Grounding
And
Shielding

RYP Masters Program
Electronics for Space

Lecture notes

Swedish Inst. of Space Physics

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Grounding

Ground = sink for electrical charge

Grounding:
Connecting return conductors of electrical circuits to a reference potential.

Bonding:
Connection of two conducting surfaces in order to provide a good electrical contact.

Space systems also use the term "ground". They are electrically referenced to the vehicle skin, which acts as reference potential plane.

A grounding concept for electronic circuits, assemblies or even systems, serves the purpose

- to avoid circulating EMI due to potential differences between mutually connected electrical units of a system
- to provide an equipotential reference plane
- to prevent common mode coupling
- to avoid low impedance ground loops
- to protect against shock hazards owing to high voltages appearance ESD on a frame or box housing by harness damage

Goal:
Realize and control a low (zero) impedance plane for all connections including material, bondings, contact pressures, contact area.

- All impedances are frequency dependant
- A ground distance (electric length) of λ/4 yields isolation
- Maximum extension of a ground plane must be less than λ/20-λ/15

Ground plane impedance $$Z = 0.26 * 10^{-6} \sqrt{\frac{\mu f}{\sigma}} \left(1 + \left|\tan 2\pi \frac{l}{\lambda}\right|\right)$$

$$l < \frac{\lambda}{20} \qquad Z = 0.26 * 10^{-6} \sqrt{\frac{\mu f}{\sigma}}$$

$$l = \frac{\lambda}{8}, 3\frac{\lambda}{8} \qquad Z \approx 0.52 * 10^{-6} \sqrt{\frac{\mu f}{\sigma}}$$

$$l = \frac{\lambda}{4}, 3\frac{\lambda}{4} \qquad Z \to \infty$$

	Thickness		
Frequency	0.1 mm	1 mm	10 mm
60 Hz	172μΩ	17.2μΩ	1.83μΩ
1 kHz	172	17.5	11.6
10 kHz	172	33.5	36.9
100 kHz	175	116	116
1 MHz	335	369	369
10 MHz	1.16 mΩ	1.16 mΩ	1.16 mΩ
100 MHz	3. 69	3.69	3.69

Impedance Values for Ground Plane Impedance (ohms-per-square)

System Level Grounding

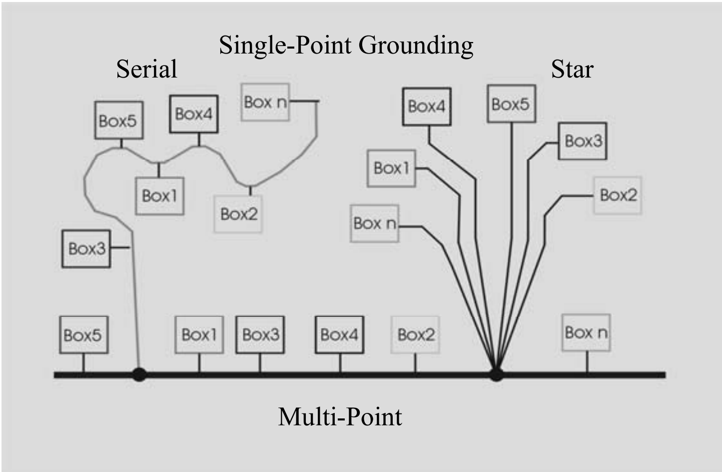
There are three main system grounding methods

- Single-Point Grounding
  - Either Series or Parallel
  - Best for frequencies below 1 MHz
  - Has the largest amount of ground loop currents
- Multi-point Grounding
  - Preferred for frequencies above 1 MHz.
  - Minimizes loop currents and ground impedance of planes.
  - Lead Lengths must be kept extremely short
  - Provides for maximum EMI suppression
- Hybrid
  - A mixture of both Single-Point and Multi-Point Grounding in the same system.
- Ground loops cause RF energy to be radiated when high inductance returns are provided.

Note: Do not count on mounting screws to provide low inductance connections. They are highly inductive and can act as helical antennae at high frequencies (100 MHz-1 GHz)!! (Use conductive gaskets in addition to the screws.)

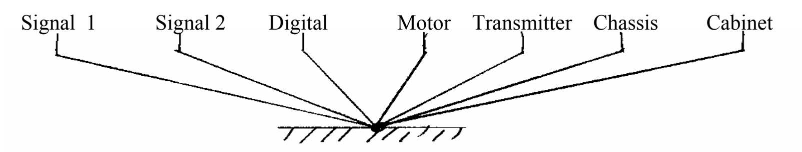
- In a Multi-point ground system, the distance between the screws should not exceed λ/20 of the shortest wave length in the system.

Grounding Concepts



Ground concept frequency limitations

Ground separation concept



System level control of grounding concept

The only way to implement a system grounding concept during the development phase (harness manufacturing!) and to keep an access open for any EMC design input during system integration and test verification, is to establish an overall **grounding and shielding diagram** early in the development phase of a space project and to continuously update it further on. This grounding and shielding diagram

- will reflect the special requirements of sensitive equipment (experiments, sensors),
- considers the geometrical box configuration of the system
- shows harness routing, cable shielding, shield grounding, interface circuitry, cable twisting for equally treated groups of wires as they enter or leave a box i.e. for

-power lines (primary and secondary if applicable)
-digital lines)
-analog lines) for control, interconnection,
-RF lines) house-keeping or scientific purposes
-pyro lines)
-clock lines)

Ground concept for modern scientific satellites.

Distributed Single Point Grounding (DSPG)

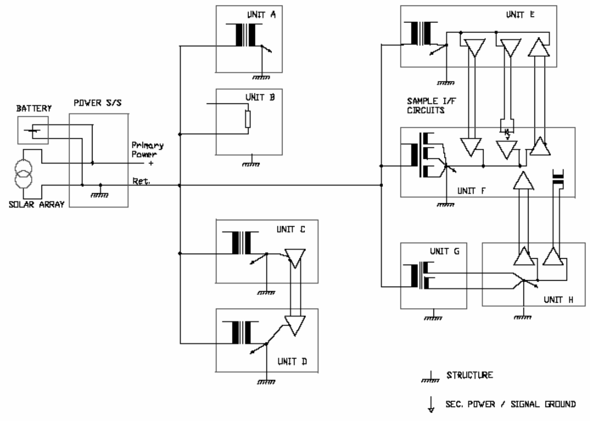
- The primary power shall be connected to the spacecraft ground structure at one point only (star point) within the instrument subsystem.
- This grounding of primary power shall be done within the PCU of the power S/S.
- All primary power return for users shall be routed to this star point.
- The S/C structure shall not be used as return path for power and signals (except low level signals) to avoid common mode loop effects.
- Secondary power shall be grounded to structure only once in each unit / experiment

The principle is to realize a single point grounding of each independent power network and galvanic isolation between those networks.

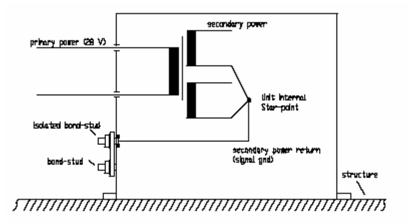
,

The internal grounding shall fulfill the following rules:

- The ground plane shall carry only low level signal return currents in order to minimize magnetic loop effects.
- Secondary return leads shall be directly routed from the corresponding unit and be connected to the local ground reference point.
- In case that several units are supplied from the same DC/DC converter secondary power output, the distribution shall be a star point system



Unit Grounding



Isolation of Primary Power from Structure

The Structure shall not be used as a return path for DC-power. The isolation measured between power lines and the experiment housing shall be equivalent to:

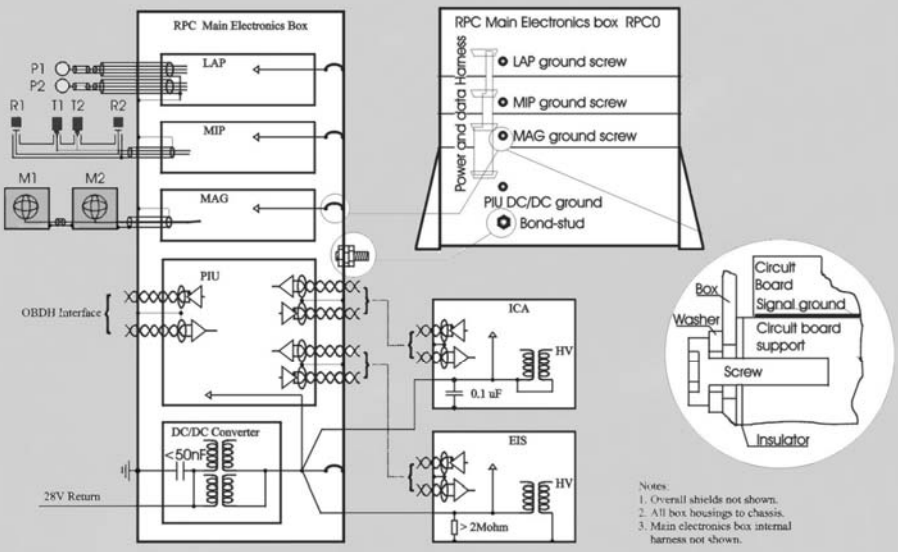
- a DC resistance of R > 1 MOhm in parallel with
- a capacitance of C < 50 nF per line.

Isolation between Primary and Secondary Power

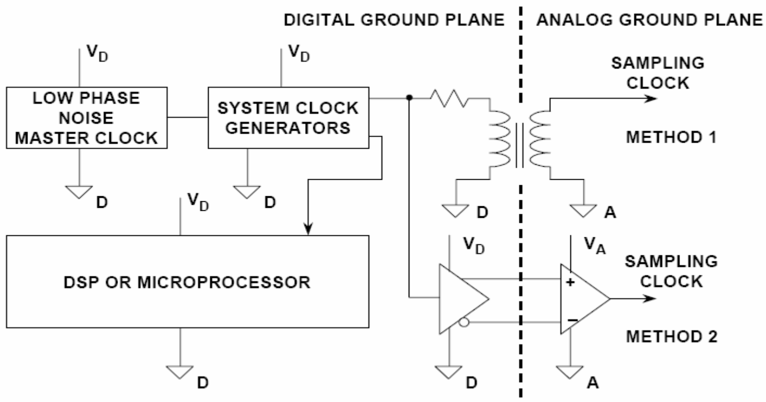
To reduce common mode noise coupling, primary power lines shall be isolated from the secondary power and signal lines. The isolation impedance shall be equivalent to:

- a DC resistance R > 1 MOhm in parallel with
- a capacitance C < 5 nF when one of the two ground

RPC Distributed Grounding Concept



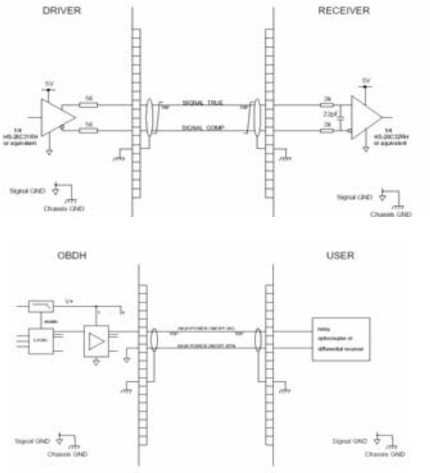
SAMPLING CLOCK DISTRIBUTION FROM DIGITAL TO ANALOG GROUND PLANES

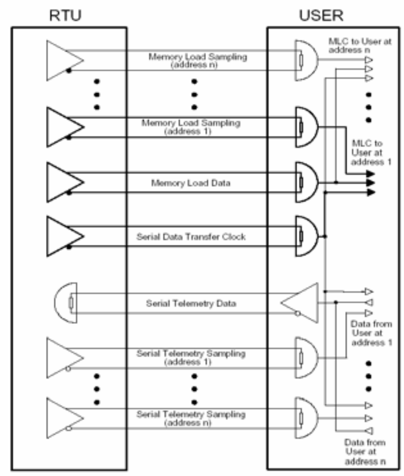


$$SNR = 20\log_{10}\left[\frac{1}{2\pi f t_j}\right]$$

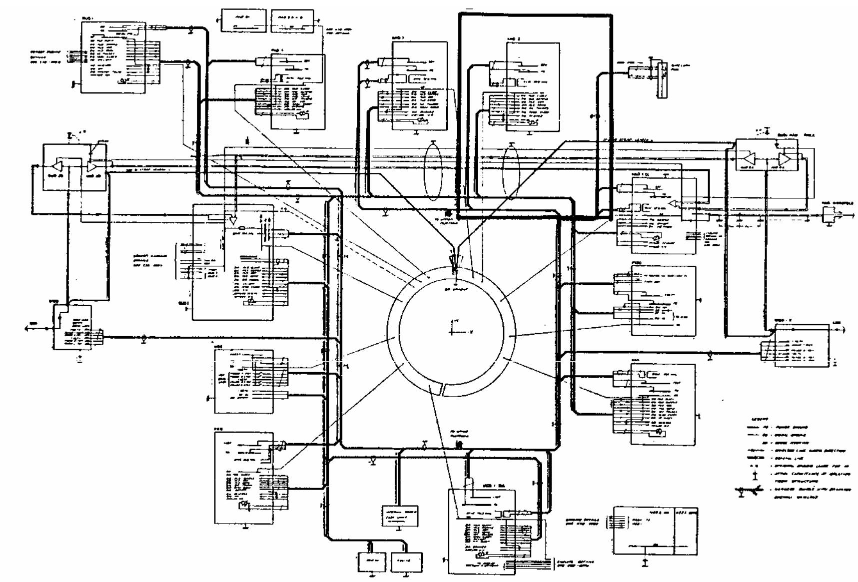
t_j = Sampling Clock Jitter
f = Analog Input Frequency

Experiment interface

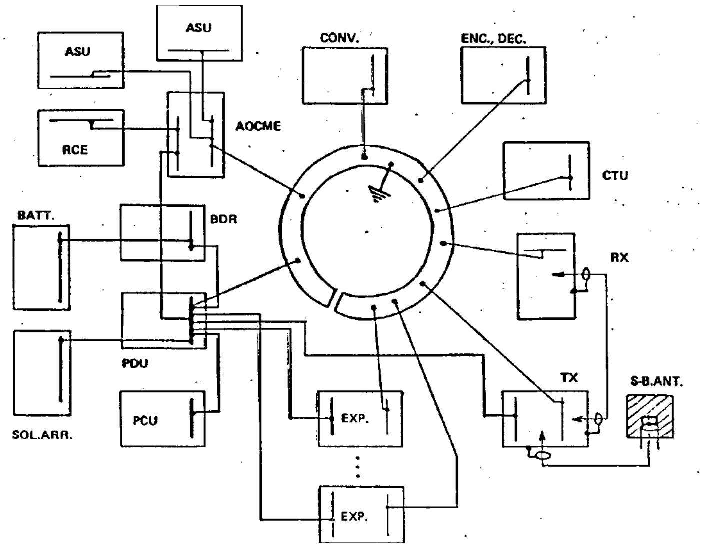




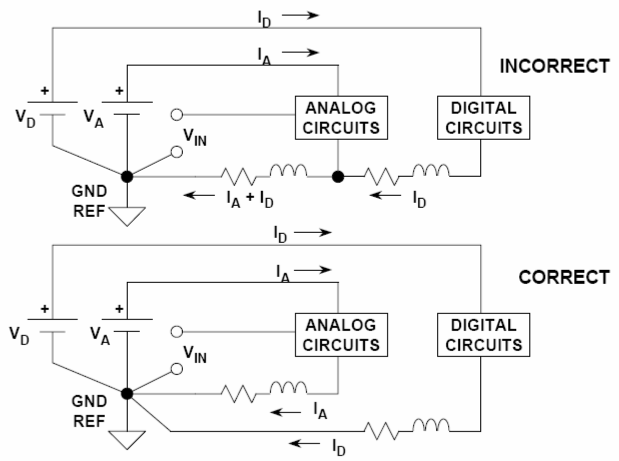
Cluster Grounding Block Diagram

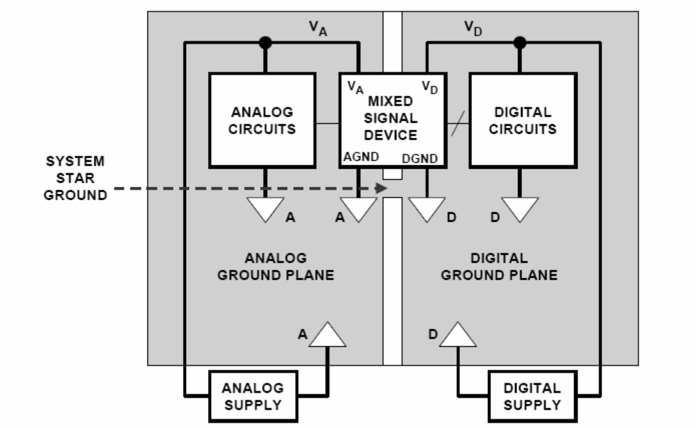


Cluster principal block diagram



DIGITAL CURRENTS FLOWING IN ANALOG RETURN PATH CREATE ERROR VOLTAGES





Spacecraft blanket charging

As a rule of thumb the following equation can be used to determine the required bonding and the surface conductivity:

$$\boxed{\Delta V = \frac{I \cdot R_{\Omega/square} \cdot d^2}{2}}$$

where
ΔV = differential voltage [Volt],
I = plasma current [A/cm^2]
d = max. distance from any point on surface to next grounding point [cm]
$R_{\Omega/square}$ = surface resistance [Ω/square]

Relative skin depth

Material	k
Aluminum	1.26
Copper	1.00
Lead	3.52
Silver	0.94
Tin	2.55
Tungsten	1.76
Brass	1.99 '(depends on alloy and temp.)
Phos-bronze	2.1
Bronze	3.1

$$\delta = 66.1 * k \Big/ \sqrt{f}$$

Copper $k = 1$

Shielding

$$S\left(dB\right) = 20\log_{10} \frac{incident \quad field}{transferred \quad field}$$

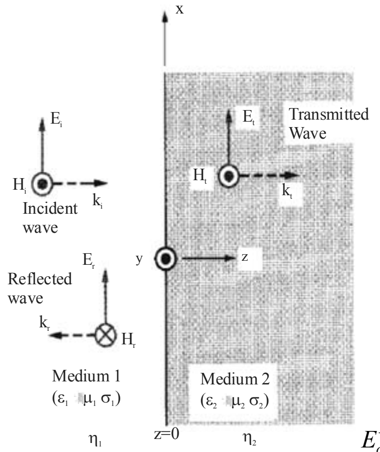
$$S = S_R \, x \, S_A \, x \, S_{MR}$$

S_R reflections
S_A absorption
S_{MR} multiple reflections

Barrier impedance of metals

$$\eta_b = \frac{\eta_m}{\left(1-e^{-t/\delta}\right)} \quad \text{for } t/\delta \text{ any value}$$

$$\eta_b \simeq \frac{\eta_m}{\left[1-\left(1-t/\delta\right)\right]} = \frac{\eta_m \delta}{t} \quad \text{for } t/\delta \ll 1$$



$$E_{x1}(z) = \underbrace{E_{x1}^{+}(z)}_{\text{incident wave}} + \underbrace{E_{x1}^{-}(z)}_{\text{reflected wave}} = E_{o1}^{+}e^{-jk_1z} + E_{o1}^{-}e^{jk_1z}$$

$$H_{y1}(z) = H_{y1}^{+}(z) + H_{y1}^{-}(z) = \frac{E_{o1}^{+}}{\eta_1}e^{-jk_1z} - \frac{E_{o1}^{-}}{\eta_1}e^{jk_1z}$$

$$E_{x2}(z) = \underbrace{E_{x2}^{+}(z)}_{\text{transmitted wave}} = E_{o2}^{+}e^{-jk_2z}$$

$$H_{y2}(z) = H_{y2}^{+}(z) = \frac{E_{o2}^{+}}{\eta_2}e^{-jk_2z}$$

$$\begin{cases} E_{x1}(0) = E_{x2}(0) \quad \Rightarrow \quad E_{o1}^{+} + E_{o1}^{-} = E_{o2}^{+} \\ H_{y1}(0) = H_{y2}(0) \quad \Rightarrow \quad \dfrac{E_{o1}^{+}}{\eta_1} - \dfrac{E_{o1}^{-}}{\eta_1} = \dfrac{E_{o2}^{+}}{\eta_2} \end{cases}$$

$$k = \omega\sqrt{\varepsilon\mu\left(1 - j\,\sigma/\omega\varepsilon\right)}$$

$$\eta = \sqrt{\mu/\left(\varepsilon\left(1 - j\,\sigma/\omega\varepsilon\right)\right)}$$

$$\frac{E_{o2}^{+}}{E_{o1}} = \frac{2\eta_2}{\eta_1 + \eta_2} \overset{\textit{Defined as}}{=} \tau_E \qquad \frac{E_{o1}^{-}}{E_{o1}^{+}} = \frac{\eta_2 - \eta_1}{\eta_2 + \eta_1} \overset{\textit{Defined as}}{=} \Gamma_E$$

$$\frac{H_{o2}^{+}}{H_{o1}} = \frac{2\eta_1}{\eta_1 + \eta_2} = \tau_H \qquad \frac{H_{o1}^{-}}{H_{o1}^{+}} = \frac{-E_{o1}^{-}/\eta_1}{E_{o1}^{+}/\eta_1} = -\Gamma_E = \frac{\eta_1 - \eta_2}{\eta_2 + \eta_1} = \Gamma_H$$

$$\tau_E = 1 + \Gamma_E \qquad SWR = \frac{1+\Gamma}{1-\Gamma}$$

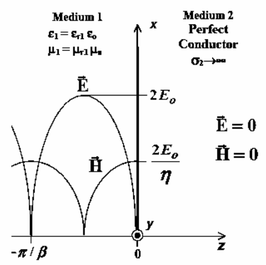
Matching impedances $\eta_1 = \eta_2 \Rightarrow \Gamma_E = 0 \Rightarrow \tau_E = 1 \Rightarrow \textit{No reflection} \quad SWR = 1$

