (GSE) from manufacturing to test to operations and maintenance. Overstress from ground power and underspecified GSE can reduce operational reliability and longevity.

It is assumed herein that all of the indirect conducted lightning environments are common mode - in line, line-to-chassis ground, or ground potential as depicted in Figure 2, below. Differential mode wire-to-wire coupling is more of a mode conversion issue.

A nagging problem for lightning protection is the scarcity of electronics' susceptibility to the WF4 I·R-drop ground potential difference between boxes. One reason is that the DO-160 tests do not properly simulate this phenomenon.[15] Inspection hints that semiconductor substrate latchup is a potential susceptibility; substrates are -0.5V pn junctions connected to every component on a chip and to chip ground.

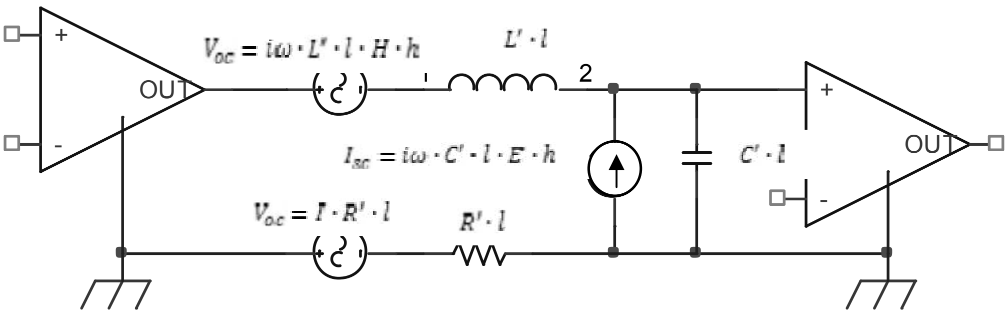


Figure 2. EMI & Lightning Common Mode Voltage & Current Sources

$\boldsymbol{V_{ov} = i\omega \cdot L' \cdot l \cdot H \cdot h}$ in wires from magnetic fields coupling directly or through a shield.
$\boldsymbol{V_{oc} = I \cdot R' \cdot l}$ from current running through a common chassis ground and/or a cable shield.
$\boldsymbol{I_{sc} = i\omega \cdot C' \cdot l \cdot E \cdot h}$ from an electric field between the cable and nearest chassis ground, coupling directly or through a shield.

1.2. Models. Regarding lossless antennas small compared to the wavelength of the incident radiation, the following figure developed by Baum[6,18,19] is a lesson about the relationships between the different electromagnetic fields, their coupling to different conductors, and the geometric and electromagnetic parameters of the conductors. E, dB/dt, H, and dD/dt each couple to different conductor geometries.

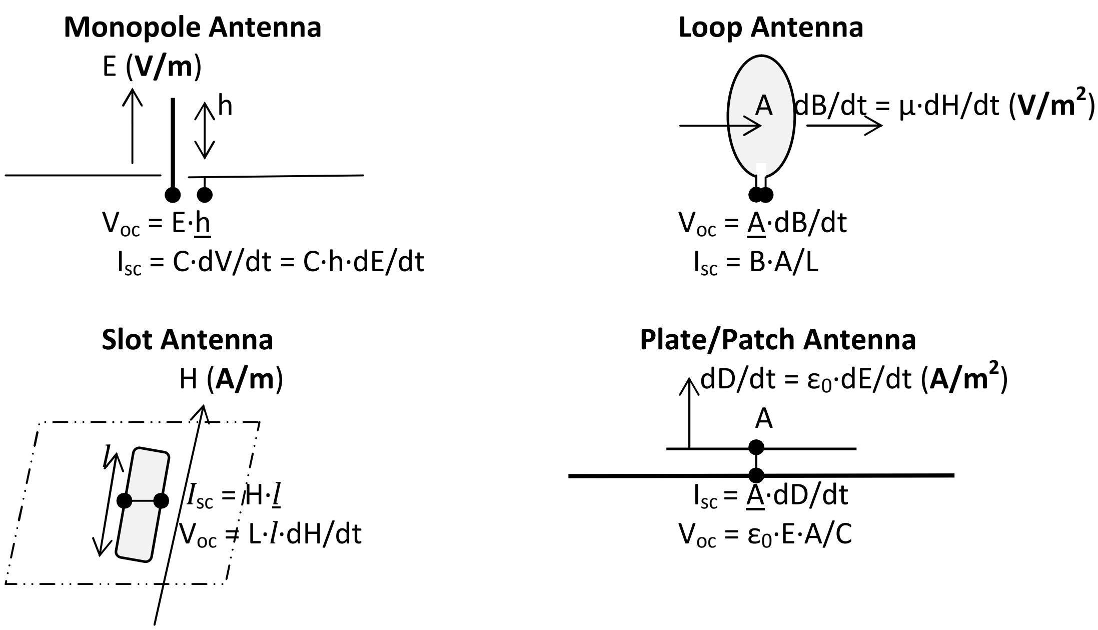


Figure 3. Four Basic Fields and Four Basic Antennas[5,18,19]
A Lesson on Maxwell's Equations & Low Frequency Coupling to Electromagnetic Fields

1.3. Parameters. The RLGC circuit parameters that are used to model interactions electromagnetic fields around conductors and voltages and currents within conductors are summarized below.[6,7,9] <u>Understanding and estimating the resistance and inductance of each conductor from in-flight to system level ground-tests to bench level box tests is fundamental to understanding the indirect lightning phenomena</u>. It is also necessary for designing, manufacturing, and controlling indirect lightning protection and ultimately for certification.

(1) $C = \frac{\varepsilon \cdot A_{eq}}{l} = \frac{\iint \vec{D} \cdot \vec{dA}}{\oint \vec{E} \cdot \vec{dl}} = \frac{I}{\dot{V}}$, capacitance in terms of geometric, fields, & a circuit parameter;

(2) $L_{ext} = \frac{\mu \cdot A}{l_{eq}} = \frac{\iint \vec{B} \cdot \vec{dA}}{\oint \vec{H} \cdot \vec{dl}} = \frac{V}{\dot{I}}$, external inductance in terms geometry, fields, & a circuit parameter;

(3) $L_{int} = \frac{\mu_0 \cdot \iiint |H|^2 \cdot dV}{I^2}$, internal inductance by the energy method;

(4) $R'_{dc} = 1/\pi \cdot r^2 \cdot \sigma$, the DC resistance per unit length of a wire;

(5) $R'_{dc} = 1/2 \cdot \pi \cdot r \cdot t \cdot \sigma$, the DC resistance per unit length of a cylinder, radius, r; thickness, t;

(6) $L'_{int} = \frac{\mu_0}{8 \cdot \pi}$, the low frequency internal inductance per unit length of a solid cylinder;

(7) $L'_{int} = \frac{\mu_0}{4 \cdot \pi} \cdot \frac{t}{r}$, the low frequency internal inductance of a cylindrical shell, radius, r; thickness, t;

(8) $R'_{ac} = 1/2 \cdot \pi \cdot r \cdot \delta \cdot \sigma$, the AC resistance of a wire or a cylinder;

(9) $L'_{int} = \frac{\mu_0}{4 \cdot \pi} \cdot \frac{\delta}{r}$, the high frequency internal inductance per unit length of a solid cylinder and a cylindrical shell;

(10 $G'_{dc} = \frac{\sigma}{\varepsilon_0} \cdot C'$, the DC conductance per unit length of wire insulation; typically, $G_{dc} << 10^{-7}$;

(11) $G'_{ac} \cong \omega \cdot C' \cdot \tan \delta_d$, the AC conductance per unit length of wire insulation; $\tan\delta_d \equiv$ loss tangent.

1.4. EM Laws The relations, $V = I \cdot R$, $H = I/2 \cdot \pi \cdot R$, and $V_{oc} = i\omega \cdot B \cdot A$ derive from Ohm's Law, Ampere's Law, and Faraday's Law, respectively.[9] Kirchhoff's voltage and current laws then complete the arsenal.

Faraday's Law:	$\oint \vec{E} \cdot \vec{dl} = \iint \dot{\vec{B}} \cdot \vec{dA}$	or	$\nabla \times \vec{E} = -\dot{\vec{B}}$
Ampere's Law + $\dot{\vec{D}}$:	$\oint \vec{H} \cdot \vec{dl} = \iint \vec{J} \cdot \vec{dA} + \iint \dot{\vec{D}} \cdot \vec{dA}$	or	$\nabla \times \vec{H} = \vec{J} + \dot{\vec{D}}$
Coulomb's Law:	$\iint \vec{D} \cdot \vec{dA} = \iiint \rho \cdot dV$	or	$\nabla \cdot \vec{D} = \rho$
Source-free B-field:	$\nabla \cdot \vec{B} = 0$		
Permittivity:	$\vec{D} = \varepsilon \cdot \vec{E}$,	$\varepsilon_0 = 8.85pF/m$ in free space	
Permeability:	$\vec{B} = \mu \cdot \vec{H}$	$\mu_0 = 1.26\mu H/m$ in free space	
Ohm's Law:	$\vec{J} = \sigma \cdot \vec{E}$		

Figure 4. Maxwell's Equations & Constitutive Relations[9]

1.5. Structural I·R Voltage Drop

The most unique indirect lightning environment in composite aircraft is the large voltage induced along the path of the lightning current through the system. The corresponding cable current is different, also, due to the unique in-flight interaction, and is the focus of Paper #1[13].

1.5.1. Voltage IR-Drop across the Airframe and Induced Cable Currents and Voltages

In IN608[14], Paper #3[13], and Paper #4[15], we employed the heuristic RL model in Figure 5, below, as much as possible. Now, we must parameterize the resistance and inductance so that designers can construct their own system unique models and add a groundplane to reduce the cable excitation. First, we must look at airborne versus ground test in a never-before-seen way.

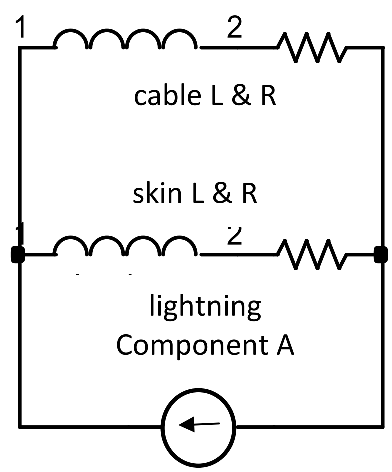


Figure 5. Simple Heuristic Parallel RL System Circuit Model[13]
Model can be constructed arbitrarily more complicated.

1.5.2. Resistance and Internal inductance of a Cylinder[7]

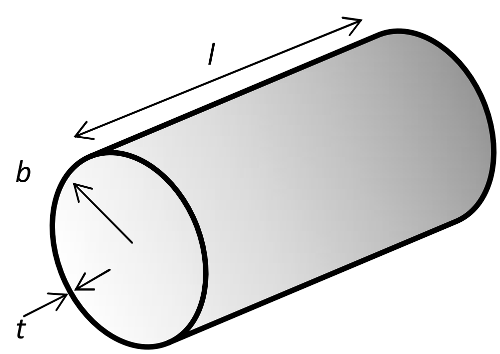


Figure 6. Cylindrical Shell Geometric Parameters
Cylinder of length *l*, radius *b*, skin thickness *t*, permeability, *μ*, and conductivity *σ*

Schelkunoff's approximation for the internal impedance of a conducting coaxial cylinder is:[7]

(12) $\boldsymbol{Z_{int} \cong R_{dc} \cdot \sqrt{i\omega \cdot \tau_d} \cdot \coth\left(\sqrt{i\omega \cdot \tau_d}\right)}$, where $\boldsymbol{\tau_d = \mu \cdot \sigma \cdot t^2}$ is the diffusion time.

The detailed parameters are given in Section 1.3, above.

A solid groundplane of length, l, thickness, t, and width, w, has a DC resistance of

(13) $$\boldsymbol{R_{gndplane}} = \frac{l}{w \cdot t \cdot \sigma}.$$

Expanded foil mesh is usually specified in terms of the DC ohms per square, R_{square}, therefore the groundplane DC resistance for each segment of length, l, and width, w, is

(14) $$\boldsymbol{R_{gndplane}} = \boldsymbol{R_{square}} \cdot \frac{l}{w}$$

1.5.2. I·R-Voltage Drop and Corresponding Cable Currents, Models of In Flight & Ground Test[13,14]

The following circuit formulas govern the induced current and voltage from a 200kA Component A lightning current in a 10m CFC airframe at 10mΩ/m, R_s, with and without a 15mΩ groundplane, R_g, and with R'_{TSP} = 40mΩ/m TSPs and AWG 36 wire braid R'_{OVB} = 10mΩ/m cable shields. The DC resistance is inversely proportional to both the radius and thickness of the skin and cable shields. The numerical examples below are just that, examples; each system requires use of its own numbers. The in-flight interaction renders the internal inductance of each component too small to consider.

Parallel resistance of the skin and groundplane the same length:

(15) $$\boldsymbol{R_p} = \frac{R_s \cdot R_g}{R_s + R_g},$$

Parallel resistance of n cables the same length:

(16) $$\boldsymbol{R_{nc}} = \frac{R_c}{n};$$

Voltage across the CFC skin and n cables the same lengths, no groundplane:

(17) $$\boldsymbol{V_{no-g}} = \boldsymbol{I_L} \cdot \frac{R_s \cdot R_c}{n \cdot R_s + R_c}, \text{ where } I_L = 200kA;$$

Voltage across the CFC skin and n cables, with groundplane:

(18) $$\boldsymbol{V_{ng}} = \boldsymbol{I_L} \cdot \frac{R_p \cdot R_c}{n \cdot R_p + R_c};$$

Current per cable with a CFC skin and no groundplane:

(19) $$\boldsymbol{I_{c-no-g}} = \frac{V_{no-g}}{R_c} = \boldsymbol{I_L} \cdot \frac{R_s}{n \cdot R_s + R_c}, \text{ where } R_c = R_{TSP} \text{ or } R_{OVB};$$

Current per cable with a CFC skin and with a groundplane:

(20) $$\boldsymbol{I_{c-g}} = \frac{V_{ng}}{R_c} = \boldsymbol{I_L} \cdot \frac{R_p}{n \cdot R_p + R_c}, \text{ etc. and so on.}$$

The following examples use these parameters: $R_{skin} = R_s = 100m\Omega$, $R_{gndplane} = R_g = 15m\Omega$, $R_{TSP} = \frac{40m\Omega}{m} \cdot 10m = 400m\Omega$, and $R_{OVB} = \frac{10m\Omega}{m} \cdot 10m = 100m\Omega$, all scaled by the ratio of length/(width or radius x thickness x conductivity) for the reader that has different parameters.

1.5.2.2. Waveforms & Current Division, In-Flight versus Ground-Test with No Groundplane

The division of lightning current between the CFC skin and one cable is illustrated in Figure 7, below, for the in-flight case with internal inductance for all common mode currents.

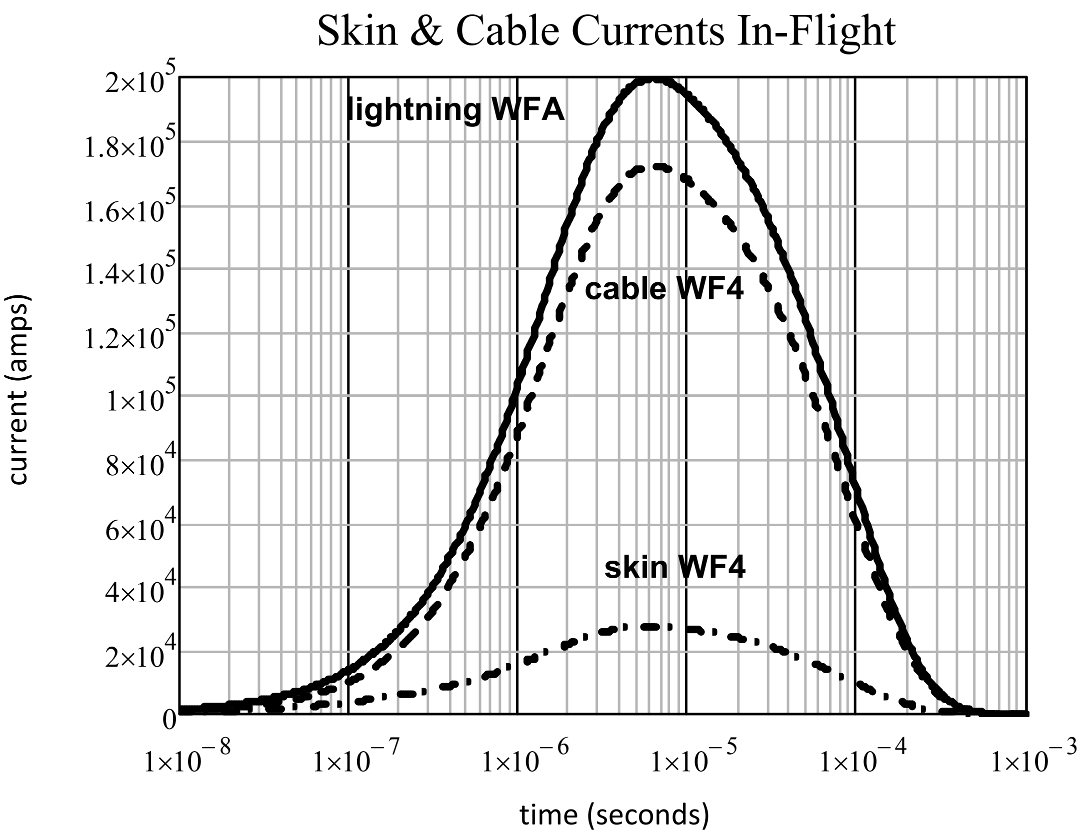


Figure 7. In-Flight Induced Current on a CFC Skin and One Cable with No Groundplane
Cable current, I_t = 172kA, Skin Current = 28kA

We will show below that with a groundplane, the current induced on one out of ten TSPs is 4kA, a 13dB reduction. Also, we will show how dependent this result is on cable shield thickness. The groundplane reduces the induced transients more on poor conductors like the 40mΩ/m flat foil braid TSP, above, than on better conductors like a 10mΩ/m 36AWG wire braid shield. When the number of cables exceeds ten, the groundplane no longer reduces induced voltages.

The positive side to this phenomenon is that the groundplane only has to meet DC resistance requirements for the in-flight interaction. In the ground-test interaction, below, the groundplane had to also meet low inductance requirements in order to reduce the induced transients which means it had to have considerable width independent of resistance requirements.

The division of lightning current between the CFC skin and one cable for the ground-test case is illustrated in Figure 8, below, with external inductance for all common mode currents. The external inductance has the most effect on the rise time of the cable currents, 18µs ground-test WF5A versus 1.5µs in-flight WF4.

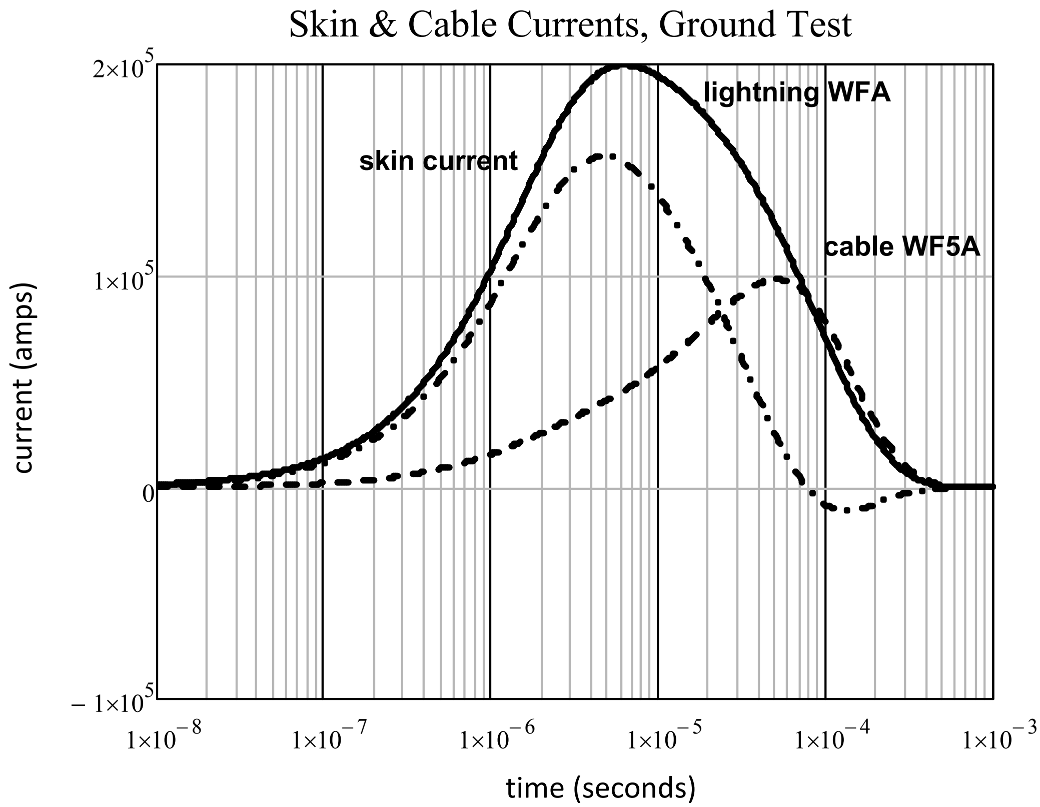


Figure 8. Ground-Test Induced Current on a CFC Skin and One Cable with No Groundplane
Cable current, I_t = 100kA, Skin Current = 170kA

1.5.2.3. Induced Current and Voltage vs. Number of Cables with & without a Groundplane

Figure 9 is the current division between the 100mΩ skin currents I_4 and I_3 and the interior I_4 with no groundplane on n = 1-100 40mΩ/m TSPs. The WF4 skin current ranges from 170kA to 12kA and the WF3 skin current ranges from 20kA to 1kA. The total WF4 current running through the interior on the TSPs ranges from 30kA to almost 200kA and with a rise time of 1.5µs not the 18µs of WF5A.

Figure 10 is the current division between the 100mΩ skin I_4 and I_3 and the interior I_4 with a 15mΩ groundplane and n = 1-100 40mΩ/m TSPs. The WF4 skin current ranges from 26kA to 8kA and the 1MHz WF3 skin current ranges from 3kA to 800A. The total current through the interior stays close to 200kA.

Figure 11 is the total voltage induced end-to-end with only TSP cables and with and without a 15mΩ groundplane. As the number of cables increases, the effect of the groundplane diminishes, i.e. $R_c/n \approx R_g$. The WF4 voltage without the groundplane ranges from 14kV to 1.1kV. With the groundplane, the WF4 voltage ranges from 3kV to 900V.

The large "n" extrapolation assumes the voltage is inversely proportional to "n" meaning that the groundplane is assumed to be far less resistive than any number of cables.

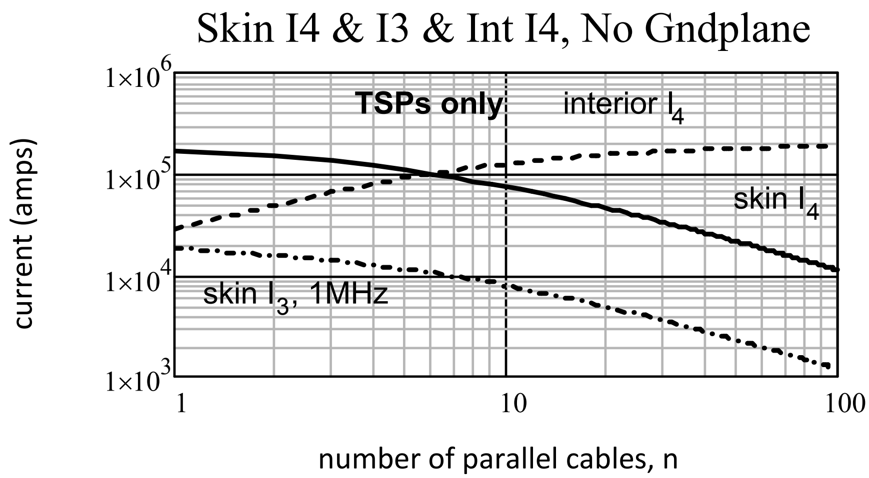


Figure 9. Skin WF4 & WF3 Current and Cable Current for n = 1-100 Cables, No Groundplane

This illustrates that the in-flight interaction, as opposed to the present-day ground-test, results in far more current through the interior and far less on the exterior, subject to a proper test.

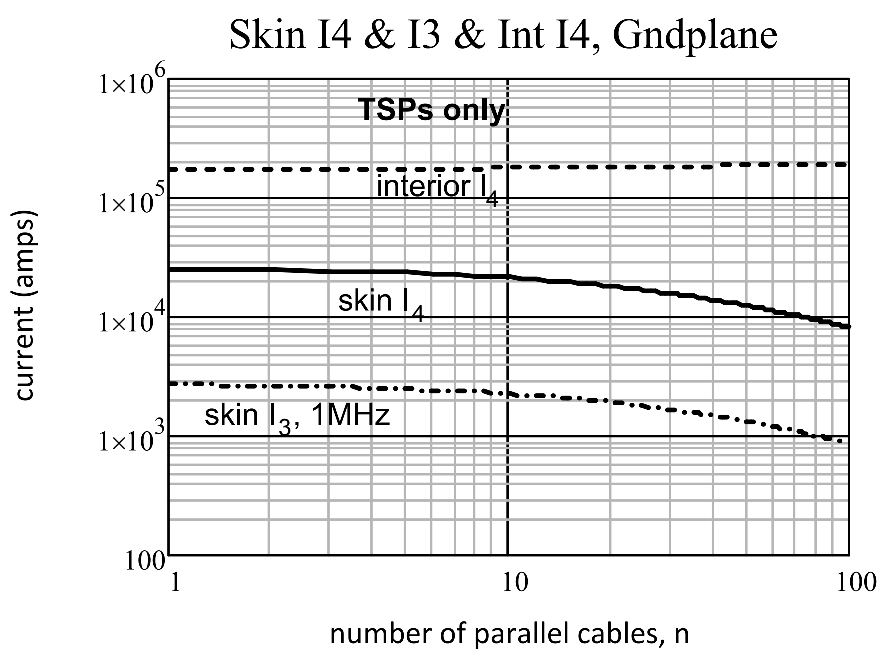


Figure 10. Skin WF4 & WF3 Current and Cable Current for n = 1-100 Cables, with Groundplane

A groundplane increases the interior current and reduces the exterior current by almost an order of magnitude, good reduction of induced voltage but the opposite for induced current.

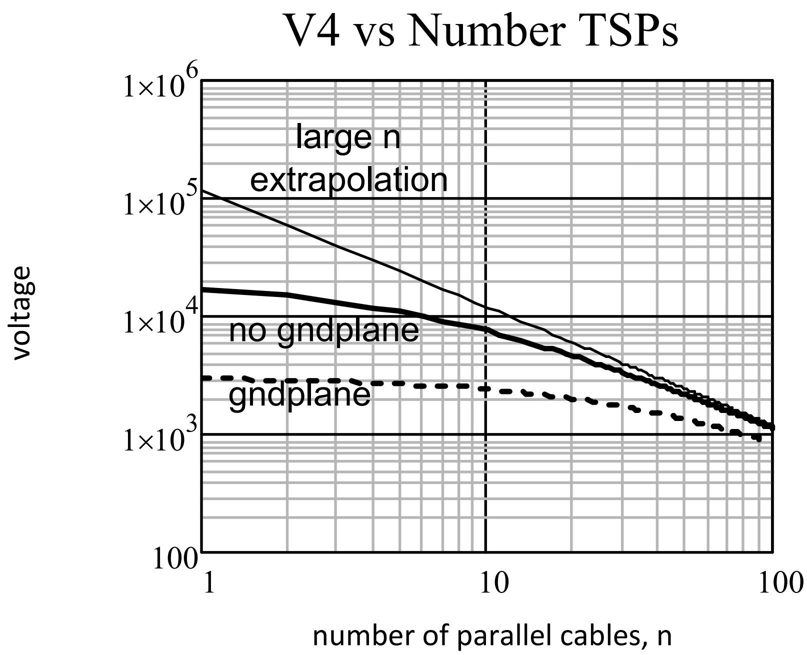


Figure 11. I·R Voltage Drop between Two Points 10m Apart for n = 1-100 TSPs with & without a Groundplane

$$\boldsymbol{V_{no-g} = I_L \cdot \frac{R_s \cdot R_c}{n \cdot R_s + R_c}} \ \& \ \boldsymbol{V_{ng} = I_L \cdot \frac{R_p \cdot R_c}{n \cdot R_p + R_c}}$$

Figure 12 is the total voltage induced end-to-end with only AWG 36 overbraid cables and with and without a 15mΩ groundplane. As the number of cables increases, the effect of the groundplane diminishes. The WF4 voltage without the groundplane ranges from 10kV to 200V. With the groundplane, the WF4 voltage ranges from 2.6kV to 200V. The voltage is dependent on the distance between inter-connected boxes. The resistance is proportional to the ratio of length/(width·thickness).

Figure 13 shows the induced WF4 current on one out of n = 1-100 TSP cables (40mΩ/m), with and without the 15mΩ groundplane. Without the groundplane, the individual cable current ranges from 20kA to 2kA. With the groundplane, the cable current ranges from 5kA to 1.4kA. The current is independent of length between inter-connected boxes.

Figure 14 shows the induced WF4 current on one out of n = 1-100 AWG 36 wire braid shielded cables (10mΩ/m), with and without the 15mΩ groundplane. Without the groundplane, the individual cable current ranges from 100kA to 2kA. With the groundplane, the cable current ranges from 28kA to 2kA. The current is independent of length between inter-connected boxes.