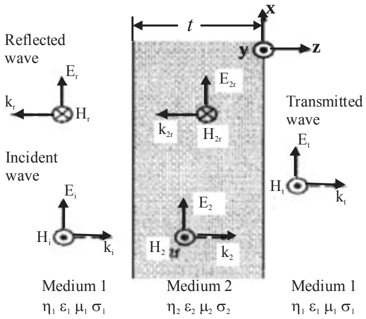
Medium 2 = perfect conductor $\sigma_2 = \infty \Rightarrow \eta_2 = 0 \Rightarrow \Gamma_E = -1 \quad \textit{and} \quad \tau_E = 0$

Total reflection. The electrical field at the interface is

$$E_{o1}^{+} + E_{o1}^{-} = E_{o1}^{+} + \Gamma_E E_{o1}^{+} = E_{o1}^{+} - E_{o1}^{+} = E_{o2}^{+} = 0$$

Reflection of UPW in a perfect conductor





$$\frac{E_i}{E_t}=\frac{2\eta_2}{\eta_1+\eta_2}\frac{2\eta_1}{\eta_1+\eta_2}=\frac{4\eta_1\eta_2}{\left(\eta_1+\eta_2\right)^2}=S_E$$

$$\frac{H_t}{H_i}=\frac{2\eta_1}{\eta_1+\eta_2}\frac{2\eta_2}{\eta_1+\eta_2}=\frac{4\eta_1\eta_2}{\left(\eta_1+\eta_2\right)^2}=S_H$$

$$\eta_1 \gg \eta_2$$

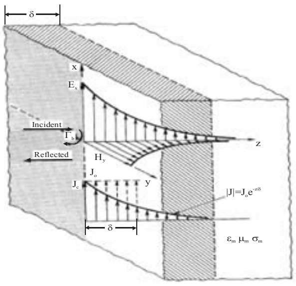
$$\frac{H_t}{H_i}=\frac{E_t}{E_i}=\frac{4\eta_2}{\eta_1}=S_R$$

The shielding effectiveness in dB is

$$S_{RdB}=20\log\frac{4\eta_1\eta_2}{\left(\eta_1+\eta_2\right)^2}\overset{\eta_1\gg\eta_2}{\approx}20\log\frac{4\eta_2}{\eta_1}$$

Absorption loss

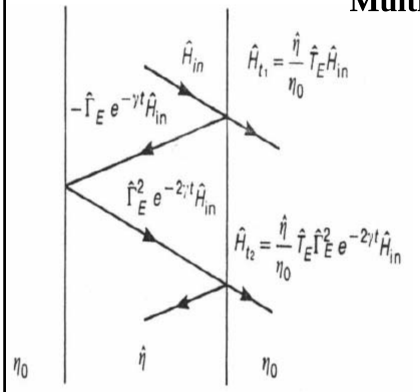
$$S_A=e^{\gamma t}=e^{\frac{t}{\delta}}=e^{t\sqrt{\pi f\sigma\mu}} \quad S_{AdB}=20\log e^{\frac{t}{\delta}}=15t\sqrt{f}\sqrt{\sigma\mu}$$



Relative conductivity and permeability of metals.

Metal	σ_r	$\mu_r @ \le 10kHz$	$\sqrt{\sigma_r\mu_r}$
Silver	1.064	1	1.03
Copper	1	1	1
Gold	0.7	1	0.88
Aluminum	0.63	1	0.78
Brass	0.47	1	0.69
Magnesium	0.38	1	0.61
Tin	0.151	1	0.39
Lead	0.079	1	0.28
Supermalloy	0.023	100.000	53.7
Purified Iron	0.17	5.000	29.2
Mumetal	0.0289	20.000	24.0
50% Nickel, Iron	0.0384	1.000	6.2
Commercial Iron (0.2 impure)	0.17	200	5.38
Steel	0.17	180	5.53
Nickel	0.23	100	4.7
Stainless Steel (1Cu,18Cr,8Ni,&Fe)	0.02	200	2.0

Multiple Reflection loss



$$\hat{H}_t = \hat{H}_{t_1} + \hat{H}_{t_2} + \hat{H}_{t_3} + \cdots$$
$$= \hat{H}_{t_1}(1 + \Delta_2 + \Delta_3 + \cdots)$$

$$\eta \ll \eta_o \qquad t \ll \delta$$

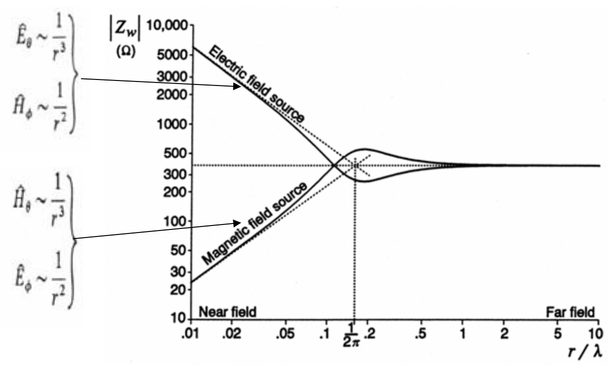
$$S_{MR(dB)} = 20\log\left[2\sinh\left(\frac{t}{\delta}\right)\right] - S_{A(dB)} = 20\log\left[2\sinh\left(\frac{t}{\delta}\right)\right] - 20\log e^{-t/\delta}$$

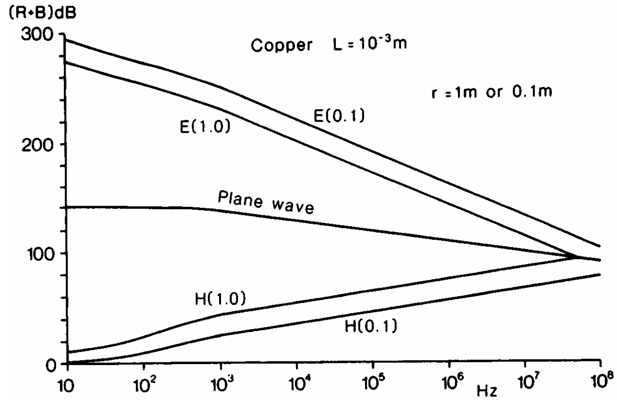
$$S_{MR(dB)} = 20\log\left(1 - e^{-2t/\delta}\right)$$

Shielding effectiveness in the near field

Electrical dipole $$Z_{wE} = \frac{j/\beta r + 1/(\beta r)^2 - j/(\beta r)^3}{j/\beta r + 1/(\beta r)^2} \xrightarrow{\beta r \ll 1} |Z_w|_e = \frac{1}{2\pi f \varepsilon_o r} = 60\frac{\lambda_o}{r}$$

$$Z_{wH} = -\eta_o \frac{j/\beta r + 1/(\beta r)^2}{j/\beta r + 1/(\beta r)^2 - j/(\beta r)^3} \xrightarrow{\beta r \ll 1} |Z_w|_m = 2\pi f \mu_o r = 2370\frac{r}{\lambda_o}$$



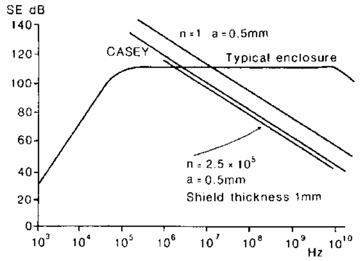


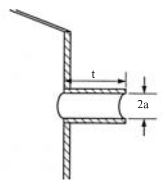
The effect of apertures

Gain G for a circular aperture of radius **a** $G = \frac{P_2}{P_1} = \left(\frac{2\pi a}{\lambda}\right)^2$

$$S_{AP} = 20\log\left(\lambda/2\pi a \quad \sqrt{n}\right) + 32t/2a \qquad t > a$$

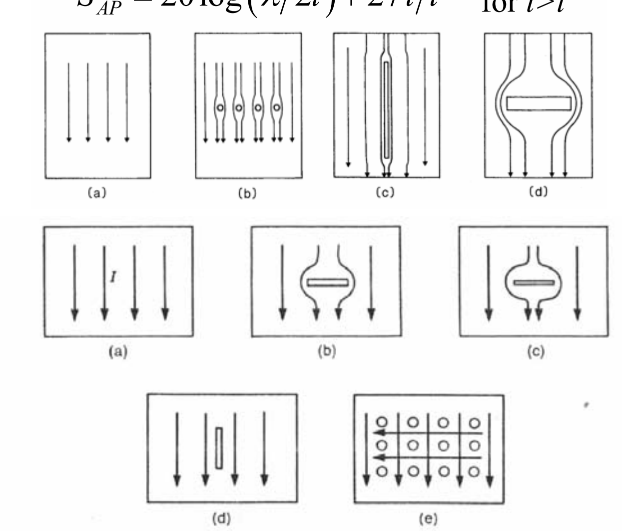
n apertures waveguide cut-off



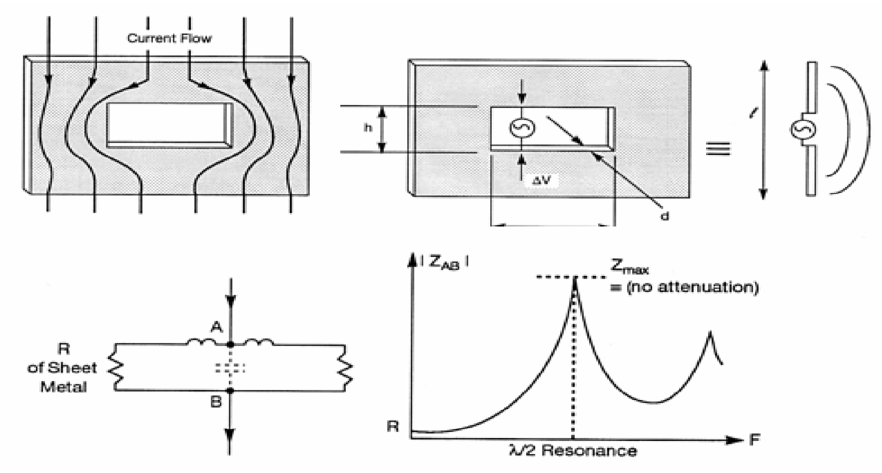


Shielding effectiveness due to the presence of a slot

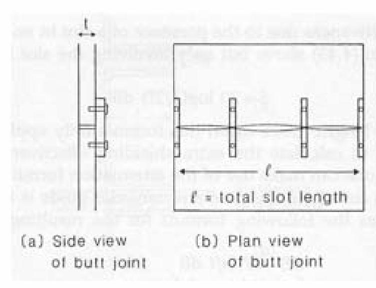
$$S_{AP} = 20\log\left(\lambda/2l\right) + 27\,t/l \quad \text{for } t>l$$

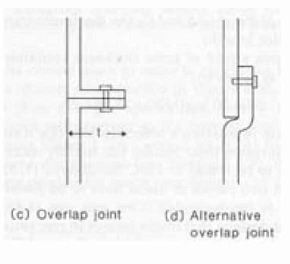


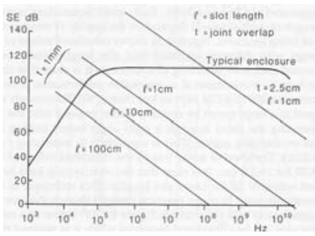
Slot



Slot

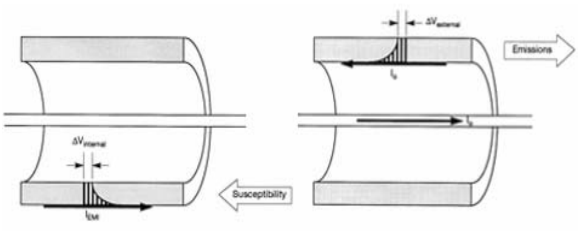






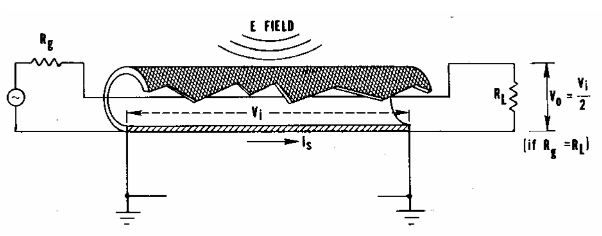
Alternative ways to describe shield quality

Alternative ways to describe shield quality $Z_t = \frac{1}{I_o}\frac{dV}{dz}$



$V = LZ_t I_o$

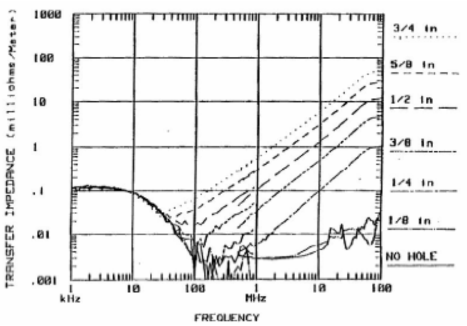
Cable of length L



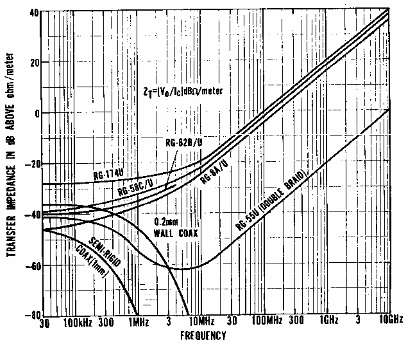
Alternative ways to describe shield quality

$$Z_t = R_o\left(1+j\right)\left(\frac{t}{\delta}\right)\Big/\sinh\left(\left(1+j\right)\left(\frac{t}{\delta}\right)\right)+j\omega M$$

Apertures

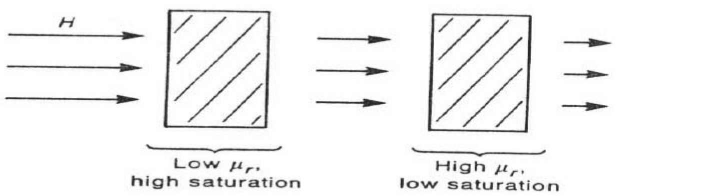


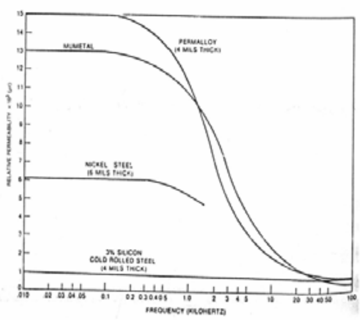
Measured surface transfer impedance of a
1-1/4 diameter cooper pipe with a single hole.



High H-fields

First reduce incident magnetic field so that it does not saturate the second shield.



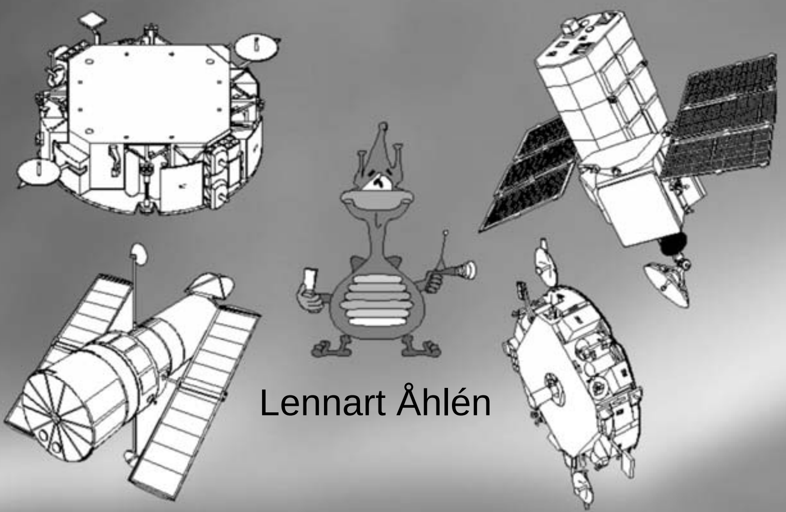


Frequency Dependence of various ferromagnetic materials

•Shielding enclosures of switching power supplies are constructed from steel(cheaper) rather than Mumetal

•For 60Hz interference Mumetal is used if field strengths are not too large to saturate the material

Satellite Grounding Concepts



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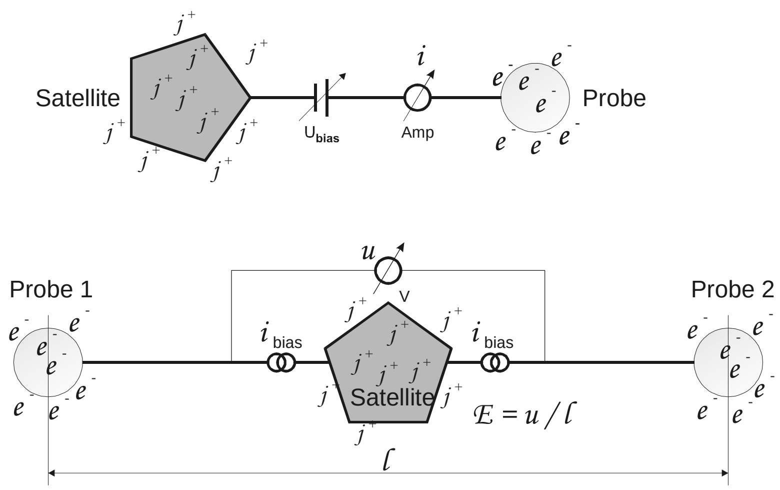
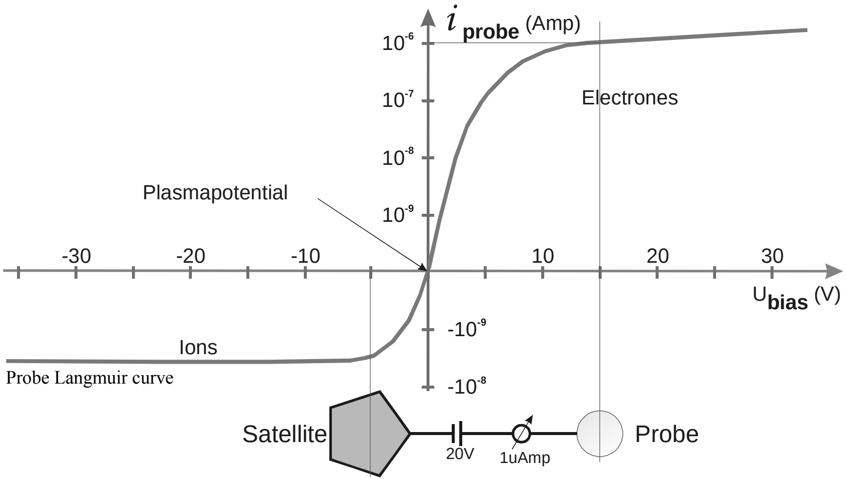
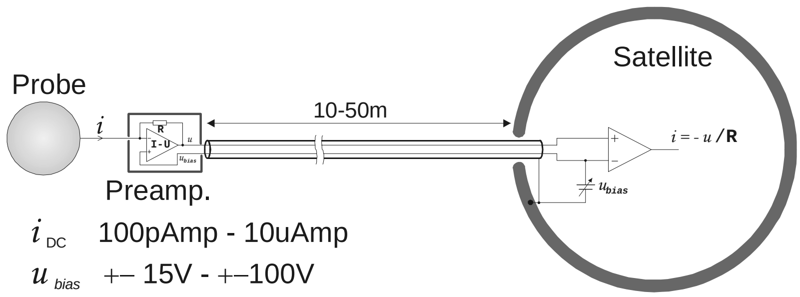
E-field and Density measurements



Two probe density measurement



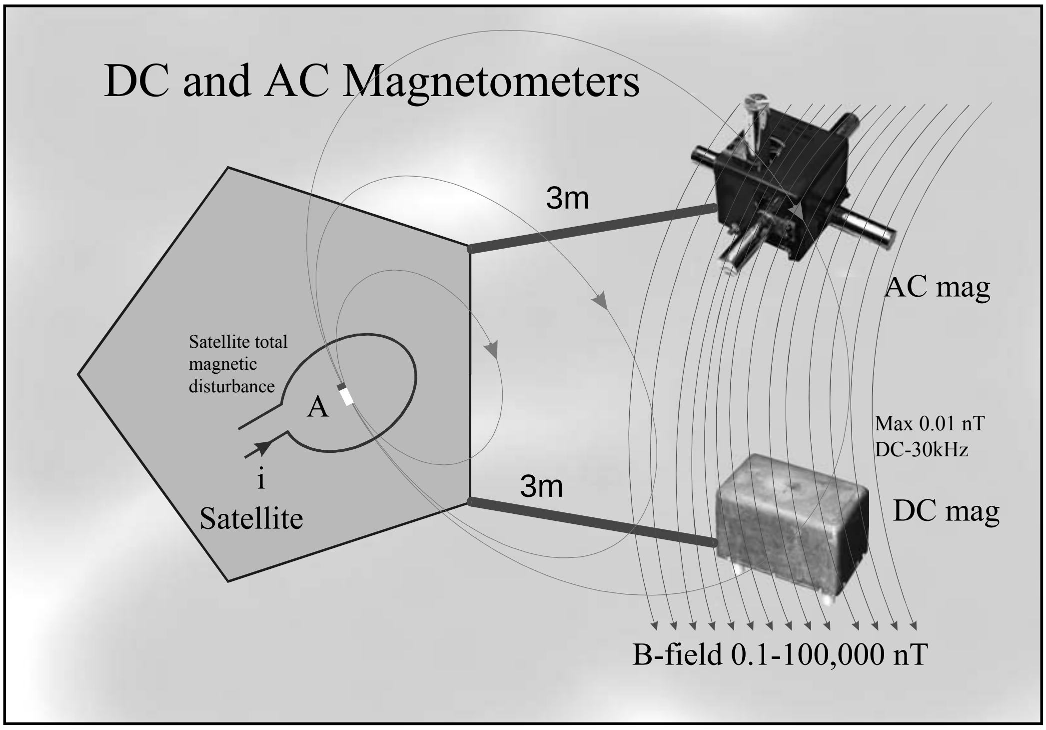
Density Probe functional block diagram



i_{DC} 100pAmp - 10uAmp

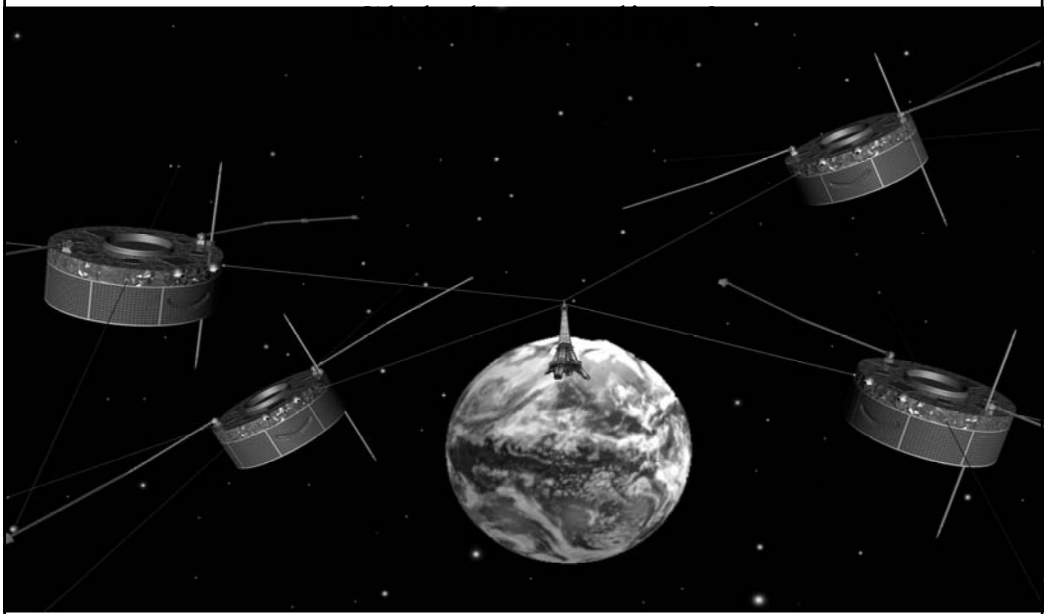
u_{bias} +– 15V - +–100V

i_{AC} 0.1 - 50 % modulation up to 10kHz

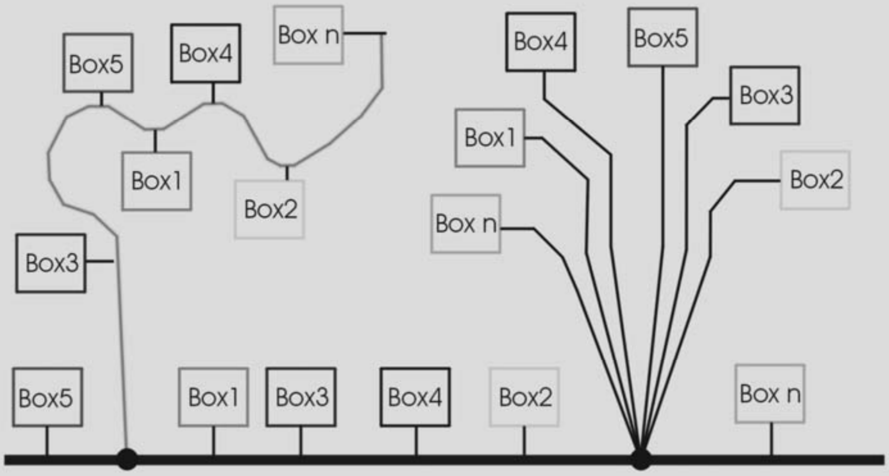


Experimenter EMC requirements

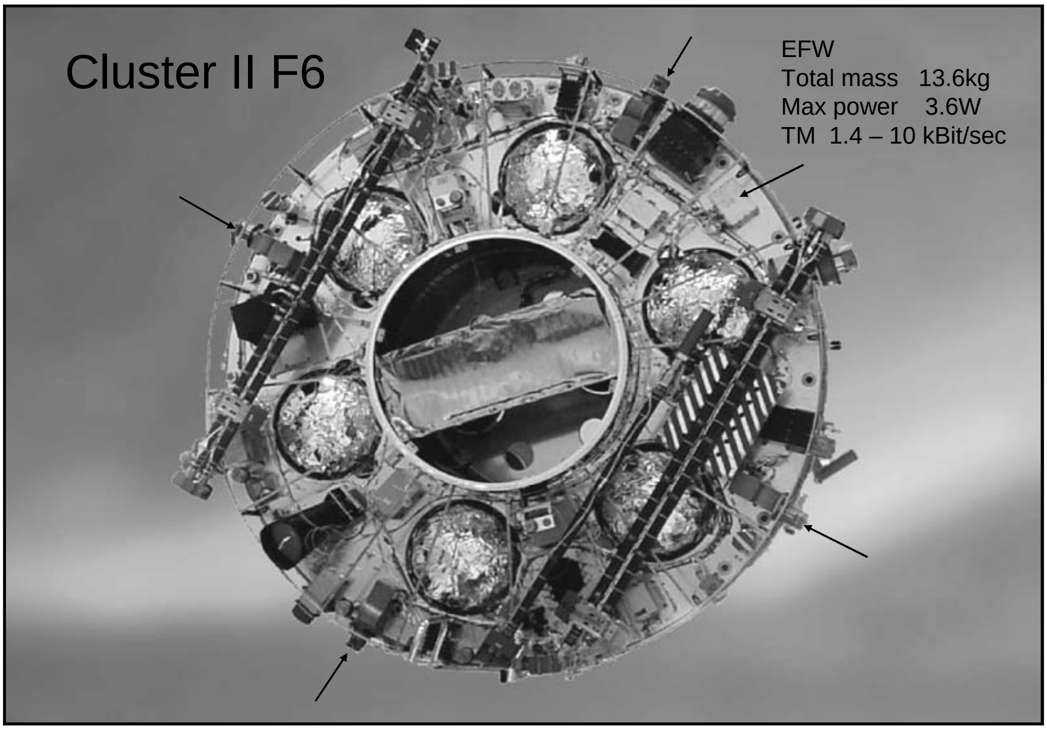
1. All spacecraft surfaces exposed to the plasma environment shall be sufficiently conductive and grounded. < 5 kohm/sq
2. Small surfaces differential charging potential shall not exceed +-10 V, assuming a plasma current of 5 nA/cm^2
3. The S/C structure shall not be used as return path for power and signals except for sensor signals to avoid common impedance coupling and magnetic disturbances.
4. Isolated receivers and balanced differential signals should be used as subsystem signal interfaces.
5. All active wires shall be twisted with its return wire and loops on circuit boards should be minimize to reduce magnetic disturbances.
6. The spacecraft system shall use a Distributed Single Point Grounding.
7. Secondary power shall be grounded to structure only once in each unit / experiment.
8. Cable shields shall be grounded to structure ground at both ends. Shields shall not be used as the return path for signal or power.
9. Non-magnetic materials shall be used wherever possible.The use of ferro-magnetics shall be avoided wherever possible.



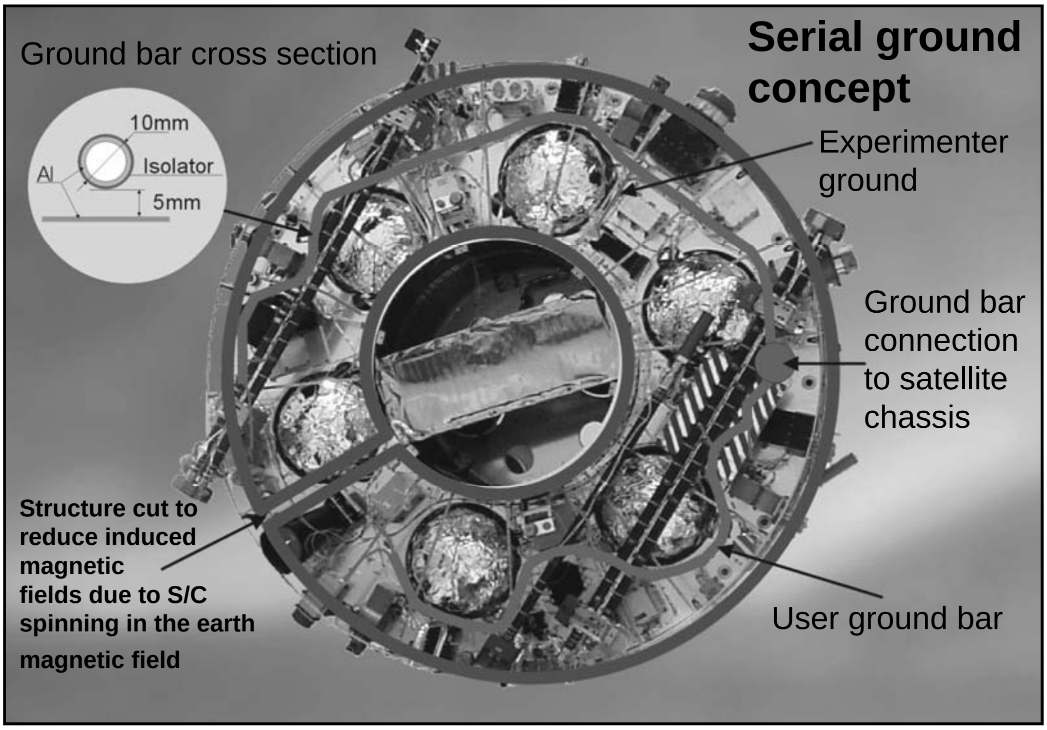
Serial, Star or Plane Ground?

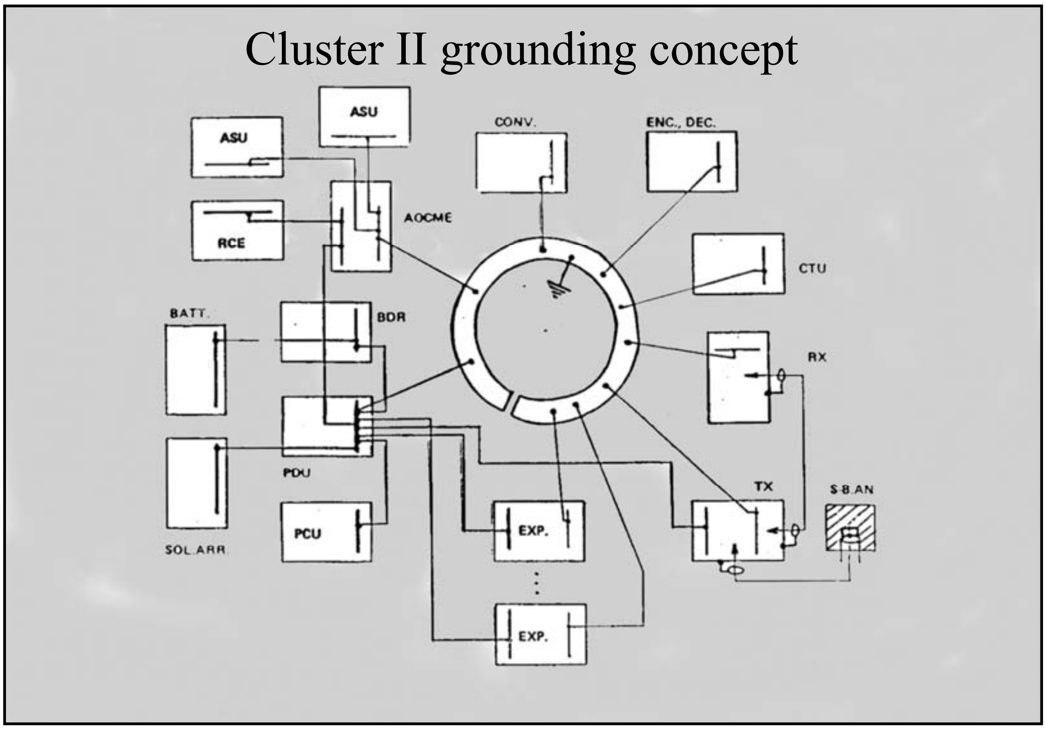


Cluster II F6



EFW
Total mass 13.6kg
Max power 3.6W
TM 1.4 – 10 kBit/sec


Serial ground concept
Ground bar cross section
10mm
Al
Isolator
5mm
Experimenter ground
Ground bar connection to satellite chassis
Structure cut to reduce induced magnetic fields due to S/C spinning in the earth magnetic field
User ground bar

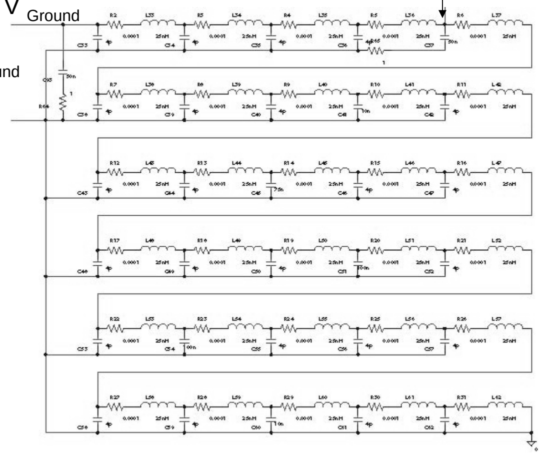

Cluster II grounding concept
ASU
ASU
CONV.
ENC., DEC.
RCE
AOCME
CTU
BATT.
BDR
RX
PDU
TX
S-B.AN
PCU
EXP.
SOL.ARR
EXP.

Cluster ground bar SPICE model

1mA experimenter Injected noise from100Hz to100MHz

Experimenter ground V_{Ground}

Z'_{Ground}

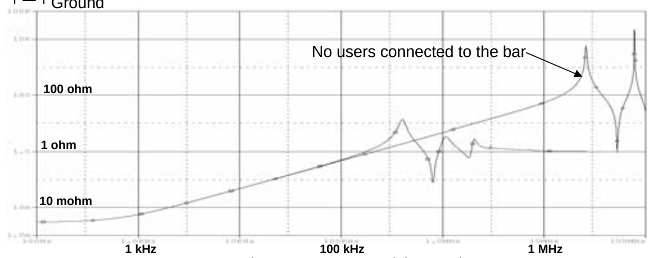


Seven Experiments (10nF-100nF) are connected to the 3m long Ground bar.

Satellite Chassis Ground

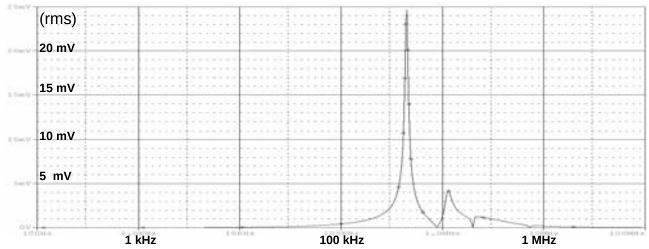
SPICE simulation results

$|Z|_{Ground}$



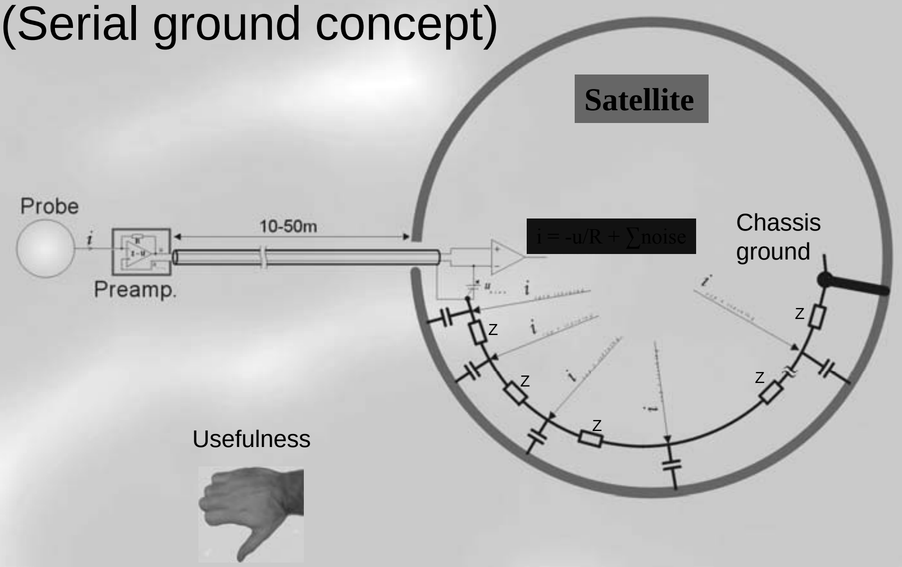
Experimenter ground impedance

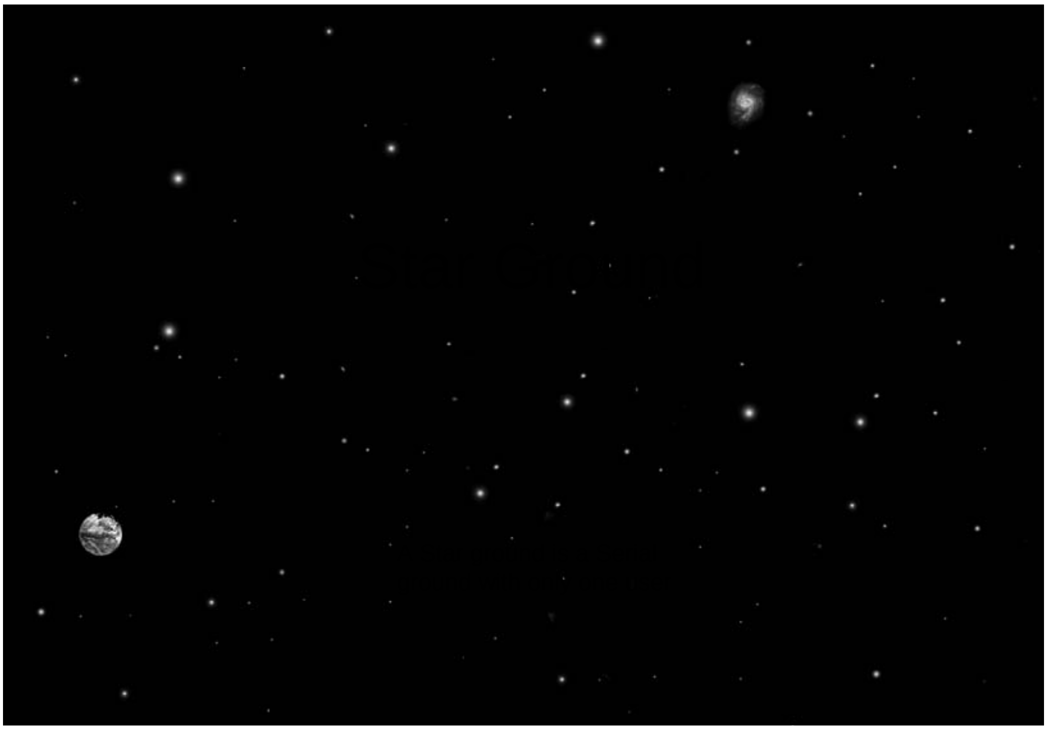
V_{Ground} (rms)