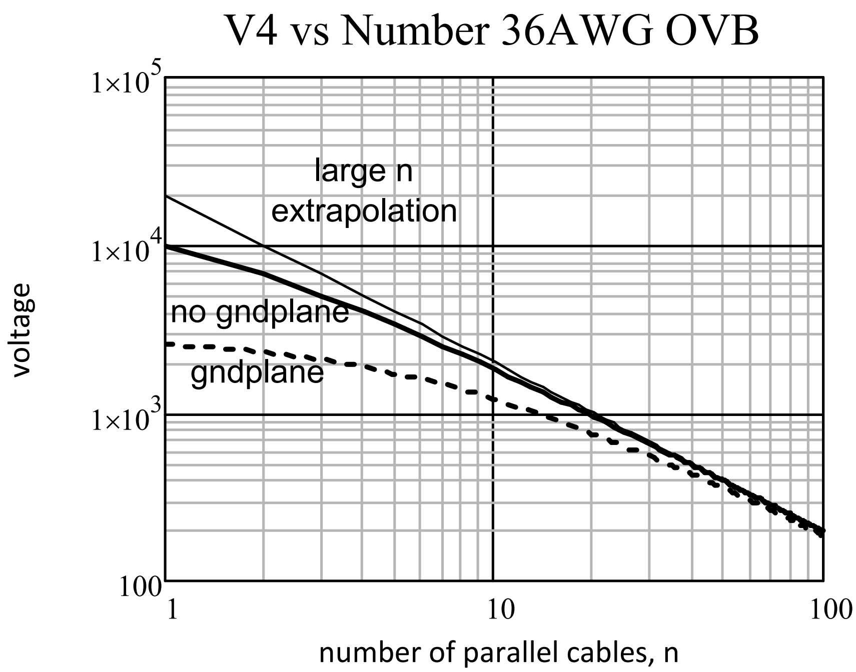


Figure 12. I·R Voltage Drop between Two Points 10m Apart for n = 1-100 36AWG OVBs with & without a Groundplane

$$V_{ng} = I_L \cdot \frac{R_p \cdot R_c}{n \cdot R_p + R_c}.$$

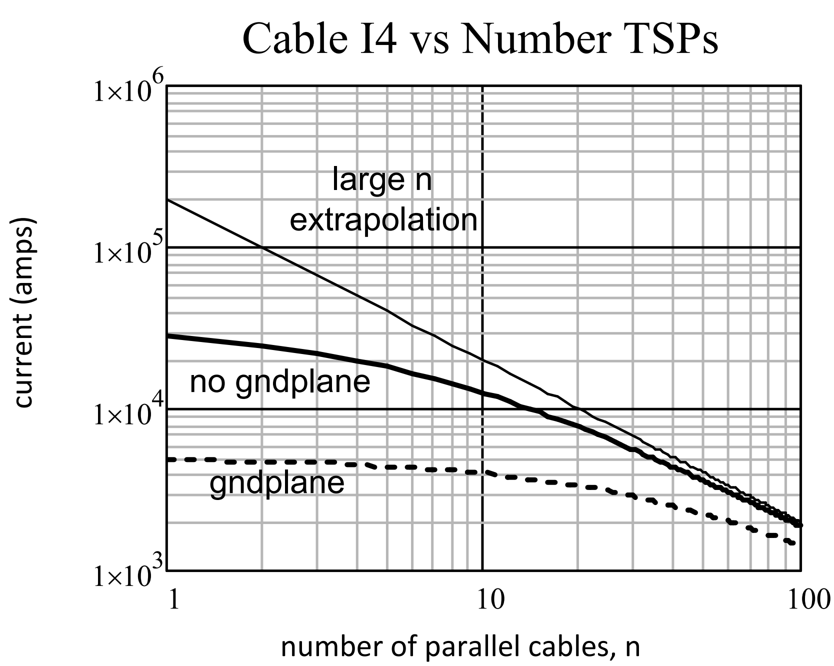


Figure 13. Individual Cable Shield Current for n = 1-100 TSPs with & without a Groundplane

$$I_{c-g} = I_L \cdot \frac{R_p}{n \cdot R_p + R_c}.$$

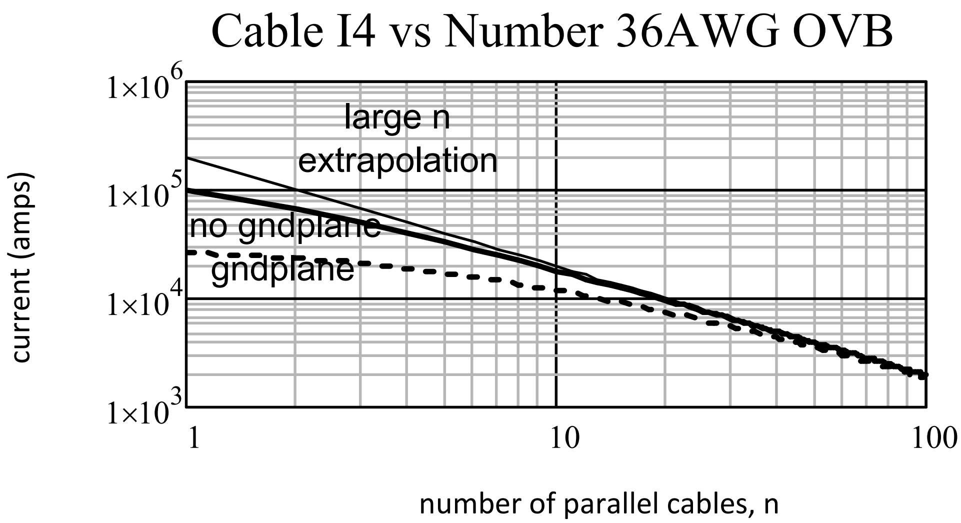


Figure 14. Individual Cable Shield Current for n = 1-100 36AWG cable Shields with & without a Groundplane

$$I_{c-g} = I_L \cdot \frac{R_p}{n \cdot R_p + R_c}.$$

The groundplane reduction of induced current and voltage for a single cable ranges from 11dB to 16dB on one cable to almost no effect for 100 cables, $V_{l \cdot R} = I_L \cdot \frac{R'_s \cdot l \cdot R'_c \cdot d}{n \cdot R'_s \cdot l + R'_c \cdot d}$, where I_L is the lightning current, R'_S is the composite skin resistance per unit length, l is the length of the system, R'_c is the cable resistance per unit length, d is the distance between boxes, and "n" is the number of parallel cables.

The corresponding current on one out of "n" parallel cables is $I_{l \cdot R} = \frac{V_{l \cdot R}}{R_c} = I_L \cdot \frac{R'_s \cdot l}{n \cdot R'_s \cdot l + R'_c \cdot d}$.

1.6. Aperture Coupling to Cables. Most EMI and lightning interactions end up as coupling to cables. At and above their resonant frequencies, cables can be modeled as distributed transmission lines, Figure 15. At and below their resonant frequencies, cables must be modeled as lumped element circuit models, Figure 16. Z' and Y' denote units per unit length although it doesn't matter.

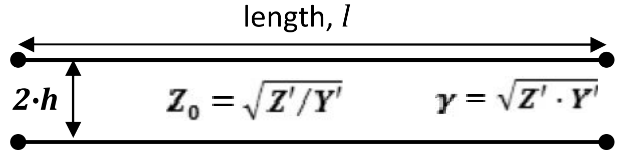


Figure 15. Distributed Transmission Line, Impedance Z_0, , Height, h, and Length, l

The open circuit voltage induced in this line at low frequencies is $V_{oc} = A \cdot dB/dt = E \cdot 2 \cdot h$, where $A = l \cdot 2 \cdot h$ and $B = E/c$. The short circuit current induced is $I_{sc} = BA/L$ where L is the transmission line inductance. The voltage is proportional to the line length; the current is not. The inductance limits the current.

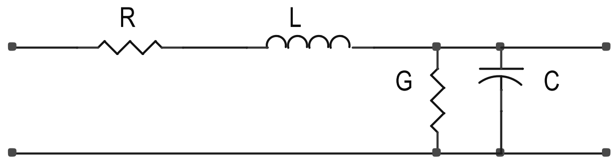


Figure 16. RLGC Discrete Model of Transmission Line Incremental Length or RLGC Model of Transmission Line up to First Resonance, $f < c/4 \cdot \pi \cdot l$

Aperture coupling to transmission lines is modeled as a point voltage source for magnetic coupling from a point magnetic dipole source in a nearby aperture.[5,24]

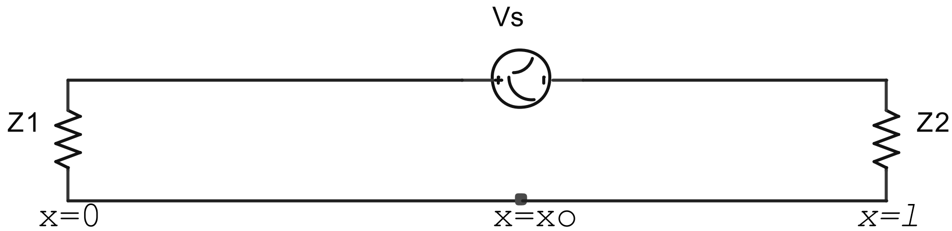


Figure 17. Discrete Voltage Source at x = x_0 in a Transmission Line

The point dipole source in the aperture is a distance, R, from its equivalent point source in the line at x = x_0 along the line. In Figure 18, below, R is the diagonal distance to the aperture and d is the perpendicular height above the plane of the aperture:

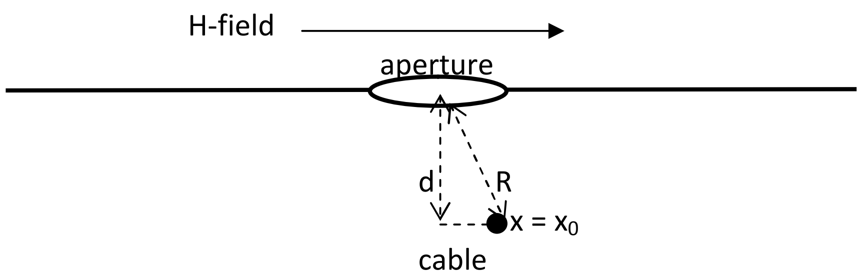


Figure 18. Aperture nearby a Cable Grounded back to Plane of Aperture[6,23]

We assume that the lightning current on the skin is parallel to the cable inside the aperture so that the magnetic field is optimally coupled to the cable; with that, we dismiss the vector algebra in the following models. The prime symbols on R', L', and C' denote units per unit length. The transmission line equations for a point inductive source, above, are as follows:[6,23]

(21) $\frac{dV}{dx} - (R' + i\omega \cdot L') \cdot I = i\omega \cdot \mu_0 \cdot \frac{d}{\pi \cdot R^2} \cdot m \cdot \delta(x - x_0)$;

(22) $\frac{dI}{dx} - (G' + i\omega \cdot C') \cdot V = 0.$

The dipole moment, m, is defined as follows minus the directional vector relationships:[6]

(23) $m = -2 \cdot \alpha_m \cdot H$, the magnetic dipole due to the exterior magnetic field and

This model tacitly assumes that the cable is grounded back to the plane of the aperture, a real but rare occurrence, a typical academic oversight. Normally, the cables are mounted over a groundplane running parallel to the cables as depicted in Figure 19. Furthermore, the voltage source term does not apply to a two wire transmission line (cable plus its image in the groundplane) as intended, just one wire behind the aperture.[5,24] The two wire line like one cable over a groundplane (the usual system configuration) has the flux linkage reduced by the cable height, h, above the groundplane.

The voltage difference between two wires or the cable and its image in the groundplane is proportional to the following taken from Figure 19, below, where h is the height above the groundplane.

(24) $\frac{1}{R^2} - \frac{1}{(R+h)^2} \cong \frac{1}{R^2} \cdot \frac{2 \cdot h}{R} = \frac{2 \cdot h}{R^3}$.

The inductive Waveform 2 (WF2) voltage term from a Component A current, I_A, on the exterior of radius, R_{syst}, is then approximated as follows where we have let $d = R$:

(25) $\boldsymbol{V_{WF2}(x) = -i\omega \cdot \mu_0 \cdot \frac{1}{\pi \cdot R} \cdot \frac{2 \cdot h}{R} \cdot 2 \cdot \alpha_m \cdot \frac{I_A}{2 \cdot \pi \cdot R_{syst}} \cdot \delta(x - x_0) \cong i\omega \cdot M_{12} \cdot I_A}$, where

the factor, $2 \cdot h/R$, reduces the flux linkage down to what's going through the cable-groundplane loop area, not between the cable and the plane of the aperture. We are going to let $d = R$ for simplicity and because we are not working an external problem with antenna patterns.

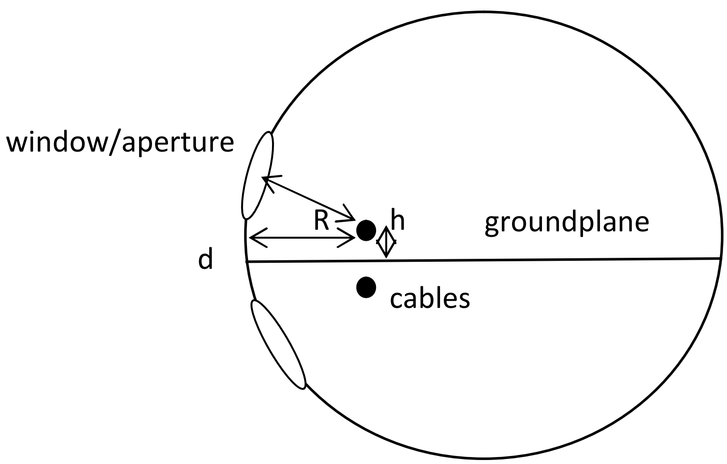
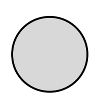
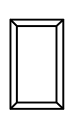


Figure 19. Cables Routed Along Groundplane Exposed to Apertures

The polarizability is explained in the references. $\boldsymbol{\alpha_m}$ is the magnetic polarizability of an aperture (units of volume):[6]

(26) $\boldsymbol{\alpha_{window} \cong \frac{4 \cdot r^3}{3}}$, the polarizability of a round window of radius, r, 

(27) $\boldsymbol{\alpha_{door} \cong \frac{\pi}{12} \cdot \frac{h^3 \cdot (1+3 \cdot w/h)}{\ln[4 \cdot (h+w)/g]}}$, the polarizability of a door of height, h, width, w, with a nonconducting gap, g, between the door and the airframe. 

Door hinges and fasteners reduce this latter polarizability however they are not controlled low impedance bond joints and therefore shouldn't be used for protection credit.

This solution is somewhat 'forced' in that we're treating an interior problem like it is an exterior problem, in other words, as a cable on one side of a groundplane with an aperture in it with the incident field on the other side of the plane, i.e. with the radiation amplitude falling off as a function of distance from a dipole, $1/R$, $1/R^2$, and $1/R^3$. A quasi-static or modal expansion of the interior fields is required and those then coupled to the aperture fields that emanate from the exterior and coupled to cables inside. The dimensional analysis remains the same so we will proceed.

The solution to the transmission line problem with a point source at $x = x_0$ for $x_o \leq x \leq l$, to the right of the source and left of the left load at $x = l$:

(28) $\boldsymbol{V}_{>}(\boldsymbol{x_0} \leq \boldsymbol{x} \leq \boldsymbol{l}) = \frac{V_s}{2} \cdot \frac{(1-\Gamma_{so})}{1-\Gamma_1 \cdot \Gamma_2 \cdot e^{-2\gamma \cdot l}} \cdot \left\{e^{-\gamma \cdot (x-x_0)} + \boldsymbol{\Gamma}_{Lo} \cdot e^{\gamma \cdot (x-x_0)}\right\}$,

where

(29) $\boldsymbol{V}_s \cong \boldsymbol{M}_{aperture} \cdot \frac{I_{lightning}}{\tau_{rise\ time}}$,

(30) $\boldsymbol{\Gamma}_1 = \frac{Z_1 - Z_0}{Z_1 + Z_0}$ is the voltage reflection coefficient at the left load, Z_1, $x = 0$, where

(31) $\boldsymbol{Z}_0 = \sqrt{\boldsymbol{Z}'/\boldsymbol{Y}'}$, the transmission line characteristic impedance,

(32) $\boldsymbol{\Gamma}_2 = \frac{Z_2 - Z_0}{Z_2 + Z_0}$ is the voltage reflection coefficient at the right load Z_2, $x = l$,

(33) $\boldsymbol{\Gamma}_{so} = \boldsymbol{\Gamma}_1 \cdot e^{-2 \cdot \gamma \cdot x_0}$ is the phase shifted voltage reflection coefficient at $x = x_0$ looking towards the left load, Z_1, where

(34) $\boldsymbol{\gamma} = \sqrt{\boldsymbol{Z}' \cdot \boldsymbol{Y}'}$, the propagation factor of the transmission line, where

(35) $\mathcal{R}e(\gamma) < 0$, establishes the stability condition,

(36) $\boldsymbol{\Gamma}_{Lo} = \boldsymbol{\Gamma}_2 \cdot e^{-2 \cdot \gamma \cdot (l-x_0)}$ is the phase shifted voltage reflection coefficient at $x = x_0$ looking towards the right load, Z_2.

Hyperbolic functions should be avoided because they allow the singularities and poles to slip to the right of the imaginary axis although most published work and canned software don't recognize this.

Note that the resonant denominator is common to all transmission line solutions whether from external or internal discrete sources, as above, external distributed EM field sources, or internally generated switching transients. Plus, solutions exist at anti-resonant frequencies that have null responses at every other harmonic frequency, the voltage peaking at one set of harmonics and the current peaking at the other harmonics. See Figures 20 and 21, below, to see this behavior on both a shielded grounded cable and on a cable terminated in virtual open circuits, both driven by a point voltage source located midway between the loads. In both plots, the current has been multiplied by an arbitrary factor in order to fit all curves in a reasonable space. Also plotted is the transmission line resonant denominator peaking at all harmonics as a double check on the others.

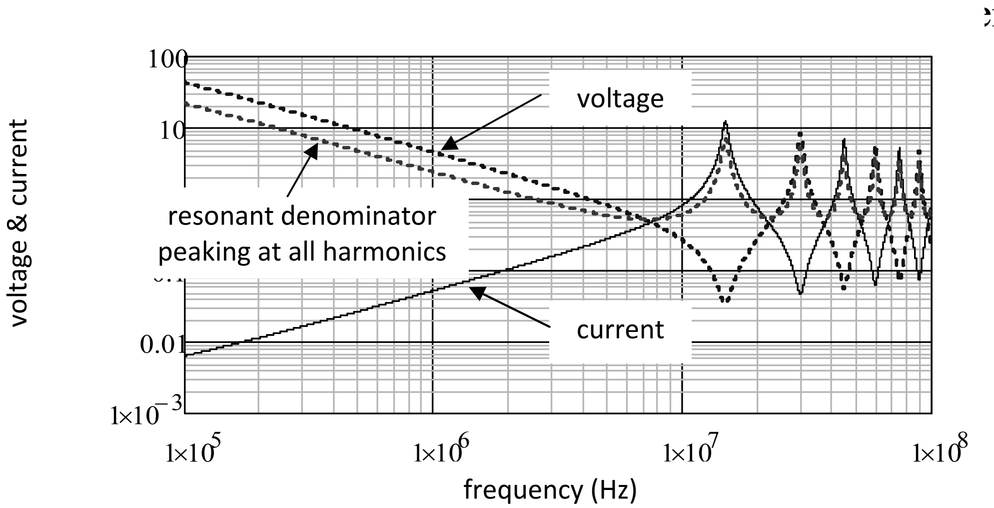


Figure 20. Voltage and Current Induced on a 10m Grounded Shielded Cable
First Current Peak & First Voltage Anti-Resonance at f = c/(2·l)
First Voltage Peak & First Current Anti-Resonance at f = c/l

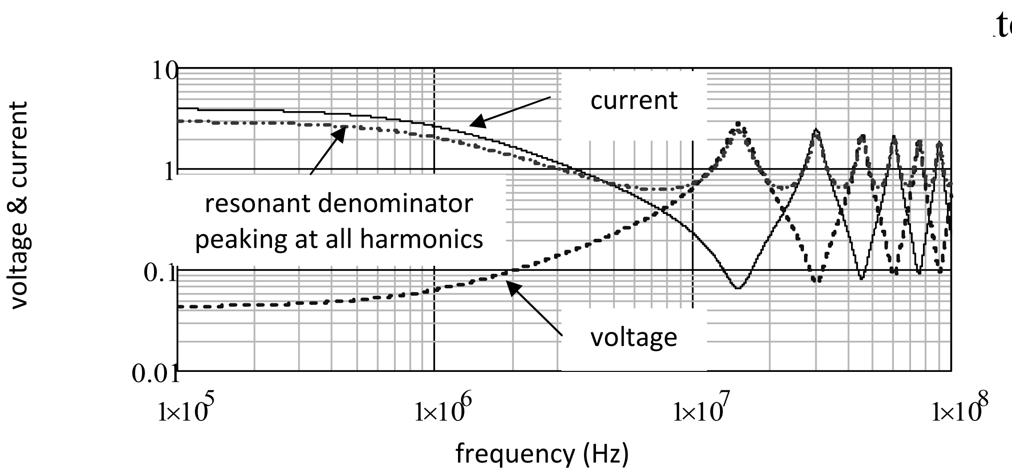


Figure 21. Voltage and Current Spectra on a 10m Unshielded Line with Large Resistive Loads
First Voltage Peak & First Current Anti-Resonance at f = c/2·l
First Current Peak & First Voltage Anti-Resonance at f = c/l

Which harmonic resonant frequencies are excited depends upon where along the line the point source is located like plucking a string; that is, the different harmonics peak at different locations.

We have spent an undo amount of space for this problem because (1) the references treat unrealistic configurations, (2) we want to properly estimate the reduction of flux linkage through the space between the cable and its groundplane, and (3) we want to illustrate the resonant and anti-resonant behavior of transmission lines excited by point sources. The reduction in flux linkage has been modeled for crosstalk between two cables above a groundplane, below, versus cables dangling in free space but not for aperture coupling.[6]

1.7. Cross Coupling between Exposed Cables and Less Exposed Cables. As we mentioned above, the crosstalk problem for two cables over a groundplane has been published correctly.[6] With over 20dB more high frequency WF4 current derivative flowing on internal cables due to the in-flight IR voltage drop, we have to see how much of this couples to adjacent cables. With the slow rising WF5A from ground-test, we didn't worry. Three expressions for the crosstalk follow, one including the cable radii for close distances, one without cable radii for cables further apart, and one for the cables very far apart:

(37) $\boldsymbol{M_{12}} = \frac{\mu_0 \cdot d}{2 \cdot \pi} \cdot \ln\left[\frac{d^2+(2 \cdot h-r)^2}{d^2+r^2}\right]$, height, h, distance apart, d, with the cable radii, r, included and

(38) $\boldsymbol{M_{12}} \cong \frac{\mu_0 \cdot d}{2 \cdot \pi} \cdot \ln\left[1+\left(\frac{2 \cdot h}{d}\right)^2\right]$ with the cable radii, r, set to zero.

(49 $\boldsymbol{M_{12}} \cong \frac{\mu_0 \cdot d}{4 \cdot \pi} \cdot \ln\left[1+\left(\frac{2 \cdot h}{d}\right)^2\right] \cong \frac{\mu_0 \cdot d}{4 \cdot \pi} \cdot \left(\frac{2 \cdot h}{d}\right)^2$, when $\boldsymbol{r << h << d}$

<u>This crosstalk determines "levels of protection" within the system</u> that is controlled by cable spacing assuming they are fastened short distances from structure at regular intervals. For example, a 2kA WF4 current running next to another cable for a distance of 2m can induce as much as 500V WF2 "crosstalk".

1.8. Inverse Laplace Techniques for Recovering Time Domain Results. The models herein start with simple low frequency domain formulas that can be transformed back to the time domain using inverse Laplace transform tables[10] and software such as Mathcad for numerics.

(40) $\mathcal{L}^{-1}\left(\frac{1}{s+\alpha}\right) = e^{-\alpha \cdot t}$,

(41) $\mathcal{L}^{-1}\left[\frac{b}{(s+a)^2+b^2}\right] = e^{-a \cdot t} \cdot \sin(b \cdot t)$.

As an example, the induced current in a grounded cable routed through a CFC (carbon fiber composite) airframe carrying a Component A lightning current (Figure 7) is expressed in the frequency domain with the Laplace "s" as follows:[13,14]

(42) $\boldsymbol{I_{cable}(s)} = \boldsymbol{I_A(s)} \cdot \frac{Z_{CFC}(s)}{Z_{CFC}(s)+Z_{cable}(s)} = \boldsymbol{I_A} \cdot \left(\frac{1}{s+\alpha} - \frac{1}{s+\beta}\right) \cdot \frac{L_{CFC}}{L_{CFC}+L_{cable}} \cdot \frac{s+b}{s+\gamma}$,[13,14] where

$I_A(s)$ is the Component A lightning spectrum, Z_{CFC} is the $R_{CFC} + s \cdot L_{CFC}$ impedance of the skin, and Z_{cable} is the $R_{cable} + s \cdot L_{cable}$ impedance of the cable(s).

The time domain cable current (42) is the inverse Laplace transform of the above expression:

(43) $\boldsymbol{I(t)} = \mathcal{L}^{-1}[\boldsymbol{I_{cable}(s)}] = \boldsymbol{I_A} \cdot \frac{L_{CFC}}{L_{CFC}+L_{cable}} \cdot \left[\frac{(b-\alpha) \cdot e^{-\alpha \cdot t}-(b-\gamma) \cdot e^{-\gamma \cdot t}}{\gamma-\alpha} - \frac{(b-\beta) \cdot e^{-\beta \cdot t}-(b-\gamma) \cdot e^{-\gamma \cdot t}}{\gamma-\beta}\right]$,

from a Table of Laplace Transforms.[10]

The current is the sum of four different time domain exponentials each with a different amplitude determined by the different lightning and system parameters, above. The system inverse time constant, $\boldsymbol{\gamma^{1}}$, competes with the lightning rise and fall times for the definition of the induced waveform, a phenomenon unique to the parallel RL branches driven by a current source.[13,14]

1.9. Cable Shielding. The low frequency transfer impedance of a coaxial shield of length, l, radius, r, and thickness, t, is approximated as follows:[6,7]

(44) $\boldsymbol{Z_T(s) = R_{dc} \cdot \frac{\sqrt{s \cdot \tau_d}}{\sinh \sqrt{s \cdot \tau_d}}}$. where R_{dc} is the DC resistance (5) $\boldsymbol{R_{dc} = l/2 \cdot \pi \cdot r \cdot t \cdot \sigma}$, above, and

(45) $\boldsymbol{\tau_d = \mu \cdot \sigma \cdot t^2}$ is the diffusion time through the shield wall thickness, t.

We will simplify this to the following so that we can use the Laplace transform tables to obtain an answer, capturing the low frequency effect of the diffusion time on the rise and fall times of the shielded transient but compromising the accuracy of the higher frequency portion which we are not concerned with anyway at this level of application:

(46) $\boldsymbol{Z_T(s) \cong R_{dc} \cdot \frac{1}{1+s \cdot \tau_d}}$.[6]

Higher frequency cable shielding includes the transfer inductance, L_T, magnetic leakage through the many holes in a braid shield quantified by the optical coverage. The transfer impedance is then defined as follows and an example plot in Figure 22:

(47) $\boldsymbol{Z_T(s) = \left[R'_{dc} \cdot \frac{\sqrt{s \cdot \tau_d}}{\sinh \sqrt{s \cdot \tau_d}} + s \cdot L'_T\right] \cdot l \cdot \sin\left(\omega \cdot \frac{v_e - v_i}{v_e \cdot v_i} \cdot \frac{l}{2}\right) / \omega \cdot \frac{v_e - v_i}{v_e \cdot v_i} \cdot \frac{l}{2}}$.

The last sinx/x term is due to the difference in the velocity of propagation between the external, v_e, and internal, v_i, waves where the velocity is inversely proportional to the square root of the dielectric constants of the two different insulation materials, $v \approx 1/\sqrt{\varepsilon_r}$. It's a high frequency effect that seldom concerns lightning shielding. Wire manufacturers use their good insulation around the wires and the cheap stuff for the jacket hence the difference.

The transfer inductance, L'_T, can vary over a range of $1pH/m$ to $1nH/m$. It is controlled for the most part by specifying the optical coverage of the braid shields, nominally 85% to 95%. Any more and the braid gets too stiff.

Table 1 summarizes cable shield parameters in several ways for the diverse set of intended readers.

Table 1. Cable Shield Parameters

shield type	braid size & type	shield thickness t(m)	cable diam. (inches)	diffusion time τ_d	DC resistance R_{dc} (Ω/m)	f & t = 1/2·π·f where R_{dc}=R_{ac}
Aracon[62,63,64]	Ni, Cu, Ag plated Kevlar[63] braid	$1.27 \cdot 10^{-6}$	¼”	35ns	223mΩ/m	100MHz 1.6ns
	Ni, Cu, Ag plated Kevlar[53] braid	$1.27 \cdot 10^{-6}$	1”	35ns	56mΩ/m	100MHz 1.6ns
Copper flat foil braid	1.5 mil flat Cu foil braid	$3.8 \cdot 10^{-5}$	¼”	105ns	36mΩ/m	3MHz 53ns
Copper wire braid	40AWG Cu wire braid	$8 \cdot 10^{-5}$	¼”	467ns	17mΩ/m	682kHz 233ns
	38AWG Cu wire braid	10^{-4}	1”	729ns	2.2mΩ/m	437kHz 364ns
	36AWG Cu wire braid	$1.27 \cdot 10^{-4}$	1”	1.18µs	1.7mΩ/m	318kHz 500ns
	34AWG Cu wire braid	$1.6 \cdot 10^{-4}$	1”	1.87µs	1.4mΩ/m	170kHz 935ns
	30AWG Cu wire braid	$2.54 \cdot 10^{-4}$	1”	4.71µs	882µΩ/m	67kHz 2.4µs
	20AWG Cu wire braid	$8.1 \cdot 10^{-4}$	1”	47.8µs	267µΩ/m	6.6kHz 24µs
Copper wire braid over mumetal foil	2 mil µ-foil 38AWG Cu OVB	$7.6 \cdot 10^{-5}$ mumetal foil	¼”	421ms mumetal foil	13mΩ/m Cu braid	1.9Hz 437kHz

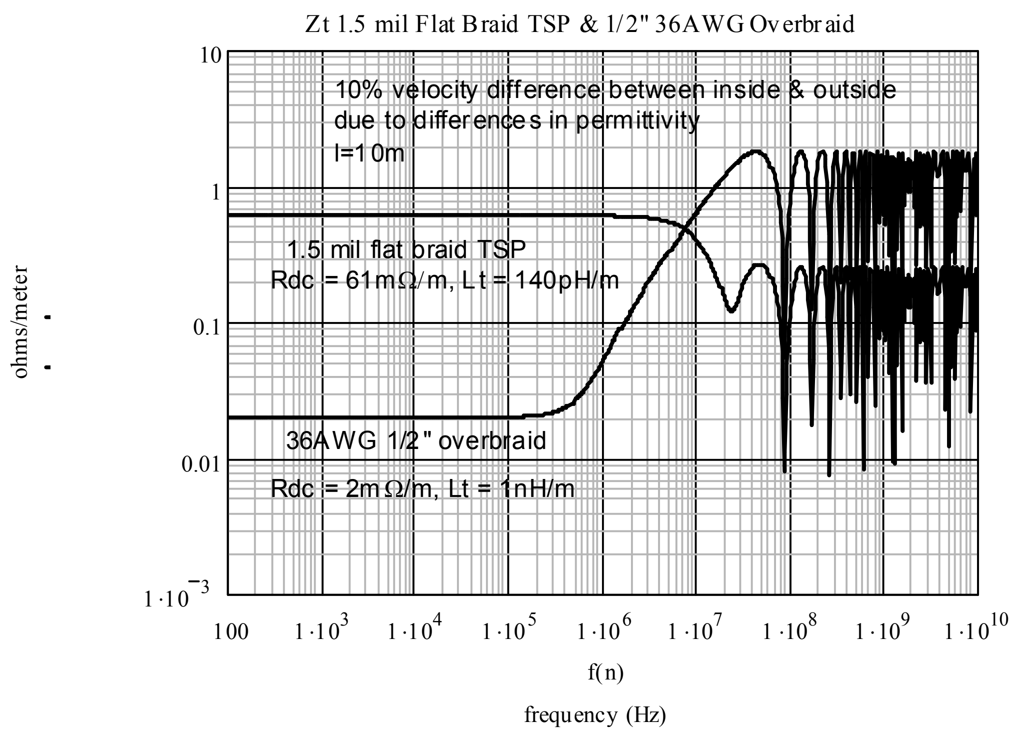


Figure 22. Transfer Impedance of 10m Long Cables with Velocity Interference Effect

The transfer inductance is important mostly for higher frequency lightning WF3 damped sinusoids. The high frequency velocity interference effect above 10MHz is an EMI issue. For lightning, the DC resistance, R_{dc}, and diffusion time, τ_d, are the most important parameters which in turn are functions of cable length, cable radius, shield thickness, shield conductivity, and shield magnetic permeability.

1.10. Scaling by Geometric, Electrical, and Spectral Parameters. All of the coupling parameters and electrical system elements are functions of their geometries and distance from the sources of coupling to the induced voltages and currents on the cables.

Examples of scaling:

(a) Aperture mutual coupling is dimensionally proportional to the aperture radius, r cubed, the cable height above a groundplane, h, and the square of the distance, R, of a cable away from the aperture, and the system radius, R_{syst}, at that location i.e. $M_{ap} \approx r^3 \cdot h/(R^2 \cdot R_{syst})$.

(b) Structural and cable resonant frequencies, f, are inversely proportional to the length, l, of both, generally as half wavelengths, i.e. $f \approx c/2 \cdot l$.

(c) Structural resonances, $f = \omega/2\cdot\pi$, occur between 1MHz and 20MHz, well above the second break frequency of the lightning spectrum at $\beta_A/2\pi = 103kHz$. When the lightning spectrum drives an RLC resonant circuit, the time domain amplitude is proportional to β/ω_3. That is, the higher the resonant frequency, ω_3, because of the smaller lengths, the lower the amplitude of the induced resonant transient, i.e. the peak WF3 current is proportional to the peak Component A current as follows:

(48) $I_{WF3} \cong \frac{M_{12}}{L_{cable}} \cdot \frac{\beta_A}{\omega_3} \cdot I_A$.

This cable inductance is the external inductance because the cable is part of a closed loop circuit of differential, $I_{DM} = I_1 - I_2$, currents, with length, l, height, h, and radius, r.

(49) $L_{cable} \cong \frac{\mu_0 \cdot l}{2\cdot\pi} \cdot \ln\left(\frac{2\cdot h}{r}\right)$.

(d) I·R-drop voltages, $V_{WF4} = I_L \cdot \frac{R_p \cdot R_c}{n \cdot R_p + R_c}$, are functions of the following:

(i) distance between interconnected boxes,
(ii) aircraft skin resistance, (scales as length/radius x thickness x conductivity)
(iii) cable resistance (scales as length/ radius x thickness x conductivity),
(iv) groundplane resistance, (scales as length/ width x thickness x conductivity) and
(v) number of cables and other conductors parallel to the common mode lightning current through the system.