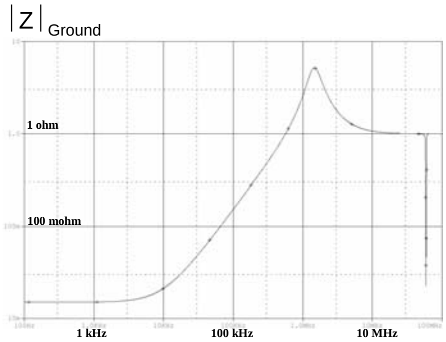
Experimenter ground voltage due to 1mA induced noise

Density probe noise due to poor grounding (Serial ground concept)

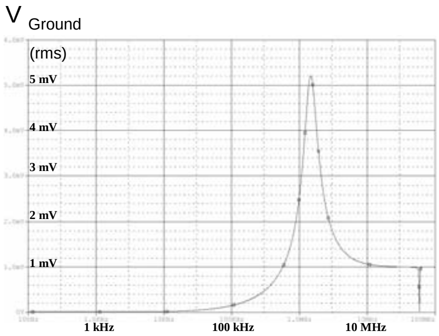




Star ground SPICE simulation



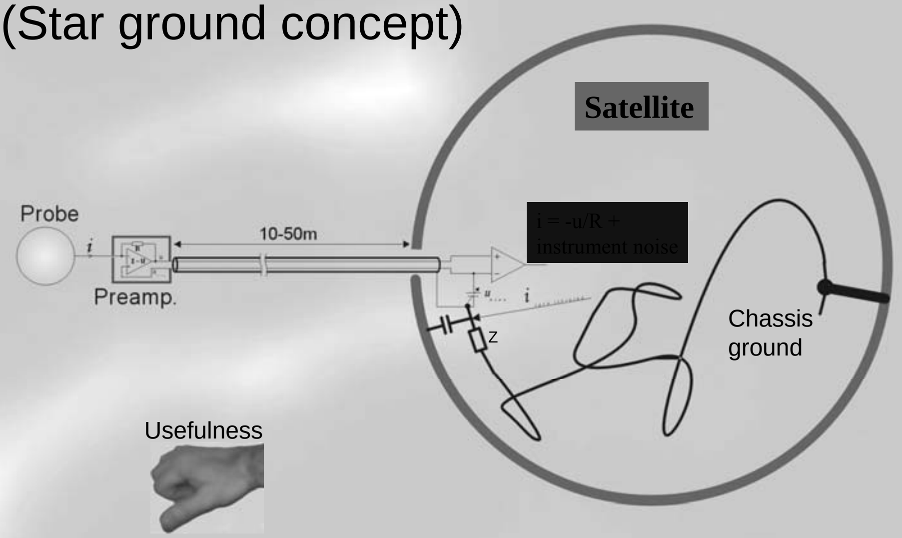
Experimenter ground impedance

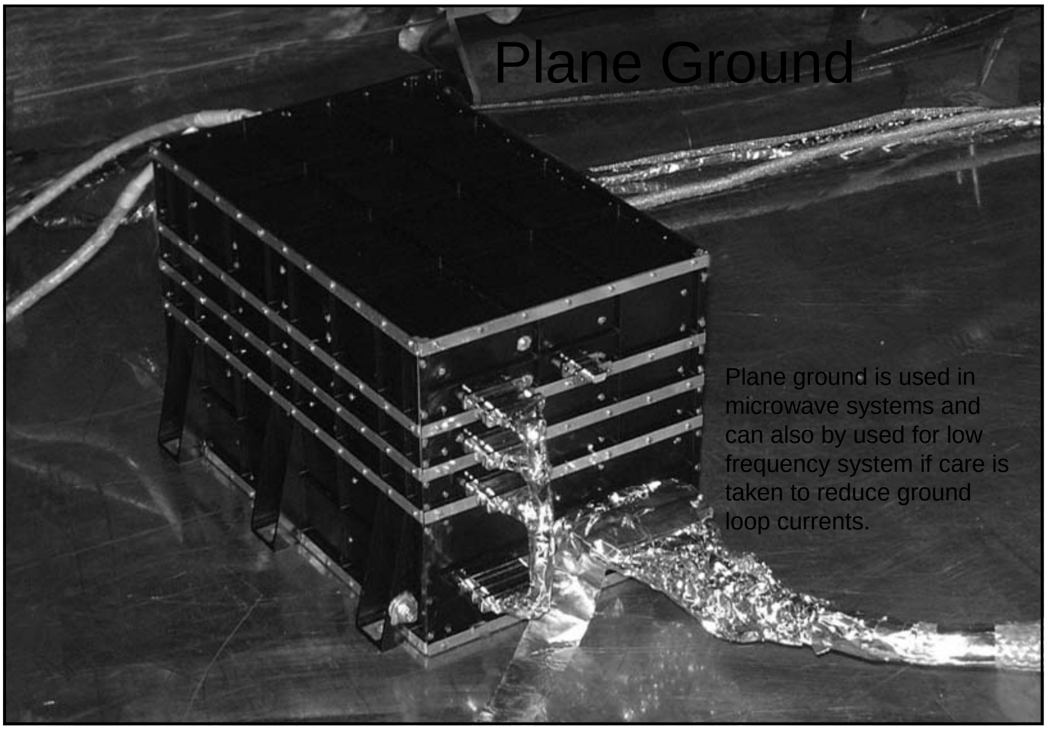


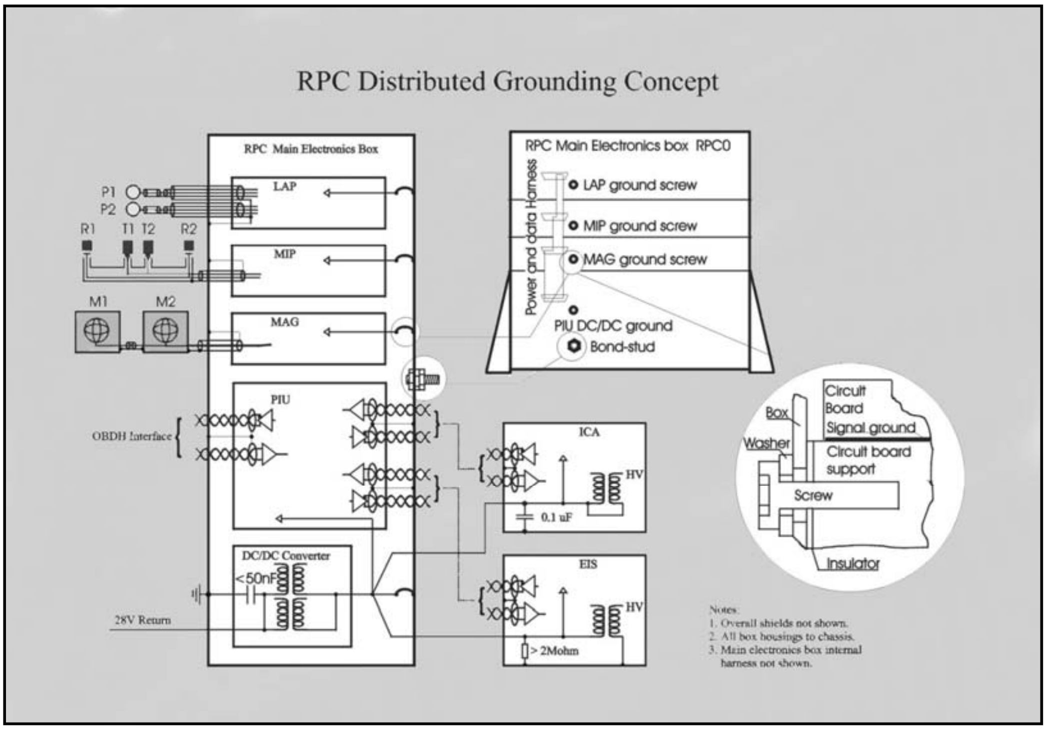
1mA injected at the end of the ground bar

The serial ground bar model including only the end experiment was used for this simulation.

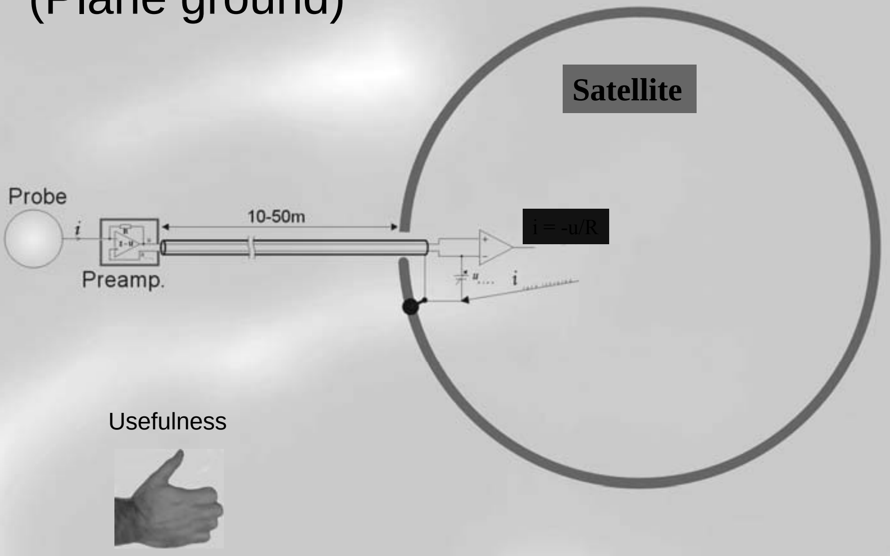
Density probe noise due to poor grounding. (Star ground concept)



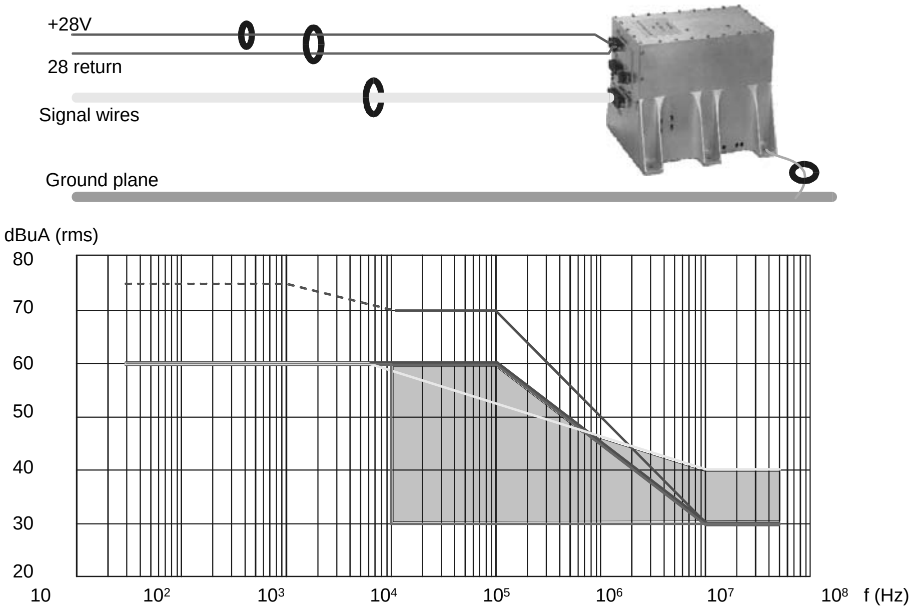

Plane Ground
Plane ground is used in microwave systems and can also by used for low frequency system if care is taken to reduce ground loop currents.


RPC Distributed Grounding Concept
RPC Main Electronics Box
LAP
MIP
MAG
PIU
DC/DC Converter
<50nF
P1
P2
R1
T1
T2
R2
M1
M2
OBDH Interface
28V Return
ICA
EIS
HV
0.1 uF
> 2Mohm
RPC Main Electronics box RPC0
Power and data Harness
LAP ground screw
MIP ground screw
MAG ground screw
PIU DC/DC ground
Bond-stud
Circuit Board
Signal ground
Box
Washer
Circuit board support
Screw
Insulator
Notes:
1. Overall shields not shown.
2. All box housings to chassis.
3. Main electronics box internal harness not shown.

Correct Density probe grounding (Plane ground)



Rosetta Common Mode requirements



Rosetta EMC test Östersund

Rosetta EMC test Östersund

Cluster WEC integration

Grounding

RYP Masters Program
Electronics for Space

Swedish Inst. of Space Physics

2002

Lennart Åhlén

Grounding

The subjects of grounding and bonding on system level have something to do with each other although they deal with different areas: A grounding concept relies upon good bonding and cannot exist without. So these terms should not be mixed in common language use.

Grounding = connecting return conductors of electrical circuits to a reference potential

Bonding = connection of two conducting surfaces in order to provide a good electrical contact

A grounding concept for electronic circuits, assemblies or even systems, serves the purpose

- to avoid circulating EMI due to potential differences between mutually connected electrical units of a system
- to provide an equipotential reference plane
- to prevent common mode coupling
- to avoid low impedance ground loops
- to protect against shock hazards owing to high voltages appearance ESD on a frame or box housing by harness damage,

The term "ground" has been adapted from former times, which understood the earth as a sink for electrical charge. Now it also applies to metallic structures, frames, crates, housings etc. when these individual parts are connected to each other to build up a common potential plane.

Space systems also use the term "ground". They are electrically referenced to the vehicle skin, which acts as reference potential plane. The physical energy budget is maintained by dissipation into space by discharge and thermal and electromagnetic radiation.

The interconnected system of wires, structure elements and boxes is called ground plane potential for many electrical units as it provides the same potential so that no EMI can locally exist at a unit or be coupled from one to another.

Goal: Realize and control a low (zero) impedance plane for all connections including material, bondings, contact pressures, contact area.

Since all impedances are frequency dependant special attention has to be spent on the frequency spectrum used onboard the vehicle and on the inductance of ground leads to the structure earthing point and structure current paths.

Even at zero dc-resistance of a ground bus, for example, there are significant impedances at RF when the geometrical extension of the bus approaches the order of magnitude of some operational signal wavelength λ, ($\lambda/4$ yields isolation!).

Ground plane impedance

The impedance Z between two small areas on a ground plane where connections are made for potential reference is:

Length of ground plane l	$Z = 0.26 * 10^{-6} \sqrt{\frac{\mu f}{\sigma}} \left(1 + \left\lvert \tan 2\pi \frac{l}{\lambda} \right\rvert \right)$
$l < \frac{\lambda}{20}$	$Z = 0.26 * 10^{-6} \sqrt{\frac{\mu f}{\sigma}}$
$l = \frac{\lambda}{8}, 3\frac{\lambda}{8}$	$Z \approx 0.52 * 10^{-6} \sqrt{\frac{\mu f}{\sigma}}$
$l = \frac{\lambda}{4}, 3\frac{\lambda}{4}$	$Z \rightarrow \infty$

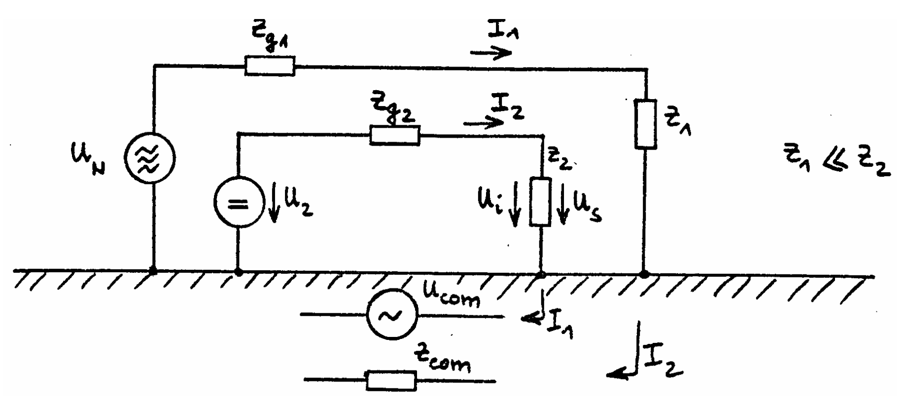
where l= max. extension of ground plane, m
λ= wavelength of interest, m
$\mu = \mu_r \mu_o$ = permeability of ground plane material, VS/Am
f = frequency of interest, Hz
σ = material conductivity of ground plane, 1/Ω m

From this it is obvious that a ground plane need not necessarily act as a common potential plane, especially for $l > \lambda/10$, approximately. Therefore it is important and established as common practice to limit the maximum extension of a ground plane to less than $\lambda/20$-$\lambda/15$. This means that the longest run of interconnection gives the highest frequency to be considered in grounding philosophy.

The ground plane can be a flat conductive area e.g. a honeycomb, interconnect boxes and enclosures to essentially the same electrical reference potential. A ground plane may also consist of a bus bar connected to the return leads of an assembly to form a substartpoint, which is once grounded to another master ground plane etc.

Common mode impedance coupling

Figure below illustrates the fact that there can be a cross talk from one circuit to another via common ground impedance:



U_S = signal voltage at Z_2 due to U_2
U_i = interface voltage across Z_2 due to common mode coupling at $Z_{com;}$ (U_i is is the

voltage across Z_2 at U_2 neglected.

Thus

$$\frac{U_N}{U_{COM}} = \frac{Z_{g1} + Z_1 + Z_{com}}{Z_{com}} \approx \frac{Z_{g1} + Z_1}{Z_{com}} \qquad \text{since } Z_{com} << Z_{g1}, Z_1$$

$$U_{com} = \frac{Z_{com} U_N}{Z_{g1} + Z_1}$$

Furthermore from U_i:

$$\frac{U_i}{U_{com}} \approx \frac{Z_2}{Z_2 + Z_{g2}}$$

Hence $U_i = \frac{Z_2}{Z_2 + Z_{g2}} * \frac{Z_{com} U_N}{Z_{g1} + Z_{14}}$ interference voltage.

Thus it follows, that, if return currents flow through a reference plane it should be of very low impedance.

If individual return wires are used with only one connection to the plane the problem still exists for higher frequencies where distributed capacitive grounding of units becomes effective: i.e. multipoint grounding!

Single point grounding

If the returns of the units of a distributed electronic system are all connected once to the same "star" point or small area of a reference ground plane so that they all are tied to the same potential by low impedance bonds this system is called "single point grounded" (for dc and low frequencies} .In most cases this principle is realized by use of a hierarchical system of substarpoints which allow to build functional assemblies. This is to avoid long ground leads and to maintain low lead impedances:

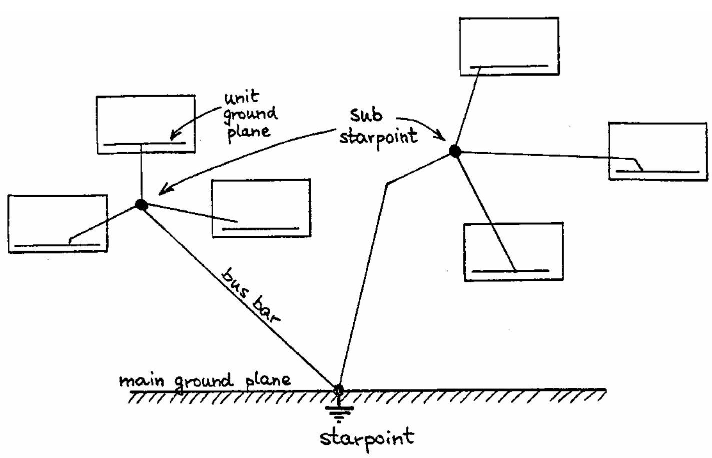


Figure 1.1 Star ground

To well understand, the star "point" in general cannot be a geometrical point but a small area because it has to carry many bonding connections by either soldering, welding, clamping, screwing or using fast-on clip connections which is known as

practical wide spread use in car manufacturing. The unique feature for all different methods is the carefully controlled low impedance quality of the joints.
There are different terms commonly used which designate the central star point and which all mean the same, in essential:

Star point, earth point, earthing plane, grounding plane, grounding point, unipoint ground reference plane, system ground plane, vehicle ground plane etc.

An ideal single point grounding system, usually, cannot be realized because

- distributed parasitic capacitances exist between units, cables and their environment which present grounding paths for higher frequencies

- all ground leads have certain impedances so that a unit may find a more suitable grounding path elsewhere but through the intentional wire
- units are normally interconnected with various types of wires, shields etc. Furthermore some units may be supplied by different manufacturers and be often off-the-shelf products with little or no chance to change the design with respect to grounding philosophy. This is especially problematic if the interconnection cable runs are longer than $\lambda/20$,

- RF equipment per se is designed multipoint grounded with shortest leads possible to earth for every subassembly and use of earth referenced unsymmetrical coax lines. .

It is a fact that for relatively low frequencies a single point approach is operating better than the multipoint version which vice versa shows better performance at higher frequencies (>0.5 MHz). There is a wide application region in between both, called hybrid grounding philosophy, which considers the different equipment and assembly parameters, cable lengths and operating frequencies to establish a compromise for the various sections of a complex electronic system.

In summary, the first approach of a grounding philosophy of a space project is multi point grounding but the hardware realization will be a sort of controlled hybrid grounding.

System level control of grounding concept

The only way to implement a system grounding concept during the development phase (harness manufacturing!) and to keep an access open for any EMC design input during system integration and test verification, is, to establish an overall **grounding and shielding diagram** early in the development phase of a space project and to continuously update it further on. This grounding and shielding diagram

- will reflect the special requirements of sensitive equipment (experiments, sensors),
- considers the geometrical box configuration of the system
- shows harness routing, cable shielding, shield grounding, cable twisting for equally treated groups of wires as they enter or leave a box i.e. for

- power lines (primary and secondary if applicable)
- digital lines)
- analog lines) for control, interconnection,
- RF lines) housekeeping or scientific purposes
- pyro lines)
- clock lines)

It will not show the connector pin allocation of the harness but it has to be established and continuously updated in close cooperation with the harness-manufacturing group, which is usually involved in bookkeeping of the overall allocation.

It will not be a system circuit diagram because it only shows types or classes of wire with respect to their physical interface layout instead of showing each individual line. In this sense a grounding diagram is considered a powerful tool in keeping the technical overview (shadow engineering) during the hardware phase of a space project and to realize the essential ideas of a hybrid-grounding concept.

Ground concept for modern scientific satellites.

To illustrate the above mentioned guidelines the Rosetta grounding concept is presented with regard to those items and features which may also be of interest for other space vehicles. This grounding concept was implemented in view of extremely tight requirements on system noise emissions within the structure and on the harness. The grounding concept is denoted as Distributed Single Point Grounding (DSPG) and an the layout scheme as follows: (see Figure 1.2)
- The primary power shall be connected to the spacecraft ground structure at one point only (star point) within the power subsystem.
- This grounding of primary power shall be done within the PCU of the power S/S.
- All primary power return for users shall be routed to this star point.
- The S/C structure shall not be used as return path for power and signals (except low level signals) to avoid common mode loop effects.

The principle is to realize a single point grounding of each in dependent power network and galvanic isolation between those networks.

Secondary Power:

Secondary power shall be grounded to structure only once in each unit / experiment, see Figure 1.3. The internal grounding star point of the secondary power return shall be as close as possible to the DC/DC converter. It also serves as the signal reference ground of the referring equipment.

This connection to structures shall be performed by a low impedance connection (removable bar, external to the unit) furnished by the experimenter.

The internal grounding shall fulfill the following rules:
- the ground plane shall carry only low level signal return currents in order to minimize magnetic loop effects.
- Secondary return leads shall be directly routed from the corresponding unit and

be connected to the local ground reference point.

- In case that several units are supplied from the same DC/DC converter secondary power output, the distribution shall be a star point system (see Figure 1.1).

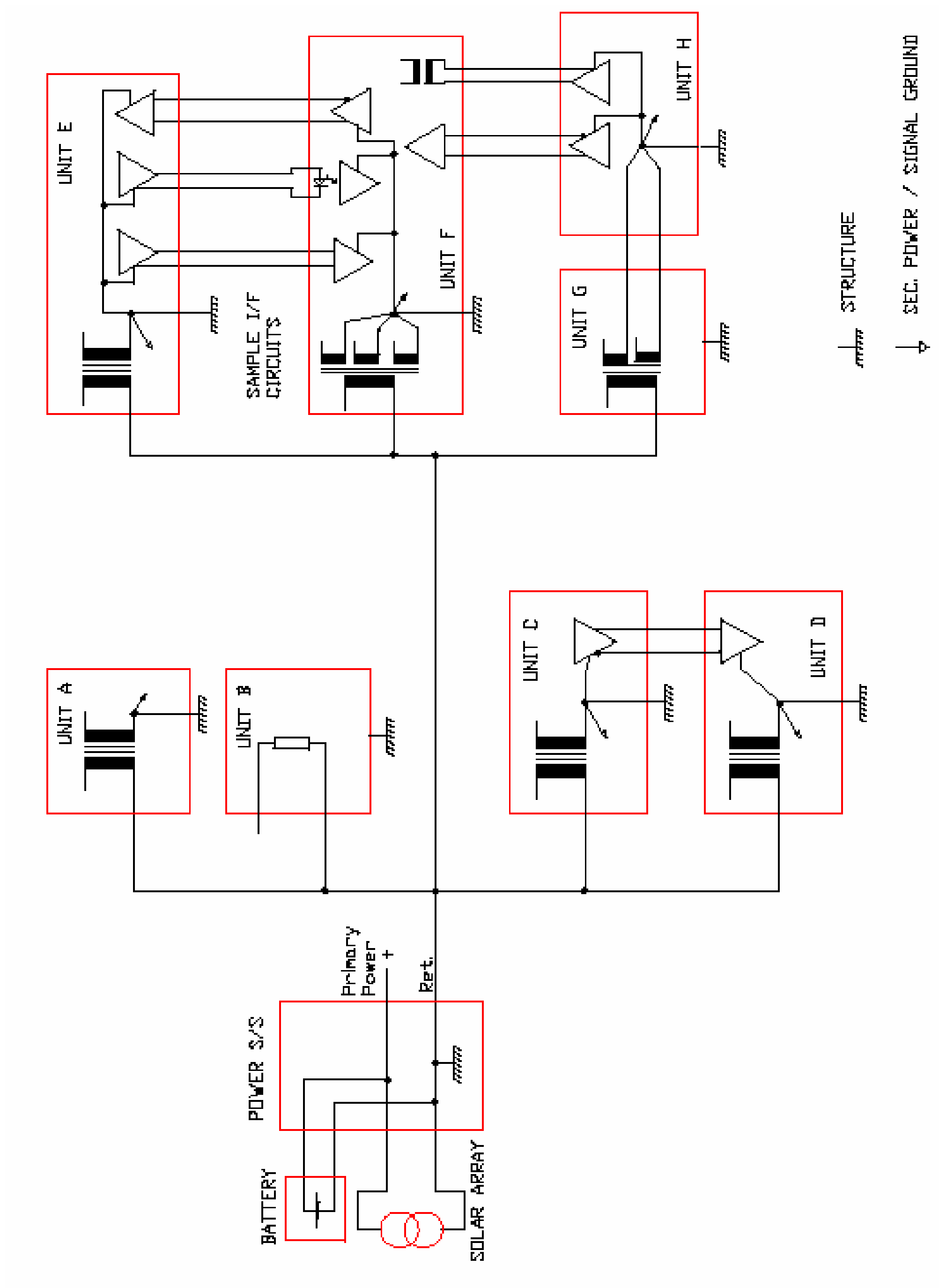


Figure 1.2. ROSETTA Grounding and Isolation Concept

- isolated receivers and balanced differential signals are preferred.
- all deviations from these general grounding requirements shall be notified to the ESA Project Office and shall be agreed prior to implementation.

To establish control of experiment grounding the Experimenter shall define the experiment-grounding diagram in the EID B (See figure 1.5).

The minimum extent of each diagram shall be:
- Primary / secondary power grounding,
- Interface circuit grounding,
- EMI filters,
- Electronic box grounding,
- Principal interface circuit diagram.

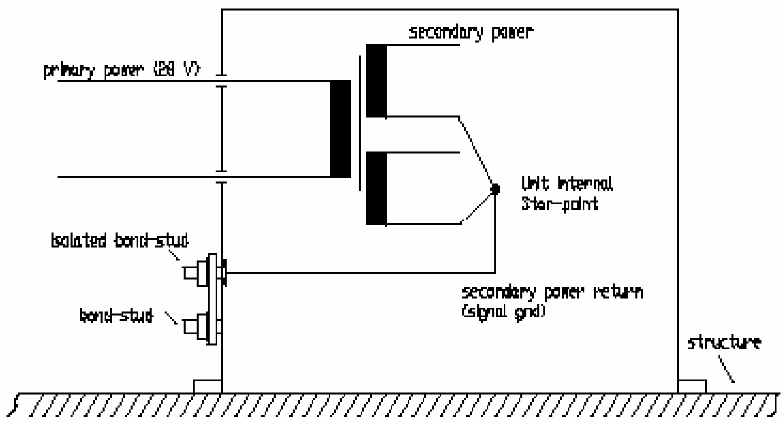


Figure 1.3 outlines the unit grounding.

Signal Interfaces Grounding

Between electrical units all signal driver outputs shall be referenced to ground and all receiver inputs shall be isolated from ground. Signal receivers shall provide common mode rejection and isolation capability. In order to prevent ground loops the differential interface circuits shall be designed to maintain the common mode isolation of Figure 1.4

Preferable solutions of the different interfaces are listed below.

TYPE OF INTERFACE	TRANSMITTER	RECEIVER	CONDITIONS
Analog, unit internal			separate secondary power supply for analog circuits
Analog, unit / subsystem external	balanced	differential	
Digital, unit internal			separate secondary power supply for digital circuits
Digital, unit / subsystem external	single-ended	differential	
	single-ended	opto-coupler	
	single-ended	isol. transformer	
Digital, unit / subsystem external OBDH bus	balanced	differential	
Sync. & clock signals, unit external	balanced	differential	
RF transmission	single-ended	single-ended	coax. cable

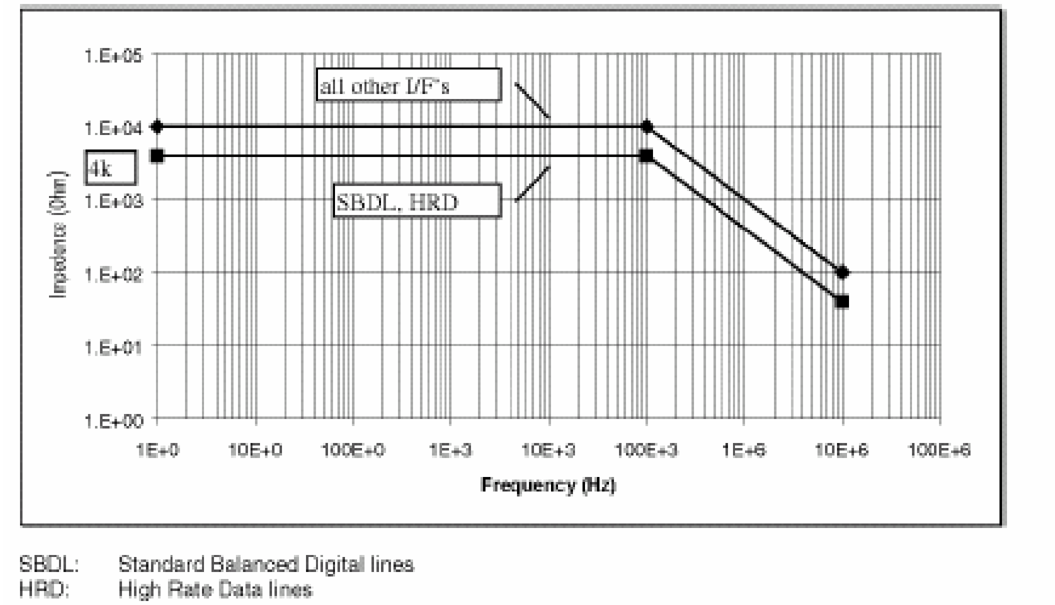


Figure. 1.4 Common Mode Signal Line Isolation

Isolation of Primary Power from Structure

The Structure shall not be used as a return path for DC-power. The isolation measured between power lines and the experiment housing shall be equivalent to:

- a DC resistance of R > 1 MOhm in parallel with
- a capacitance of C < 50 nF per line.

The 50 nF requirement applies directly at the primary power inputs.

Isolation between Primary and Secondary Power

To reduce common mode noise coupling, primary power lines shall be isolated from the secondary power and signal lines. The isolation impedance measured between Primary power return and the Secondary power/signal ground (all external links removed) shall be equivalent to:

- a DC resistance R > 1 MOhm in parallel with
- a capacitance C < 5 nF when one of the two ground reference points is disconnected

The use of static shields between primary and secondary windings of transformers is recommended, in order to reduce the capacitive coupling between primary and secondary side to low values (< 0.1 nF). This static shield should be connected to the primary power return by means of a low inductance strap.

Isolation of Secondary Power from Structure

When disconnected from the ground (external ground bar removed, see Fig. 1.3), the isolation measured between the secondary power return (signal ground) and the experiment housing shall be equivalent to:

- a DC-resistance of R > 1 MOhm in parallel with
- a capacitance of C < 50 nF per unit

Bonding and Case Shielding

Each experiment unit shall be housed in a non-magnetic metallic case, which shall form an electromagnetic shield. The case shall not contain any apertures other than those essential for sensor viewing or outgassing vents. If outgassing vents are required they should be as small as possible (less than 5 mm diameter) and should be located in the case surface which is closest to the experiment mechanical mounting plane (spacecraft structure ground).

For each particular box case all conductive parts shall be bonded to each other either by direct (metal-to-metal) or indirect bonding (via conductive jumper). The resistance between two adjacent unit case parts shall not exceed 2.5 mΩ.
Across movable parts a bonding conductor shall be used to ensure a definite electrical contact between those parts. The DC resistance across such bond straps shall be ≤ 25 mΩ.

All unit cases shall be bonded to spacecraft structure via the equipment box feet. The minimum bonding contact area shall be at least 1 cm² . The resistance between unit case and structure shall be ≤10 mΩ.

Boxes using thermal fillers shall be bonded to spacecraft structure via an adequate bond strap. The resistance between unit case and structure via this bond strap shall be ≤ 10 mΩ.

Each unit shall provide a bonding stud, to enable the bonding test during EMC testing and system integration. The bond stud shall be close to the mounting plane. The resistance between this bond stud and the unit case mounting feet shall be ≤ 2.5 mΩ.
This grounding point should be a stud.

Experiment sensors may require being electrically isolated from the structure. Such units shall have the unit case connected to the experiment secondary power (signal) ground inside the unit. In such a case the sensor may need a special overall shielding box in addition to the sensor housing that is bonded to the spacecraft structure. This shall be subject to agreement with the ESA Project Office.

All outer surfaces exposed to space shall be conductive in order to avoid differential charge build-up resulting in the risk of electrostatic discharge. With respect to electrostatic protection the DC resistance between any other conductive components, which does not perform any electrical function, i.e. CFRP, CFK, conductive coatings etc., and spacecraft structure shall be ≤ 1 kΩ.

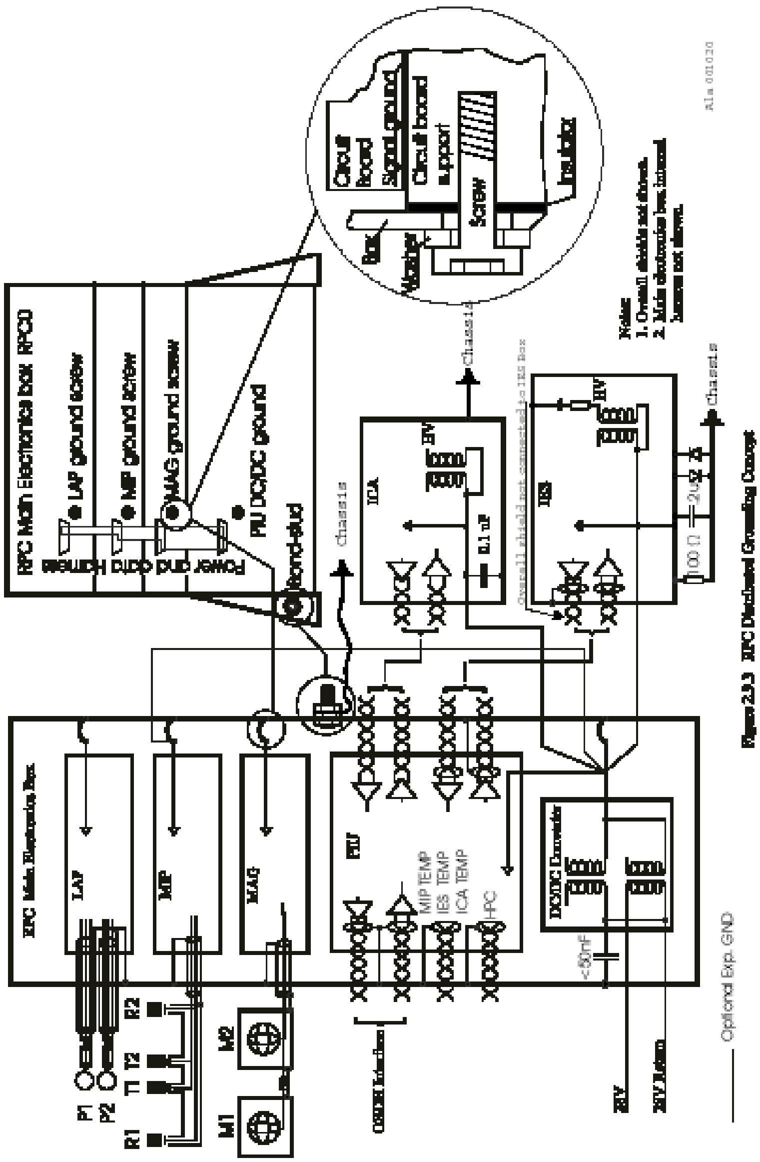
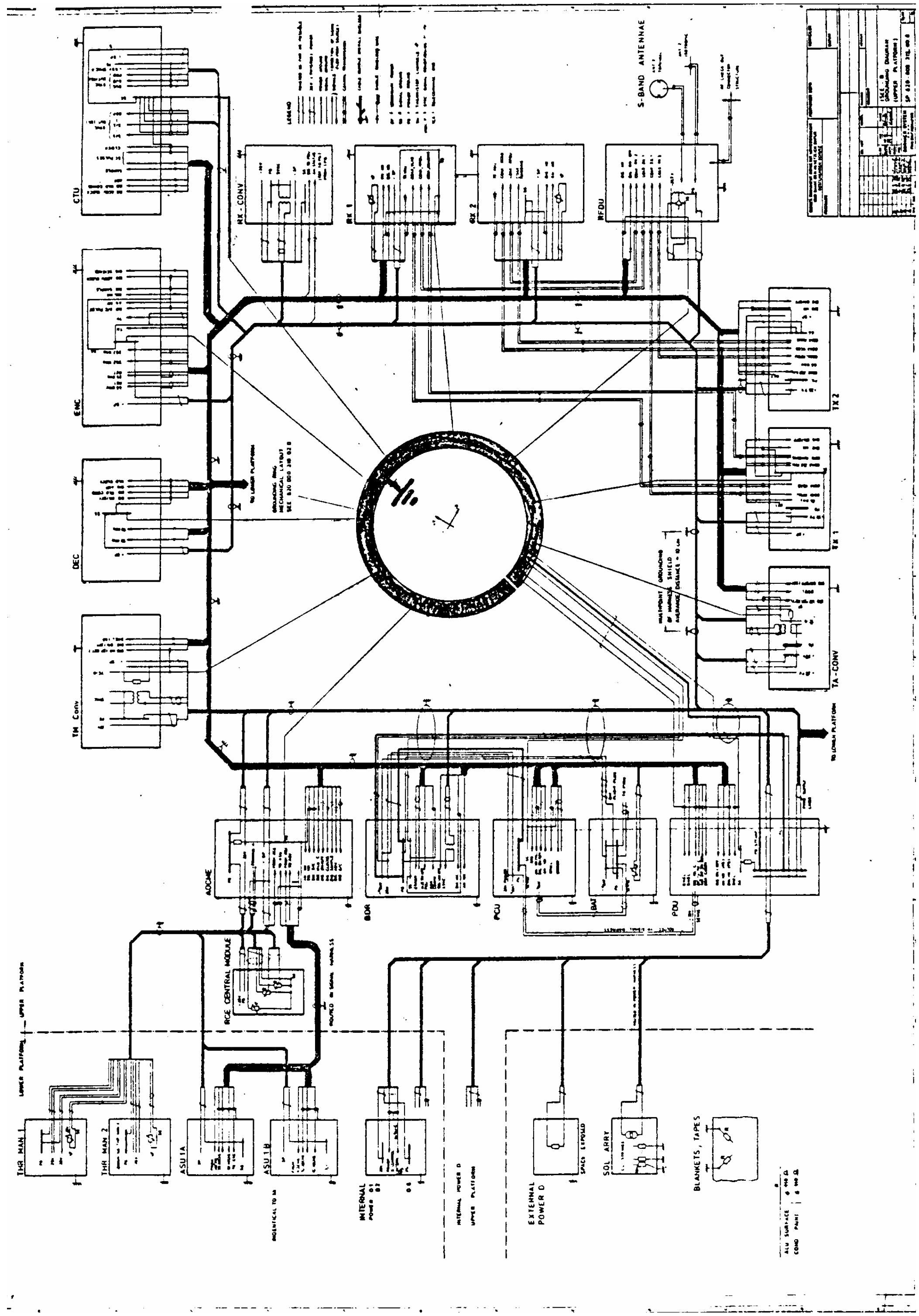
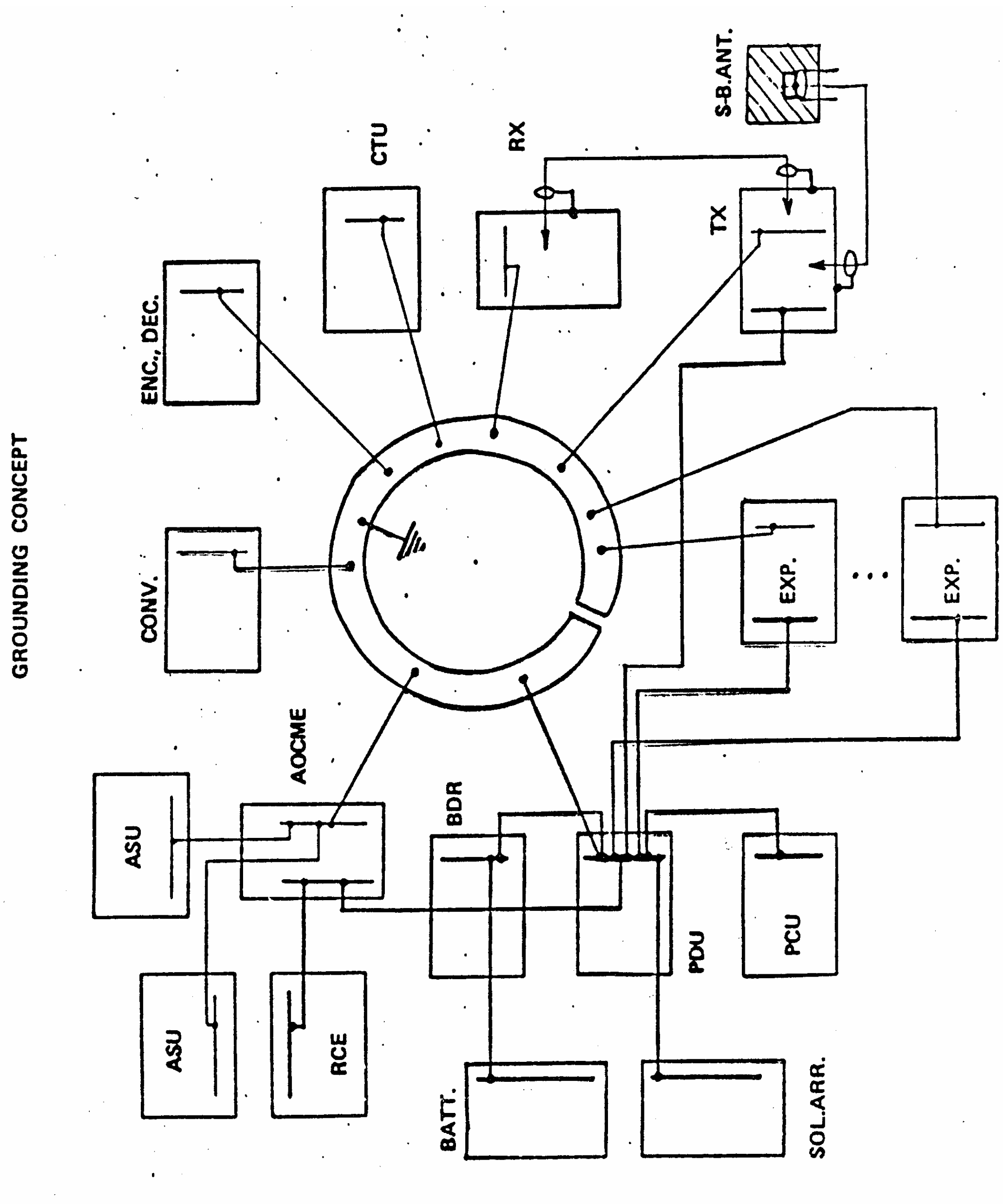