The associated cable currents, $I_{WF4} = I_L \cdot \frac{R_p}{n \cdot R_p + R_c}$, are functions of the same except distance between boxes. The induced voltage is proportional to length of shielded cables if the cables do not run in a straight line between boxes. This is tacitly included in cable resistance, above, however.

The combined voltage and current, power = V·I, mostly affects power and peak power rating of surge arrestors and voltage clamping diodes because when they trigger/clamp, they draw high currents even inside cable shields. Diode power ratings vary from 100W to 100kW and higher for special devices.

(e) Raceway cables and cables along wing spars depicted in Figures 23 and 24 are exposed to the magnetic field at the surface of those bodies. The magnetically induced voltage is proportional to the cable loop area, $A = l \cdot h$, the cable length times the height above structure and inversely proportional to the circumference/radius, R_{syst}, of the main body. The induced current is independent of the length, l, of the cable but is proportional to the height, h, above the main body. The WF2 induced voltage and WF1 induced current are as follows:

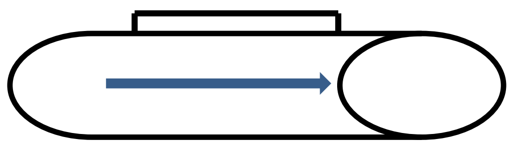


Figure 23. Raceway Cable

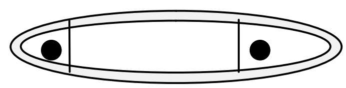


Figure 24. Cables along Wing Spars

(50) $\boldsymbol{V_{oc}} \cong \boldsymbol{\mu_0 \cdot l \cdot h} \cdot \frac{I_A}{2 \cdot \pi \cdot R_{syst}} \cdot \frac{1}{\tau_r}$.

(51) $\boldsymbol{I_{sc}} \cong \frac{M_{12}}{L_{cable}} \cdot \boldsymbol{I_A}$.

where

(52) $\boldsymbol{M_{12}} \cong \frac{\mu_0 \cdot l \cdot h}{4 \cdot \pi^2 \cdot R_{syst}}$.

(53) $\boldsymbol{L_{cable}} \cong \frac{\mu_0 \cdot l}{2 \cdot \pi} \cdot \ln\left(\frac{2 \cdot h}{r}\right)$.

(54) $\boldsymbol{\tau_r}$ = Component A, D, or H rise time and

(55) $\boldsymbol{r}$ = cable radius.

For a wing, Figure 24, elliptical coordinates provide an estimate of circumference,

(56) $\boldsymbol{C \approx \pi \cdot (a + b)}$, where a is the wing chord and b is the camber.

2. SAE ARP5412A Indirect Lightning Environments & RTCA/DO-160F Test Environments

2.1. Lightning Standard Environments. Table 2 is the definition of the indirect lightning waveforms, graphed in Figure 25. Table 3 is the definitions of the thresholds and environment levels based upon the source of their data. Figure 26 is a definition of safety margins for certification purposes. Tables 4-8 of indirect lightning to cables and pins for damage by amplitude levels and waveforms from SAE ARP5412A[2] and RTCA/DO-160F[1] Section 22 caused by an 18µs/88µs double exponential Component A 200kA negative lightning current[2] on the airframe and allocations for functional upset from the multiple strike 100kA Component D and 10kA Component H. This note does not address the effects of different lightning strike zones on indirect lightning preferring to assume all parts of the airframe are in Zone 3 Current Conduction Zone conducting 200kA Component A.[35] The exceptions are components on the skin exposed to any one or all of the different direct multiple and multi-burst strikes or swept strokes in the different zones. The entries in Tables 3 through 7 are interpreted to be the box design and test levels although other waveforms and levels are allowed. One purpose of this note is to get rid of these misleading "one-size-fits-all" tables and replace them with simple model formulas for scaling the levels and waveforms by each system's size, material composition, and cable geometries. Multiple Stroke Component D and Multiple Burst Component H levels are allegedly only for functional upset however their high frequency content is the same as Component A. The average engineer believes that these tables apply to all systems and will erroneously choose levels from them without guidance to the contrary. Companies with electromagnetic and lightning expertise develop their own tables tailored to their specific aircraft.[29,38] This note intends for everyone to do so with the tools provided herein.

Before we tabulate the indirect lightning environments in the standards, let's introduce the standardized external strike components and the resulting internal indirect lightning.

2.1.1. Direct Strike Components used to Define Indirect Transients. The external direct lightning strikes themselves are given alphabetic designations, WFA-WFH although indirect lightning is derived only from Components A, D, and H, below.[2]

(57) $\boldsymbol{I_A(t) = 218.8kA \cdot \left(e^{-t/88\mu s} - e^{-t/1.5\mu s}\right)}$, the single stroke 200kA Component A;

(58) $\boldsymbol{I_D(t) = 109.4kA \cdot \left(e^{-t/44\mu s} - e^{-t/773ns}\right)}$, Multiple Stroke 100kA Component D followed by 13 more, 50kA D/2, at half this amplitude, 10ms to 200ms apart, all within 1.5 seconds;

(59) $\boldsymbol{I_H(t) = 10.6kA \cdot \left(e^{-t/53\mu s} - e^{-t/52ns}\right)}$, Multi-Burst 10kA Component H, three bursts of twenty transients each, 50µs to 100µs between transients, 30ms to 300ms between bursts, all within 620ms.

Components D and H have been relegated to functional upset threats while Component A has been relegated to a damage threat. <u>These three exterior strike Components A, D, and H have about the same dI/dt and the same spectral content, $\tilde{I}(f)$, above 1MHz meaning that they will excite the same WF2 and WF3 indirect transients</u>. That does not relegate them to only "functional upset" as in the standards, especially when their levels in Tables 5, 6, and 7 meet or exceed the "damage levels" in Tables 3 and 4.

2.1.2. Indirect Lightning Waveforms' Definitions. Let's define the indirect waveforms and the rationale for their designations, remembering that these all have been derived from present-day system level ground tests.[13]

Waveform 1 (WF1) is induced through aperture or inductive coupling and is the same as the lightning Component A waveform; this characteristically appears as a current on a shielded cables;

(60) $\boldsymbol{I_{sc}(WF1) \cong B(Component\ A) \cdot A/L_{cable}}$

Waveform 2 (WF2) is the time derivative of lightning Component A or WF4; it appears as a voltage induced through inductive coupling from the lightning current Component A on the skin;

(61) $\boldsymbol{V_{oc}(WF2) \approx L_{mutual} \cdot dI(Component\ A)/dt}$

There also exist undefined versions of this for Components D and H although the peak WF2 voltage is the same for all three, A, D, and H.

Waveform 3 (WF3) is a resonant damped sinusoid whose amplitude and frequency depend upon the length of the resonating element, either (a) the airframe as a whole or (b) individual cables;

(62) $\boldsymbol{I_{cable}(WF3) \cong \mathcal{L}^{-1}(V(WF2\ and/or\ WF4)/[R + s \cdot L + 1/s \cdot C])}$.

Waveform 4 (WF4) is induced through the resistive voltage drop through a CFC (carbon fiber composite) skin. V = I·R, and is the same as that of the lightning current Component A (and Waveform 1);

(63) $\boldsymbol{V(WF4) \cong I(Component\ A) \cdot R}$

Waveform 5A (WF5A) is the current induced on cable shields grounded at both ends in a CFC airframe during ground-test with a nearby return current, shown below as the inverse Laplace transform of the frequency domain model; in-flight, with no return current, this cable current is WF4;[13]

(64) $\boldsymbol{I_{WF5A}(t) \cong \mathcal{L}^{-1}\left\{I_{Component\ A}(s) \cdot \frac{Z_{skin}(s)}{Z_{skin}(s)+Z_{cable}(s)}\right\}}$

Waveform 5B (WF5B)is the voltage induced on cables diffused through the transfer impedance of an aluminum skinned airframe; it is small, long, and usually ignored;

(65) $\boldsymbol{V_{WF5B} \cong \mathcal{L}^{-1}\{I_{Component\ A}(s) \cdot Z_{T,aluminum}(s)\}}$

Waveform 4 is erroneously attributed also to lightning Component D (See Table 5, above.) although there are circumstances where that confusion is not surprising. (Discussed in Section 10.)

Waveform 6_H ($WF6_H$) is the current induced on cables from the multi-burst lightning Component H;

(66) $\boldsymbol{I_{WF6H} = peak \cdot 1.057 \cdot \left(e^{-t/5.3\mu s} - e^{-t/5.2ns}\right)}$

How $WF6_H$ got onto cables in SAE ARP5412A for Component H and a corresponding $WF6_D$ didn't make it for Component D is one of the inconsistencies in the standards; it's because WF4 is used for the Component D strikes. Components D and H will not usually strike the same location on the aircraft unless it's moving very slowly therefore some of those strikes will diffuse through the CFC skin with a corresponding increase in rise time instead of striking on a spot connected to the interior cables and/or groundplane.

Table 2. Standardized Indirect Lightning Waveforms[1,2]
Extracted from Table 9 in SAE ARP5412A

• Component A and Waveforms 1 and 4 are the same 1.5µs/88µs double exponential with a x1.094 multiplier, i.e. $I_A(t) = 200kA \cdot 1.094 \cdot \left(e^{-t/88\mu s} - e^{-t/1.5\mu s}\right)$ peaks at 200kA.
• Waveform 2 is the derivative of Waveform A, 1, and 4 with a x1.00 multiplier. (See note below.)
• Waveform 3 is a damped sinusoid waveform at 1 and 10MHZ with a damping Q-value of 9-37 with a x1.059 multiplier.
• Waveform 5 consists of two waveforms:
Waveform 5A is a 23µs/79µs double exponential with a x2.334 multiplier.
Waveform 5B is a 12.5µs/631µs double exponential with a x1.104 multiplier.
• Waveform 6_D doesn't exist having been replaced with WF4.
• Waveform 6_H is a 5.2ns/5.3µs double exponential associated with Component H.

Note: WF2 doesn't graph like the derivative definition, above, but does graph nicely as the following:

(67) $\boldsymbol{WF2(t) \cong -A \cdot e^{-t/100ns} + B \cdot e^{-t/1.5\mu s} - C \cdot e^{-t/88\mu s}}$.

2.1.3. Tables of Allocated Indirect Lightning Waveforms and Levels.

The indirect lightning environments that are provided by SAE ARP5412A and RTCA/DO-160F are tabulated below in Tables 3-7 as they appear in those standards. These are intended to be both design and test levels. Other waveforms derived by the various programs are designated Waveform Z. Rationale for the different levels follows the tables. Definitions of the transient level acronyms follow next along with their intended use in estimating safety margins of the equipment in the system.

Table 2, below, appears to be an over-laborious set of acronyms simply to ensure that induced transients do not exceed box failure thresholds.[1,2,5] However, ATL has been broken into several categories depending upon the origin of the data, in agreement with FAA AC 20-136A;[5] for example (a) ATL_t, the intended extrapolated system test data and box test data, (b) ATL_a, system modeling and analyses, and (c) ATL_s, similarity through scaling by geometry and electrical parameters.

■

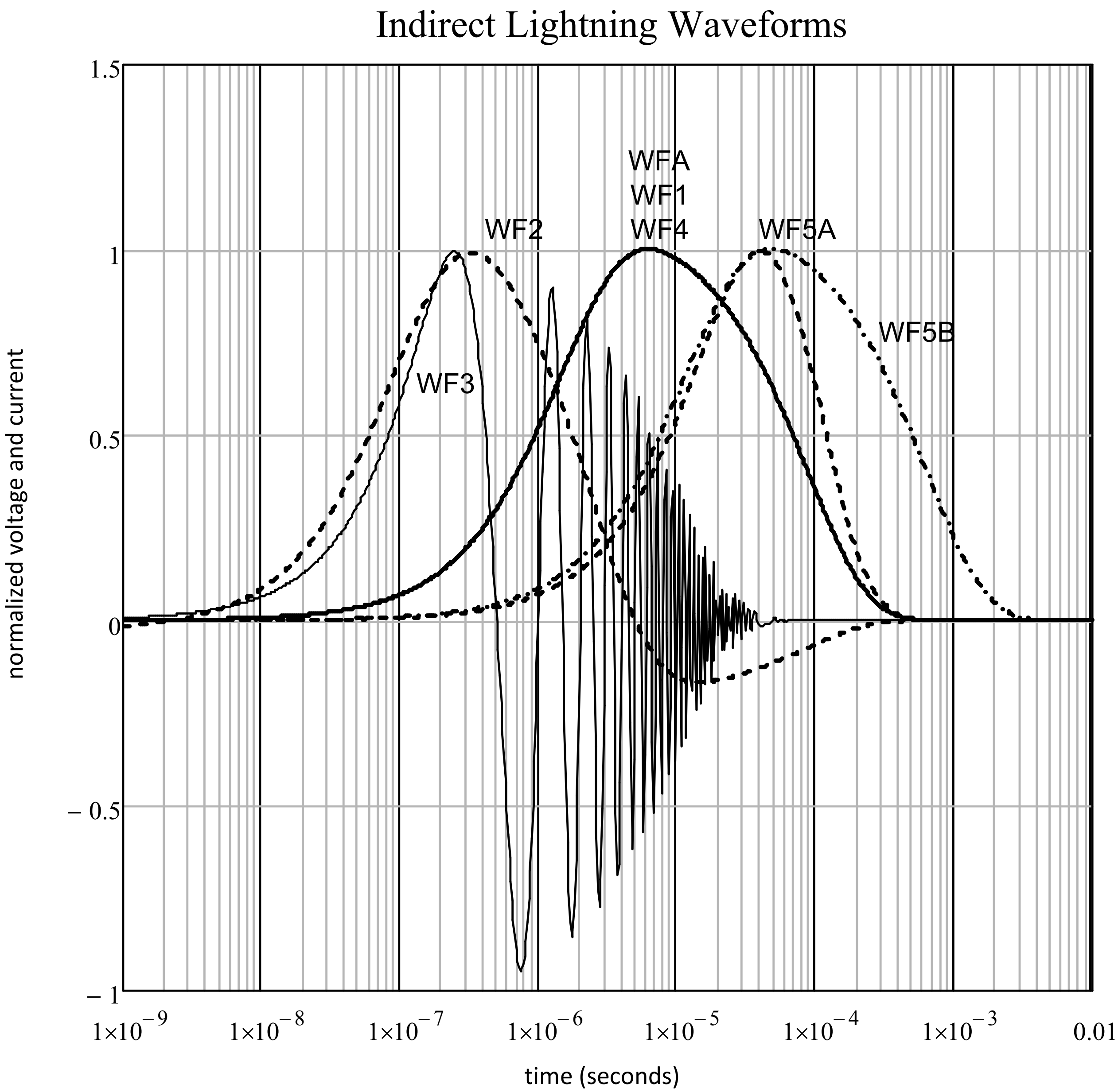


Figure 25. Indirect Lightning Waveforms 1, 2, 3 (1MHz), 4, 5A, & 5B

The sad part of analysis and similarity is that, in the hands of DERs (in-house engineers with special training and FAA certification to be FAA Designated Engineering Representative) wearing the wrong hat, they are used to justify waivers on equipment failing tests. Also, as has been pointed out in IN615[13] and In618[15], tests are too often ill-defined and, in fact, need to be redesigned. A new system level test procedure is also needed.[61] Another factor is that these levels need to change as the system design matures particularly when the cable routing is completed and the circuit designers have identified their I/O circuit damage and upset susceptibility levels. The latter is difficult because designers do not think in terms of “common mode”.

Table 3. DO-160F Equipment Environment, Design, & Susceptibility Levels[1,5]	
Levels	**Definitions**
TCL	Test Control Level
ETDL	Equipment Transient Design Level
ETSL	Equipment Transient Susceptibility Level
ATL_t	Actual Transient Level – test*
ATL_a	Actual Transient Level – analyses*
ATL_s	Actual Transient Level – scaling/similarity*

Author's Note: * Denotes data origins for ATL.

FAA Advisory Circular 20-136A[5] defines "margin (dB) = ETDL - ATL" in Figure 26, below. The author prefers margin = ETSL - ATL, however it's not worth arguing about because ETSL is difficult to obtain particularly for common mode unless diode voltage clamps define it by default.

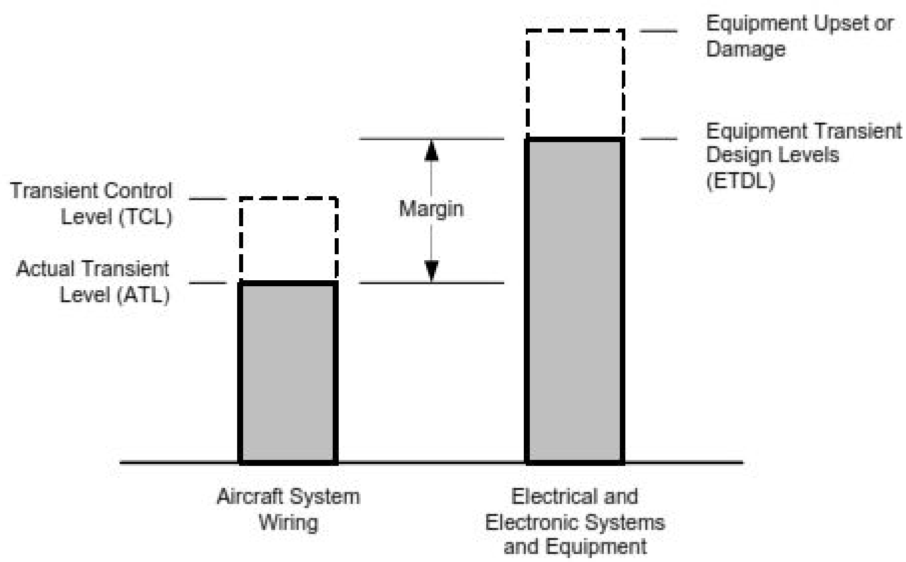


Figure 26. FAA Aircraft Circular 20-136A Definition of Margin between ETDL and ATL[5]

Table 4. SAE ARP5412A Allocation of Single Stroke Lightning Test Levels to Cables Bundles[2]
Derived from the 200kA Component A
(RTCA/DO-160F Table 22-3; * denotes WF5A not WF5.)[1]

Cable Bundle TCL, ETDL or Test Levels due to Component A					
Level	**Waveform 1**	**Waveform 2**	**Waveform 3**	**Waveform 4**	**Waveform 5***
1	50V/100A	50V/100A	100V/20A	50V/100A	50V/150A
2	125V/250A	125V/250A	250V/50A	125V/250A	125V/400A
3	300V/600A	300V/600A	600V/120A	300V/600A	300V/1000A
4	750V/1500A	750V/1500A	1500V/300A	750V/1500A	750V/2000A
5	1600V/3200A	1600V/3200A	3200V/640A	1600V/3200A	1600V/5000A

Table 5. SAE ARP5412A Allocation of Lightning Test Levels to Pins/Wires[2]
Derived from the 200kA Component A
(RTCA/DO-160F Table 22-2; * denotes WF5A not WF5.))[1]

Individual Conductor TCL, ETDL or Test levels due to Component A			
Level	**Waveform 3**	**Waveform 4**	**Waveform 5***
1	100V/4A	50V/10A	50V/50A
2	250V/10A	125V/25A	125V/125A
3	600V/24A	300V/60A	300V/300A
4	1500V/60A	750V/150A	750V/750A
5	3200V/128A	1600V/320A	1600V/1600A

Table 6. Test Levels & Waveforms for Cable Bundle Multiple Stroke Tests[2]
Derived from the 100kA & 50kA Waveform D
(RTCA/DO-160F Table 22-4)

Level		**Waveforms**				
		2/1	**2/1**	**3/3**	**4/1**	**4/5A**
		VT/IL	**VT/IL**	**VT/IL**	**VT/IL**	**VT/IL**
1	first stroke	50/50	50/50	100/20	25/50	20/60
	subsequent strokes	25/25	25/25	50/10	12.5/25	10/30
2	first stroke	125/125	125/125	250/50	62.5/125	50/160
	subsequent strokes	62.5/62.5	62.5/62.5	125/25	31.25/62.5	25/80
3	first stroke	300/300	300/300	600/120	150/300	120/400
	subsequent strokes	150/150	150/150	300/60	75/150	60/200
4	first stroke	750/750	750/750	1500/300	375/750	300/800
	subsequent strokes	375/375	375/375	750/150	187.5/375	150/400
5	first stroke	1600/1600	1600/1600	3200/640	800/1600	640/2000
	subsequent strokes	800/800	800/800	1600/320	400/800	320/1000

Table 7. SAE ARP5412A Table 4 for Component D Induced Indirect Environments[2]

Response to D and D/2 as a Fraction of A					
Transient	**Waveform 1**	**Waveform 2**	**Waveform 3**	**Waveform 4**	**Waveform 5**
Response to D	1/2	1	1	1/2	2/5
Response to D/2	1/4	1/2	1/2	1/4	1/5

Table 8. Test Levels & Waveforms for Cable Bundle Multiple Burst Tests[2]
Derived from the 10kA Component H
(RTCA/DO-160F Table 22-5 for WF3_H & SAE ARP5412A Table 7)

Levels	Waveform	
	3_H	6_H
	VT/IL	IL
1	60/1	5
2	150/2.5	12.5
3	360/6	30
4	900/15	75
5	1920/32	180

All of these detailed indirect lightning levels and waveforms make it seem as though they are the only parameters needed for certification. In fact, these voltage and current levels are directly dependent upon system parameters, most of which can be and will be 100% tested or inspected during manufacturing, that must stay within tolerances in order to keep these environments below these levels:

(1) groundplane resistance,
(2) bond joint resistance,
(3) cable shield resistance, and
(4) cable routing, separation, and lengths.

This last topic is be discussed in more detail in reference #61 in terms system certification of the development aircraft, box testing, and certification of each production vehicle.

The different transient levels or peak amplitudes in Tables 1 through 5 are summarized in RTCA/DO-160F[2] as follows in Table 9:

Table 9. SAE ARP & RTCA/DO160 Levels of Indirect Lightning[1,2]

Level 1: Equipment and interconnecting wiring in a well-protected environment.
Level 2: Equipment and interconnecting wiring in a partially protected environment.
Level 3: Equipment and interconnecting wiring in a moderately exposed environment.
Levels 4 & 5: Equipment and interconnecting wiring in severe electromagnetic environments.

2.2. Comments on the Standards. Regarding Tables 3 through 9, the following issues will be addressed herein:

The term “protected environment” is not defined but it is an opinion about relative levels of shielding of fields throughout the airframes.[20] Since most EEs believe that a tin can full of holes penetrated by ungrounded cables is a shielded enclosure, they will inevitably choose overoptimistically low levels. (See Section 5.) This note will redefine these levels and the notion of “protected environment”, develop algorithms for scaling induced environments to match the system with box and cable locations, add one or two more levels to capture the very high levels avoided in these tables, and provide design guidelines for achieving desired levels.

Except for a handful of military aircraft, there are no “shielded” areas in any airframe. There are, however, cables more exposed to intense environments than others that depend upon their routing and geometry. <u>Mutual coupling between the more exposed cables and the less exposed cables is then the real basis for “levels of protection” herein</u>. Shielded boxes and cables are the only reliable and maintainable shielding in most airframes. This note will discuss how to allocate cable shielding.

Levels 1-5 allegedly range from well protected to least protected, a total range of about x32 or 30dB. Note, however, the difference between each level is about 7-8dB. Measurement accuracy and repeatability are rarely that good.[21] Protection cannot be obtained in such small increments except when using two layers instead of one layer of cable shield, the “protection” improves by 6dB. These levels imply different degrees of shielding from the airframes where shielding is not designed, controlled, or maintained and will vary an order of magnitude within such airframes as well as between airframes. Assumptions about these levels, how to determine them[20,21], and what they mean have been a misconception across the whole spectrum of EMI and lightning engineering for decades, promoted by those least qualified to address the issue. “Protection” is achieved by (1) cable routing, (2) cable shielding, and (3) circuit protection. See section 5 for a more detailed diatribe.

WF5A cable current from ground-tests is hereby changed to WF4 for in-flight.[13]

The WF4 I·R voltage is proportional to the distance between boxes in the lightning current path. The corresponding WF4 cable current is relatively independent of the length between boxes.[13] Scaling the I·R-drop cable current the same as the I·R-drop voltage is wrong.

Voltages induced inside the cable shields with WF5A, now WF4, currents are proportional to cable lengths, not box distances. This may seem to be a trivial issue however it dictates the ratings of surge arrestors placed on the effected lines. See section 13.

The Table 3 Level 5 WF4 voltage and WF5A, now WF4, current are presumably inside a composite fuselage during a ground-test with a 15mΩ low inductance groundplane, i.e. wall-to-wall & nose-to-tail and yet it is still 3x too small for a 787[29,38]. The table is corrected herein in that WF4 voltage in an in-flight composite airframe is associated with WF4 current on the cables and the voltage is scaled by system and cable lengths.[13]

Allocating 100V to 3.2kV for functional upset in the standards is akin to allocating a steam roller to repel fleas on a dog. Protection against functional upset is more of a software and fail-safe architecture issue, anyway.

Missing are higher levels in locations without low resistance and low inductance groundplanes where voltages can reach 10-20kV for WF2 & 4. The NASA ORION CFC-skinned 100mΩ Launch Abort System (LAS) had WF4 voltage levels approaching 13kV due to a 200kA strike because of the smaller diameter and no groundplane.

Waveform 4 and 5A will not exist in-flight when struck by Waveform D lightning transients. Waveform D will appear as an I·R-voltage drop and as a subsequent cable current. Waveform 3 will appear as an external resonance or as cable resonances and be the same as Component A.

Table 6 properly connects the correct WF4 voltage and WF5A (now WF4) current waveforms not done in Tables 4 and 5. However, all of the tables on this IR-drop phenomenon miss the fact that as WF4 voltage decreases with decreasing distance between boxes along the lightning path, WF5A (now WF4) current remains the same. (This was observed in the Beech Starship during testing at LTI in 1984 and agrees with simple theory.) However, for shielded cables, the induced voltage inside the shields in-flight is proportional to the cable length between boxes and is close to a WF5A waveform (now WF4), the same as on the shield.[13,14] IN618[15] recommends ground injection testing with the "shield disconnect method" using a WF4 voltage because box testing by injecting with the shields in place will never result in the same waveform induced inside the shields as will be in-flight. (This is observed in Note 3 of Table 22-3 in DO-160F but promptly ignored.)

Tables 4-8 should have the same WF2 and WF3 peak voltages because the time derivatives of all three are the same (WF2 ≈ dI/dt) and the spectral content is the same above 2MHz (WF3 is proportional to the lightning components' spectra above 2MHz)

Tables 4-8 should have WF1 currents, WF4 I·R-drop voltages, and WF4 I·R-drop cable currents proportional to the peak values of their respective external strike Components A, D, and H with waveforms the same as the external components, say $WF6_D$ and $WF6_H$.

Missing is the fact that normal copper cable shields cannot attenuate the low frequency WF4 or WF5 currents.[13,14,15] Copper shielding must be at least as thick as a seldom used 30AWG wire braid or conduit or include a new copper and mumetal cable shield.[13,14] See Table 1. Even with extreme shielding, special circuit isolation and voltage limiting is required to mitigate the effects of the extreme I·R-drop environments. See Section 13. Even with extreme shielding and special I/O circuit designs, functional upset protection needs special error correcting software or circumvention techniques so as to prevent unsafe commands from automated flight electronics; to wit, the lightning environments are so high that they cannot be reduced to below all susceptibility threshold levels by shielding alone. <u>The predominate mode of indirect lightning the cables, wires, and boxes will experience will be the I·R-drop voltage in the inner most shields</u>. That is why designing proper ground injection tests is so important in IN618[15]. As an example, an ARINC 429 vendor, Holt[60], feels obliged to allocate p<u>in injection</u> to WF3, WF4, and WF5A, the easiest test to pass, even though in-flight all three will appear as I·R-drops in the innermost cable shield. Message? The tables allow vendors to choose the easiest environments and tests regardless of the physics.

Missing is any scaling of levels by system size, aperture size and distance to cables, and cable geometry. See all subsequent sections to see the geometric dependence of the allocated waveforms, frequencies, and levels/amplitudes.

Missing is scaling of resonant WF3 amplitudes & frequencies by system length and cable geometry.

Missing is any algorithm to establish any level in any system, the subject of this note. Most designers believe that Tables 2 through 6 are the extent of their choices. SAE ARP5415A does little to clarify the issues. The result is a tragicomedy.

This note addresses these issues by developing simple formulas for estimating lightning induced transients at the beginning of programs without knowing details about the system. Managers of programs that develop composite systems are extremely weight conscious, therefore the lightning engineer must obtain reasonable early results in order to be allocated the weight and space necessary for lightning protection. The method is not precise but does allow reasonable estimates for early design allocations. Adjustments can be made to the allocations as design details become available, e.g. box locations, cable routing, cable lengths, and circuit susceptibility levels.

Warning: The radial distribution of lightning around the skin is unknown except in the present low level system level lightning ground-test. There the system is made to be the inner conductor of a coaxial test fixture and have a relatively uniform current distribution. Assuming the current to be uniform around the skin is then forcing the allocations to be such as to pass the extrapolated low level test. Nonuniform distributions can be treated with more sophisticated models, albeit later in a program, and by adding margins to these allocations. Nonuniform distributions often result in physical damage or degradation such as holes in radomes, ailerons being damaged, antennas blown off, and even flight control cables welded to a rib, based upon past experience. An example of nonuniform currents is the phenomenon of swept stroke lightning that will move from one attachment point to another from the first attachment towards the tail as the vehicle moves through the lightning stroke that remains more or less stationary in the atmosphere depending upon the speed of the vehicle.[35]

In reality, at the lowest levels, the allocated voltages are dictated by I/O circuit susceptibilities, ETSL, which in turn define the allocated groundplane, cable and wire shielding, and circuit protection. In modern systems with sensitive low voltage electronics, only primary power is unshielded. Everything else has at least shielded twisted pair (TSP) level of shielding with 20-70mΩ/m low frequency transfer impedance. The shielding is therefore designed against the allocated shield current and the allowed induced voltage, $Z_T \cdot l \cong I_{shld}/V_{I/O}$ with cable length and connector shielding included.

3. Comparisons, WF4 & WF5A in SAE ARP5412A vs. Present In-Flight Results

Let's compare these numbers against those for cable bundles in SAE ARP5412 and DO-160 although no cable lengths or numbers are mentioned in those documents:

3.1. IR-Drop WF4 Voltage & WF4 Cable Current. The following Tables 10 and 11 summarize the above graphical results for n = 1, 10, and 100 10m long cables, Figures 11-14. Tables 19-23 in Section 12 include the length of cables, also. They show a comparison the maximum and minimum induced WF4

IR-drop voltage with DO-160F. The present results are about an order of magnitude higher on the high end.

Table 10. V_4 with and without a Groundplane for n = 1, 10, & 100 Cables				
cable	**groundplane**	**n = 1**	**n = 10**	**n = 100**
TSP	yes	3kV	2.3kV	1kV
TSP	no	14kV	8kV	1.2kV
OVB	yes	2.5kV	2.1kV	200V
OVB	no	10kV	2kV	200V
RTCA/DO-160F Table 22-3		50V to 1.6kV		

Table 11 is the comparison of the induced WF4 IR-drop cable current with DO-160F WF5A. The present results are about an order of magnitude higher on the high end.

Table 11. I_4 with and without a Groundplane for n = 1, 10, & 100 Cables				
cable	**groundplane**	**n = 1**	**n = 10**	**n = 100**
TSP	yes	5kA	4kA	1.3kA
TSP	no	30kA	12kA	2kA
OVB	yes	27kA	12kA	2kA
OVB	no	100kA	20kA	2kA
RTCA/DO-160F Table 22-3		150A to 5kA		

It is well known among lightning engineering experts that SAE ARP5412A does not include the largest voltages and currents because it's never documented outside of proprietary corporate documents. This note exposes those large numbers. Truth is that the largest are so large that protection is moot; mechanical damage is quite likely. Properly designed tests can produce good numbers for comparison. The large induced environments usually occur at the extremities of the system like the wings and empennage where groundplanes are difficult to install.

3.2. Induced Pin Levels

We have two kinds of pin levels, shielded and unshielded, neither one easy.

3.2.1. Unshielded Pin/Wire Voltages and Currents. Open-circuit pin voltages will be the same as the induced voltages on cables, above.

Short circuit pin currents depend upon the load at the other end. Pin currents connected to the loads at both ends depend upon the sum of the loads. I/O circuits protected by diode voltage clamps will almost short out if the induced voltage exceeds their breakdown voltage, V_{BD}. Then, the source impedance is well defined and the short circuit current is high.

Summarizing, pin voltages should be the same as the cable voltages. Pin currents are any one's guess however we should specify short circuit current for power calculations and ratings on surge suppression devices. Ideally, the lightning engineer, if he had the time and skills, should give each designer a Thevinin equivalent lightning source circuit for each of the designers' I/O circuits.

3.2.2. Shielded Pin/Wire Voltages and Currents. Shielded pin voltages depend upon (1) the groundplane resistance, if there is one, (2) the shield length, (3) the shield's DC resistance, and (4) the shield's diffusion time. Because indirect lightning is such a low frequency phenomenon, the shield voltage will be the $I_{shld} \cdot R_{dc} \cdot l = I \cdot R$-drop in the shield, not the $i\omega \cdot L_T \cdot l_{shld}$ voltage in series within the shielded wires.

Bottom line: Shielding the new in flight WF4 cable current is just as near impossible as the WF5A current. Also, the induced voltage is larger, above, because the induced WF4 current on the shield is larger than the old WF5A.

3.3. Waveform 3 Resonant Damped Sinusoid Excitations

Two facts up front, (1) the WF3 frequency is $f_3 = c/2l$, where l is the length of the system and the cables and (2) the amplitude/level of WF3 is proportional to the length (inversely proportional to the frequency).

3.3.1. Sources of WF3

WF3 transients have two sources, (1) the resonant airframe and (2) resonant cables. The WF3 airframe resonance is pronounced on aluminum skinned systems but damped far more on CFC skinned systems. The Components A, D, and H external currents will induce the same WF3 resonant currents. The largest WF3 cable currents since IN614[13] are those excited on cables with WF4 currents on them. Those, like the external WF3 currents, will have peak values β/ω_3 less than the WF4 currents, where β = 1/(1.5μs rise time) of the WF4 currents.

3.3.2. Waveform 3 Comparisons

The following Table 12 compares the present results for WF3 cable 1MHz and 10MHz resonances excited by WF4 IR-drop currents on the same cable to those in SAE ARP5412A. We focus on cable current here because we're dealing with shielded grounded cables for the most part. As usual, the present results run over an order of magnitude higher at the lower 1MHz frequency, however that low of a frequency will seldom be observed except on long unshielded lines loaded down with LC filters. At 10MHz, the present results are very close to the levels in 5412.

Figure 27, below, shows the natural frequencies of a Boeing 707/USAF EC-135.[6] These are inherent to the geometry of the aircraft. Any lightning strike will excite about the same natural frequencies no matter where it attaches or exits however which resonant mode is largest depends on where the system is "plucked". These diagrams also show where WF3 resonance peak and null as a function of the WF3 frequency.

Table 12. Comparison of Max Min 5412A WF3 to Present Results			
	5412A WF3	1MHz WF3 without/with gndplane	10MHz WF3 without/with gndplane
Level	**5412**	**present**	**present**
1	20A	200A/130A	20A/13A
5	640A	10kA/2.7kA	1kA/270A

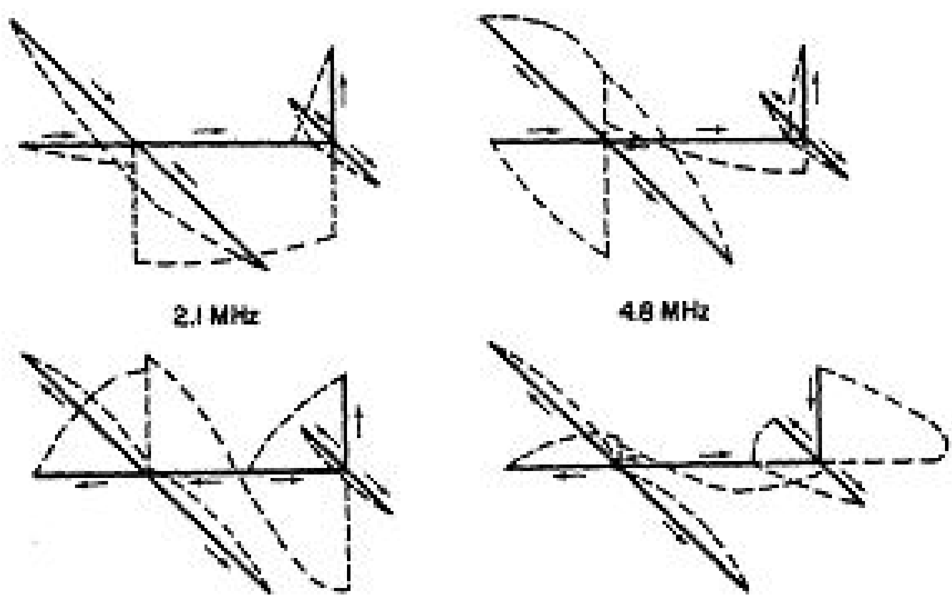


Figure 27. Natural Resonant Frequencies of a Boeing 707/ USAF EC-135[6]

4. Shielded versus Unshielded Wires in Composite Airframes

Aircraft usually do not shield power wires and usually use chassis/structure/groundplane for the power return connection. Such wires running any distance will have surge arrestors at both ends large enough to clamp the WF4 voltage and conduct the maximum WF4 current. Such a wire or set of wires will therefore respond almost the same as shielded cables grounded at both ends because when the surge arrestors conduct, the wires are effectively grounded into structure. In a composite business aircraft, such surge arrestors have to be rated for 12kVA.

5. Perceptions of Shielding Effectiveness of Unshielded Enclosures

This section addresses the issue of shielding of electromagnetic fields by airframes not designed to do so and the wrong-headed test technique used to obtain such shielding data.

The average EMC engineer believes that a tin can full of holes is a shielded enclosure and has the "data" to show it!? Boeing tried to turn one of those tin cans (commercial Boeing 747-100) into a shielded enclosure at a cost of millions and barely got 25dB. The Navy had the same luck (Lockheed Martin C-130). Of course, that 25dB was REAL – EMP hardened USAF E4-B and USN TACAMO, with every hole and penetration shielded, filtered, and/or bonded/grounded. The belief that such levels can be achieved without protection is horribly wrong-headed. Even worse is the belief that such field data has any bearing on induced transients on cables running through those areas.

The underlying point is that shielding data is always acquired with the wrong test techniques, usually results in 20-30dB of fictitious shielding effectiveness, and is therefore very popular. Bad test data lives forever. This concept is promoted by those least qualified to do so.

The “wrong test technique” is the use of traveling wave antennas (log periodic and horn antennas) inside metal enclosures like an aircraft fuselage where the EM environment is predominately standing waves. Explaining this error to the uninformed is fruitless. The measurements are always low with no correction factor(s) hence very popular. A strong argument against using any such shielding values is that they are totally uncontrolled, never maintained, and the variation system to system is never checked. The costs to do so are exorbitant.

Figure 28 depicts how directional antennas are calibrated in free space with free space impedance of 377Ω and specific angle(s) of incidence and polarization. The calibration product is a gain curve as a function of frequency, $G(f)$, usually on bore sight.

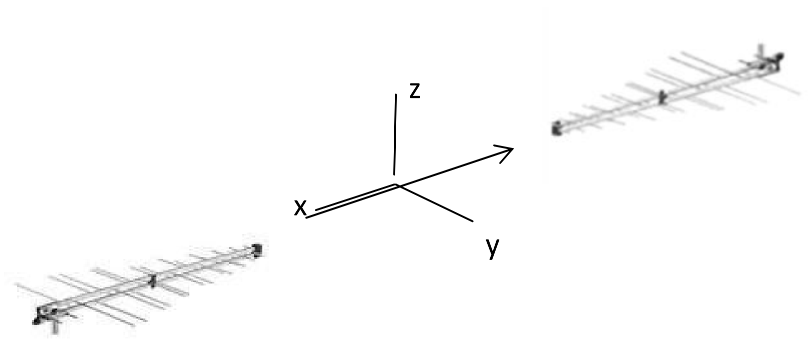


Figure 28. Direction Antenna Calibration, Free Space $Z_0(x, y, z)$ = 377Ω, $G_{cal}(f)$ Calibrated

The received power, $E^2/377$, is a function of the gain G, power P to the transmit antenna, and the distance apart, R:

(68) $\frac{E_y^2}{377} = \frac{G_{cal} \cdot P}{4 \cdot \pi \cdot R^2}$.

Figure 29 depicts misusing a directional antenna in a conducting enclosure dominated by standing waves with impedance varying orders of magnitude, peaks and nulls over the whole volume, no specific angles of incidence or polarization. The above calibrated gain, $G(f)$, cannot be used to relate received power in an enclosure to field strength and there is no correction factor. The antenna will pick up a signal but it cannot be related to field strength.

Figure 29 depicts a notional test set up from the often copied MIL-STD-285[20] and is used universally to make measurements inside airframes so that designers can lower their allocated EMI and lightning levels on their cables and boxes. The notional equation demonstrates the high degree of variability as a function of position and frequency parameters neither of which is checked in the usual quick and dirty *ad hoc* testing. The notional equation also makes the point that the high degree of variability is not in the shield's parameters. The results are usually a stream of standing waves varying two to three orders of magnitude with little reducible information about the subject shielding.

The antenna inside is not calibrated for the standing wave environments anyway so any discussion of the merits or deficiencies of this type of test are irrelevant.

Reference 21 also voices this author's concerns about of MIL-STD-285 testing in general, "The accuracy and repeatability of results are questionable across laboratories." The term "questionable" misses the point that this test can never be accurate or repeatable. It has been or is being replaced by IEEE STD 299. Nevertheless, directional antennas inside conducting enclosures are the wrong antennas and always obtain perceived field levels lower than what are really there.

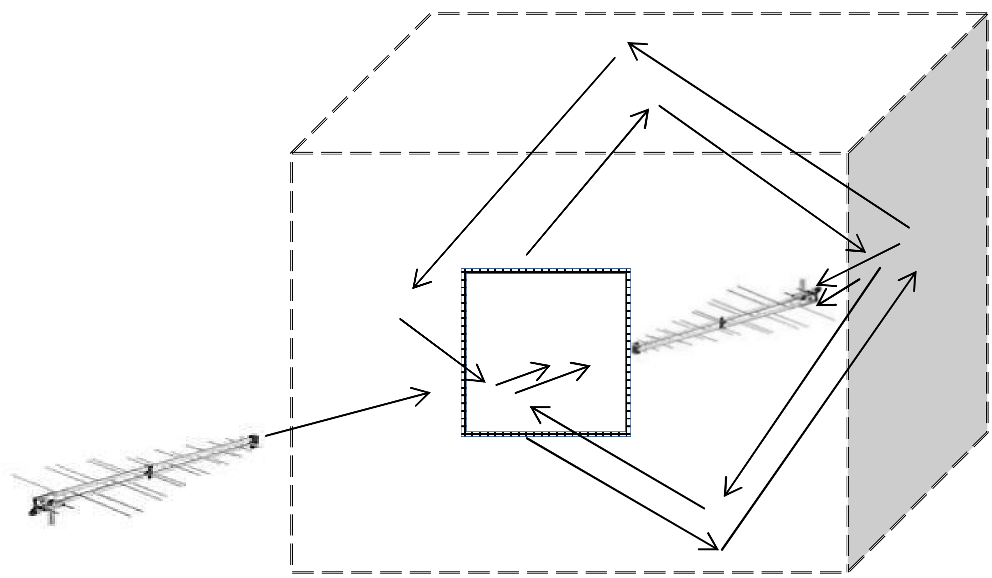


Figure 29. Directional Antenna Misuse, Cavity Z_0(x, y, z) ≈ $+Q\eta/0$, *G(f)* ≈ $\pm ? < G_{cal}(f)$[20]

(69 $$\boldsymbol{E}(x, y, z) \approx \boldsymbol{E}_0 \cdot \frac{R_T + i\omega \cdot L_T}{377} \cdot \frac{\sum\sum\sum \sin\left(\frac{l\cdot\pi\cdot x}{L_x}\right)\cdot\sin\left(\frac{m\cdot\pi\cdot y}{L_y}\right)\cdot\sin\left(\frac{n\cdot\pi\cdot z}{L_z}\right)}{\sqrt{\omega^2 - \omega_{lmn}^2 + 4\cdot\omega_{lmn}^2 / Q_{lmn}^2}}$$, where *l*, *m*, *n* = 1, 2, 3, …

The same goes for composite CFC airframes with a special nuance. You will be shown data on attenuation of electric fields which is usually good. You will not be shown data on the attenuation of magnetic fields which is usually poor. Why? Electric fields terminate on any conductor, hence don't penetrate, while magnetic fields pass through CFC like it's almost transparent.[6,16] The FAA sees through the fiction and requires HIRF testing at max levels without the CFC fuselage.[4]

5.1 Bond Joints & Shielding

Aluminum-to-aluminum electrical bond joints achieve ≤ 2.5mΩ DC resistance across a 1" square faying surface/joint. Carbon composite–to-carbon composite joints achieve 1-20Ω across similar joints. That difference is about 50-78dB.

That is a measure of the relative shielding effectiveness of enclosures made of those materials since an enclosure's shielding is no better than its joints/seams (plus apertures, etc.).

The 140dB mumetal cable shield introduced in paper #2[14] turns into an 80dB shield by the non-ferrous connectors.

An analyst's observation about a cap over a cylinder with a small gap is worthy of notice: "The comparison shows that a perfectly conducting cap with an annular slot, even though the slot is extremely narrow, seems to have not much effectiveness in shielding against the external magnetic field."[6] In other words, unbonded seams render an enclosure unshielded.

5.2 Low Frequency Diffusion

A typical solid (no seams, etc.) aluminum enclosure's magnetic field diffusion curve will start to fall off from 0dB around 10Hz while a CFC enclosure will start to fall off around 10kHz. Throw in poor joints and seams and apertures and the shielding effectiveness of CFC enclosures is effectively zero.

Confusion about the shielding of a panel versus that of an enclosure often causes the shielding to be overestimated. First, a panel will have much higher apparent shielding effectiveness at low frequencies. In Figure 30 below, the difference for any material is about 10^7 or 140dB at the lowest frequencies.[6]

Second, an enclosure will be constructed with many seams, joints, apertures, and conductive penetrations that reduce the shielding effectiveness. [As an aside, once, while working on a program that included many truck mounted shielded enclosures, the system engineers came to me complaining that the standard shielded shelter leaked rain water and asked what I could do to stop it. I subsequently specified the low frequency magnetic shielding so strict that the only way to meet it was to use shelters with all welded seams. A young EMC engineer that used to work for me was employed at one of the shelter vendors. When his boss came to him asking about the unusual requirement, the young engineer said that it came from a physicist and he knew who it was. My telephone immediately rang with the young engineer and his boss on the line asking me about the requirement. When I told them that it wasn't for electromagnetic shielding but rather to stop water leaks, we all enjoyed a laugh. The requirement was eventually dropped after word spread, leaks or no leaks.] The point is that the seams and joints in a shielded enclosure dictate both the low frequency and high frequency shielding and the requirements placed on an enclosure dictate how it is fabricated. High performance EMI and TEMPEST test enclosures almost all use welded seams and double walled construction. Even then, the ambient will often include nearby radio, TV, or radar emitters. (It's the door(s).)

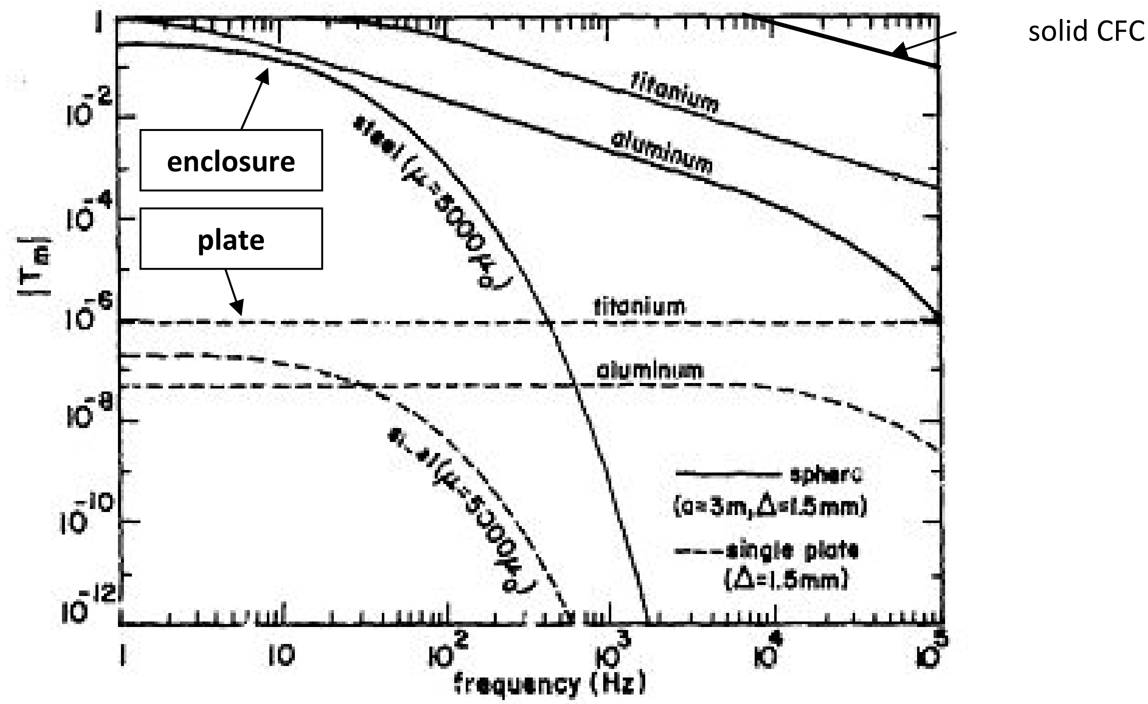


Figure 30. Infinite Plate versus an Enclosure Shielding Attenuation[6]

5.3 Copper Mesh Effects on CFC Shielding[42]

A typical copper mesh in the outer layer of a CFC surfaces has resistance of about 0.4-3mΩ/square and inductance of 300-1000pH/square. On a 10m long CFC cylinder with a 2m radius, the total resistance of the mesh is as large as the CFC. Squelch the argument that the mesh adds appreciably to the CFC shielding effectiveness. Figure 31, below, shows Dexmet shielding data on their expanded foil. Note that as low as 10MHz it is still inductive, rolling off at 20dB/decade. The shielding effectiveness is normally modeled as the ratio of the sheet impedance to that of free space:[6,16]

(70) $$SE \cong 20 \cdot \log \frac{|R_{sh} + i\omega \cdot L_{sh}|}{120 \cdot \pi}.$$

Choosing the sample with SE = 40dB at 1GHz, the sheet inductance is L_{sh} = 637pH. A SE = 80dB at 10MHz assures that the sheet resistance is no more than R_{sh} = 38mΩ.

Two points on such mesh, (1) the mesh used in the outer skin is as light and thin as possible therefore a poorer shield and (2) the mesh used in the internal ground planes under cable runs and boxes is more robust because it has to carry power and power fault return currents plus around 100kA of lightning. The most practical bond joints that can be made with the external fine mesh is Class S static bond for p-static bleed off purposes whether in or on CFC or fiberglass. It is necessary to obtain those bonds to reduce noisy p-static discharges. The internal groundplane bond joints have to be as low as possible, e.g. 2.5mΩ in order to keep the system I·R-drop as low as possible.

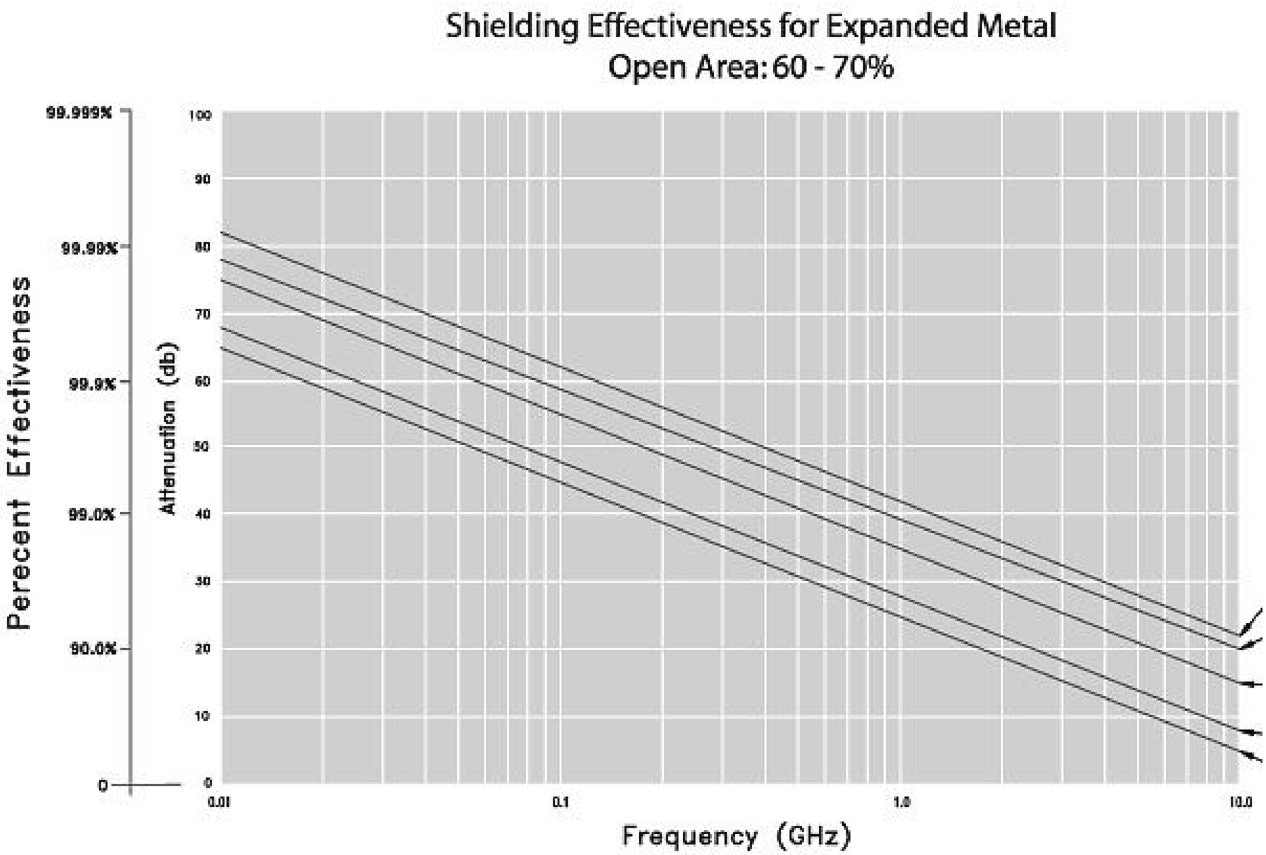


Figure 31. Dexmet Data on Shielding Effectiveness of their Expanded Foil[42]
(No information on test method or the material resistance, thickness, and strand width)
(Remember from Figure 30, this shielding is for a panel, not an enclosed volume.)

5.4 Aircraft Fuel Sensor Protection against Lightning & EMI

Fuel quantity probes are typically vertical coaxial capacitors whose outputs vary with the height of fuel inside the probe. They are connected back to the electronics with a shielded twisted pair (TSP) of wires that fasten differentially across the two coaxial cylinders of the probe. The probes are unshielded from the outside except for the TSP shielding and whatever unintentional shielding the fuel tanks provide. Fuel tanks are seldom constructed to be shielded enclosures because the designers don't want to degrade the corrosion control of the anodized fittings and hatches. Bonding jumpers are the norm for achieving a common electrical connection between all of the fuel system components and aircraft structure. Bonding jumpers are only good for static bonds because their inductance is more than the plumbing they're fastened to.

The fuel probes capacitance to the fuel tanks is due to their proximity to the top and bottom of the tanks. The induced voltage across that capacitance induced by lightning and EMI is the safety concern. Common mode and differential induced voltages can cause wrong fuel readings if present long enough.

Figure 32, below, illustrates the wiring between the fuel probe and the fuel system electronics. The twisted pair is designated as one line. The TSP shield is designated as a dashed line. The capacitance to both the top and bottom of the fuel tank is designated as one capacitor. The induced voltage across the capacitor is due to the voltage induced across the fuselage and the wing, $V_{fuselage} + V_{wing}$, between the probe and the end-point electronics. This problem is complicated by the probe not being inside a shielded enclosure like the usual electronic load.

The voltage induced across the fuselage and wing is due to the V = I·R-drop from lightning current through the two structures and the structure resistance. That voltage divides across the capacitance and the load at the other end. That voltage also includes the voltage drop across the cable shield in parallel. In composite-skinned airframes, this voltage can reach 10kV. The cable shield does little because the probe is unshielded. The voltage goes down as the number of other cables and plumbing that parallel it increase in number, shown crudely below; N_f is the number of parallel conductors in the fuselage and N_w is the number in the wing.

(71) $$\boldsymbol{V}_{lightning} \cong \boldsymbol{I}_{lightning} \cdot \left(\frac{R_{fuselage} \cdot R_{shield}}{N_f \cdot R_{fuselage} + R_{shield}} + \frac{R_{wing} \cdot R_{shield}}{N_w \cdot R_{wing} + R_{shield}} \right).$$

A cable shield ungrounded at either end will offer zero shielding effectiveness up to the quarter-wave resonant frequency of the cable shield where it will then amplify the voltage by say Q ≤ 10.

The probe is a capacitive load therefore the induced voltage across the probe will decrease with increasing frequency (ignoring resonances), $V_{probe} = I_{cable}/i\omega \cdot C_{probe}$. That makes the induced voltage larger at the electronic load on the other end taking away the safety issue of HIRF-induced voltage on fuel probes.

Summary. The structural IR voltage drop induces the largest voltage between the probe and the tank walls.

In this simple example, the induced voltage ranges from a max of 100kV down to 200V depending upon the groundplane, the cable shield thickness, and the number of parallel conductors. Installing wide groundplanes in the wings is difficult because there is no room. Other environmental conditions may require a minimum voltage standoff rating for the probes *in situ*.

Clearly, transformer isolation (of digital signals from an A/D device) and/or optical isolation both with several pF capacitance, series current-limiting resistors, and at least 1kV standoff rating is required to make this susceptibility truly safe.

<u>The fuel tank would have to possess shielding effectiveness equaling or exceeding the cable shield in order to make this problem depend upon the cable shield only</u>. Fuel system designers are too protective of their corrosion resistant non-conducting anodized joints to allow sealing the tanks with conductive seams and joints. They will, however, attach a spider's web of bonding jumpers across almost all non-conducting joints. This is OK for static discharge grounding but efficient for EMI and lightning only in your dreams. There are also a lot of moveable conductive disconnects in fuel lines subject to high vibration; these are sealed with fitting s that employ gaskets for fuel leaks and have finger stock that make electrical contact across these moveable joints.

What we're trying to get across here is that even though there is a lot of conductive plumbing paralleling cable runs in aircraft, taking credit for them in reducing lightning allocations on the cables is questionable since they do not possess EMI and lightning bonding requirements of, say, 2.5mΩ, just static, 1Ω.

The Fully Automated Electronic Control (FADEC) on the engines has a similar problem except that it is fully shielded.

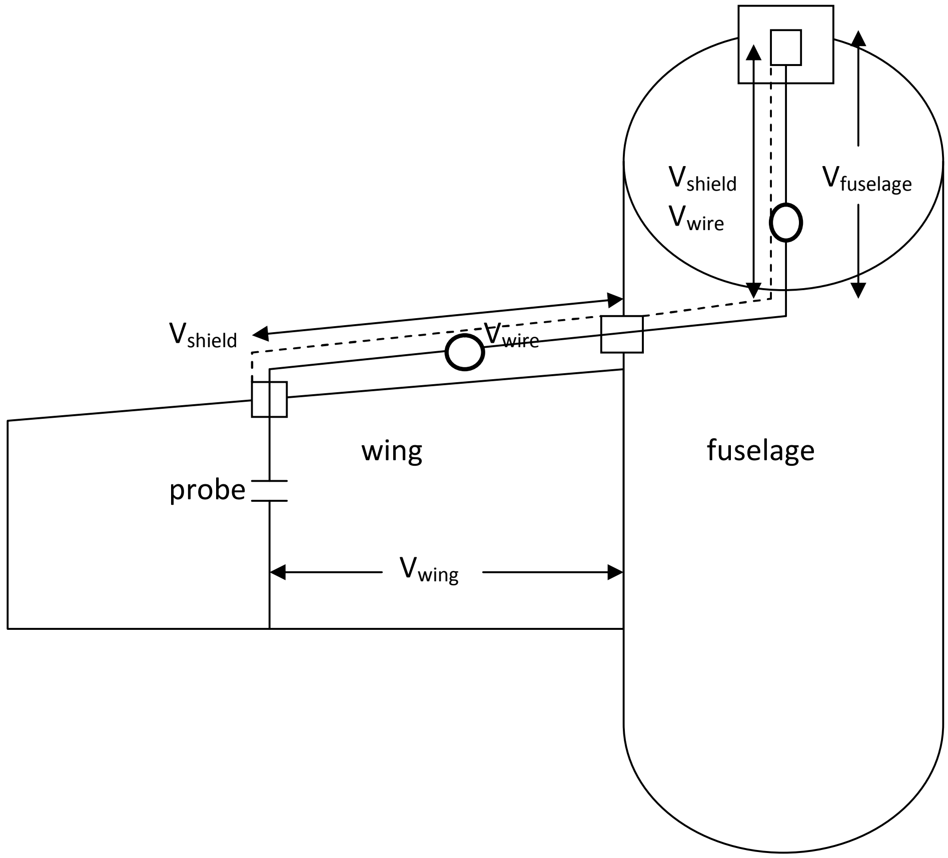


Figure 32. Diagram of Fuel Tank Level Sensor Connection to the Aircraft with Different induced Voltages

5.5. Redefinition of Protected Environments

1. There are no controlled "protected environments" within modern airframes other than shielded boxes and shielded cables.

2. There are cable runs that do not pass close by apertures or into the wings, empennage, or wing body fairing.

3. Cables running from outside the pressure hull will normally be terminated on a bulkhead connector at the pressure hull while the wires will run through, shielded or unshielded.

4. Cables are routed close and far from other cables with some separation because of EMC, safety, or mechanical damage and other separation for convenience. Cables that do run by apertures will be routed within some distance from those that do not.

5. Most of the electromagnetic energy in the so-called protected volumes comes from cables running into those volumes from other parts of the system. Those cables carry the currents induced from apertures, I·R-drop, etc. in the more exposed parts of the airframes. The more energized cables will then crosstalk onto the cables that are less energized. The pretense that the fields inside the protected areas control the excitation of the cables is false.

6. As a heuristic example, Figure 33, a shielded cable running by an aperture will have a current, I_1, induced on it due to the lightning current, I_{ext}, and fields on the outside of the aperture. Let's call that cable #1. That cable runs close to another cable #2 that is not near such an aperture and it will "crosstalk" onto cable #2 through a mutual inductance, M_{21}.

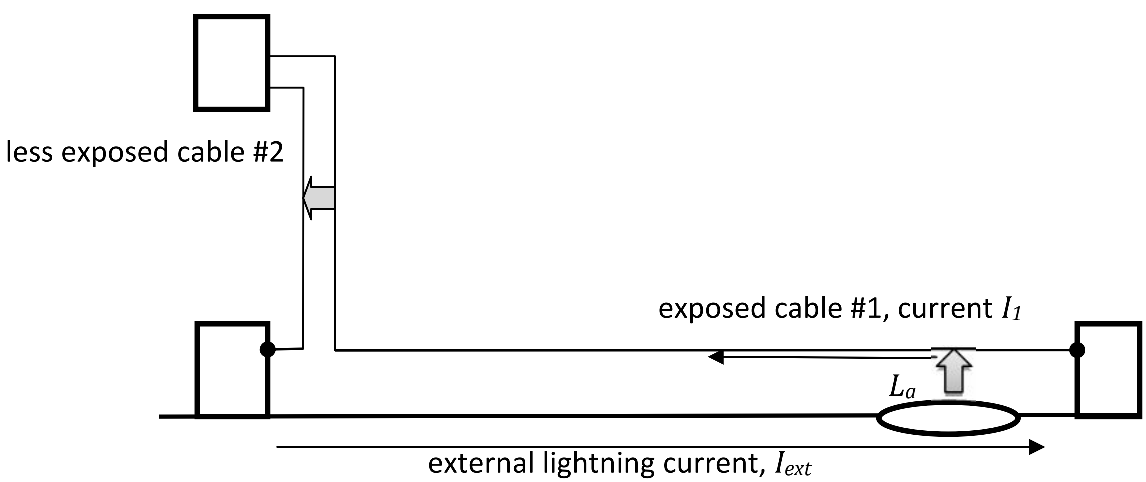


Figure 33. Mutual Coupling from One Cable to Another cable

The current, I_1, on cable #1 due to its proximity to an aperture will be approximately

(72) $$I_1 \approx \frac{L_{ap}}{L'_1 \cdot l_1} \cdot I_{ext},$$

where L_{ap} is the mutual coupling through the aperture and $L'_1 \cdot l_1$ is the inductance of cable #1.