Figure 1.5: RPC Grounding Concept



Cluster Ground Concept



Cluster Grounding Concept

Frequency Control

Some experiments and/or subsystems require defined frequencies and/or guard bands free from on-board generated interferences. To fulfill this requirement a number of specific restricted frequency bands may be identified and will have to be avoided by operating onboard equipment.
Therefore the selection of frequencies for experiments should avoid the critical bands specified in the frequency control plan under all conditions of operation also including identifiable failure modes. All frequencies of primary converters and switching regulators shall be controlled.
A Frequency Plan shall be established as part of the EMC program, with the assistance of the Experimenters. For this purpose the Experimenter shall specify the susceptible and potential emission frequencies in the EID B.

Unit	Source	Frequency	Comments
RPC-1.1 (IES)	Microprocessor	12.5 MHz	
RPC-1.1 (IES)	HV Power Supply	65.5 kHz	
RPC-2.1 (ICA)	Microprocessor	6 or 12 MHz	TBC
RPC-2.1 (ICA)	DC/DC HV supplies	20-60kHz (oscillator at 200-600kHz)	TBC
RPC-3.1 (LAP)	Probe	100Hz-10kHz ±32V, 7kHz-147kHz 150dBuV on short boom probe	
RPC-3.1(LAP)	Microprocessor	6, 8, 12, 24 MHz	TBC
RPC-3.1(LAP)	ADC	3-16 MHz, 6 MHz	TBC
RPC-4.1(MIP)	Transmitter (antenna)	28-224kHz (7kHz res.) 238-448kHz (14kHz res.) 476-896kHz (28kHz res.) 952-1792kHz (56kHz res.) 1904-3472kHz (112kHz res.)	Level below 40mV rms
RPC-4.1 (MIP)	Long Debye Length Mode	7-168kHz (7kHz res.)	Level 32 V rms of 7 kHz sine
RPC-4.1 (MIP)	Microprocessor	14336kHz	
RPC-4.1 (MIP)	Sampling frequency	7168kHz	
RPC-4.1 (MIP)	Data transfer clock	400kHz	
RPC-5.1(MAG1), RPC-5.2(MAG2)	Drive frequency	12.63345kHz up to 5^{th} harmonic (pulses)	
RPC-5.1(MAG1), RPC-5.2(MAG2)	Clock frequency	4.194304 MHz and 419.4394 kHz	
RPC-5.1(MAG1), RPC-5.2(MAG2)	FPGA & ADCs	32768Hz	
RPC-6.0 (PIU)	Microprocessor	4-8MHz	
RPC-6.0 (PIU)	1355-interface	100-400kHz	
RPC-6.0 (PIU)	DC/DC-conv.	65.5kHz	
RPC-6.0 (PIU)	Switch Signal	100 kHz	

RPC operating frequency bands

Shielding

RYP Masters Program
Electronics for Space

Lecture notes

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1. Waves in materials.

For a better understanding of shielding effects, we now start upon looking at electromagnetic waves in material media. In conductive medium free charges are present, which generate a current under the influence of the wave electrical field. The current $\overrightarrow{J_C}$ is related to $\vec{E}$ through ohms law.

$$\overrightarrow{J_C} = \sigma \vec{E} \tag{1.1}$$

The material my also have specific relative values for ε and μ.

$$\varepsilon = \varepsilon_r \varepsilon_o \quad \mu = \mu_r \mu_o \tag{1.2}$$

Maxwell's equations in phasor notation become.

$$\nabla \times \vec{E} = -j\omega\mu\overrightarrow{H} \tag{1.3}$$

$$\nabla \times \overrightarrow{H} = \sigma\vec{E} + j\omega\varepsilon\vec{E} \tag{1.4}$$

The wave equation for the electrical field is given by.

$$\begin{aligned} &\nabla \times \nabla \times \vec{E} = \nabla\nabla \bullet \vec{E} - \nabla^2 \vec{E} = -j\omega\mu\nabla \times \overrightarrow{H} = -j\omega\mu\left(\sigma + j\omega\varepsilon\right)\vec{E} \\ &\Rightarrow \quad \nabla^2 \vec{E} = j\omega\mu\left(\sigma + j\omega\varepsilon\right)\vec{E} \end{aligned} \tag{1.5}$$

We have set the net charge density to zero, despite the conductivity in the media, so that the divergence of the electrical field is zero.

In one dimension, Uniform Plane Wave (UPW) with $\frac{\partial \overrightarrow{E_x}}{\partial x} = \frac{\partial \overrightarrow{E_x}}{\partial y} = \frac{\partial \overrightarrow{H_y}}{\partial x} = \frac{\partial \overrightarrow{H_y}}{\partial y} = 0$ and $\vec{E} = E_x\left(z,t\right)$, the wave equation is simply

$$\frac{d^2 \overrightarrow{E_x}}{dz} = j\omega\mu\left(\sigma + j\omega\varepsilon\right)\overrightarrow{E_x} = \gamma^2 \overrightarrow{E_x} \tag{1.6}$$

The general solution to that equation is of a simple form.

$$E_x\left(z\right) = \overline{E}_m^+ e^{-\gamma z} + \overline{E}_m^- e^{\gamma z} \tag{1.7}$$

$H_y(z)$ can now easily be obtained from Eq. (1.4)

$$H_y(z) = -\frac{1}{j\omega\mu}\frac{dE_x}{dz} = \sqrt{\frac{\sigma + j\omega\varepsilon}{j\omega\mu}}\left(\overline{E}_m^+ e^{-\gamma z} + \overline{E}_m^- e^{\gamma z}\right) = \frac{1}{\overline{\eta}}\left(\overline{E}_m^+ e^{-\gamma z} + \overline{E}_m^- e^{\gamma z}\right) \tag{1.8}$$

These solutions are in agreement with the voltage and current in lossy transmission lines.

The intrinsic impedance η of the medium is defined as.

$$\overline{\eta} = \eta e^{j\tau} = \sqrt{\frac{j\omega\mu}{\sigma + j\omega\varepsilon}} \tag{1.9}$$

The intrinsic impedance is complex as long as the conductivity is not zero.

For the propagation constant γ, one can obtain the Re and Im part as.

$$\gamma = \sqrt{j\omega\mu\left(\sigma + j\omega\varepsilon\right)} = \alpha + j\beta \tag{1.10}$$

$$\alpha = \frac{\omega\sqrt{\mu\varepsilon}}{\sqrt{2}}\sqrt{\sqrt{1+\left(\frac{\sigma}{\omega\varepsilon}\right)^2} - 1} \tag{1.11}$$

$$\beta = \frac{\omega\sqrt{\mu\varepsilon}}{\sqrt{2}}\sqrt{\sqrt{1+\left(\frac{\sigma}{\omega\varepsilon}\right)^2} + 1} \tag{1.12}$$

The quantity α is referred as the attenuation constant, and the quantity β is referred to as the phase constant. Considering only the forward wave solutions E_x and H_y can be written as.

$$E_x(z) = \overline{E}_m^+ e^{-\alpha z} e^{-j\beta z} \tag{1.13}$$

$$H_y(z) = \frac{1}{\eta}\overline{E}_m^+ e^{-\gamma z - j\tau} = \frac{1}{\eta}\overline{E}_m^+ e^{-\alpha z} e^{-j\beta z} e^{-j\tau} \tag{1.14}$$

Writing the complex undetermined constant $\overline{E}_m^+$ as a magnitude and angle as $\overline{E}_m^+ = E_m^+ e^{j\theta}$ and using the time domain form $E_x = \mathrm{Re}\left\{\overline{E}_x e^{j\omega t}\right\}$ gives.

$$E_x\left(z,t\right) = E_m^+ e^{-\alpha z}\cos(\omega t - \beta z + \theta) \tag{1.15}$$

$$H_y\left(z,t\right) = \frac{1}{\eta}E_m^+ e^{-\alpha z}\cos(\omega t - \beta z + \theta - \tau) \tag{1.16}$$

1.1 Classification of materials.

Lossless Media.

It is important to study the properties of these equations. To simplify the analysis we will start with UPW in a lossless media $\sigma = 0$. For this case the propagations constants become.

$$\alpha = 0 \quad \beta = \omega\sqrt{\varepsilon_r \varepsilon_o \mu_r \mu_o} \tag{1.17}$$

Since $\sigma = 0$, the wave sees no attenuation as it propagate in the medium. The intrinsic impedance becomes.

$$\eta = \sqrt{\frac{j\omega\mu}{j\omega\varepsilon}} = \sqrt{\frac{\mu_r \mu_o}{\varepsilon_r \varepsilon_o}} \tag{1.18}$$

In free space $\mu_r = \varepsilon_r = 1$ the intrinsic impedance is $\eta_0 = \sqrt{\mu_o / \varepsilon_o} = 120\pi = 377\Omega$.

For lossless media the field vector becomes.

$$E_x\left(z,t\right) = E_m^+ \cos(\omega t - \beta z + \theta)) \tag{1.19}$$

Thus we observe that a point on the waveform must move in the z direction for increasing time, so that.

$$\omega t - \beta z + \theta = \text{constant} \tag{1.20}$$

The derivative of (1.20) with respect to *t* will give the phase velocity of the wave:

$$v_p = \frac{dz}{dt} = \frac{\omega}{\beta} = \frac{1}{\sqrt{\varepsilon_r \varepsilon_o \mu_r \mu_o}} \tag{1.21}$$

The propagations constant $\beta = \omega\sqrt{\varepsilon_r \varepsilon_o \mu_r \mu_o}$ also referred to as the phase constant, which units are radians per meter. As the wave propagates in the media β is the change in phase. The distance between adjacent points is the wavelength λ. We know that $\beta\lambda = 2\pi$. Since $\beta = \omega\sqrt{\varepsilon_r \varepsilon_o \mu_r \mu_o}$ and $v = 1 \big/ \sqrt{\varepsilon_r \varepsilon_o \mu_r \mu_o}$ for this lossless medium, the wavelength becomes.

$$\lambda = \frac{2\pi}{\beta} = \frac{v_p}{f} = \frac{1}{f\sqrt{\varepsilon_r \varepsilon_o \mu_r \mu_o}} \tag{1.22}$$

Lossy Media.

Imperfect dielectrics with $\sigma \neq 0$ but $(\sigma/\omega\varepsilon) \ll 1$ gives the following key parameter equations.

$$\gamma = \sqrt{j\omega\mu(\sigma + j\omega\varepsilon)} = j\omega\sqrt{\varepsilon\mu}\sqrt{1 - j\frac{\sigma}{\omega\varepsilon}} \approx \frac{\sigma}{2}\sqrt{\frac{\mu}{\varepsilon}} + j\omega\sqrt{\mu\varepsilon} + \ldots.. \tag{1.23}$$

$$\alpha \approx \frac{\sigma}{2}\sqrt{\frac{\mu}{\varepsilon}} \quad \beta \approx \omega\sqrt{\varepsilon\mu} \tag{1.24}$$

$$v_p = \frac{\omega}{\beta} \approx \frac{1}{\sqrt{\varepsilon\mu}} \quad \lambda = \frac{2\pi}{\beta} \approx \frac{1}{f\sqrt{\mu\varepsilon}} \tag{1.25}$$

$$\eta = \sqrt{\frac{j\omega\mu}{\sigma + j\omega\varepsilon}} = \sqrt{\frac{j\omega\mu}{j\omega\varepsilon}}\left(1 - j\frac{\sigma}{\omega\varepsilon}\right)^{-1/2} \approx \sqrt{\frac{\mu}{\varepsilon}} \tag{1.26}$$

The rule of thumb is that the above approximations for imperfect dielectric can be applied when.

$$\frac{\sigma}{\omega\varepsilon} \leq 0.1$$

When the condition above is verified, the imperfect dielectric behaves as a perfect dielectric, except for a small attenuation term in the fields.

Good conductor with $\sigma \neq 0$ but $(\sigma/\omega\varepsilon) \gg 1$ gives the following key parameter equations.

$$\gamma = \sqrt{j\omega\mu(\sigma + j\omega\varepsilon)} \approx \sqrt{j\omega\mu\sigma} = \sqrt{\omega\mu\sigma}\left(\frac{1}{\sqrt{2}} + j\frac{1}{\sqrt{2}}\right) = \sqrt{\pi f\mu\sigma}(1 + j) \tag{1.27}$$

$$\alpha = \sqrt{\pi f\mu\sigma} \quad \beta = \sqrt{\pi f\mu\sigma} \tag{1.28}$$

$$v_p = \frac{\omega}{\beta} \approx \sqrt{\frac{4\pi f}{\mu\sigma}} \quad \lambda = \frac{2\pi}{\beta} \approx \sqrt{\frac{4\pi}{f\mu\sigma}} \tag{1.29}$$

$$\eta = \sqrt{\frac{j\omega\mu}{\sigma + j\omega\varepsilon}} \approx \sqrt{\frac{j\omega\mu}{\sigma}} = \sqrt{\frac{\pi f\mu}{\sigma}}(1 + j) \tag{1.30}$$

The rule of thumb is that the above approximations for good conductors can be applied when.

$$\frac{\sigma}{\omega\varepsilon} \geq 10$$

For a good conductor α and β are approximately equal. The medium impedance η has nearly equal Re and Im parts, therefore its phase angel is approximately 45^o. This means that E and H fields have always a phase difference $\tau = 45^o$.

As the wave propagate through the lossy medium, its amplitude $E_m^+ e^{-\alpha z}$ will decrease. Over a distance of $\delta = 1/\alpha$ it will be reduced by $1/e$ or 37%. The quantity δ is named the skin depth of the medium at that frequency.

$$\frac{1}{\alpha} = \delta = \frac{1}{\sqrt{\pi f \mu \sigma}} \tag{1.31}$$

Values of the skin depth for some materials are given in the table 1.1.The formula for the skin depth is:

$$\delta = 66.1 * k / \sqrt{f}$$

Where δ is the skin depth in mm, f is the frequency, and k is a function of material with copper as reference $k = 1$,

Table 1.1 Relative skin depth.

Material	k	
Aluminum	1.26	
Copper	1.00	
Lead	3.52	
Silver	0.94	
Tin	2.55	
Tungsten	1.76	
Brass	1.99	(depends on alloy and temp.)
Phos-bronze	2.1	
Bronze	3.1	

2. Shielding:

A shield is a conductive enclosure that fully or partly encloses an electrical system. For a shield to be effective, it must completely enclose the system that is to be protected. A shield serves as a barrier to electromagnetic fields. That in order to prevent both emissions from inside the shield end external fields to effect system inside or outside the shield. Figure 2.1 illustrate both cases of interference. The shielding effectiveness is the ratio of the incident field to a shielding barrier to the magnitude of fields that is transmitted through the barrier. The shield effectiveness S can be expressed in dB as.

$$S\left(dB\right)=20\log_{10}\frac{incident \quad field}{transferred \quad field} \tag{2.1}$$

This definition is good for boxes and large systems, but is not used for configurations like shielded cables. The total shielding effectiveness is given by $S = S_R \bullet S_A \bullet S_{MR} \bullet S_{AL}$, where S_R, S_A, S_{MR} are shielding factors due to reflections, absorption, and multiple reflections.

2.1 Barrier impedance of metals.

The barrier impedance of a good conductor is given by Eq. (1.30)

$$\eta_b = \eta_m = \sqrt{\pi f \mu/\sigma}\left(1+j\right)$$

this equation is predicted upon the material barrier thickness, t , being many skin depths, i.e., $t \gg \delta$. In order to establish the applicability of that equation, a somewhat arbitrary situation is selected in which $t \le 3\delta$. At $t = 3\delta$, 95% of the current flows in the material and 5% of the current flows beyond the thickness of the material. Thus the barrier impedance must be 5% greater than Eq. (1.30). To accommodate any metal thickness, a generalization of Eq. (1.30) is required.

$$\eta_b = \frac{\eta_m}{\left(1-e^{-t/\delta}\right)} \qquad \text{for } t/\delta \text{ any value} \tag{2.2}$$

$$\eta_b \simeq \frac{\eta_m}{\left[1-\left(1-t/\delta\right)\right]} = \frac{\eta_m \delta}{t} \qquad \text{for } t/\delta \ll 1 \tag{2.3}$$

2.2 Reflection losses.

A general plane wave reflection/transmission problem consists of an incident plane wave impinging on a multilayer structure. An important special case is the two region problem shown in Figure. 2.1. Both media are considered being infinite in thickness and any of the media may have arbitrary amount of loss and $\eta_1 \gg \eta_2$. The vector representing the direction for the incident wave in Figure 2.1 is a real vector.