The crosstalk of that current onto another shielded cable #2 is approximately

(73) $\boldsymbol{I_2} \approx \frac{M'_{12} \cdot l_{12}}{L'_2 \cdot l_2} \cdot \boldsymbol{I_1}$,

where $M'_{12} \cdot l_{12}$ is the mutual magnetic coupling between cables #1 and #2 along a length of cables l_{12} common to both and $L'_2 \cdot l_2$ is the inductance of cable #2.

The current induced on cable #2 due to the exterior current, I_{ext}, is then

(74) $\boldsymbol{I_2} \approx \frac{M'_{12} \cdot l_{12}}{L'_2 \cdot l_2} \cdot \frac{L_{ap}}{L'_1 \cdot l_1} \cdot \boldsymbol{I_{ext}}$.

A measure of the "protection" of cable #2 due to its location away from an aperture yet near to cable #1 that runs by an aperture is then written as an attenuation, A,

(75) $\boldsymbol{A} \approx \frac{M'_{12} \cdot l_{12}}{L'_2 \cdot l_2} \cdot \frac{L_{ap}}{L'_1 \cdot l_1}$.

The cables are both at height, h = 1cm, above the groundplane, a distance, d, apart, and have radii, r = 1cm. The mutual inductance is that of two cables above a groundplane:[6]

(76) $\boldsymbol{M_{12}} = \frac{\mu_0 \cdot l}{2 \cdot \pi} \cdot \ln\left[\frac{d^2+(2 \cdot h-r)^2}{d^2+r^2}\right]$ with the cable radius, r, included and

(77) $\boldsymbol{M_{12}} \cong \frac{\mu_0 \cdot l}{2 \cdot \pi} \cdot \ln\left[1+\left(\frac{2 \cdot h}{d}\right)^2\right]$ with the cable radius, r, set to zero.

The WF4 cable current is set to 10kA and the results can be scaled according to actual levels. The lengths of parallel routing of the source and receptor cables are $l_{12} = 1m, 2m, and\ 3m$. The induced WF1 current and WF2 voltage are given as follows:

(78) $\boldsymbol{I_1} \cong \frac{M'_{12} \cdot l_{12}}{L_1} \cdot \boldsymbol{I_4}$;

(79) $\boldsymbol{V_2} \cong \boldsymbol{M'_{12}} \cdot \boldsymbol{l_{12}} \cdot \frac{I_4}{\tau_r}$, where the WF4 rise time, $\tau_r = 1.5\mu s$.

These simple formulas are provided so that the lightning engineer can scale his or her results according to the system situation at hand. I_1 (WF1) is plotted in Figure 34 and V_2 (WF2) in Figure 35.

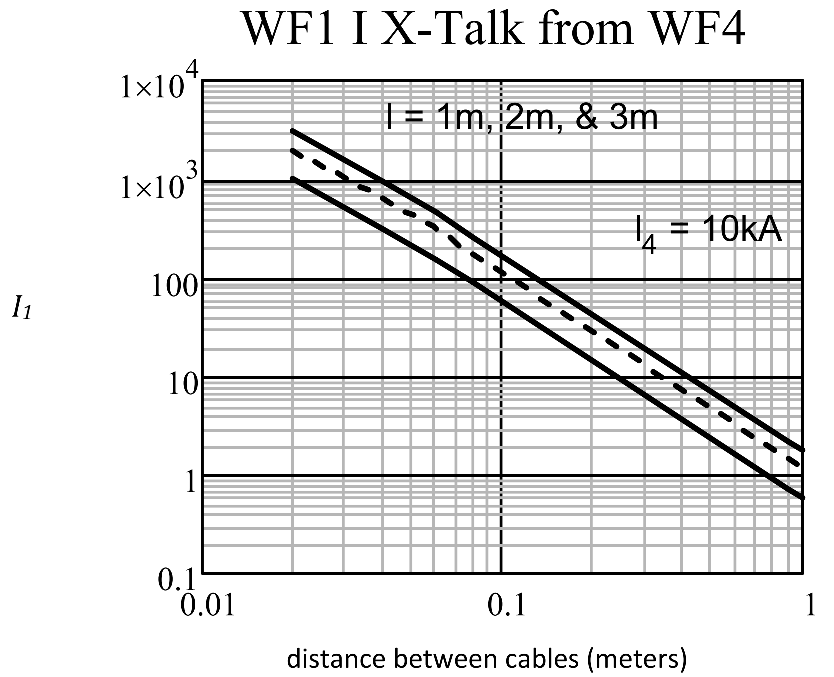


Figure 34. WF1 Current Cross Coupled from WF4 Current for Parallel Paths of 1m, 2m, & 3m

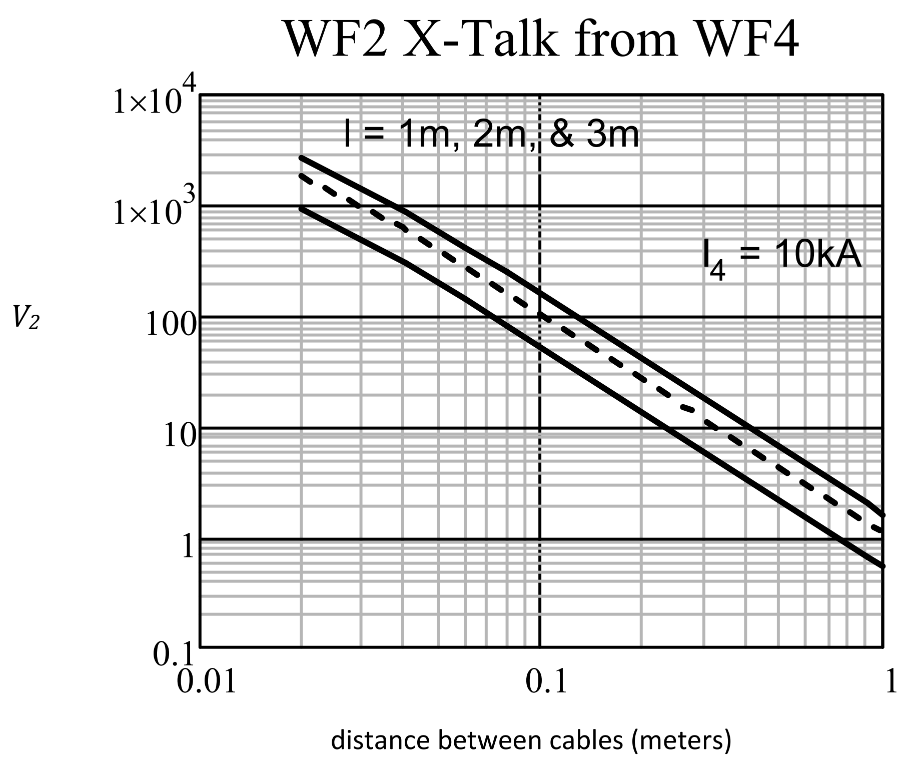


Figure 35. WF2 Voltage Cross Coupled from WF4 Current for Parallel Paths of 1m, 2m, & 3m

Clearly, each cable's inductance and lengths play the important role of current-limiting. The length of cable runs, l_{12}, where cable #1 is close to cable #2 is a big factor, too, along with the distance between cables compared to their height above structure.

Allow both cable to have the same inductance, L_1 = L_2 = 3µH. Cable #1 running 6" from a 1' radius aperture has an aperture coupling inductance of L_{ap} = 30nH. The mutual inductance between two cable 3" apart for a length of 1m is about M_{12} =300nH. Therefore, the ratio of current, I_2, on cable #2 to the external current, I_{ext}, is about A = 70dB down from the external 200kA, 20dB down from the first cable. In other words, if a 2kA current is induced on cable #1, about 200A could be induced on cable #2. Assuming cable shields with a transfer impedance of 100mΩ each, the induced voltages could be V_1 = 200V and V_2 = 20V, respectively. Clamping these levels of common mode voltage down to say 15V is easily done with small diodes, one for each wire/pin, however it requires a small space (0.24"x0.12") on the respective circuit cards plus a chassis groundplane on the cards, space usually not available 6 months into a program. A small isolation transformer would take up more room, say 0.6"x0.4". In order to preclude weak components, specify that each TVS must withstand 10,000 pulses at the max levels. Normal transient traffic can wear out cheap components unless this is specified.

One purpose of this note is to point out that the above resources of (1) a decent cable shield and (2) surge suppressor parts and (3) space on circuit cards have to be identified at the beginning of a program, not too late into it, otherwise they will be difficult to obtain. Over-specifications can be reduced as information becomes available but under-specifications are difficult to increase.

6. Critical versus Non-Critical Circuit Protection

In the interests of cost cutting, critical circuits like cabin pressurization will be well protected. Right next to those circuits, often in the same connector and often on the same circuit card, lie less-critical circuits such as cabin temperature and are less protected and don't have to pass the same levels of EM environments. Guess what? EMI and lightning on the non-critical lines will crosstalk into the critical circuits inside the avionics boxes and induce failures.

Heavily protected critical circuits must therefore be isolated to a degree from less protected less-critical circuits all the way into the circuit cards.

Warning: The I·R-drop voltages cannot be protected against one circuit at a time in a box. The phenomenon effects the entire box.

7. Electrical Components Mounted on the Exterior Exposed to Direct Strikes

Components mounted on the exterior are considered sacrificial when directly struck by lightning. However, we need to prevent the resulting lightning-induced transients from getting inside and causing damage to other hardware. The first job is to determine the zones these components are located. Then, apply surge protection as close to the external components as possible so as to stop the lightning from penetrating into the system any further, examples shown in Table 13.

Table 13. Typical Protection on External Components to keep Lightning from the Interior

External Component	Protection Suggestions	Notes
Antennas	¼-wave stubs, gas surge arrestors	residual voltage ≈ 15V
Pitot & temp	diode clamps	100kW
Lights	diode clamps	100kW
Open connectors	TBD pending zoning assessment	(usually not a concern)
Control surface sensors	diode clamps	100kW
WOW*, brakes, & sensors	diode clamps	100kW
Windshield deicing conducting film	diode clamps plus inductive filter, if possible	100kW

Note: * WOW are the weight on wheel switches in the landing gear that disengage the autopilot and other automatic controls upon touchdown.

8. Nearby Lightning Strikes, Fields and Coupling

Magnetic fields close to a lightning strike follow Ampere's law.

(80) $$\boldsymbol{H} = \frac{I}{2\cdot\pi\cdot R}.$$

The open circuit voltage induced in nearby vertical loops of area $A = l\cdot h$ follows faraday's law.

(81) $$\boldsymbol{V}_{loop} = \boldsymbol{A}_{loop}\cdot\boldsymbol{\mu}_0\cdot\frac{d\boldsymbol{H}}{dt} \cong \boldsymbol{A}\cdot\boldsymbol{\mu}_0\cdot\frac{H_{peak}}{\tau_{rise\ time}}.$$

A 200kA lightning strike 100m away can induce 200 volts in a 10m long cable 3" off of structure, 2kV when 10m away. This is enough to cause damage if not mitigated. This is both an airborne and a ground problem. See Section 13, below, for ground surge protection.

9. Ground Operations Surge Requirements

Aerospace EMI/EMC and lightning requirements documents are characteristically silent on ground operations. MIL-STD-461C showed a healthy respect for ground surge environments by making CS06 power surge on GSE power connections x2 larger than that for operational systems unconnected to GSE.

Besides lightning, normal ground power transients can wear out components. The USN put CS106 back into part of MIL-STD-461F[43] with the following rationale, far too practical and experienced for most pedants:

> *"The Navy submarine community has found the obsolete CS06 of MIL-STD-461 (through revision C) requirement to be an effective method to minimize risk of transient related equipment and subsystem susceptibility. This type of transient susceptibility test has been successful in early identification of transient related EMI problems in naval equipment and subsystems. The Navy has found good correlation between transient related shipboard problems, including longevity, degraded performance, premature failures, and CS106 susceptibilities."*

IEEE and ANSI have more experience to add:[32]

> *"Surge voltages occurring in low-voltage ac power circuits originate from two major sources: lightning effects (direct or indirect) on the power system and system switching transients. The rate of occurrence of surges varies over wide limits, depending upon the particular power system. Prediction of the rate for a particular system is always difficult and frequently impossible. The rate is related to the level of the surges; low-level surges are more prevalent than high-level surges. The frequency of surges is a key indicator of their damage potential."*

Typical ground surges are specified as follows in Table 14 depending upon location and the degree of building lightning protection:

Table 14. Comparing Lightning Standards inside a Facility
NFPA 780,[31] ANSI/IEEE C62.41,[32] & IEC 61000 4-5[33]

NFPA 780 A.4.18.3.1	ANSI/IEEE C62.41	IEC 61000-4-5
Service entry – 40kA	Category C – 3-10kA, 6-20kV	Class 4 – 4kV
Branch panels – 20kA	Category B – 1-3kA, 2-6kV	Class 3 – 2kV
Point of utilization – 10kA	Category A – 70-200A, 2-6kV	Class 2 – 1kV

The very different levels are indicative of the different committees and their charters. The ANSI/IEEE C62.41 documents appear to be the most researched and complete. The point is that any system on the ground must be able to withstand such surges on ground connections depending upon the protection provided by the host facility. A list of US and European standards that address power surges and lightning protection of facilities and the equipment inside are the following:

NFPA 780, Standard for the Installation of Lightning Protection Systems[31]

ANSI/IEEE C62.41-1991, IEEE Recommended Practice in Low Voltage AC Power Circuits[32]

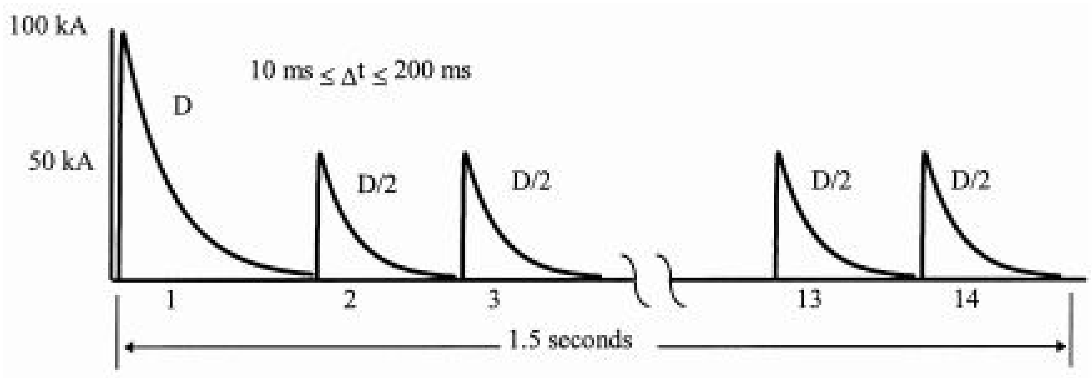
IEC 61000 4-5, Equipment Classification of Power Surge and Lightning Protection[33]

All too often, everyone focuses on in-flight issues and turn a blind eye towards ground operations, ground support equipment including power sources, and test.

10. Functional Upset, Multiple Stroke and Multi Burst Environments

The primary purpose of the Multiple Stroke and Multiple Burst Waveform sets, below, is to evaluate system functional upset susceptibilities.[1,2] This section will address the allocation of the electrical environments plus allocation of bit error and burst error requirements on relevant circuits and software.

10.1. Multiple Stroke. Multiple Stroke is defined as 14 Component D transients, the first stroke at max value and the subsequent strokes at half the initial transient, 10ms-200ms apart, not to exceed 1.5s in total duration with waveforms and level listed in Tables 5-7 and depicted below in Figure 36.[1,2]



One current component D followed by thirteen current component D/2s distributed over a period of up to 1.5 seconds.

Figure 36. Component D Multiple Strokes[1,2]

The Component D waveform is defined as follows:

(82) $$I_D(t) = 109.405kA \cdot \left(e^{-t/44\mu s} - e^{-t/773ns}\right)$$

This current divides between the skin and internal conductors as shown below in Figure 37. The voltage induced inside various shielded cables is shown below in Figure 37. The faster rise time is not affected by the internal inductance any more than the slower Component A.

Again, normal copper cable shields attenuate too little for damage let alone functional upset. <u>This is why this paper recommends that functional upset be solved with circuit and software design that disallows false commands from automated flight equipment due to the lightning and EMI environments</u>. Avionics companies like Rockwell Collins and Honeywell are well experienced with the tricks-of-the-trade, e.g. fail-passive, fail-operational, etc., all architectures that have multiple redundant sensors repetitively sampled and correlated before issuing automated commands.

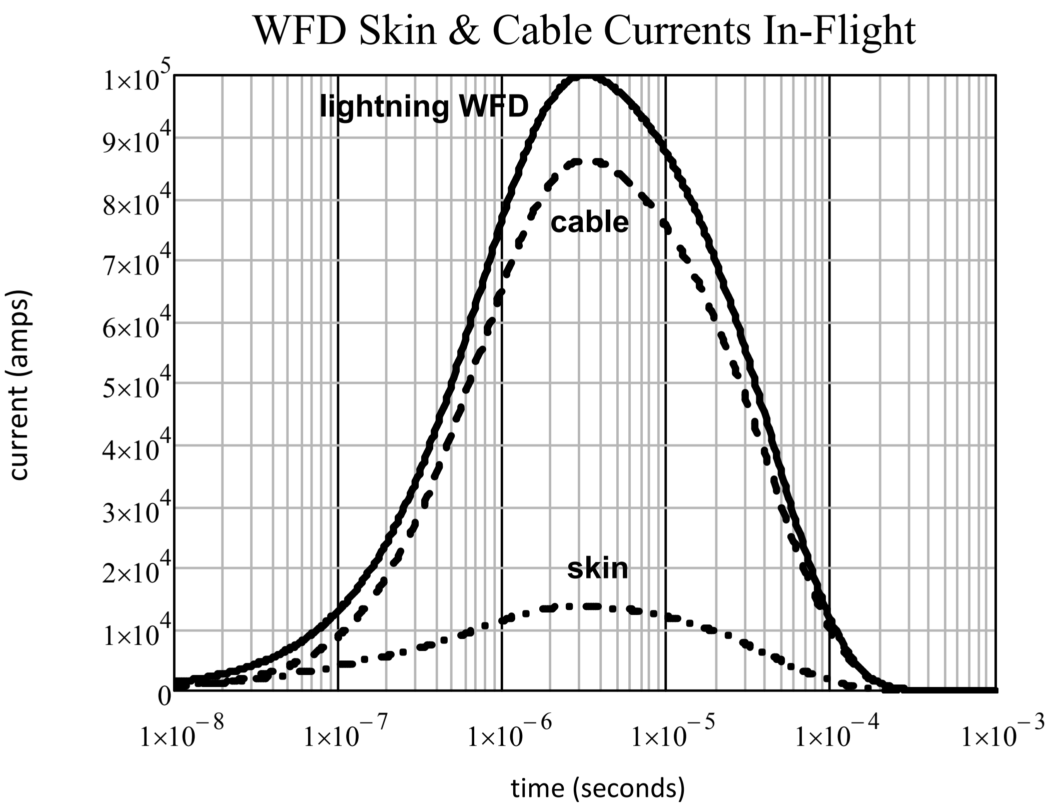


Figure 37. Waveform D Current Division between CFC Skin & Ten Cables

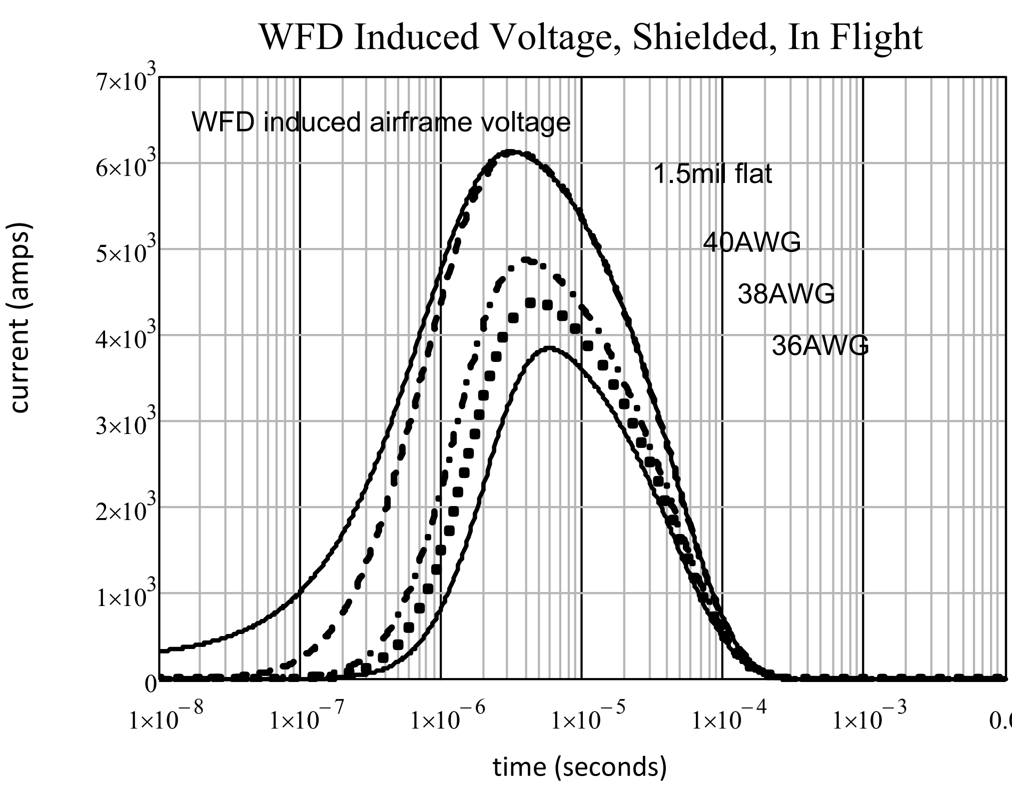


Figure 38. Waveform D Induced I·R Voltage inside Shielded Cables
See Table 1 for Shield parameters.

10.2. Multi Burst. Multi Burst is defined as 3 bursts of 20 Component transients, 50µs-1ms between individual transients, 30ms-300ms between bursts, not to exceed 620ms in total duration with waveforms and levels defined in Table 7 and depicted below in Figure 39.[1,2]

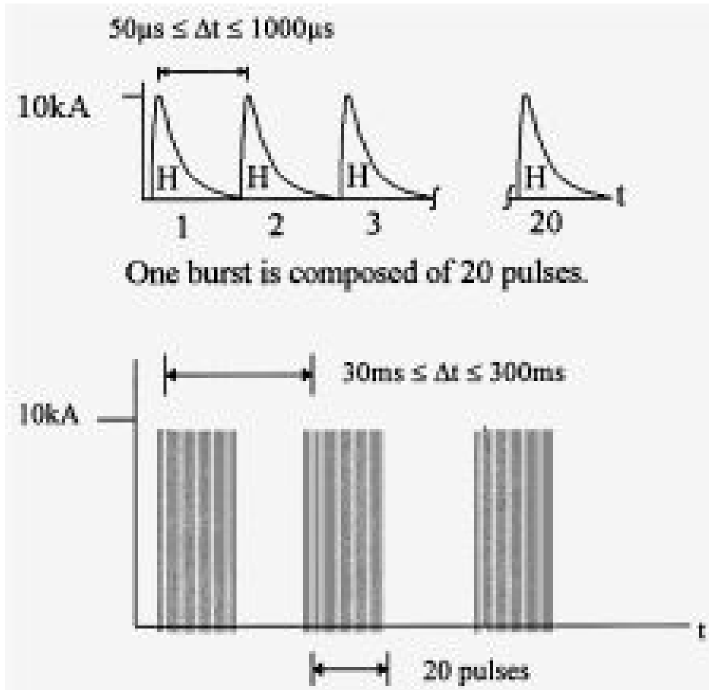


Figure 39. Component H Multi-Burst Strokes

The Component H waveform is defined as follows:

(83) $$I_H(t) = 10.572kA \cdot \left(e^{-t/53\mu s} - e^{-t/52ns}\right)$$

This current divides between the skin and internal conductors as shown below in Figure 39. The voltage induced inside various shielded cables is shown below in Figure 41. Note in Figure 40 the odd low frequency IR-drop voltage behavior. The NTSB and the author agree on the cause, "pilot error". The faster rise time is not affected by the internal inductance any more than the slower Component A.

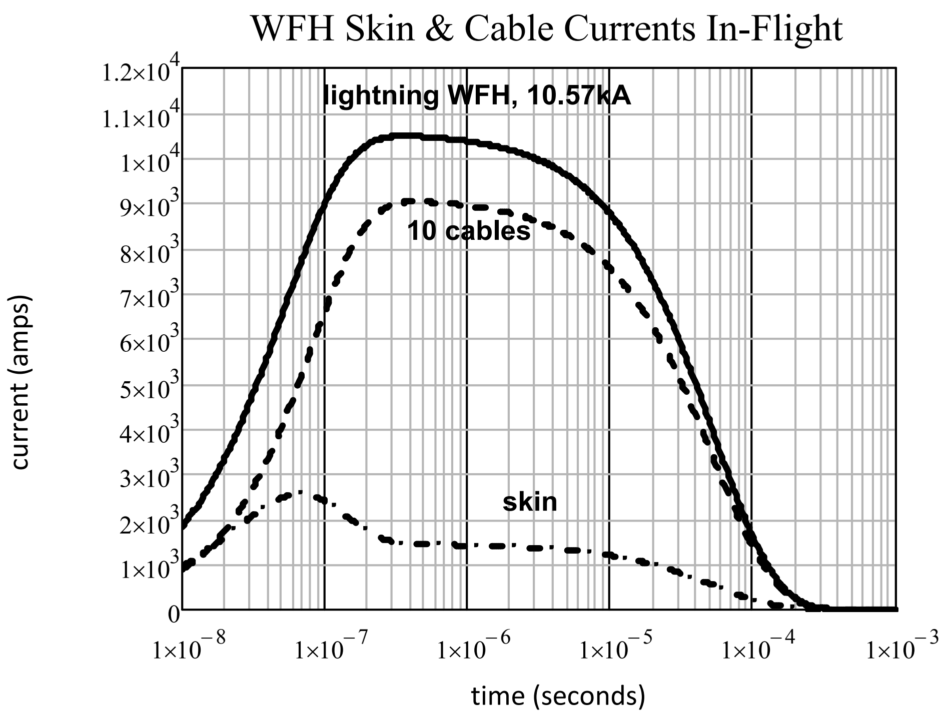


Figure 40. Waveform H Current Division between CFC Skin & Ten Cables

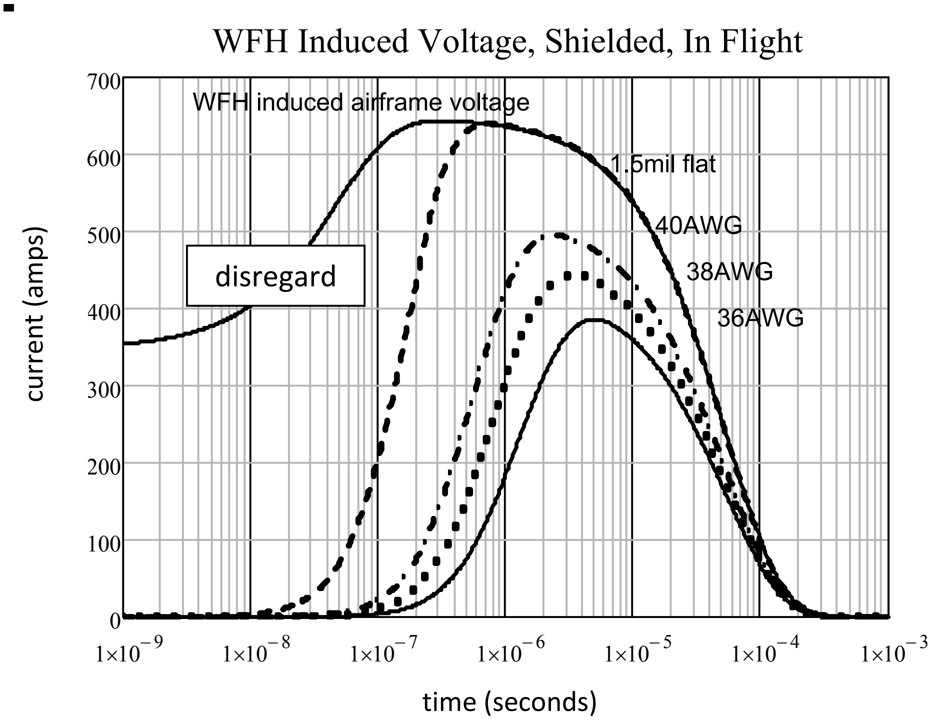


Figure 41. Waveform H Induced I·R Voltage inside Shielded Cables
See table 1 for shield parameters.

For both Components D and H, the higher frequency early time part of the transients is being conducted on the interior cables.

10.3. Waveform 3 Resonant Damped Sinusoids induced on Cables by Components A, D, and H

The spectra of Components A, D, and H are plotted in Figure 42.

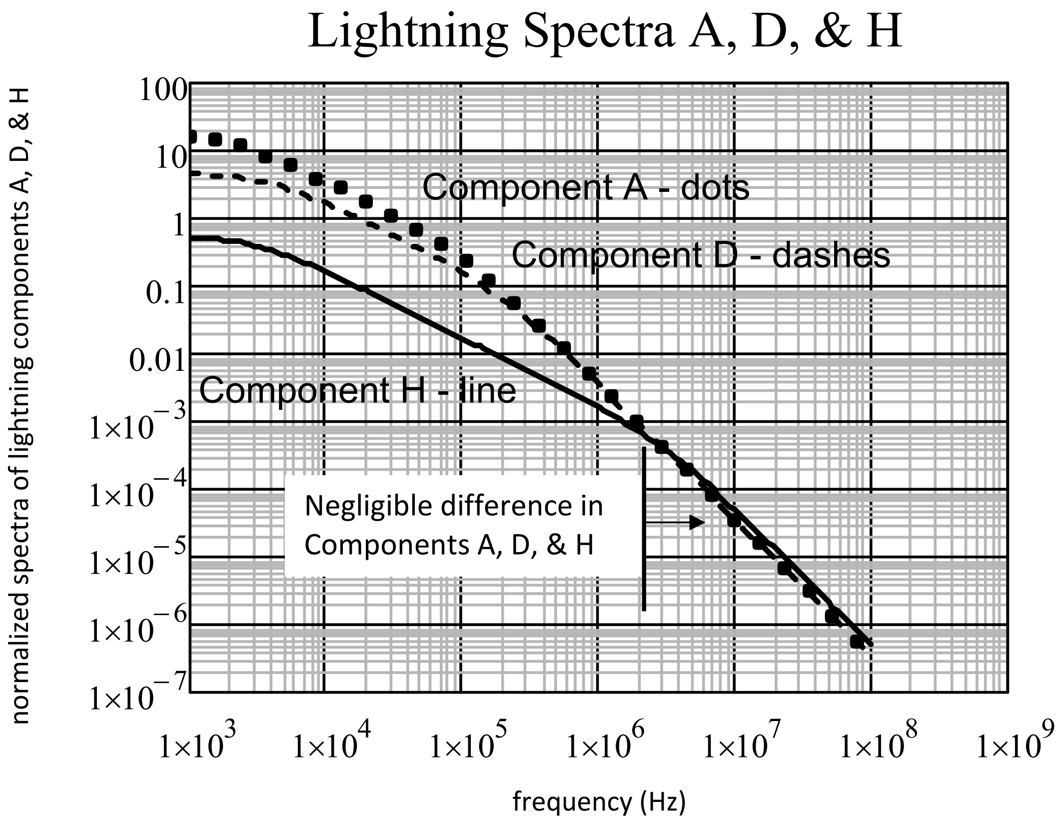


Figure 42. Spectra of Lightning Components A, D, & H

Clearly, in terms of frequency, Components A, D, and H are comparable above 2MHz. This makes WF3 damped sinusoid cable resonances comparable for all three. In other words, a 1kA WFA, D, or H cable current will all produce the same 69A WF3 current at 15MHz, more at lower frequencies and less at higher frequencies, inversely proportional to the frequency.

As explained in Section 1.5 and above, Component A peak cable currents can run from 100kA down to 2kA, Component D peak cable currents can run from 8kA down to 133A, and Component H peak cable currents can run from 900A down to 18A. However, to repeat the above, the resonant WF3 transients will be the same for all three above 2MHz or for cables 75m or shorter. Note that Tables 3, 5, and 7 from DO-160F have WF3 levels different, x20 in current and x1.67 in voltage, an indication of problems in those tables. The thinking was on peak transient amplitudes, not spectral amplitudes.

These external lightning waveforms end up as Waveform 3 damped sinusoid currents on cables. The WF3 frequencies depend upon (1) the length of the system and (2) the lengths of the cables; the longer the system and cables, the higher the amplitude (proportional to length) and the lower the frequency (inversely proportional to length).

10.5. Functional Upset Protection. All of these phenomena contribute to more problems for functional upset protection of the electronics. For that reason, this author recommends three protection features common to high speed digital busses, (1) common mode protection on both the line driver and line receiver ends of the lines mostly for damage protection, (2) isolation devices on both ends, and (2) software mitigation to ignore or correct the bit errors and burst errors that these types of transients induce. More design guidance is offered in Section 13.

10.6. Integration of Lightning Functional Upset Environments with EMI Environments

While we're at it and because the lightning engineer and the EMI engineer are usually one and the same, EMI also appears as bit errors and burst errors with characteristics such as the example described below in Figures 43 and 44.

EMI seldom is a single noise signal but contains multiple exposures depending upon the emitters. Radars emit some of the largest emissions and are repetitive pulses. Signal bit error and burst error analyses should use the following ranges of pulsed RF types of signals with amplitudes large enough to cause errors but not damage components simultaneously incident upon all box circuit interfaces as common mode (line-to-chassis) noise approximately as the following suggestion:

1. pulse width, pw = 0.4µs-150µs,

2. pulse repetition frequency, prf = 1kHz-200kHz,

3. RF frequencies within each pulse, 1GHz-32GHz, and

4. amplitudes large enough to induce errors but not cause damage to circuit components.

5. Inability to operate through these environments should be reported back to the circuit designers and EMI engineers in order to improve the common mode suppression/attenuation and/or to the data engineer to see if the error correction software can fix these types of unwanted signals.

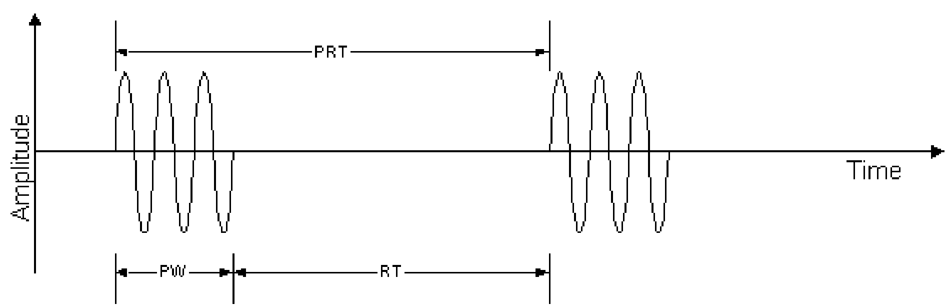


Figure 43. Simplified Radar Pulse Train

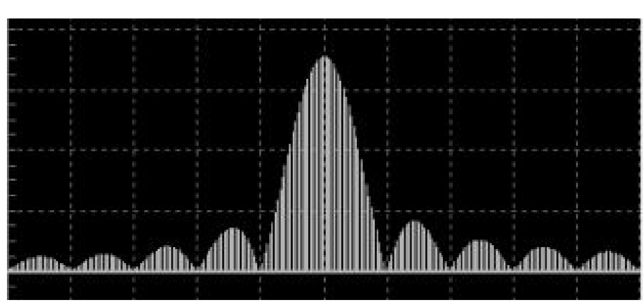


Figure 44. Spectrum (absolute magnitude) of Simplified Radar Signal Centered on the RF Frequency

11. HIRF versus Lightning

Lightning consultants and engineers are being asked about high intensity radio frequency (HIRF) environments as compared to lightning induced environments.[25] A representative set of HIRF environments are shown in Table 15.[4,22] Do not take these specific levels seriously as they are subject to change with every new revision of the requirements.

Quite simply, the peak induced voltage on a cable of height, h, above structure due to an incident electric field, E, is $V = E \cdot h$ plus resonances, $V \approx Q \cdot E \cdot h$, where Q is the cable resonant Q-factor, $Q \approx 10$. Induced currents are that voltage divided by the transmission line surge impedance, $Z_0 \approx$ 100-300Ω or $I \approx Q \cdot B \cdot A/L$, where B is the incident magnetic flux, $B = E/c$. See Figure 3. The so called “antenna mode' with the E-field incident parallel to the cable is not considered because all cables in systems are mounted close to structure; that means that the sum of the incident and reflected E-fields is about 60dB down from the incident E-field. The magnetic field is considered, however.

For an exposed cable 1” above structure and a HIRF E-field strength of 3kV/m, the induced voltage is $V = E \cdot h = 76V$ times the resonant Q-values, $V = 760V$ at the first resonant peak, and the induced current is about $I = BA/L = 750mA$ times the resonant Q-values, $I = 7.5A$ at the first resonant peak, both independent of cable length and both roll off exponentially with increasing frequency above the first resonance. HIRF induced currents can be shielded much better than low frequency broadband indirect lightning transients, i.e. a 100mΩ cable shield transfer impedance will reduce 7.5A to 750mV induced in the shielded wires, 10mΩ transfer impedance will result in 75mV induced voltage, common mode. Lightning excitation will be mostly a $V = I \cdot R$-drop in the shields, not a $V = i\omega \cdot L$ in the wires.[15] (See Figure 22.) Cable shield transfer impedances level off to about 200mΩ to 2Ω above a few tens of MHz (See Figure 22.), therefore each situation must be evaluated separately. Common mode mitigation in the I/O circuitry can reduce both EMI and lightning another 20dB or more. However, isolation plus voltage clamping is the more likely successful lightning fix for the $I \cdot R$-drop in chassis ground. One more fact: good EMC control will keep internal noise levels low so that externally generated noise has more margin.

Cable resonances rarely exceed Q ≈ 10 at lower frequencies and decrease with frequency. Likewise, HIRF high frequencies do not propagate very far down electrical cables. Therefore, in terms of cable coupling, HIRF does not compete with indirect lightning with kilo-amps of current. The worst HIRF threat to electronics is through line-of-sight coupling to RF antennas and apertures in equipment with sensitive electronics inside, e.g. burn out of RF front ends like GPS, ILS, etc. and disruption or damage to optical sensors and “glass cockpit” electronics with many large LED or plasma displays.

Lightning protection will improve the low frequency noise mitigation and HIRF protection will improve the high frequency noise mitigation. Lightning still has to work harder to obtain protection; for instance, adding a second layer of copper wire braid cable shield will improve HIRF protection by 20db while lightning protection only gets 6dB. In difficult lightning shielding situations, adding a layer of mumetal foil is needed; HIRF benefits because of the near 100% optical coverage and the lossy nature of the mumetal.

Table 15. Representative FAA HIRF Environments[4,22]

Certification HIRF Environment[22]		
Frequency	E-Field (V/M)	
	Peak	Ave.
10 kHz - 100 kHz	50	50
100 kHz - 500 kHz	50	50
500 kHz - 2 MHz	50	50
2 MHz - 30 MHz	100	100
30 MHz - 70 MHz	50	50
70 MHz - 100 MHz	50	50
100 MHz - 200 MHz	100	100
200 MHz - 400 MHz	100	100
400 MHz - 700 MHz	700	50
700 MHz - 1 GHz	700	100
1 GHz - 2 GHz	2000	200
2 GHz - 4 GHz	3000	200
4 GHz - 6 GHz	3000	200
6 GHz - 8 GHz	1000	200
8 GHz - 12 GHz	3000	300
12 GHz - 18 GHz	2000	200
18 GHz - 40 GHz	600	200

Comment: EMI and HIRF are usually specified in terms of the electric field not the magnetic field. The polarization of both are critical to coupling to cables even though EMI tests only address the E-field.

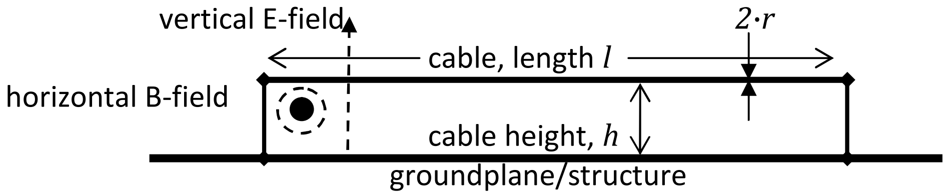


Figure 45. HIRF Field Excitation Polarization for Cable

Area, $A = h{\cdot}l$.

Inductance, $L_{cable} \cong \frac{\mu_0 \cdot l}{2\cdot\pi} \cdot \ln\left(\frac{2\cdot h}{r}\right)$

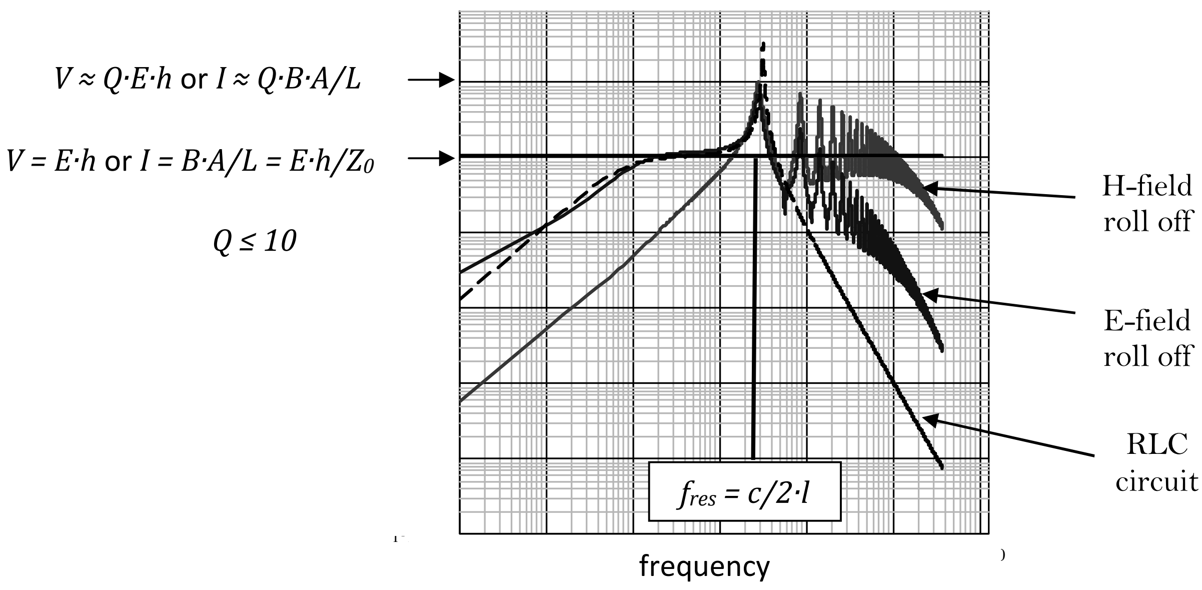


Figure 46. Example HIRF Cable Excitation Levels, Realistic RLGC Cable Parameters E-Field Perpendicular to Cable, Incident at 90⁰ (Broadside)to Cable*

(* L. West unpublished analyses.)

12. New Indirect Lightning Allocation Guidelines

The tables in DO-160F and 5412A run from WF1 on the left to WF5A on the right. We will follow that order. Pin allocations will have the same voltage with currents divided by the number of pins/wires both of which provide TVS ratings.

12.1. Inductive Coupling WF1 & WF2. WF1 currents and WF2 voltage induced environments from inductive coupling are shown in Tables 16 and 17. WF1 current and WF2 induced voltage due to cross coupling with WF4 currents on nearby cables are shown in Tables 18 and 19.

Table 16. WF1 Current from Inductive Coupling from Windows, Doors, and Wing Spars Cable 1cm above Groundplane				
Levels	**20cm window**	**2mx1m door**	**2mx2m door**	**wing spar**
1	10A (1.2m away)	10A (3.2m away)	10A (4m away)	4kA ≉ length
2	100A (40 cm away)	100A (1m away)	100A (1.3m away)	"
3	250A (23cm away)	250A (70cm away)	250A (80cm away)	"
4	600A (14cm away)	600A (43cm away)	600A (50cm away)	"
5	1.5kA (10cm away)	1.5kA (30cm away)	1.5kA (32cm away)	"
6	--	3.2kA (19cm away)	3.2kA (22cm away)	"
7	--	6.4kA (12cm away)	6.4kA (14cm away)	"

Table 17. WF2 Voltage from Inductive Coupling from Windows, Doors, and Wing Spars Cable 1cm above Groundplane				
Levels	**20cm window**	**2x1m door**	**2x2m door**	**wing spar**
1	10V (50cm away)	--	--	--
2	50V (20cm away)	50V (4m away)	50V (5m away)	--
3	125V (19cm away)	125V (1.6m away)	125V (2.4m away)	--
4	300V (15cm away)	300V (1.3m away)	300V (2m away)	314V (1m long)
5	750V (12cm away)	750V (1m away)	750V (1.2m away)	750V (2.2m long)
6	1.6kV (10cm away)	1.6kV (60cm away)	1.6kV (80cm away)	3kV (7m long)
7	--	3.2kV (52cm away)	3.2kV (56cm away)	4.5kV (10m long)
8	--	6.4kV (30cm away))	6.4kV (33cm away)	9kV (20m long)
9	--	60kV (10cm away)	60kV (12cm away)	14kV (30m long)

As applied to the WF1 and WF2 cable coupling, above, "Levels of Protection" are related to the distance of the cable from the apertures and to the cable length and height away from wing spars.

Table 18. WF1 Current from Cable Cross Coupling to 10kA WF4 Groundplane Present		
Levels	**required cable separation range: l_{12} common length = 3, 2, & 1m**	**Induced WF1**
1	d = 8-12cm	100A
2	d = 5-10cm	250A
3	d = 2-6cm	600A
4	d = 2-3cm	1.5kA
5	d = 1cm	3.2kA

Table 19. WF2 Voltage from Cable Cross Coupling to 10kA WF4 Groundplane Present		
Levels	**required cable separation range: l_{12} common length = 3, 2, & 1m**	**Induced WF2**
1	d = 7-13cm	50V
2	d = 6-12cm	125V
3	d = 3-8cm	300V
4	d = 1.3-5cm	750V
5	d = 1-2cm	1.6kV

For each Level of Protection from the DO-160F and SAE ARP5412A tables, the required cable separation distances are comparable for induced WF1 currents and WF2 voltages. These are the real Levels of Protection, cable cross coupling instead of bogus shielding attenuation; that is, the less exposed cables are excited by nearby exposed cables, NOT fields in the region allegedly penetrating from outside the system (and badly measured, also). Transport aircraft have more space to route cables; general aviation business and private aircraft have far less space.

12.2. Waveform 3 Environments

WF3 is generated by resonances on cables excited by (1) aperture coupling from external Components A, D, & H, (2) wing spar coupling from the same, and (3) cables with current from the WF4, D, & H I·R-drop that resonate at $f_3 = c/2 \cdot l_{cable}$ with amplitudes proportional to the length, summarized in Table 20. Scaling the environments is supplied below. Any WF3 excited on the CFC skin will not be treated in this note because of its complexity and small Q. See Tables 17 & 19 for first order estimated of WF3 voltages.

Table 20. WF3 Current from Windows, Doors, & Wing WF1 & Wing & Fuselage WF4 I·R-Drop											
l(m) cable	f_3	window WF1		door WF1		wing WF1		wing WF4		fuselage WF4	
		WF1	WF3	WF1	WF3	WF1	WF3	WF4	WF3	WF4	WF3
1	150MHz	14A	9mA	100A	670mA	4kA	2.7A	12kA	8A	12kA	8A
2	75MHz	"	18mA	"	1.3A	"	5.4A		16A	"	16A
3	50MHz	"	27mA	"	2A	"	8A		24A	"	24A
4	37MHz	"	36mA	"	2.7A	"	11A		32A	"	32A
5	30MHz	"	45mA	"	3.4A	"	14A		40A	"	40A
10	15MHz	"	90mA	"	6.7A	"	28A		80A	"	80A
20	7.5MHz	"	180mA	"	13A	"	56A		160A	"	160A
30	5MHz	"	270mA	"	20A	"	80A		240A	"	240A
60	2.5MHz	"	540mA	"	40A	"	160A		480A	"	480A
100	1.5MHz	"	900mA	"	67A	"	280A		800A	"	800A

Note: DO-160F WF3 currents ranges from 20A to 640A.

The WF1, WF4, and WF3 levels, above in Table 19, scale up or down as follows where r is the aperture size, d is the distance of cables from apertures, h is the height of the cables above the groundplane, and R_{system} is the radius of the system at the aperture location:

(84) $\boldsymbol{WF3 \approx WF1/4 \cdot \frac{l_{cable}}{150m}}$;

(85) $\boldsymbol{window\ WF1 \cong 14A \cdot \left(\frac{r}{20cm}\right)^3 \cdot \left(\frac{1m}{d}\right)^2 \cdot \frac{h_{cable}}{1cm} \cdot \frac{2m}{R_{system}}}$;

(86) $\boldsymbol{door\ WF1 \cong 100A \cdot \left(\frac{h_{door}}{2m}\right)^3 \cdot \left(\frac{1m}{d}\right)^2 \cdot \frac{h_{cable}}{1cm} \cdot \frac{2m}{R_{system}}}$;

(87) $\boldsymbol{wing\ WF1 \cong 4kA \cdot \frac{h_{cable}}{1cm} \cdot \frac{2m}{\pi \cdot (a+b)}}$, where a = wing chord & b = wing camber;

(88) $\boldsymbol{I \cdot R\ drop\ WF4 \cong 200kA \cdot \frac{R'_s}{n \cdot R'_s + R'_c}}$, where

(89) $\boldsymbol{R'_s \approx 100m\Omega \cdot \frac{length/10m}{width/2m \cdot thickness/2mm \cdot conductivity/10^4}}$ with no groundplane

(90) $\boldsymbol{R'_s \approx R'_{gnd} \approx 15m\Omega \cdot \frac{length/10m}{width/2m \cdot conductivity/5.8 \cdot 10^7}}$ for the groundplane

(91) $\boldsymbol{R'_c} \approx \frac{length/10m}{radius/1/2"\cdot thickness/36AWG\cdot conductivity/5.8\cdot 10^7}$ and
n = number of parallel cables.

12.3. Waveform 4 induced IR-drop Voltage and Cable Currents

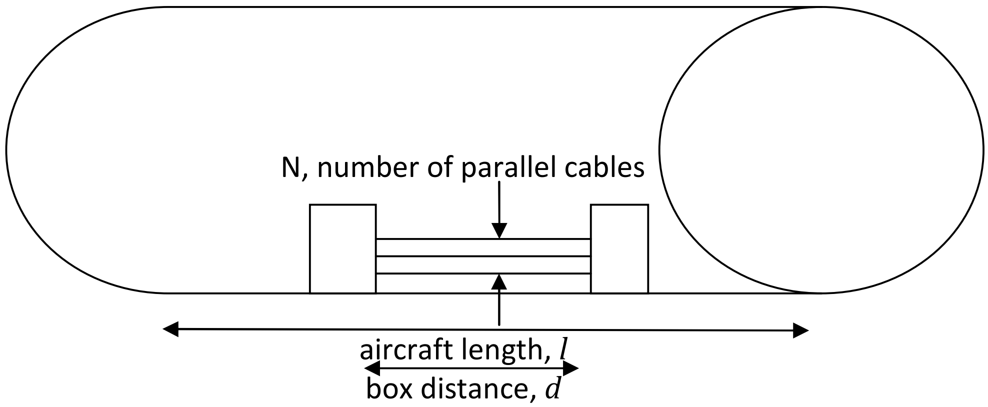


Figure47. WF4 V & I Allocations vs. Length & Number of Parallel Cables

$$\boldsymbol{V_4 = I_A} \cdot \frac{R'_s \cdot l \cdot R'_c \cdot d}{N \cdot R'_s \cdot l + R'_c \cdot d}.$$

$$\boldsymbol{I_4 = I_A} \cdot \frac{R'_s \cdot l}{N \cdot R'_s \cdot l + R'_c \cdot d}.$$

The following Tables 21-24 have derived from a 10m long 2m wide CFC system with 100mΩ end-to end DC resistance with and without a 15mΩ groundplane. Two types of cables are used, (1) high resistance ¼" TSP and (2) 1" 36AWG overbraid (OVB). <u>All of the resistances scale as the ratio of length</u>/(<u>width x thickness x conductivit</u>y).

Table 21. WF4 – V/I No Groundplane, Aircraft 10m Long, 2m Wide, TSP Cables			
Level/ Box Distance	**Number of Parallel Cables**		
	1	**10**	**100**
1m	1.6k/30k	800/12k	130/2k
2m	3.2k/30k	1.6k/12k	260/2k
3m	5k/30k	2.4k/12k	390/2k
4m	6k/30k	3.2k/12k	520/3k
5m	8k/30k	4k/12k	650/2k
6m	10k/30k	4.8k/12k	780/2k
7m	11k/30k	5.6k12k	910/2k
8m	13k/30k	6.4k/12k	1k/2k
9m	14k/30k	7k/12k	1.2k/2k
10m	16k/30k	8k/12k	1.3k/2k

Table 22. WF4 – V/I Groundplane, Aircraft 10m Long, 2m Wide, TSP Cables			
Level/ Box Distance	Number of Parallel Cables		
	1	10	100
1m	300/5k	230/4k	100/1.3k
2m	600/5k	460/4k	200/1.3k
3m	900/5k	690/4k	300/1.3k
4m	1.2k/5k	920/4k	400/1.3k
5m	1.5k/5k	1.65k/4k	500/1.3k
6m	1.8k/5k	1,38k/4k	600/1.3k
7m	2.1/5k	1.6k/4k	700/1.3k
8m	2.4k/5k	1.8k/4k	800/1.3k
9m	2.7k/5k	2.1k/4k	900/1.3k
10m	3k/5k	2.3k/4k	1k/1.3k

Table 23. WF4 – V/I No Groundplane, Aircraft 10m Long, 2m Wide, OVB Cables			
Level/ Box Distance	Number of Parallel Cables		
	1	10	100
1m	1k/100k	200/20k	20/2k
2m	2k/100k	400/20k	40/2k
3m	3k/100k	600/20k	60/2k
4m	4k/100k	800/20k	80/2k
5m	5k/100k	1k/20k	100/2k
6m	6k/100k	1.2k/20k	120/2k
7m	7k/100k	1.4k/20k	140/2k
8m	8k/100k	1.6k/20k	160/2k
9m	9k/100k	1.8k/20k	180/2k
10m	10k/100k	2k/20k	200/2k

Table 24. WF4 – V/I Groundplane, Aircraft 10m Long, 2m Wide, OVB Cables			
Level/ Box Distance	Number of Parallel Cables		
	1	10	100
1m	250/27k	120/12k	20/2k
2m	500/27k	240/12k	40/2k
3m	750/27k	360/12k	60/2k
4m	1k/27k	480/12k	80/2k
5m	1.25k/27k	600/12k	100/2k
6m	1.5k/27k	720/12k	120/2k
7m	1.75k/27k	840/12k	160/2k
8m	2k/27k	960/12k	170/2k
9m	2.25k/27k	1.1k/12k	180/2k
10m	2.5k/27k	1.2k/12k	200/2k

13. Lightning Protection and Allocation

Indirect lightning protection in composite airframes consists of four features:

groundplane
isolation
cable shielding
surge suppression
current-limiting resistors
software error mitigation

13.1. Groundplane(s)

Groundplanes are installed in CFC aircraft for reasons of current return, fault protection, and common electrical reference. We need to milk every reduction in the resistance as possible. We do that in two ways:

groundplane segments' resistance per square
bond joints and ground straps between groundplane segments

Bond joints can contribute as much as 2/3 of the end-to-end resistance of a many-segmented groundplane. In the simple example in Figure 48, 28 series and parallel 2.5mΩ bond joints **alone** contribute 1.5kV I·R voltage drop nose-to-tail and nose-to-wing tip.

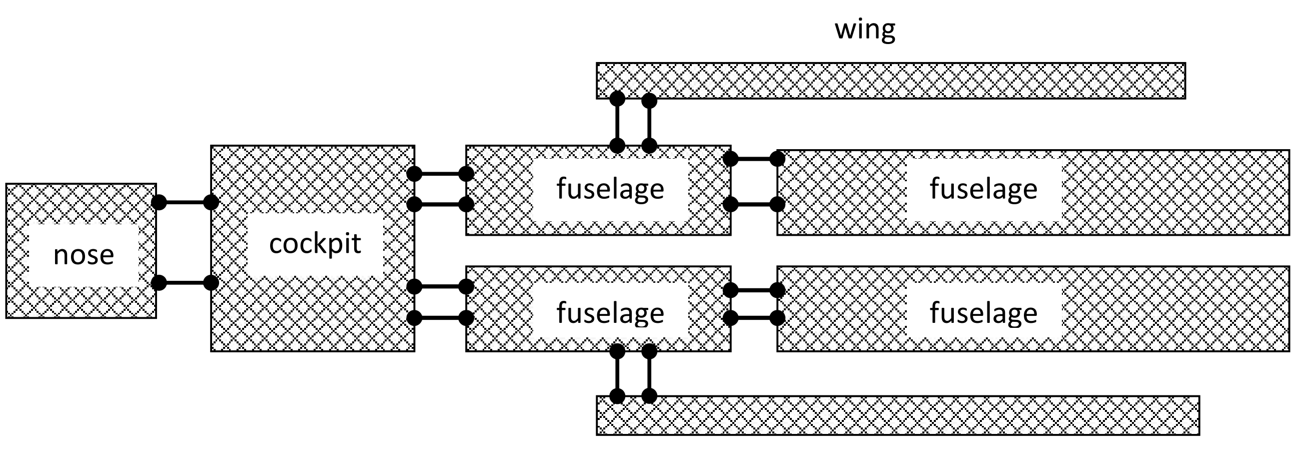


Figure 48. Notional Groundplane in Segments with Bonding Jumpers

13.2. Impedance of a Mesh Groundplane

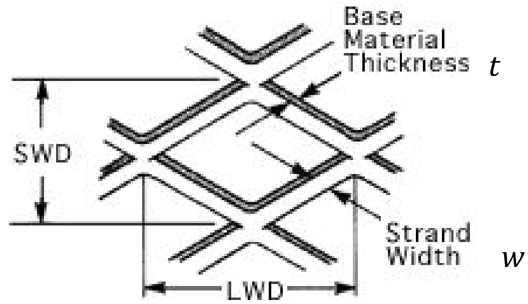


Figure 49. Expanded Foil Mesh Close Up with Geometric Parameters[42]

For exterior direct strike protection: The metal mesh products of aluminum (Dexmet Product code 4AL8-080F) and copper (Dexmet Product Code 2CU4-100A) both have 78g/m^2 (0.016lb/ft2) density, which resulted in a thickness of approximately 0.10mm (0.004in). See Table 25.

Table 25. Dexmet Example Mesh Parameters[42]	
Dexmet Product Code	**2CU4-100A**
Weight – LBS/SF (±10%)	0.015
Weight – GMS/SM (±10%)	73.3
Base Metal Thickness (±10%)	0.002 inch 0.051 mm
LWD (±5%)	0.100 inch 2.54 mm
Overall Thickness (±.001 inch/±.025 mm)	0.004 inch 0.102 mm
Open Area (±5%)	84%

Example: 3 Ni 5-077[42]
First number represents nominal original thickness 3 = .003" (.076 mm)
Letters are chemical symbol for material Ni = Nickel
Number immediately following letters represents strand width 5 = .005" (.127 mm)
Last number indicates the long dimension of the diamond 077 = .077" (1.96 mm)

To make the groundplane expanded foil mesh resistance less than cable & airframe 80mΩ, say R(gndplane) = 20mΩ, then with the mesh groundplane segment parameters, R_{sq}, resistance per square, length, & width → $R_{gnd} = R_{sq} \cdot l/w$, (1) 1mΩ/square & l = 10m requires a width = 50cm and (2) 100mΩ/square & l = 10m requires a width = 5cm

13.3. Bond joints

A typical set of DC resistance values for individual bond joints is tabulated in Table 26, below.

Table 26. Typical DC Resistance Requirements for Electrical bond Joints
(1 square inch faying surface

Bond Joint Materials	**DC Resistance per Joint**
Aluminum-to-aluminum joints (use MIL-C-5541E, class type 3 coating for Class R/L bonds. Avoid disallowed types of Alodyne (use 600 not 1200) and the corresponding Iridite.)	2.5 milliohms
Aluminum-to-stainless steel joints	0.1 ohm
Aluminum-to-titanium joints	0.1 ohm
Aluminum-to-nickel joints	5 milliohms
Nickel -to- nickel joints	5 milliohms
Stainless steel-to-stainless steel joints	0.1 ohm
Aluminum-to-graphite joints – disallowed/insulate	~~1 ohm~~ ∞
Electronic boxes to structure (system reference groundplane)	2.5 milliohms
Large area metal-to-metal structural surfaces	2.5 milliohms
Non-electronic boxes to structure	25 milliohms
Across hinges (flight control surfaces, access covers)	25 milliohms
Harness shield to structure	2.5 milliohms
Mechanical equipment to structure	100 milliohms
Bonding CFC composite to composite components	20 ohms

A four probe tester such as a Biddle 247000 Digital Low Resistance Ohmmeter Dual-Pak General-Purpose Model should be used to verify any coupons not previously tested.[45]

A Loop Resistance Tester (LRT) is needed to make bond joint resistance measurements in situ.

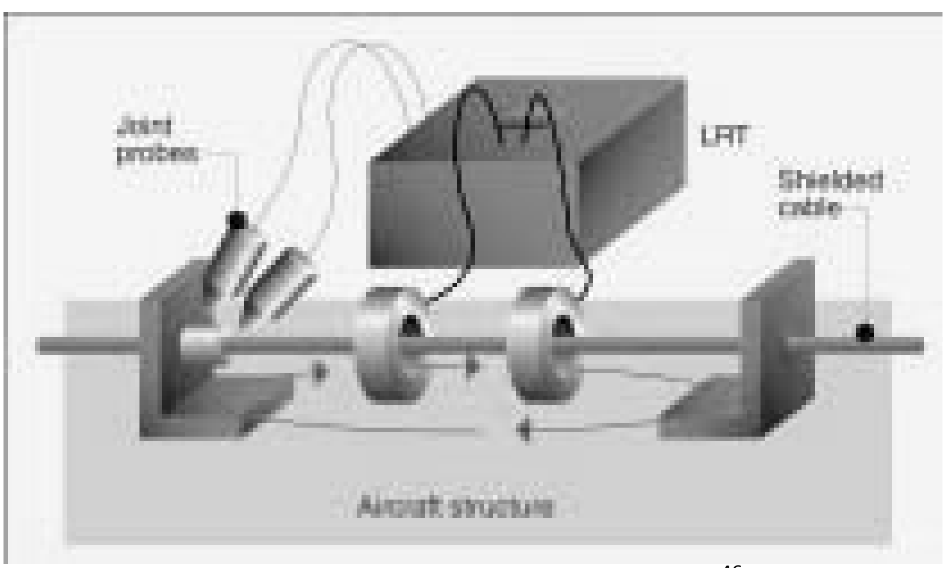


Figure 49. Loop Resistance Tester (LRT)[46]
(non-invasive, used by commercial aircraft companies)

Riveted joints make up most of the aluminum-to-aluminum joints.