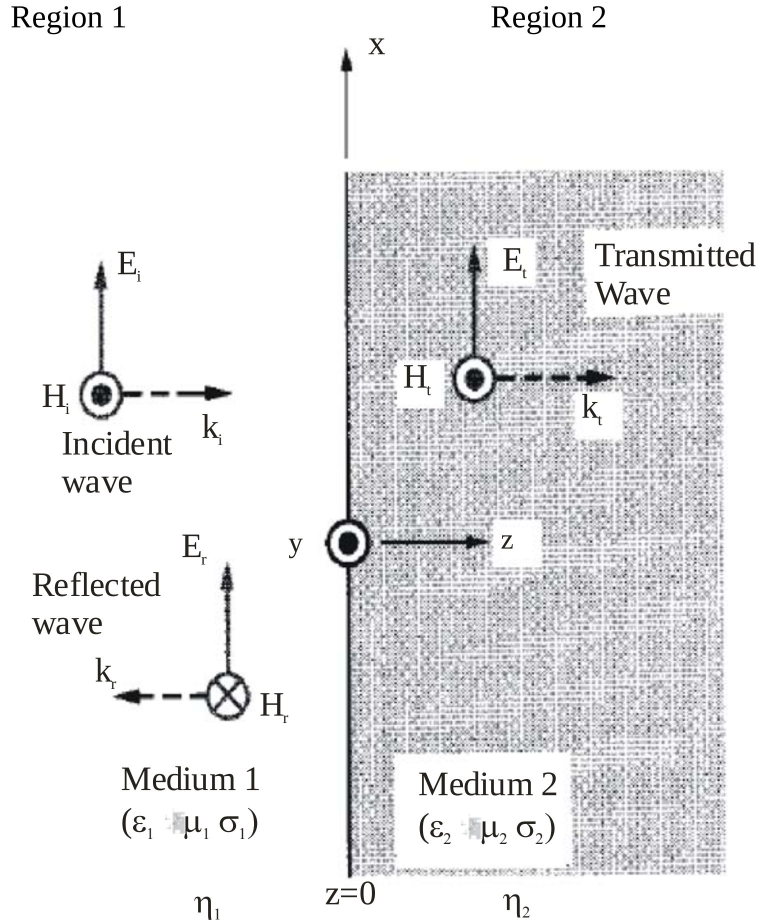

Figure. 2.1 Reflection of UPW with normal incidence.

The wavenumber vector is defined as $k = \omega\sqrt{\varepsilon\mu(1 - j\,\sigma/\omega\varepsilon)}$ and the intrinsic impedance is $\eta = \sqrt{\mu/(\varepsilon(1 - j\,\sigma/\omega\varepsilon))}$

In medium 1.

$$E_{x1}(z) = E_{x1}^{+}(z) + E_{x1}^{-}(z) = E_{o1}^{+}e^{-jk_1 z} + E_{o1}^{-}e^{jk_1 z} \tag{2.4}$$

$$\begin{array}{cc} \uparrow & \uparrow \\ \text{incident} & \text{reflected} \\ \text{wave} & \text{wave} \\ \downarrow & \downarrow \end{array}$$

$$H_{y1}(z) = H_{y1}^{+}(z) + H_{y1}^{-}(z) = \frac{E_{o1}^{+}}{\eta_1}e^{-jk_1 z} - \frac{E_{o1}^{-}}{\eta_1}e^{jk_1 z} \tag{2.5}$$

In medium 2.

$$E_{x2}(z) = E_{x2}^{+}(z) = E_{o2}^{+}e^{-jk_2 z} \tag{2.6}$$

$$\begin{array}{c} \uparrow \\ \text{transmitted wave} \\ \downarrow \end{array}$$

$$H_{y2}(z) = H_{y2}^{+}(z) = \frac{E_{o2}^{+}}{\eta_2}e^{-jk_2 z} \tag{2.7}$$

At the boundary at $z = 0$, tangential electric and magnetic fields must be continuous, we get.

$$\begin{cases} E_{x1}(0) = E_{x2}(0) & \Rightarrow \quad E_{o1}^{+} + E_{o1}^{-} = E_{o2}^{+} \\ H_{y1}(0) = H_{y2}(0) & \Rightarrow \quad \dfrac{E_{o1}^{+}}{\eta_1} - \dfrac{E_{o1}^{-}}{\eta_1} = \dfrac{E_{o2}^{+}}{\eta_2} \end{cases} \tag{2.8}$$

If we assume the incident fields are known, we can solve the above equations to get.

$$\frac{E_{o2}^{+}}{E_{o1}} = \frac{2\eta_2}{\eta_1 + \eta_2} \overset{Defined\ \ as}{=} \tau_E \quad \text{(Transmission coefficient)} \tag{2.9}$$

$$\frac{E_{o1}^{-}}{E_{o1}^{+}} = \frac{\eta_2 - \eta_1}{\eta_2 + \eta_1} \overset{Defined\ \ as}{=} \Gamma_E \quad \text{(Reflection coefficient)} \tag{2.10}$$

We find.

$$\tau_E = 1 + \Gamma_E \tag{2.11}$$

Rewrite $E_{o1}^{+} + E_{o1}^{-} = E_{o2}^{+}$ in terms of H and we get.

$$H_{o1}^{+}\eta_1 + H_{o1}^{-}\eta_1 = H_{o2}^{+}\eta_2 \tag{2.12}$$

By using Eqs (2.8) and (2.12) we will arrive to the following equations for the reflected and transmitted H fields.

$$\frac{H_{o2}^{+}}{H_{o1}} = \frac{2\eta_1}{\eta_1 + \eta_2} = \tau_H \tag{2.13}$$

$$\frac{H_{o1}^{-}}{H_{o1}^{+}} = \frac{-E_{o1}^{-}/\eta_1}{E_{o1}^{+}/\eta_1} = -\Gamma_E = \frac{\eta_1 - \eta_2}{\eta_2 + \eta_1} = \Gamma_H \tag{2.14}$$

Standing wave ratio (SWR) is defined as.

$$SWR = \frac{1+\Gamma}{1-\Gamma} \tag{2.15}$$

Special cases

Matching impedances $\eta_1 = \eta_2 \Rightarrow \Gamma_E = 0 \Rightarrow \tau_E = 1 \Rightarrow No \quad reflection \quad SWR = 1$

Medium 2 = perfect conductor $\sigma_2 = \infty \Rightarrow \eta_2 = 0 \Rightarrow \Gamma_E = -1 \quad and \quad \tau_E = 0$

The wave undergoes a total reflection. This is in analogue with a short circuit transmission line. The electrical field at the interface is

$$E_{o1}^{+} + E_{o1}^{-} = E_{o1}^{+} + \Gamma_E E_{o1}^{+} = E_{o1}^{+} - E_{o1}^{+} = E_{o2}^{+} = 0$$

Because of interference between incident and reflected wave, there will be a standing wave in medium 1 see Figure. 2.2.

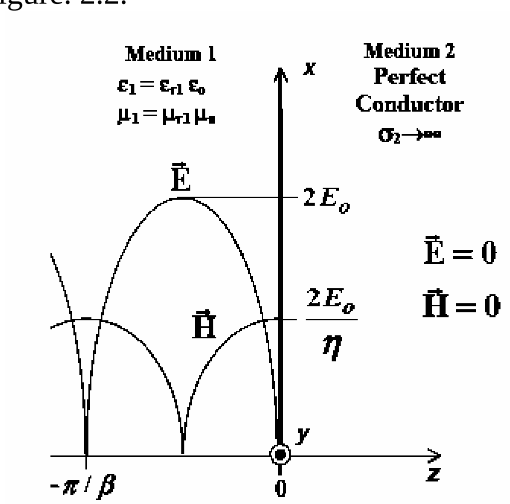

Figure. 2.2 Reflection of UPW in a perfect conductor.

2.2.1 Reflection losses of fields by a metal plate.

We will now find an expression for the ratio of fields transmitted through a metallic plat in between two identical dielectrics see Figure 2.3 below. Only single reflection is used in this analyze $t \gg \delta$. This expression gives the shielding effectiveness S_R .

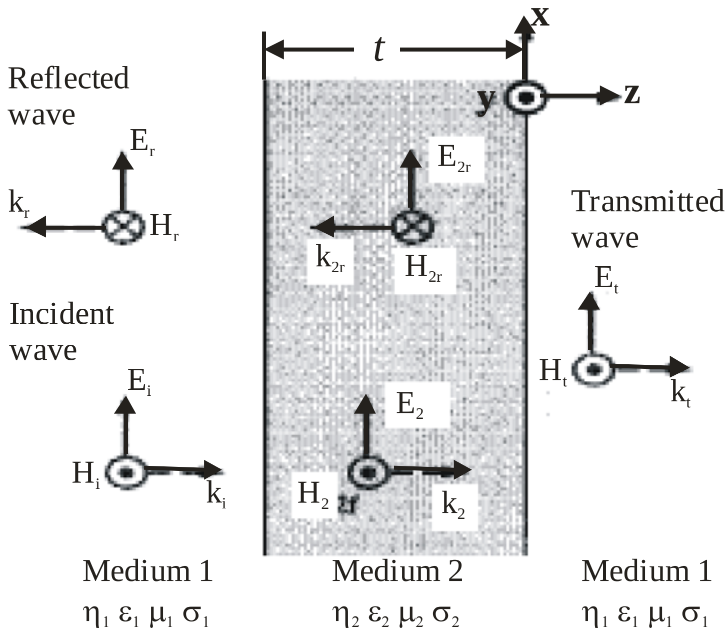

Figure. 2.3 Transmission of fields through a shielding plate with thickness t.

For the $E-$field Eq (2.9) can be used twice to get the transmitted field.

$$\frac{E_i}{E_t} = \frac{2\eta_2}{\eta_1+\eta_2}\frac{2\eta_1}{\eta_1+\eta_2} = \frac{4\eta_1\eta_2}{(\eta_1+\eta_2)^2} = S_E \tag{2.16}$$

For the magnetic field

$$\frac{H_t}{H_i} = \frac{2\eta_1}{\eta_1+\eta_2}\frac{2\eta_2}{\eta_1+\eta_2} = \frac{4\eta_1\eta_2}{(\eta_1+\eta_2)^2} = S_H \tag{2.17}$$

If $\eta_1 \gg \eta_2$ we get

$$\frac{H_t}{H_i} = \frac{E_t}{E_i} = \frac{4\eta_2}{\eta_1} = S_R \tag{2.18}$$

The shielding effectiveness in dB is

$$S_{RdB} = 20\log\frac{4\eta_1\eta_2}{(\eta_1+\eta_2)^2} \overset{\eta_1\gg\eta_2}{\approx} 20\log\frac{4\eta_2}{\eta_1} \tag{2.19}$$

Ratio of transmitted and incident magnetic and electrical fields is identical. Primary transmission of magnetic field is left boundary and for the electrical field it is the right boundary.

2.3 Absorption loss:

A field is reduced at the right interface of a conductive plate from its value at the incident interface by the absorption factor. Figure 2.4 illustrate the field and current density decrease in a lossy material.

$$S_A = e^{\gamma t} = e^{\frac{t}{\delta}} = e^{t\sqrt{\pi f \sigma\mu}} \quad S_{AdB} = 20\log e^{\frac{t}{\delta}} = 15t\sqrt{f}\sqrt{\sigma\mu} \tag{2.20}$$

This is valid for $\eta_2 \ll \eta_1 \quad and \quad t \gg \delta$

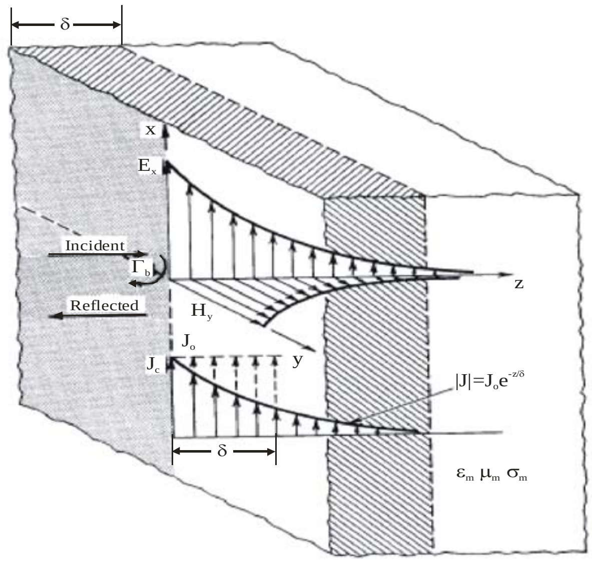

Figure 2.4 Electric and magnetic field intensities, and electric current density distributions in lossy media.

Some metals and their associated conductivity and permeability are listed in Table 2.1. The column, entitled, $\sqrt{\mu\sigma}$ is ranking of the latent absorption loss of metals relative to copper. For non magnetic metals, except silver, all $\sqrt{\mu\sigma}$ values relative to copper are less than 1. Those metals provide relatively poor absorption loss. All magnetic metals, on the other hand, have relative absorption loss values exceeding two and are relatively good absorbers of energy at low frequencies compared to non magnetic metals. On the other hand, since the relative permeability degrades with frequency, magnetic metals offer absorption losses less than most non magnetic metals at higher frequencies (above approximately 100kHz).

Table 2.1 Relative conductivity and permeability of metals.

Metal	σ_r	μ_r @ $\leq 10kHz$	$\sqrt{\sigma_r \mu_r}$
Silver	1.064	1	1.03
Copper	1	1	1
Gold	0.7	1	0.88
Aluminum	0.63	1	0.78
Brass	0.47	1	0.69
Magnesium	0.38	1	0.61
Tin	0.151	1	0.39
Lead	0.079	1	0.28
Supermalloy	0.023	100.000	53.7
Purified Iron	0.17	5.000	29.2
Mumetal	0.0289	20.000	24.0
50% Nickel, Iron	0.0384	1.000	6.2
Commercial Iron (0.2 impure)	0.17	200	5.38
Steel	0.17	180	5.53
Nickel	0.23	100	4.7
Stainless Steel (1Cu,18Cr,8Ni,&Fe)	0.02	200	2.0

2.4 Multiple reflection loss:

Multiple reflection loss affects magnetic fields more than electric fields. If the shield is thin, $t \ll \delta$, the reflected magnetic component from the second boundary is re-reflected off the first boundary and again returns to the second boundary to be reflected, as shown in Figure 2.5. For electric fields almost of the incident wave is reflected at the first boundary and multiple reflections can be neglected. For magnetic fields, on the other hand, most of the incident wave passes into the shield and from Eq (2.13) we see that the amplitude is almost doubled and in that case multiple reflections inside the shield must be considered.

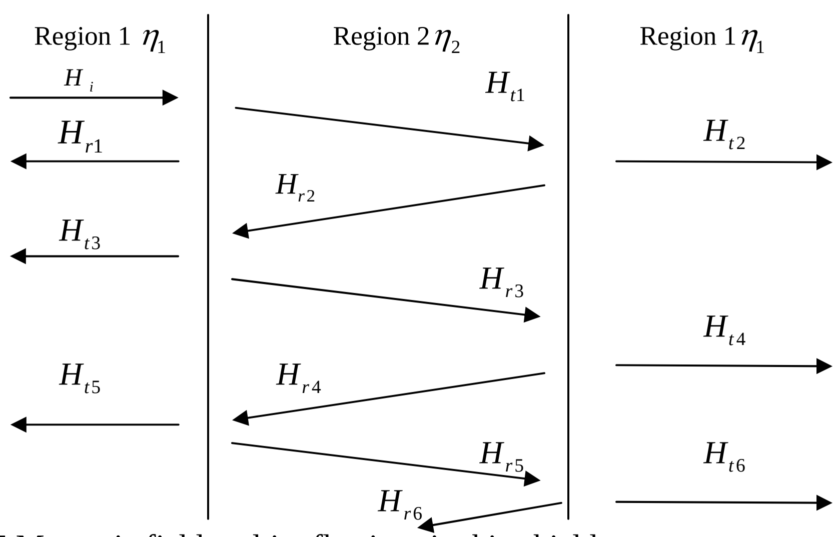

Figure 2.5 Magnetic field multi reflections in thin shields.

We assume that $t \ll \delta$, $\eta_1 \gg \eta_2$ and the phase shift is in the shield can be neglected. Under these conditions, the total transmitted wave can be written as

$$H_{t(total)} = H_{t2} + H_{t4} + H_{t6} + \ldots\ldots\ldots \tag{2.21}$$

From Eq (2.13) we get

$$H_{t2} = \frac{2\eta_1 H_i}{\eta_1 + \eta_2}\left(e^{-t/\delta}\right)\tau_H \tag{2.22}$$

We can now write for H_{t4}

$$H_{t4} = \frac{2\eta_1 H_i}{\eta_1 + \eta_2}\left(e^{-t/\delta}\right)\tau_H\left(1-\tau_H\right)\left(e^{-t/\delta}\right)\left(1-\tau_H\right)\left(e^{-t/\delta}\right) \tag{2.23}$$

which reduce to

$$H_{t4} = \frac{2\eta_1 H_i}{\eta_1 + \eta_2}\left(e^{-3t/\delta}\right)\left(\tau_H - 2\tau_H^2 + \tau_H^3\right) \tag{2.24}$$

For a metallic shield $\tau_H \ll 1$ and $\tau_H^2 \ll \tau_H$ and $\tau_H^3 \ll \tau_H$, etc. The total transmitted wave can be written as

$$H_{t(total)} = 2H_i\tau_H\left(e^{-t/\delta} + e^{-3t/\delta} + e^{-5t/\delta} + \ldots\ldots\right) \tag{2.25}$$

The infinite series has it limit.

$$e^{-t/\delta} + e^{-3t/\delta} + e^{-5t/\delta} + \ldots\ldots = \frac{1}{2\sinh(t/\delta)} \tag{2.26}$$

We now get

$$\frac{H_i}{H_{t(totalt)}} = \left(\frac{\eta_1}{4\eta_2}\right)2\sinh\left(\frac{t}{\delta}\right) \tag{2.27}$$

The shield effeteness is

$$S_{(dB)} = 20\log\left(\frac{\eta_1}{4\eta_2}\right) + 20\log\left(2\sinh\left(\frac{t}{\delta}\right)\right) \tag{2.28}$$

the first term is the reflection loss S_R. To calculate the correction factor S_{MR} we must transfer the equation in to the form of $S_{(dB)} = S_{A(dB)} + S_{R(dB)} + S_{MR(dB)}$. The second term must therefore be equal to $S_{A(dB)} + S_{MR(dB)}$. Thus, we can write

$$S_{MR(dB)} = 20\log\left[2\sinh\left(\frac{t}{\delta}\right)\right] - S_{A(dB)} = 20\log\left[2\sinh\left(\frac{t}{\delta}\right)\right] - 20\log e^{-t/\delta} \tag{2.29}$$

Expressing the $\sinh\left(t/\delta\right)$ as an exponential, gives S_{MR} as

$$S_{MR(dB)} = 20\log\left(1 - e^{-2t/\delta}\right) \tag{2.30}$$

2.5 Shielding effectiveness in the near field.

In order to see the importance for the shield effectiveness between fare and near fields we will start to look at the wave impedance for an electric (Hertzian) dipole and a magnetic (Loop) dipole. The E and H field components are given below Figure 2.6.

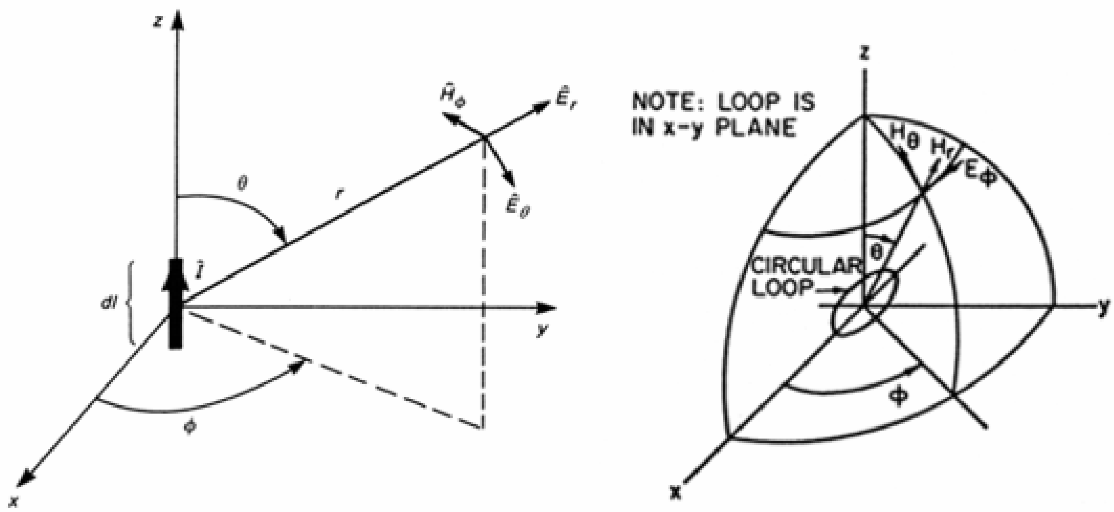

Figure 2.6 E and H fields from a small electrical and magnetic dipole.

$$E_r = 2\frac{Idl}{4\pi}\beta^2\eta_o\cos\theta\left(\frac{1}{(\beta r)^2} - \frac{j}{(\beta r)^3}\right)e^{-j\beta r} \qquad H_\theta = j\frac{\omega\mu_o Idm}{4\pi\eta_o}\beta^2\sin\theta\left(\frac{j}{(\beta r)} + \frac{1}{(\beta r)^2} - \frac{j}{(\beta r)^3}\right)e^{-j\beta r}$$

$$E_\theta = \frac{Idl}{4\pi}\beta^2\eta_o\sin\theta\left(\frac{j}{(\beta r)} + \frac{1}{(\beta r)^2} - \frac{j}{(\beta r)^3}\right)e^{-j\beta r} \qquad H_r = j2\frac{\omega\mu_o Idm}{4\pi\eta_o}\beta^2\cos\theta\left(\frac{1}{(\beta r)^2} - \frac{j}{(\beta r)^3}\right)e^{-j\beta r}$$

$$H_\phi = \frac{Idl}{4\pi}\beta^2\sin\theta\left(\frac{j}{(\beta r)^2} + \frac{1}{(\beta r)^3}\right)e^{-j\beta r} \qquad E_\phi = -j\frac{\omega\mu_o Idm}{4\pi\eta_o}\beta^2\sin\theta\left(\frac{1}{(\beta r)^2} - \frac{j}{(\beta r)^3}\right)e^{-j\beta r}$$

$$E_\phi = H_r = H_\theta = 0 \qquad E_r = H_\phi = E_\theta = 0$$

Where $\beta = 2\pi/\lambda_o$ and $\eta_o = \sqrt{\mu_o/\varepsilon_o}$

The wave impedance is obtained from the ratio of E to H and for an electric dipole it is

$$Z_w = \frac{j/\beta r + 1/(\beta r)^2 - j/(\beta r)^3}{j/\beta r + 1/(\beta r)^2} \xrightarrow{\beta r \ll 1} |Z_w|_e = \frac{1}{2\pi f\varepsilon_o r} = 60\frac{\lambda_o}{r} \qquad (2.31)$$

The electric dipole near field wave impedance is grater than the intrinsic impedance of the media. In the near field the electrical field is proportional to $1/r^3$ while the magnetic field is proportional to $1/r^2$. The magnitude of the wave impedance for a small electrical and magnetic dipole is shown in Figure 2.7. In the fare field the impedance is proportional to $1/r$ for both components, giving $Z_w \cong \eta_o$.