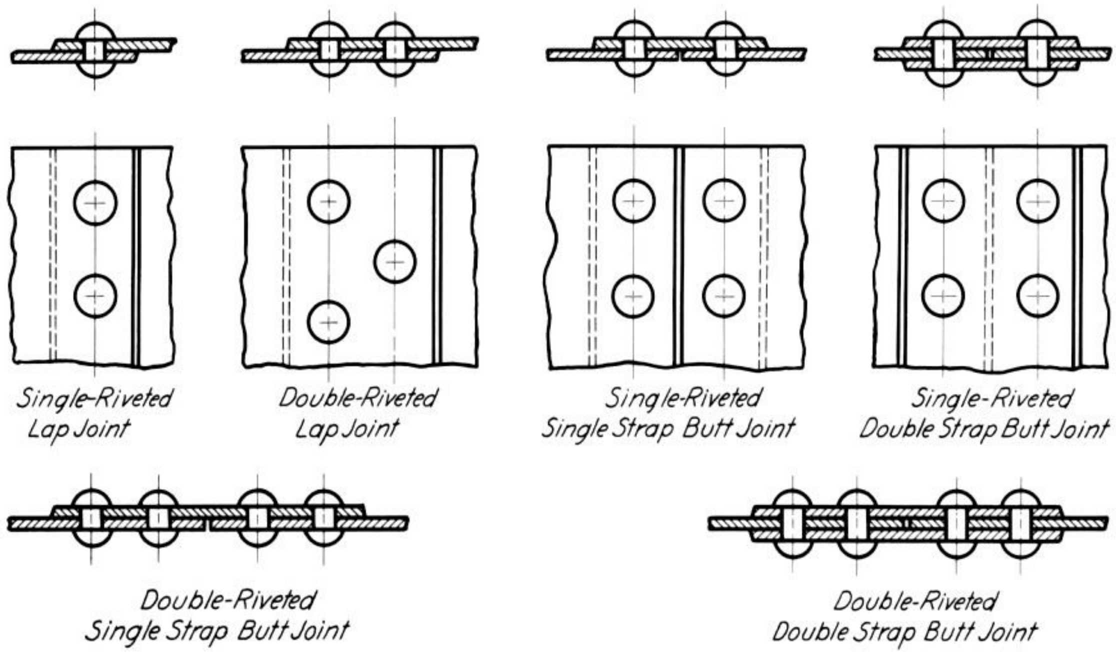


Figure 50. Riveted Joints

13.4. Ground Straps. When good metal-to-metal joints cannot be made (usually because of anodized fittings) or for moveable joints, ground straps are used.

Figure 51. Different Ground Straps

All ground straps resonate. The 4th one over from the left will resonate at between 80MHz and 120MHz. Placing one or more on flight control surfaces can introduce an 80-100MHZ EMI susceptibility. [This author had to fix one such "fix" by removing the ground strap and replacing a nonconducting gasket with a conducting one; that killed the 100MHz susceptibility and increased the margin by 60dB, the decrease in the inductance.]

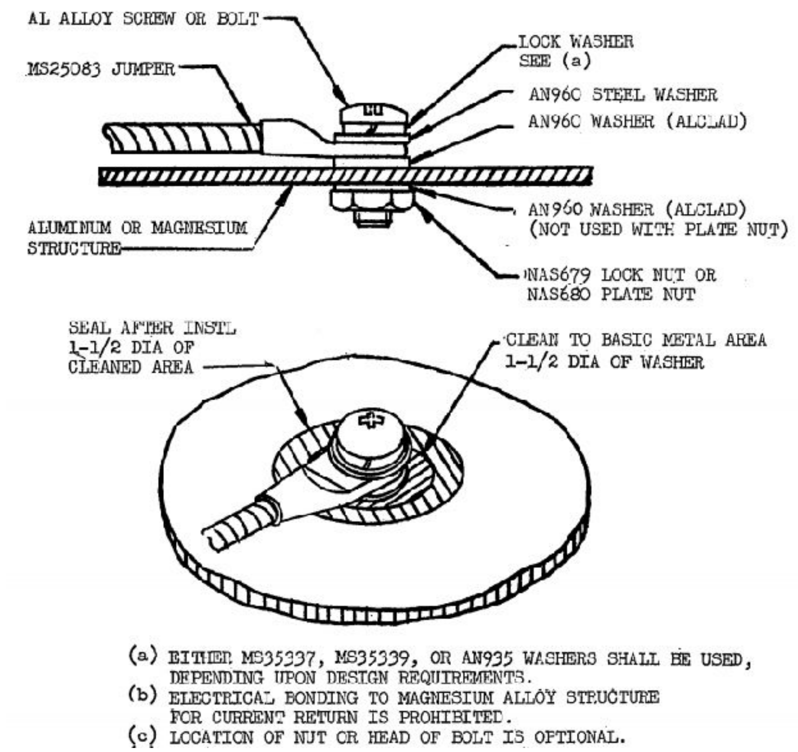


Figure 52. Groundplane Bonding Jumper Joint from MIL-B-5087B

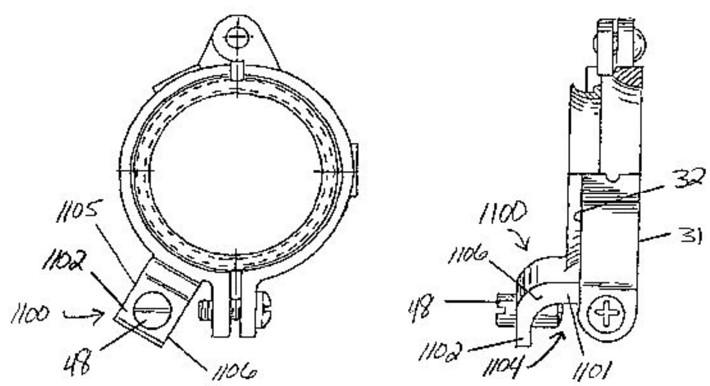


Figure 53. Tube Clamp with a Lug for Fastening a Ground Strap

Figure 54. Less Robust Static Ground Strap Connection to a Tube

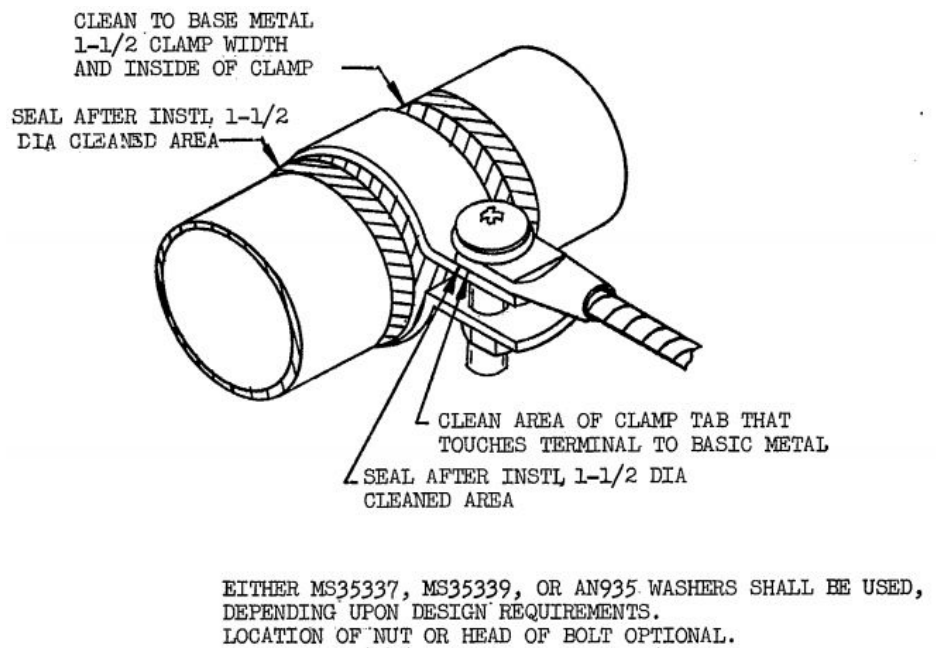


Figure 55. MIL-B-5087 Depiction of a Ground Strap across a Joint in a Tube

13.5. Cable Shielding. Cable shields are used to attenuate electromagnetic energy from wires and their I/O circuits. The standard model for wire shielding is the coaxial model, below. More complex circuit configurations will be addressed later. Transfer impedance is defined in terms of the ratio of the induced voltage across a load to the current on the shield.

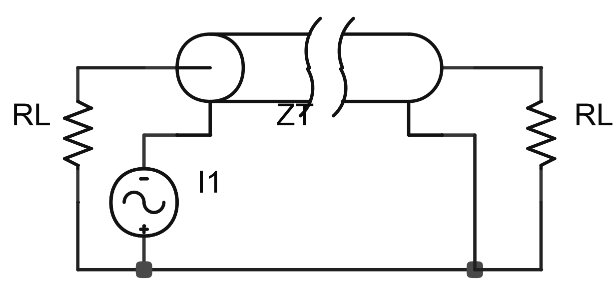


Figure 56. Simplified Coaxial Shielded Wire/Cable Example

The transfer impedance of the shield is written simplistically as:

(92) $$Z_T \cong (R'_{dc} + i\omega \cdot L'_T) \cdot l$$

where R'_{dc} is the DC resistance of the shield and L'_T is the transfer inductance, both per unit length denoted by the prime on the parameters.

The voltages induced in the shielded wire and the shield itself is illustrated in the circuit schematic below. These are usually both placed in series in the wire but that is wrong and will be explained later.

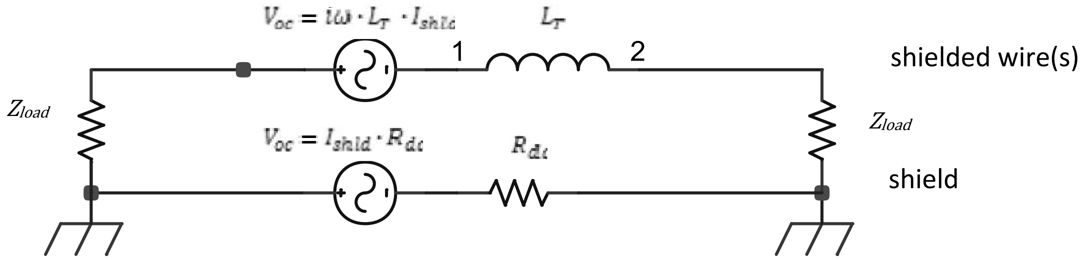


Figure 57. Circuit Schematic of Induced Voltages induced within a Shielded Wire
Sources of Common Mode EMI induced through a Cable Shield
(1) Ground Potential Difference, $V_{oc} = I_{shield} \cdot R_{dc}$
(2) Inductive Coupling to Wiring, $V_{oc} = i\omega \cdot I_{shield} \cdot L_T$

When the two loads have the same impedance and ignoring the polarity differences end-to-end, the induced voltage divides equally across both loads.

(93) $$V_{load} \cong \frac{Z_T \cdot l \cdot I_{shield}}{2}$$

A typical range of braid cable shield transfer impedances are 4m/m to 70m/m DC resistance and 1pH/m to 1nH/m transfer inductance.

The transfer inductance, L_T, will be demonstrated here for a single aperture/hole of radius 'a' in a solid coaxial shield of radius *'b'* as shown below.[6]

(94) $$L_T \cong \frac{\mu_0}{(2 \cdot \pi \cdot b)^2} \cdot \frac{4 \cdot a^3}{3}$$

2. Empirical Background
From Raychem Memo, April 15, 1992, To; Neal Enault, From: Robert Moore, Subject: Surface Transfer Impedance: Flat Braids vs. Round Braids (data on TSPs):[40]

(1) TSP diameter = 0.240"
braid wire gauge = 38AWG = 4 mil
flat braid thickness = 1.5 mil
wire braid DC resistance, R'_{dc} = 15.6mΩ/m
flat braid DC resistance, R'_{dc} = 30.4mΩ/m
wire braid transfer inductance, derived from data, L'_T ≈ 264 pH/m
flat braid transfer inductance, derived from data, L'_T ≈ 477 pH/m

(2) TSP Raychem part numbers, 55A2221-22 (flat) and 55A1221-22 (round), diameter = 0.077"
(the -22 denotes 22AWG signal wires within the braid)
(cables made per MIL-C-27500)
braid wire gauge = 38AWG = 4 mil
flat braid thickness = 1.5 mil
wire braid DC resistance, R'_{dc} = 42.3mΩ/m
flat braid DC resistance, R'_{dc} = 70.8mΩ/m

wire braid transfer inductance, derived from data, $L'_T \approx 67$ pH/m
flat braid transfer inductance, derived from data, $L'_T \approx 37$ pH/m

MIL-C-38999 connectors have Z_T of less than 1mΩ DC resistance and 0.5pH to 5pH transfer inductance. However, that is not the end of story; the joints from the connector to the shield and to the box are worse than that. Figure on 2.5mΩ to 10mΩ and then go measure some ASAP.

Composite connectors and cable shields all have higher DC resistance but good transfer inductance. It's been reported that small composite connectors have problems with high lightning currents.

Take it from an old timer; don't be afraid of grounding wire and cable shields with pigtails. They work far better than anyone thinks, just not in a transfer impedance fixture. Put them in a test with EM fields and they work very well. The pigtail inductance current limits, i.e. $I_{shld} \cong B_{field} \cdot A_{shld\ circuit}/L_{pigtail}$, in an exposure to fields, not in a constant current transfer impedance test. A modern day pigtail technique is depicted in Figure 58. Figure 60, below, is the old time method.

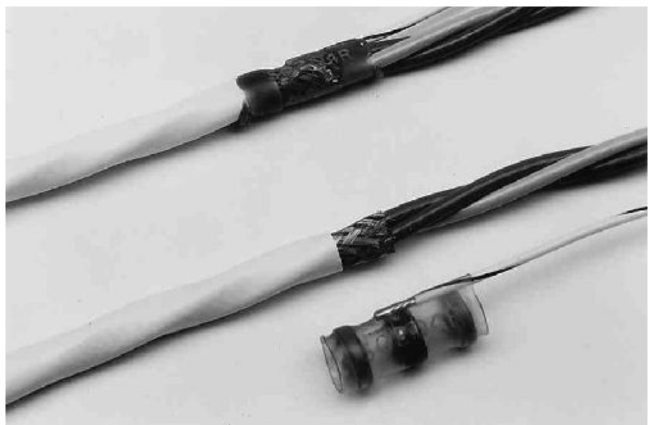


Figure 58. Tycoelectronics/Raychem Solder Sleeve Shield Terminators TSP Pigtail Wire Connection[59]

Figure 59 shows the pigtail effects in a transfer impedance fixture versus exposure to fields. The pigtail has to be ridiculously long before the effects are the same. For a 3" pigtail in the example shown, the transfer impedance of the cable shield exposed to EM fields is better than the control model and about 20dB better than in the Z_T fixture. The control cable shield has only the transfer impedance of the braid shield with zero impedance grounding at both ends. The reason for this blasphemous result is that pigtails' higher impedance current-limits the shield current in the fields' exposure thereby giving the appearance of a better shield which it isn't, e.g. $I_{shld} = B \cdot A/L$. (Reference unpublished analysis of L. West)

▬

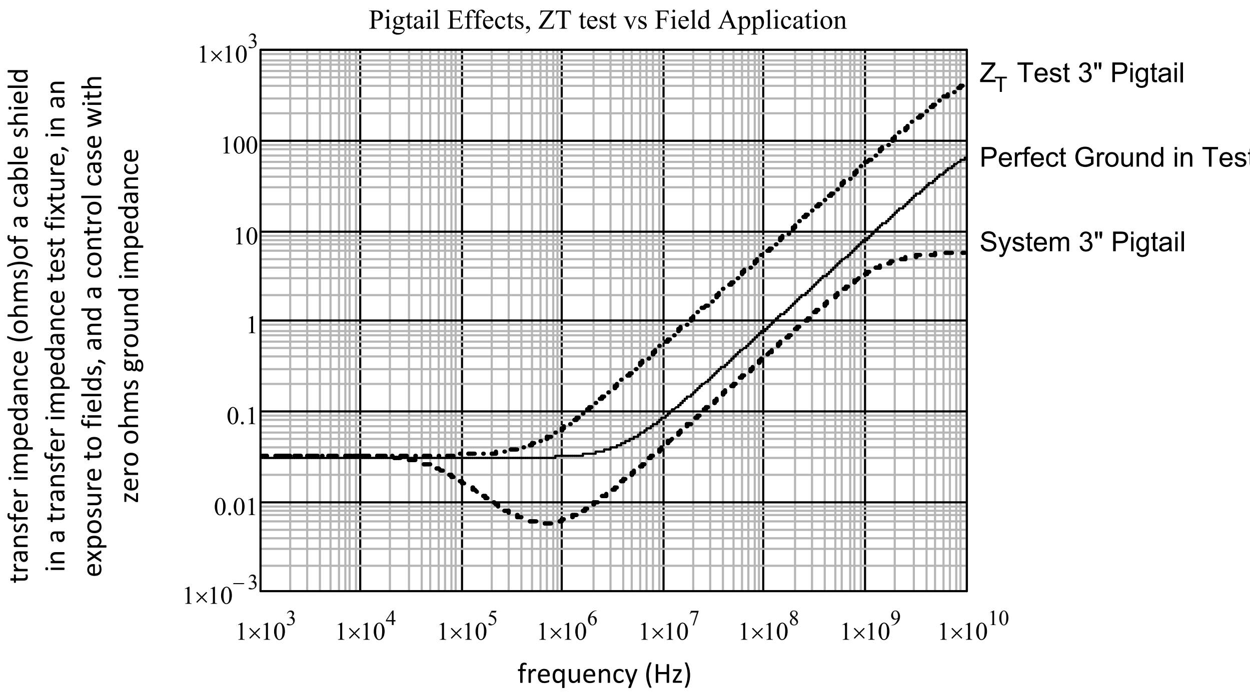


Figure 59. 3” Pigtail Effects in Transfer Impedance (Z_T) Fixture vs. Exposed to Fields (System)

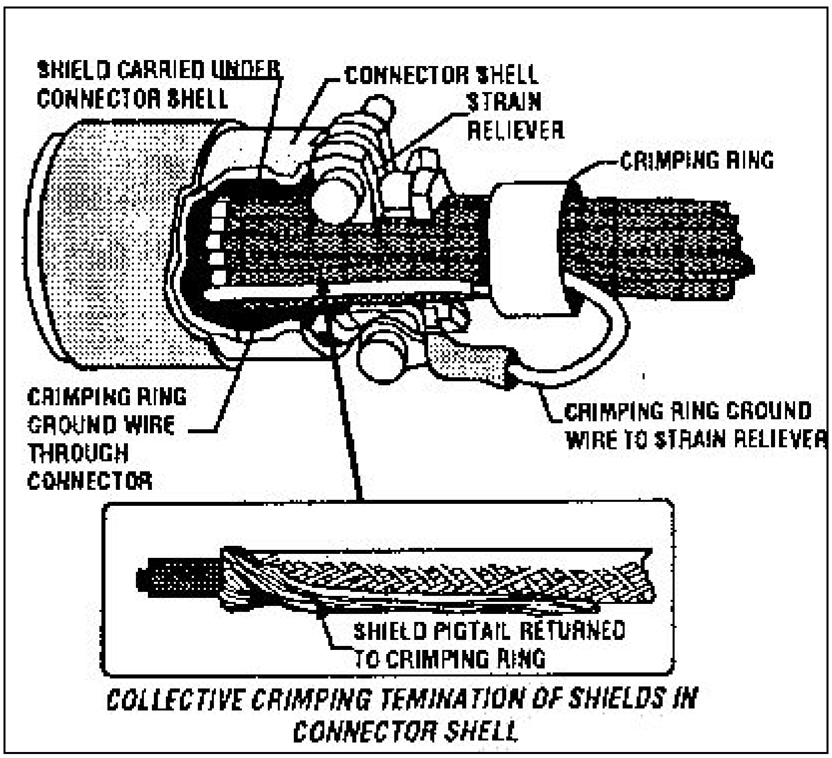


Figure 60. Old Time Cable/Wire Shield Ground Termination by Pigtail to Strain Relief (When there is no room for an EMI Backshell)

13.6. Cable Separation. Isolation through cable separation is necessary to keep high current cables from inducing noise into low voltage cables. With cables of length, l, a distance, d, apart and a height, h, above a groundplane, the mutual coupling is

(95) $\boldsymbol{M_{12}} = \frac{\mu_0 \cdot l}{2 \cdot \pi} \cdot \ln\left[\frac{d^2+(2 \cdot h-r)^2}{d^2+r^2}\right]$ with the cable radius, r, included and

(96) $\boldsymbol{M_{12}} \cong \frac{\mu_0 \cdot l}{2 \cdot \pi} \cdot \ln\left[1+\left(\frac{2 \cdot h}{d}\right)^2\right]$ with the cable radius, r, set to zero.

Mutual coupling of WF4 currents into WH1 currents and WF2 voltages in adjacent cables is depicted in Figures 34 and 35.

13.10. Diode Voltage Clamps/Transient Voltage Suppressor (TVS) Devices. There are two types of shunting devices to choose from. The least expensive type is single stage, which usually consists of a single TVS device on each line. Three stage devices are also available. The first stage of a three-stage device is a gas discharge tube, which can handle extremely high currents, but has a high threshold voltage and is too slow to protect solid state circuits. The second stage is a small series impedance which limits current and creates a voltage drop between the first and third stage and, more importantly, allows the gas tube to fire first so as not to burn out the diode clamp. The final stage is a TVS device that is fast enough to protect solid state devices and brings the clamping voltage down to a safe level for data circuits.[39]

ARINC 429 operate from 5V to 15V. I/O impedances are relatively high, 75Ω on line drivers and 12kΩ on line receivers plus line matching resistors.

In order to lower the TVS capacitance across the data lines, diodes are ganged in series.

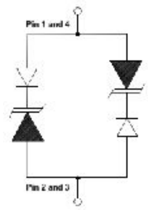


Figure 61. Modern Transient Voltage Suppressor with Series Diodes to reduce Capacitance[53,57,58]

13.11. MOV. Multi Oxide Varsistors (MOV) devices are discouraged particularly on power lines because they tend to fail in a short circuit mode under normal transient traffic. [That's why plug-in devices for home systems have a fuse in series with the MOV – cheap, effective, and replaceable; not so in-flight.]

13.12. Gas Surge Arrestors. Gas surge arrestors/gas tubes are used for shunting high currents.

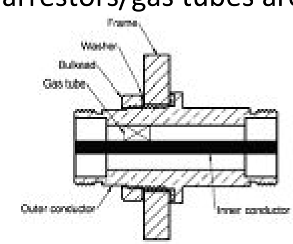


Figure 62. Gas Tube Surge Arrestor[50]

20kA-40kA, 15V pass through, lower power RF, 20-300W

13.13. Quarter-Wave Shorted Stubs. Quarter-wave stub is also used for high current and voltage surge suppression on RF lines.

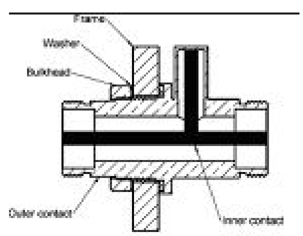


Figure 63. Quarter-Wave Shorted Stub[50]

15V pass through, RF power up to 1200W

A quarter-wave stub is not transparent to broadband PM signals. The phase goes through a monotonic transition from +π to -π over the stub's bandwidth which is usually much larger than the RF bandwidth. The further away from the antenna, the more complicated the phase change.

(97) $Z_{stub} = Z_0 \cdot \frac{1-e^{-i \cdot 2 \cdot \beta \cdot l}}{1+e^{-i \cdot 2 \cdot \beta \cdot l}}$

13.14. Vendor Ideas, MAXIM. MAXIM has some thoughts about our problem among others.[47]
When cables are long (as RS-485 data cables can be), the originating signal's common or ground may not have the same electrical potential as that of the receiving location. The RS-485 specification says to connect the drive-circuit common to frame ground, either directly or through a 100Ω resistor depicted in Figure 63.

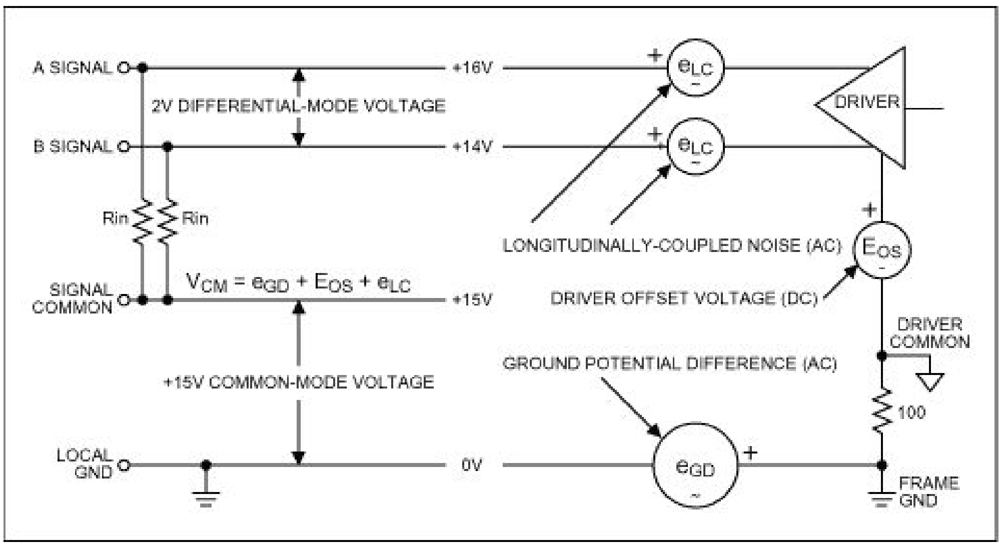


Figure 64. Three types of common-mode signal (e_{GD}, e_{LC}, and E_{OS}) present in a 2-wire data-transmission system.

Signal common can assume a common-mode voltage equal to the vector sum of the ground-potential difference, the driver-offset voltage, and any longitudinally coupled noise voltage generated along the signal path between transmitter and receiver:

(98) $V_{cm} \cong e_{gd} + E_{os} + e_{LC}$

The Origins of Common-Mode Signals:
Three sources of common-mode voltage are represented in Figure 59 as e_{GD}, e_{LC}, and E_{OS}:
E_{OS} is typically a DC offset introduced by a differential-mode driver operating from a single supply.
e_{GD} is a noise signal representing the difference in ground potentials at the transmitting and receiving locations. It is usually an AC signal containing the fundamental and possibly several harmonics of the power-line frequency.
e_{LC} is a longitudinally coupled noise signal occurring equally on both transmission lines due to inductive coupling from extraneous sources.

Minimizing Common-Mode Signals:

E_{OS} can be made quite small, or even zero, by operating a differential-mode driver from balanced supplies. In contrast, e_{GD} can be minimized only by maintaining a relatively short distance between the transmitting and receiving locations. e_{LC} can be minimized by using a shielded twisted pair: noise introduced within the cable arises equally on each of two tightly twisted wires. Otherwise, a normal-mode signal would be present due to the line's asymmetry with respect to disturbing fields. The load must also be symmetrical; the resistive and capacitive load impedances on both lines of the twisted pair must be matched. Inductively coupled signals can be prevented only by using magnetic shielding. (Author's note: The term "magnetic shielding" means that the shield is grounded to chassis ground at both ends. Forget the erroneous motherhood about ground loops.) Note that any wire carrying signal current is a source of magnetic radiation.

Suppressing Common-Mode Signals:
Common-mode signals (V_{CM}) must be rejected in the receiving circuit. That rejection is easily accomplished when the receiving circuit is passive (headphones or loudspeaker), transformer coupled, isolated and battery operated, or otherwise not referenced in any way to the transmitting-circuit common (either capacitively or resistively connected). The configurations noted here are inherently immune to common-mode signals, but receiving circuits referenced to the transmitting-circuit common must be designed to accept the full range of V_{CM} presented to them. All such designs involve differential receivers with high common-mode-rejection (CMR). If the V_{CM} is of relatively low amplitude, a high-CMR receiver alone may be adequate.

13.15. Vendor Ideas, B&B Electronics. RS422/485 Application Note, Chapter 4: Transient Protection of RS-422 and RS-485 Systems:[37]
Connecting Signal Grounds**:** Since a local ground connection is required at each node implementing shunt type protection, the consequences of connecting remote grounds together must be considered. During transient events a high voltage potential may exist between the remote grounds. Only the impedance in the wire connecting the grounds limits the current that results from this voltage potential. The RS-422 and RS-485 specification both recommend using 100 ohm resistors in series with the signal ground path in order to limit ground currents. Figure 65 illustrates the ground connection recommended in the specification.

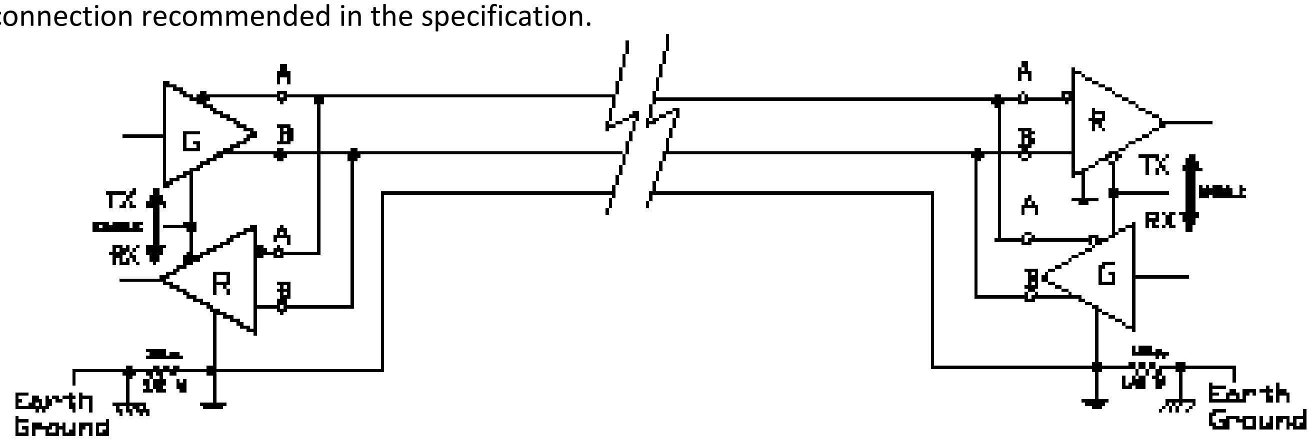


Figure 65. Signal Ground Connection between two nodes with 100 ohm resistor

(Author's note: The 100Ω resistor to ground has crept into many applications. This author has seen them burnt up because they weren't rated for the real ground potential. Jet Propulsion Lab (JPL) has two Mars Rovers operating on the Mars surface with floating grounds because they, too, underestimated the ground potential surge and blew their 100Ω resistors-to-chassis-ground. Ethernet designers have capacitors in series with resistors to chassis ground with the capacitors rated at 2kV. The ½ watt rating on the above resistors to chassis ground in Figure 65 demonstrates a lack of experience with large ground potentials in premise wiring and aerospace vehicles.)

Isolation Devices**:** Optical isolation can be implemented in a number of ways. If a conversion from RS-232 to RS-422 or RS-485 is being made, optically isolated converters are available. Optically isolated ISA bus serial cards can replace existing ports in PC systems. For systems with existing RS-422 or RS-485 ports, an optically isolated repeater can be installed.[54]

B&B Electronics had the ground potential I·R-drop in mind when they designed the I/O circuit in Figure 66, below.[37] This is more damage protection than upset but the design has merit if mode conversion can be controlled. When the TVSs clamp, the data stream will be held at the clamping voltage. We recommend series current-limiting resistors in Figure 67.

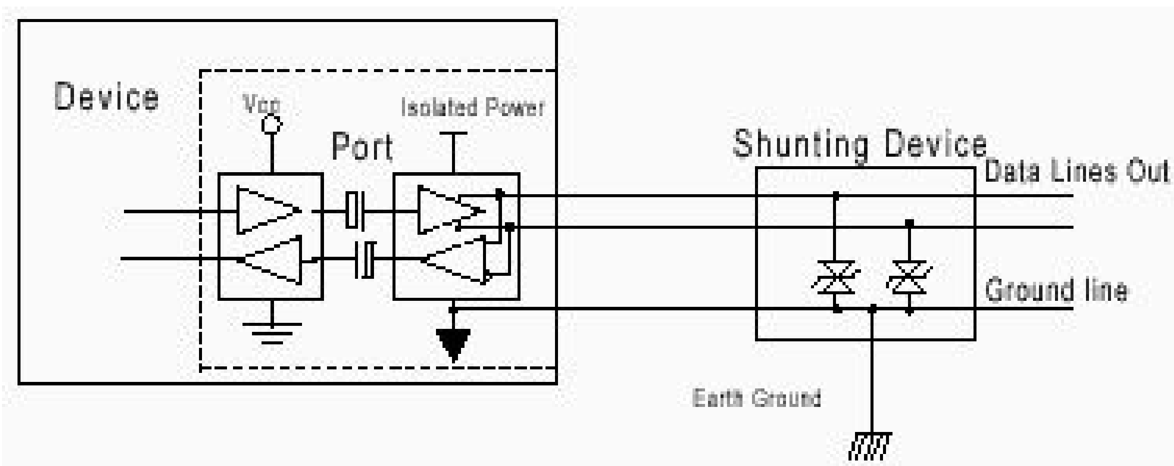


Figure 66. Isolated Node with Shunt Protection to Earth/Chassis Ground[37]
Author's note: The ground line and different ground symbols are somewhat amusing.

13.16. Fault Tolerant Architecture and Design. Design and software requirements should attenuate common mode indirect lightning transients and correct, and/or ignore the "common cause" lightning induced bit errors and/or burst errors in digital systems without causing errant commands. Analog sensors and circuits and their electronics should mitigate and/or ignore the same error signals without causing errant commands. Modern avionics and software are designed for requirements of (1) fail operational, (2) fail passive, (3) fault tolerant, and (4) distributed fault-tolerant systems. One result is that a noise signal on one or more sensors cannot throw the aircraft into an unsafe flight condition; multiple sensors in multiple locations are regularly sampled and compared before any command is given other than fail operational or fail-passive.

<u>It is literally impossible to apply enough shielding, suppression, isolation, filtering, etc. to reduce lightning transients below modern low voltage circuit noise tolerance levels</u>. Indeed, the better high speed data links such as Ethernet have error detection and correction in order to operate in their own copper media noise. We offer the following example quote about bit error mitigation to show the reader the sophistication of this software technology:[49]

Modern high speed digital data is encoded in order to reduce bit error rate, e.g. HDTV. Forward Error Correction (FEC) eliminates the need to resend messages to a degree. Reed-Solomon codes are block-based error correcting codes which are particularly good at correcting bursts (sequences) of bit errors, the type resulting from EMI and lightning. 10/100 Mbps Ethernet with 32 bit frame check sequence (CRC-32) does a cyclic redundancy check, detects 99.9999977% of all errors, and detects all burst errors of length 31 bits or less.

The standards and this note mostly address damage susceptibility to the negative Component A current. Every time a voltage limiter clamps the line voltage for lightning, a stream of bit errors can follow as long as 88µs to 1.5s. Functional upset is discussed herein not so much in terms of the levels of specific waveforms[1,2], as above and in references 1 and 2, but as design and software requirements for mitigating, correcting, and/or ignoring lightning induced "common cause" bit errors and/or burst errors in digital systems without causing erroneous automated commands. Similar recommendations are made regarding lightning induced noise on analog sensors and circuits and their electronics' capability to mitigate it or ignore it without causing errant commands. At least one launch rocket and payload was lost after launch at KSC because of one bit error that occurred late in a series of multiple lightning strikes.

Common cause failure analysis is important in reliability and safety studies, as common cause failures often dominate random hardware failures. Systems affected by common cause failures are systems in which two or more events have the potential of occurring due to the same cause. Environmental causes such as EMI and lightning effects fall into this category of failure modes.[51] Almost all the existing reliability assessment methods either fail to consider common cause failures (CCF) or fail to consider imperfect fault coverage (IPC). Both cases result in exaggerated system reliability.[52] The avionics and software designers must be aware of the type and duration of lightning induced common cause burst errors in order to design appropriate fault tolerance and recovery.

Lightning and EMI are in that class of failures known as "common cause failures", meaning that they can induce signal errors or noise-induced failure modes everywhere in the system simultaneously. These, of course, are not considered explicitly in the above architectural schemes although the architecture helps. We have to put in requirements that lightning and EMI analog errors, digital bit errors, and digital burst errors (that are assumed to be at sufficient levels to cause errors follow the prescribed repetitious environments) will not result in unsafe commands from automated control systems. This requires redundant (not identical) sensors in different locations, repetitively sampled, compared, and processors capable of correcting, tolerating, or circumventing the environments for their prescribed time periods. Make no mistake; all of the above is complicated although different avionics companies are building up experience. (The Hawker 4000 with the Honeywell Primus Epic Avionics Suite is the first aerospace system this author has observed with close to these capabilities, i.e. "smart electronics" that cannot be easily fooled by electrical noise. The Rockwell Collins Pro Line avionics has similar fault tolerant architecture. Until then, I spent 34 years diagnosing and fixing such errors that affected safety, flight control, accuracy, reliability, nuclear survivability, and nuclear safety.)

13.17. Problem: Ground Potential Lightning-Induced I·R-Drop

The notional circuit in Figure 67 depicts more clearly why the ground potential lightning I·R-drop is a box-level problem, not just an I/O circuit problem. Every chip is affected. The -0.5V substrates are viewed as the weakest component. Every chip's max ratings on V_{cc} and GND involve the -0.3V to -0.5V substrate, e.g. -0.3V < V_{cc} < 7V and the same and all other pins except the I/O pins with ±29V.

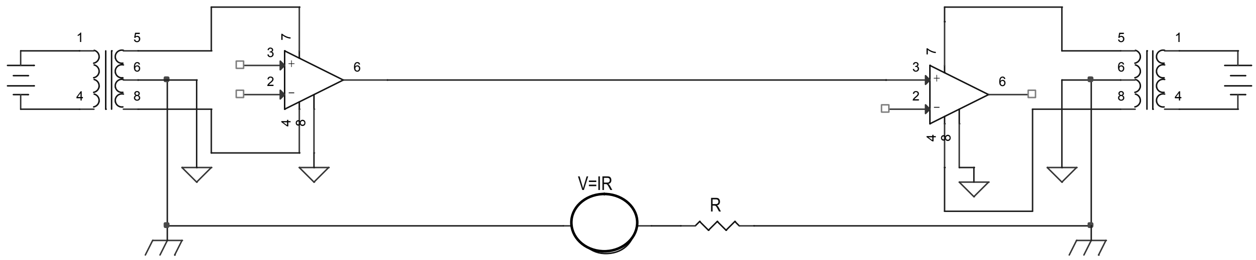


Figure 67. Ground Potential Lightning-Induced I·R-Drop
WF3 falls into this mode for frequencies ≤ 10MHz.

Circuit I/O Fix: Current-limiting + TVS +Isolation + Slow Data Rates

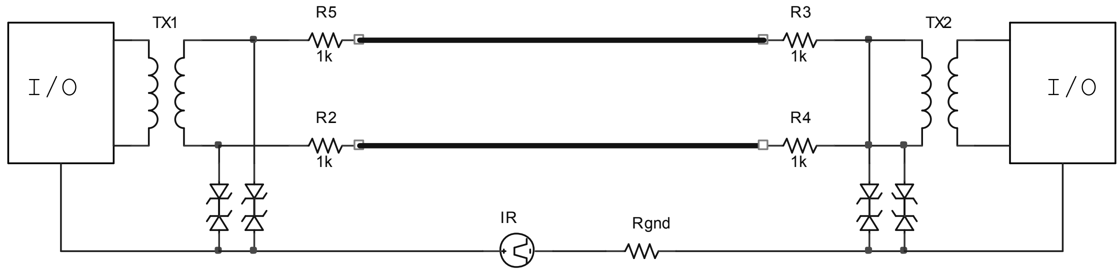


Figure 68. I·R-Drop I/O Protection
OK for Low Frequency ARINC 429; questionable on ARINC 664

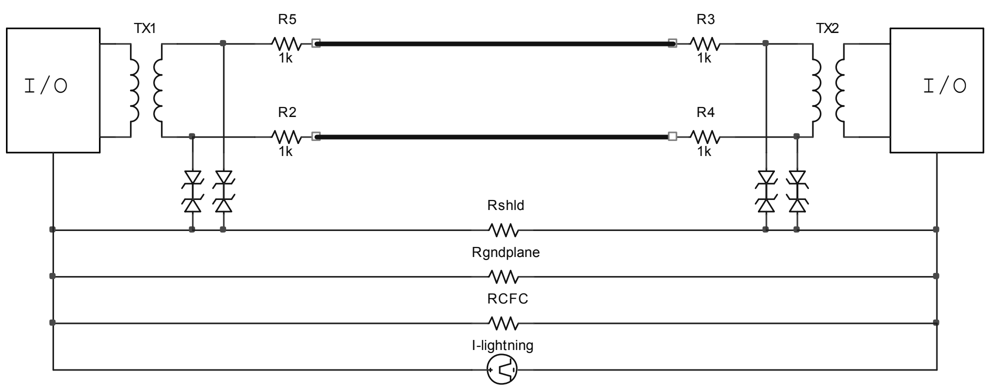


Figure 69. Simplified System I·R-Drop Model with I/O Fix

Isolation components are not 100% effective. In particular at the high voltages in lightning surges, they are voltage dividers between the isolator, C_i, and the load impedance, Z_L.

Example: (a) Z_L = 1kΩ, C_i = 5pF, R_s = 1kΩ → $V_L \approx$ 1mV

(b) Z_L = 100pF. C_i = 5pF, R_s = 1kΩ → $V_L \approx$ 10mV

Note: Pulse power rating of resistors is about x10-20 the DC power rating.

I/O Circuit(s): Spec the largest available and practical CMR on line drivers and receivers.

Isolation: All isolators considered for combating the lightning I·R-drop should have a minimum voltage standoff rating of 1kV.

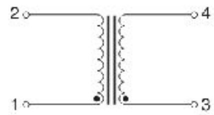


Figure 70. Isolation Transformer[X36]

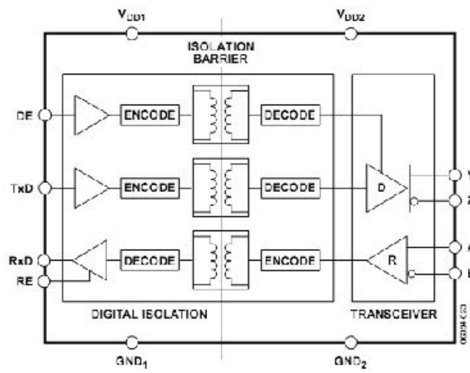


Figure 71. Isolation Transformers within Chips[55]

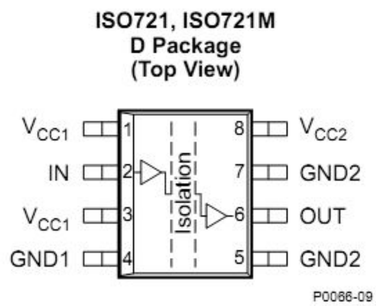


Figure 72. Typical Capacitive Isolation[56]

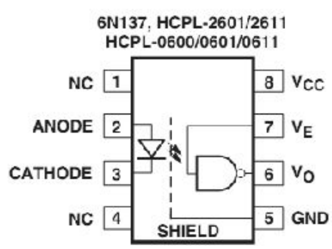


Figure 73. Photo-Transistor/Photo-Diode Isolation[53,54]

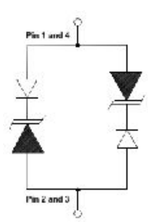


Figure 74. Typical Transient Voltage Suppressor (TVS)[53,57,58]

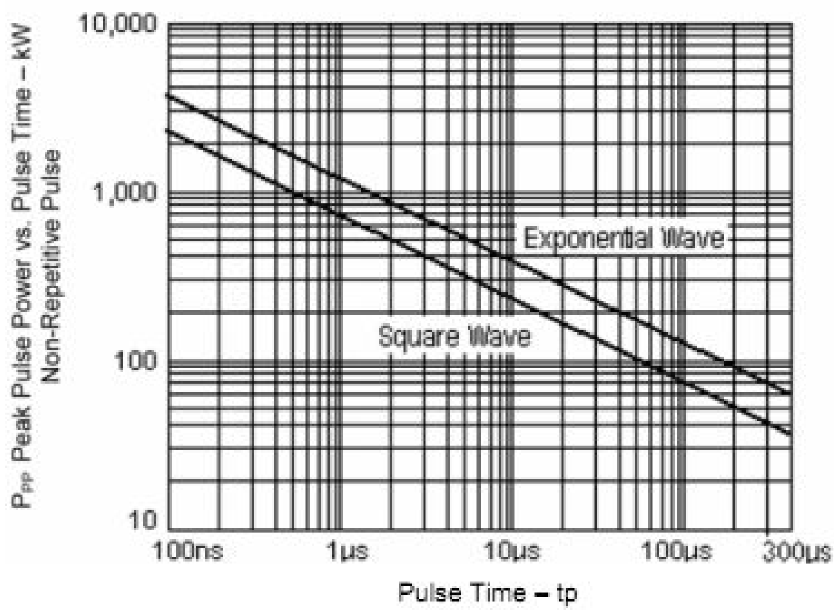


Figure 75. Typical Peak Pulse Power vs. Pulse Width[53,67,58]

Current-Limiting: Current limiting resistors in series between the TVS on the I/O circuits are necessary when TVS are installed on both ends. Current limiting keeps the diodes current below their max current ratings and keeps the clamping voltage below the isolators and I/O circuits limits. See Figure 67. Which end or both and how big the resistances are the designers' choice.
Cu/Mumetal Shield: See paper #2.[14] Actual shielding comes from the non-ferrous connectors with typically $Z_T \approx$ 5-10mΩ. If the cable is so short that its transfer impedance isn't below this level, there is no need to use the mumetal option.

The method shown in Figure 68 is recommended for the more extreme I·R-drop environments and for exposed safety critical circuits. In this case isolation protects the circuit from any voltage drops in the chassis ground connection. The shunt devices will prevent a surge from exceeding the breakdown voltage of the isolators as well as handling any differential surges on the cable. Current-limiting between the TVS will limit the TVS current, help the TVS minimize the clamping voltage which will help not overstress the isolators which will help not overstress the box circuits. Slower data rates will allow robust TVS with 50-90pF capacitance to chassis ground not degrade the signals. Author's note: Without current-limiting, the extremely large lightning I·R-drop voltages and currents would be too high; 100kW TVS have max current ratings of 1kA plus a TVS doesn't clamp the I·R-drop; it only reduces it by the clamping voltage.

If the lightning levels are enough to trigger the voltage clamps, the data stream will be affected for the duration of by that clamping voltage; functional upset in terms of a stream of bit errors will occur. In this case, the bits will all be clamped to one bit level for the duration of the transient.

It needs to be pointed out that the isolation above, be it transformer, capacitive, or optical, is really a capacitance in series with multiple series and parallel parasitic impedances within the protected circuitry. Making the designed-in isolation a small capacitance ensures that most of the externally induced voltage is induced across the isolator and not the junction capacitances of the transistors inside the isolation barrier, e.g. 5pF vs. 100pF or a x20 or 40dB reduction. That would reduce a 1kV

IR-drop to 50V across typical parasitic capacitances in the circuitry. Bottom line: isolation, too, is only relative and with multiple low and high frequency electromagnetic environments, the isolators must be measured against what they're isolating. The following Table 28 is a version of the parasitic capacitance on the circuit side of the isolators:

Table 28. Order-of-Magnitude Parasitic Capacitances in Circuits	
power FET-to-case	800pF (seems large)
transformer primary-to-secondary	600pF
transformer windings-to-shields	1400pF
circuit-to-chassis ground (including diodes)	100pF (transorbs 10,000pF)

By reducing such internal common mode noise, there is more margin for the externally induced common mode herein as well as EMI. This point is almost always overlooked by circuit designers and EMI specialists as well, i.e. lowering internal noise results in more margin against externally induced noise. All components in such circuits must be rated against the largest environment they are exposed to, quite often up to 1-2kV. The capacitance of surge protection devices must also be taken into account, e.g. avalanche diodes can have capacitance of as much as 10,000pF. Air gap arrestors can have a few pF capacitance but they are slower. In high overstress situations an air gap arrestor is followed by a diode clamp with a delay equivalent to a 30' in cable length or about 4mH inductance so the diode will not fire first and burn out.

Finally, the following Figure 76 is an example of how some Ethernet drivers and receivers are terminated and isolated from 1kV premise ground potential differences.[36] Understand this configuration and you understand common mode noise and mitigation.

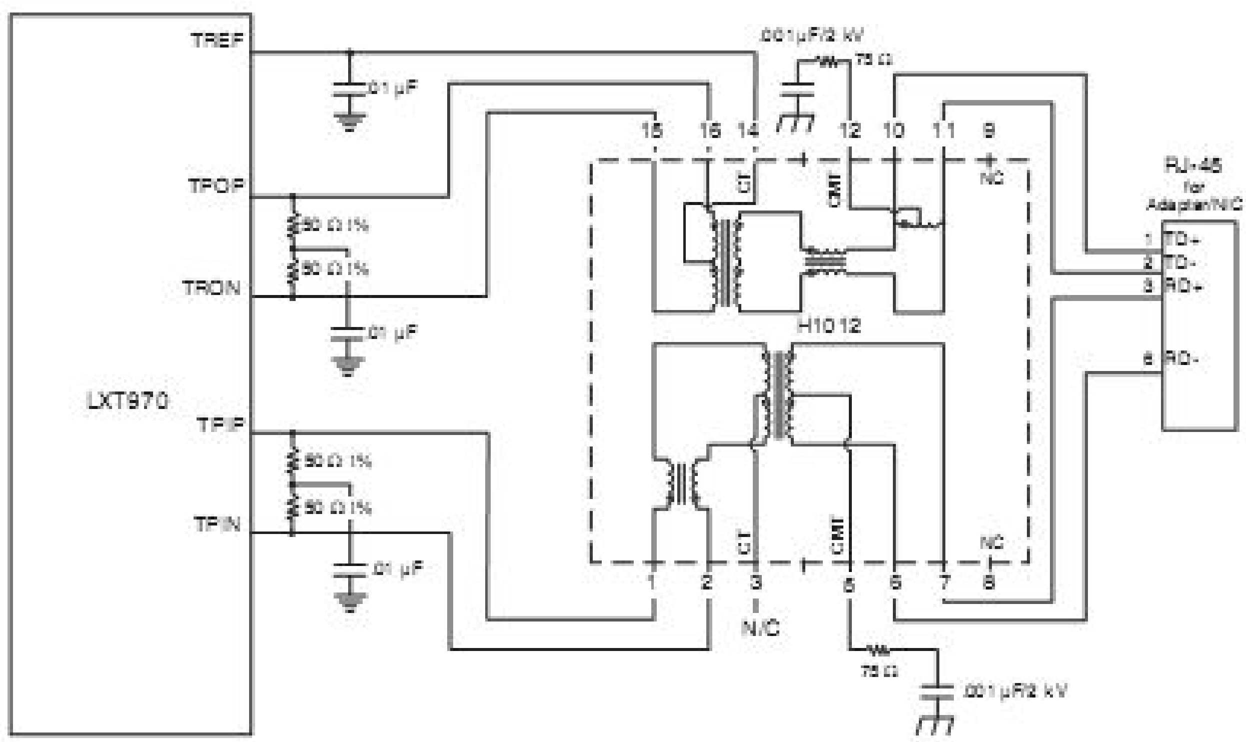


Figure 76. Example of Ethernet I/O Circuitry for Controlling Common Mode Noise[36]

Note the 2kV voltage rating of the common mode DC isolation/AC termination capacitors. This is because the ends of the Ethernet cables terminate in boxes powered from different power mains with different grounding. The result is large ground potential differences that themselves can wreak havoc with the electronics.

These circuits represent the ultimate in mitigation of common mode problems with circuit design. And, yet, software mitigation is necessary to lower the bit error rate. When TVS clamp, the bit errors and burst errors can last as long as 88µs to 1.5ms. Signal mode noise thresholds are around 100mV and CMR ranges from -0.5V to ±30V depending upon chip design.

As good as ±7V sounds to a designer in a benign environment, ±7V is a low threshold for lightning I·R-drop. That tells us how much we must suppress the lightning induced IR-drop for RS-422 circuits. The above I/O fix applied to one wire only drops the voltage across the loads to 1-10mV. The problem is that the power dissipated in the current-limiting resistors is large but the pulse rating is about x10-20 the DC power rating.

14. Concluding Remarks

1. Tables of transient levels in SAE ARP5412A and in DO-160F are totally misleading. In the hands of untrained engineers, the environments chosen are usually ludicrous. The levels of protection between the largest and smallest allocations are fictitious. The total lack of guidance and simple algorithms for the system designers is appalling.

2. This note, while not all inclusive, provides the lightning engineer with simple model formulas for allocating reasonable waveforms and levels early in a system program. Circuit designers must be educated about the common mode nature of the lightning and EMI environments at their I/O circuits.

3. The most important allocations early in a program are cable shields, I/O circuit protection, groundplane parameters, and bond joints' resistances. These all add weight and space that weight conscious manager's hate, i.e. their weight bogies were made up before you showed up. These features can be 100% tested or inspected on every tail number and form the basis for certificates of indirect lightning protection in production. The aircraft that undergoes system level lightning testing and the boxes that undergo lightning testing will provide a sample of safety margins and error budget for production tolerances. Similarity will provide the rest based upon theoretical models that can scale untested components to their anticipated safety margins. The system level test is an excellent system level in situ bond joint diagnostic tool and, in fact, that may be the real reason for running the test but don't advertise that kind of reality.

4. Environments have been scaled to in-flight interactions instead of ground-test.

5. Environment allocations have been scaled by system and cable geometry and electrical parameters instead of one size fits all.

6. Lightning Components A, D, and H have the same dI/dt and have the same spectra above 2MHz therefore the same WF2 and WF3 induced transients.

7. The IR-drop induced voltage is proportional to the distance between boxes along the airframe current path; the induced cable current is independent of that distance. Induced voltage s through cable shields is proportional the cable length. This is important for the power rating of surge suppressors.

8. Levels of Protection have been shown to involve cable-to-cable mutual coupling not attenuation of external fields. For purposes of standardization, the number of transient levels should increase on both the low and high sides. The only controlled shielding is box and cable shielding therefore the only shielding available for protection credit.

9. Induced environments are directly dependent upon parameters 100% inspectable and nearly 100% testable during airframe fabrication, a key to production certification by tail number.

10. Copper cable shielding with less than 30AWG thickness cannot attenuate low frequency WF1 and WF4 more than a few dB. The shielding is necessary for EMI and WF3 lightning.

11. Shielding and other physical protection cannot attenuate lightning induced noise below electronic noise thresholds.

12. TVS components between I/O lines and chassis ground cannot clamp the I·R-drop to TVS clamping voltage without current-limiting; they can only reduce the I·R-drop by the clamping voltage. Plus, a 100kW 40V_{BD} TVS has little over 1kA max current I_{PP} rating and a 500W 12V_{BD} TVS has I_{PP} = 20A. For a 1kA WF4 cable current, 50 12V_{BD} parallel TVSs would have to fire simultaneously in order to not exceed these limits.

13. Functional upset criteria should include software mitigation of bit errors and burst errors the duration of the transients and disallow erroneous commands from automated flight control electronics due to 88µs to 1.5s of transient noise. Common-cause failure modes from EMI and lightning beg architecture for fail-operational, fail-passive, distributed fault tolerance, and fault tolerance and recovery in general.

14. Protection of electronics against the lightning induced I·R-drop is considered to be high risk because of unknown susceptibilities. DO-160 ground injection tests have not been performed correctly. The I·R-drop and ground injection have been considered an I/O circuit susceptibility. I/O chip max ground potential ratings seldom exceed tens of volts including transformer isolated MIL-STD-1553. All chip max ratings other than I/O input/output pins are based upon substrate biases of less than negative one volt.

15. Isolation rated for ≥ 1kV installed in the I/O circuitry is required for the extreme I·R-drop environments. Isolation installed in series with I/O circuits is relative, not 100%. In terms of low frequency lightning transients, isolation is a capacitive voltage divider the success of which depends upon the impedance to chassis ground within the protected electronics in series with the isolator(s).

16. Indirect lightning protection has been broken down into the following:

a) Groundplane resistance,
b) Cable shield resistance,
c) Bond joint resistance,
d) TVS common mode protection,
e) Current-limiting,
f) Isolation between the TVS and the box I/O circuits including the possibility of isolated power supplies,
g) Combined copper and mumetal cable shields (limited by the shielding of the non-ferrous connectors),
h) Software fault tolerance of common-cause transient burst errors and recovery afterwards, and
i) Slower data rates like ARINC 429 allow 50-90pF TVS pin-to-chassis.

17. Engineering tests are needed throughout to establish parameters and tweak designs.

References

1. Radio Technical Committee on Aeronautics, RTCA/DO-160F, “Environmental Conditions and Test Procedures for Airborne Equipment, Section 22, Lightning Induced Transient Susceptibility”, Dec 2007

2. Society of Automotive Engineers Aerospace Recommended Practices, SAE ARP5412A, “Aircraft Lightning Environment and Related Test Waveforms”, Revised 2005-02

3. SAE ARP 5415A, “User’s Manual for Certification of Aircraft Electrical/Electronic Systems for the Indirect Effects of Lightning”, Revised 2002-04

4. FAA Advisory Circular on HIRF, AC 21-16C, 2/14/1990

5. FAA Advisory Circular on Indirect Lightning, AC 20-136A, 12/21/2006

6. Lee, K. S. H., Editor, *EMP Interaction: Principles, Techniques, and Reference Data*, Taylor & Francis, NY, 1995, a summary reference book based mostly on the Note Series of Carl E. Baum found on http://www.ece.unm.edu/summa/notes/index.html

7. Schelkunoff, S. A., “The Electromagnetic Theory of Coaxial Transmission Lines and Cylindrical Shields”, The Bell System Technical Journal, Volume XIII, 1934, 532-579

8. Meriwether, Cooper, & Parker, *Electromagnetic Pulse Handbook for Missiles and Aircraft in Flight*, EMP Interaction 1-1, Sandia Laboratories, 1968 (Prepared for AFWL)

9. Ramo, Whinnery, & Van Duzer, *Fields and Waves in Communication Electronics*, 3rd Edition, John Wiley & Sons, NY, 1994

10. Goldman, Stanford, *Laplace Transform Theory and Electrical Transients*, Dover Publications, Inc., NY, 1966 (copyright 1949)

11. Vance, E. F., *Coupling to Shielded Cables*, Wiley, NY, 1978

12. Vance, E. F., “Comparison of Electric and Magnetic Coupling through Braided-Wire Shields”, SRI Technical Memorandum 18, February, 1972 (Prepared for AFWL)

13. West, Larry, “In-Flight vs. Ground-Test Lightning Interactions in Composite Airframes, Effects of External vs. Internal inductance, An Errata to Everything Previously Published,” IN615, April 2011

14. West, Larry, Interaction Note 608, “Lightning Induced Waveform 5 in Composite Airframes, the Inability of Copper Braid to Shield It, and A New Layered Copper Braid and High-mu Foil Shield”, February 2009, www.ece.unm.edu/summa/notes/In/0608.pdf, Rev A, IN616, April 2011

15. West, Larry, IN618, “Assessment of & Recommendations for RTCA/DO-160F Section 22 Lightning Induced Transient Susceptibility”, September 2010

16. Casey, K. F., “Electromagnetic Shielding by Advanced Composite Materials”, Interaction Note 341, June 1977

17. West, Larry, Invited presentation to IEEE Region 5 Galveston Bay Section EMC Chapter Meeting Wednesday September 22, 2010, “Lightning Induced Transients in Composite Airframes - Analysis, Design, & Test”

18. Baum, Carl E., “Parameters for Some Electrically-Small Electromagnetic Sensors”, Sensor and Simulation Note 38, 21 Mar 67, Air Force Weapons Laboratory

19. Baum, C. E., “Electromagnetic Pulse Interaction Close to Nuclear Bursts and Associated EMP Environment Specification”, Interaction Note 76, 2 Jul 71, Air Force Weapons Laboratory

20. MIL-STD-285, Notice 1, “Attenuation Measurements for Enclosures, Electromagnetic Shielding, for Electronic Test Purposes, Method of”, 24 October 1997

21. Ogunsola, Ade, “Harmonization of Shielding-Effectiveness Standards for Enclosures”, http://www.ce-mag.com/archive/01/Spring/Ogunsola.html

22. Carter, Nigel, “The Background to the HIRF Requirements for Civil Aerospace”, Compliance Magazine, http://www.compliance-club.com/archive/old_archive/980212.html

23. Kajfez, D., “Excitation of a Terminated TEM Transmission Line through a Small Aperture”, Interaction Note 215, Jul 74, University of Mississippi, AFWL-TR-74-195, AD A021488

24. Baldacim, Crisofani, & Lautenschlager, “Lightning Effects in Aircraft of the Composite Material”, 17th CBECIMat – Congresso Brasileiro de Engenharia e Ciência das Materials, 15 a 19 de Novembro de 2006

25. Private Conversation with Ed Rupke of LTI, 2009

27. Plumer, J. A., “A Proposed Addition to the Lightning Environments Standards Applicable to Aircraft to Account for Effects of Positive Lightning Strokes of Long Duration and Moderate Intensity”, Lightning Technologies, Inc., http://www.lightningtech.com/pdfs/A_PROPOSED_ADDITION_TO_LIGHTNING_ENVIRONMENT_STDS.pdf, date unknown

28. Fullekrug, Price, Yair, & Williams, “Intense Oceanic Lightning”, Annales Geophysicae (2002) 20: 133-137 © European Geophysical Society 2002

29. Wright & Revesz, “New Technology Drives Lightning Test Standards”, Conformity, May 2008

30. Weston, David, *Electromagnetic Compatibility: Principles and Applications*, CRC Press, Taylor & Francis Group LLC, NY 2001 (2nd edition)

31. NFPA 780, Standard for the Installation of Lightning Protection Systems

32. ANSI/IEEE C62.41-1991, IEEE Recommended Practice in Low Voltage AC Power Circuits

33. IEC 61000 4-5, Equipment Classification of Power Surge and Lightning Protection

34. West, Larry, “Wiring and Cabling from a Modern Perspective, A Qualitative Discussion of Modern Circuit, Wiring, & Software Practices for EMI/EMC Control”, 2008, unpublished

35. Rupke, Edward, “Lightning Direct Effects Handbook”, Lightning Technologies, Inc, March 2002

36. Pulse Engineering, 12220 World Trade Drive, San Diego, CA 02128, http://www.pulseeng.com

37. RS422/485 Application Note, Chapter 4: Transient Protection of RS-422 and RS-485 Systems, http://www.bb-elec.com/tech_articles/rs422_485_app_note/transient_protection.asp, B & B Electronics Manufacturing Co.

38. Wright, N. and Makowski, L., “New Airliners Influence Lightning Tests”, EE Evaluation Engineering, June 2006

39. Clark, M., “Lightning Protection for Aircraft Electrical Power and Data Communications Systems”, Micronote™ 127, Rev 2, Microsemi Corporation, Scottsdale, AZ, February 2006

40. Raychem Memo, April 15, 1992, To; Neal Enault, From: Robert Moore, Subject: “Surface Transfer Impedance: Flat Braids vs. Round Braids”

41. RS-422, www.national.com/an/AN/AN-214.pdf

42. Dexmet Expanded Foil Mesh: http://www.dexmet.com/Die-Charts.html

43. MIL-STD-461F, Requirements for the Control of Electromagnetic Interference Characteristics of Subsystems and Equipment, 10 DEC 2007

44. MIL-B-5087B, Bonding, Electrical, and Lightning Protection for Aerospace Systems, 15 Oct 1964

45. http://www.atecorp.com/Equipment/Biddle/247000.asp

46. BAE Systems Platform Solutions, 3131 Story Road West, Irving, Texas 75038

47. MAXIM, http://www.maxim-ic.com/

48. Johnson & Graham, *High Speed Signal Propagation*, Prentice Hall, NJ, 2003

49. http://ecee.colorado.edu/~ecen4242/ethernet/ethernet.html#_Error_Control_Codes

50. http://www.hubersuhner.com/ie70/hs-index.htm

51. http://www.weibull.com/hotwire/issue54/relbasics54.htm

52. http://www.computer.org/portal/web/csdl/doi/10.1109/DSN.2003.1209984

53. www.vishay.com

54. www.avagotech.com

55. www.analog.com

56. www.ti.com

57. www.semtech.com

58. www.Littlefuse.com

59. http://www.tycoelectronics.com/catalog/feat/en/c/15736#

60. Holt Integrated Circuits, "Lightning Protection of ARINC 429 Devices" Application Note AN-300, Rev F, February 2011, www.holtic.com

61. West, Larry, *Indirect Lightning in Composite Aircraft*, Self Published Public Domain, TX, 2011

62. ARACON® and MICRO-COAX® are registered trademarks of MICRO-COAX, Inc. http://www.micro-coax.com/pages/technicalinfo/ARACON/aracon_downloads.asp

63. KEVLAR® is a registered trademark of E.I. du Pont de Nemours and Company.

64. Personal conversation with Bill Gaffney of Micro-Coax, Inc., 6/11/10: Thickness of Aracon[62] plating is rounded off to 50 microns; the fibers have about 11 microns of Ni, 24 microns of Cu, followed by 11 microns of Ni or 13 microns of Ag, all on Kevlar[63] fibers about 16 microns in diameter. Total diameter is about 108-110 microns.

The author thanks Roxanne Arellano for many discussions about scaling incalculable parameters, Tom Pierce, John Norgard, and Carl E. Baum for his support and help with this subject.