The magnetic loop wave impedance is

$$Z_w = -\eta_o \frac{j/\beta r + 1/(\beta r)^2}{j/\beta r + 1/(\beta r)^2 - j/(\beta r)^3} \xrightarrow{\beta r \ll 1} |Z_w|_m = 2\pi f \mu_o r = 2370 \frac{r}{\lambda_o} \qquad (2.32)$$

Magnetic field in the near field region is proportional to $1/r^3$ while the electric field is proportional to $1/r^2$. Also, in the near field of a magnetic loop the wave impedance is less than the intrinsic impedance. There for the magnetic loop is referred to as a low impedance source.

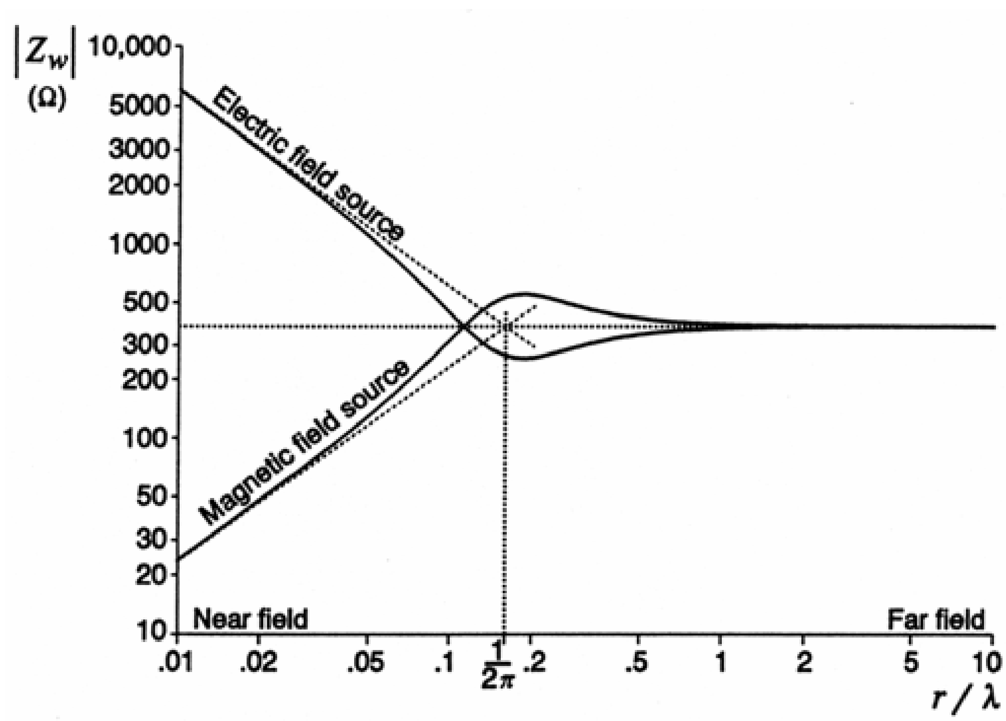

Figure 2.7 Wave impedance vs. Distance/Wavelength

The wave impedance plays a crucial role for the reflection losses as the loss is a function of the ratio between the wave impedance and the shield impedance. A low impedance magnetic field, therefore, has lower reflection loss than a plane wave. The changes caused by alternating the source – shield distance are illustrated in Figure 2.8. The main practical inference that can be drawn is that it is worthwhile situate the shield as far away as practical in the case where magnetic shielding is desired. A high impedance electric field has higher reflection loss than a plan wave and both E curves, in Figure 2.8, show that high impedance waves are easy to screen. Thin layer $(L = 0.01 - 1\mu m)$ of conducting material have useful shielding properties provide a shielding effectiveness of between 20-100dB is all that is required.

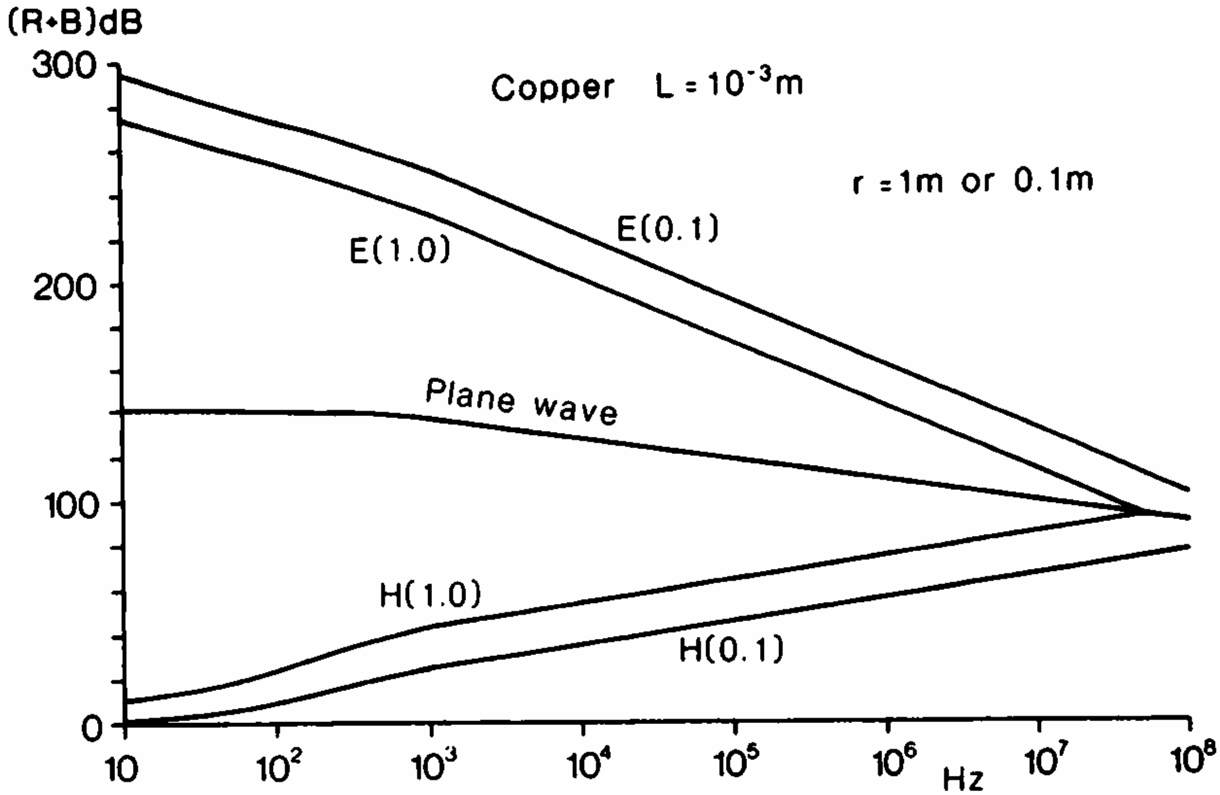

Figure 2.8 Net reflection losses from the inner surface of a 1mm thick copper sphere as the function of frequency , showing the effect of changing the source – shield from 0.1m to 1m . In the figure the curve labeled E(0.1) and E(1.0) refer to the net reflection for a high impedance E wave for the two source distances, and the curves labeled H(0.1) and H(1.0) refer to the net reflection for a low impedance H wave.

2.6 The effect of apertures

Making shielding predictions using reflection and absorption losses indicate that 200dB attenuation is easily achievable using reasonable thicknesses of common materials. In fact, the practical shielding effectiveness is not entirely determined by material characteristics but is limited by necessary apertures and discontinuities in the shielding. You will need apertures for ventilation, for control and interface access, and for viewing indicators; seams, that is, discontinuities at the joints between individual conductive members, act also as apertures. Also, shielding is almost invariably applied in the near field of the circuits inside an enclosure. The theoretical material- and field impedance-related attenuation is merely an upper bound on what is achievable, and much lower values are found in practice.

The problem of calculating the fields penetrating an aperture in a conducting plate is usually dealt with by computing the equivalent magnetic and electric dipoles which when situated in the aperture generate the required fields. These dipoles depend on the shape of the aperture and the nature of the exciting fields. Detailed studies indicate that the in most situations magnetic field penetration is the dominant process creating the far field generated by the excited aperture. It is possible to estimate the effect of holes in an otherwise impermeable conducting shield by using the standard formula for the gain G of an aperture illuminated by a plane wave.
For a circular aperture of radius a:

$$G = \frac{P_2}{P_1} = \left(\frac{2\pi a}{\lambda}\right)^2 \tag{2.33}$$

Then the shielding effectiveness SE is

$$S_{AP} = 10\log\left(P_1/P_2\right) = 20\log\left(\lambda/2\pi a\right) \tag{2.34}$$

As the G is proportional to area we can writ for n apertures

$$S_{AP} = 20\log\left(\lambda/2\pi a\sqrt{n}\right) \tag{2.35}$$

In order to allow for the fact that real shields have finite thickness it is necessary to calculate the extra attenuation caused by each aperture when it is considered to be a length of waveguide operating beyond cut-off. The total shielding effectiveness due to n circular waveguide of length t is

$$S_{AP} = 20\log\left(\lambda/2\pi a\sqrt{n}\right) + 32t/2a \tag{2.36}$$

This formula only really applies for guides of length t > a.

The first term can be considered to represent the reflection loss and the second term the absorption loss of the aperture. Figure 2.9 shows this function (for t =1 mm) superimposed on an S curve for a typical screened box. The curve for n = 2.5 x 10^5 simulates what might be expected for a room with mesh walls (a = 0.5 mm, holes on a 2 mm pitch, wall size 1 m x 1 m). Here it is assumed that n is set by the number of apertures in one side only (the side facing the impinging radiation). Results are also given in Figure 2.9 for an infinite wire mesh screen calculated using an equivalent circuit theory for the shielding effectiveness due to Casey. In this case it is assumed that the wires are 1 mm in diameter and that their axes are 2 mm apart.

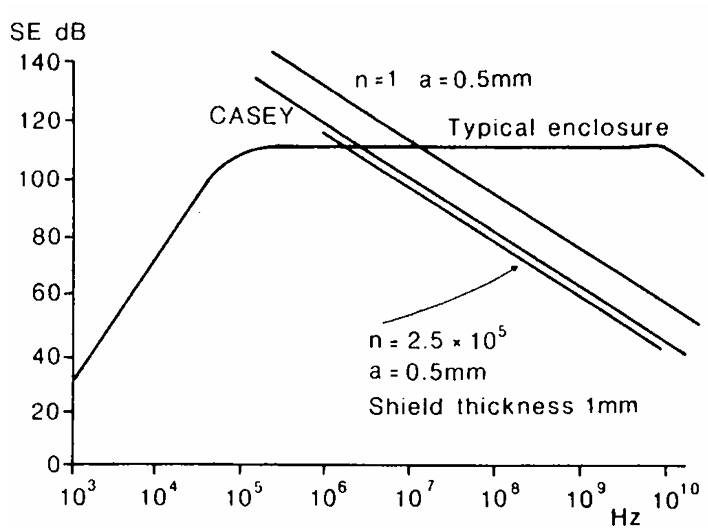

Figure. 2.9 Effect of a single n =1, or an array, n = 2.5 x 10^5, circular apertures of 0.5 mm radius on the shielding effectiveness of a typical (screened) enclosure, calculated using Equation 4.43. The line labelled CASEY refers to results for an infinite wire mesh screen of 1 mm wires having 2 mm spacing.

Any rectification aperture can be made to have a higher screening effectiveness by turning it into a length of waveguide operating beyond cut-off. Such a system is shown in Figure 2.10.

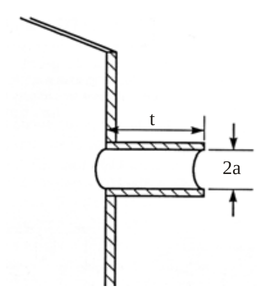

Figure 2.10 Cross section of a hole formed into a waveguide with diameter 2a and length t.

So far only circular apertures have been considered. An important type of aperture in practice is a slot (length 1 > width w). Simple arguments show that a slot is much more important than a circular aperture of the same area. By realizing that magnetic leakage is usually most important one can use the circuit model of shielding to picture what is happening. For effective magnetic shielding the induced shield currents have to flow in precisely the correct place in order to cancel out the impinging field. Thus as is shown schematically in Figure 2.11, a slot causes far more current deviation than a single aperture or even a row of apertures. The orientation of the slot with respect to the current flow is also important as is clear from comparing Figure 2.11(c) with 2.11(d). At high frequencies the slot can act as an efficient slot antenna (see Figure 2.12). For the least favorable orientation (the one needed for worst case design**),** the shielding effectiveness due to the presence of a slot in an impervious shield only involving the slot length rather than the area and the with a shield thickness of t is

$$S_{AP} = 20\log\left(\lambda/2l\right) + 27\,t/l \qquad \text{for } t>l \tag{2.37}$$

where 1= the slot length and t is the shield thickness.

Once again the first term represents a reflection loss (for a slot in a thin sheet) and the second the absorption loss.

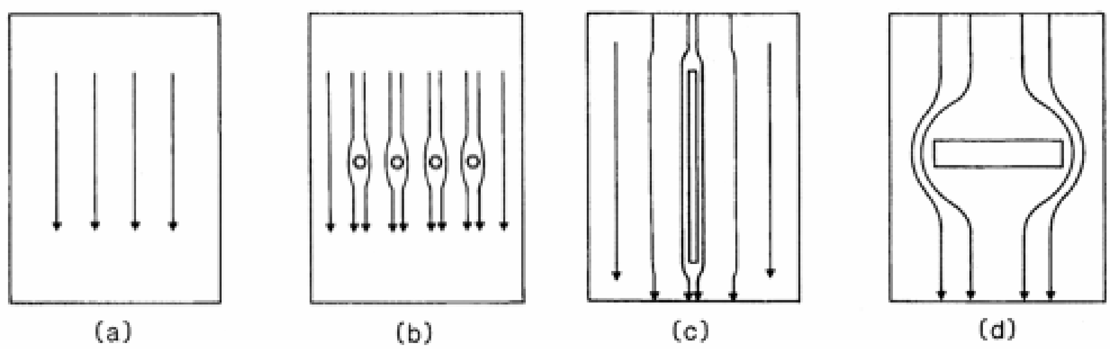

Figure 2.11 Schematic diagrams showing the effects of slots and holes on shield surface currents.

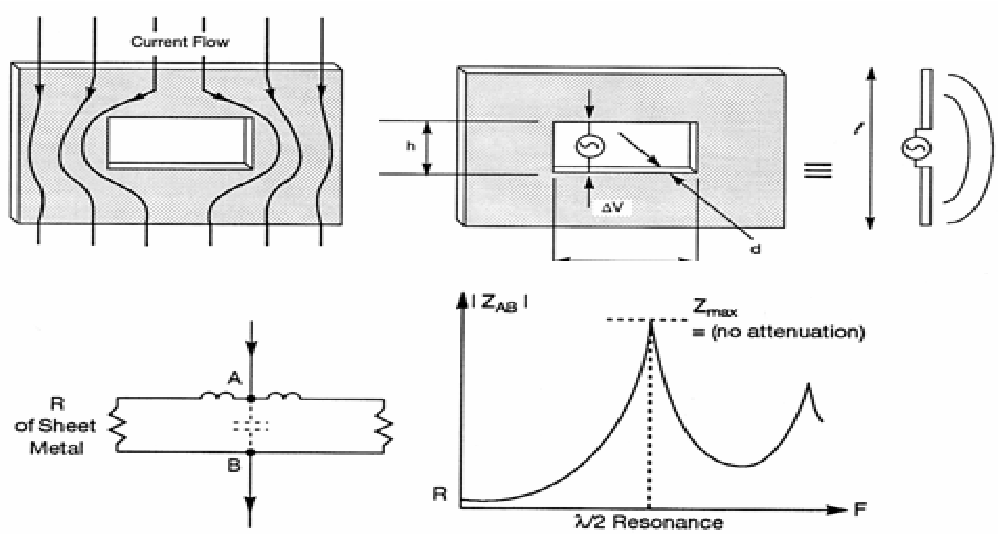

Figure 2.12 Effects of a slot in a shield. At low frequencies, the slot is approximately a short: reflection is significant.

One can use Eq (2.37) to examine the behavior of some joint designs that might be met in real situations. Slots are often found in practice when two pieces of metal have to be joined and welding (or continuous soldering) is impracticable.

Figure 2.12 shows two idealized designs, the first of which in Figure 2.12 (a) consists of just butting two thin sheets (t=1 mm) together and holding them in that position by screw fixings every so often along the seam. Even if the sheets are very carefully made the regions away from the fixing points will tend to separate as shown in Figure 2.12(b). Such small separations constitute slots. The total length of such slots will depend on the quality of construction and the frequency of fixing locations. Figure 2.13 shows calculated values of S for such a joint design with different assumed final slot lengths compared with the typical screened enclosure shielding effectiveness. It is clear that slot apertures have to be kept to a minimum if a reasonable performance is required (it is usually recommended that 1 be kept to less than $\lambda/50$). Fortunately the problem can be helped to a large extent by creating thicker regions near the joint by, for example, bending the sheet through a right angle before bolting or screwing through the overlapping region. This is equivalent to increasing t as is shown in Figure 2.12 (c). The effect of doing this on the calculated values of SE is shown in Figure 2.13 for t = 2.5 cm.

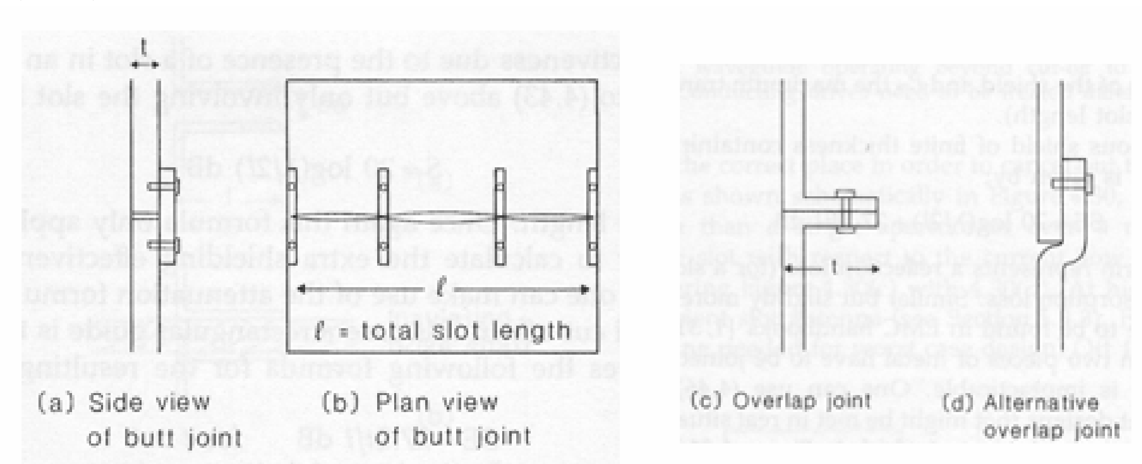

Figure 2.12 Critical dimensions of some joint designs.

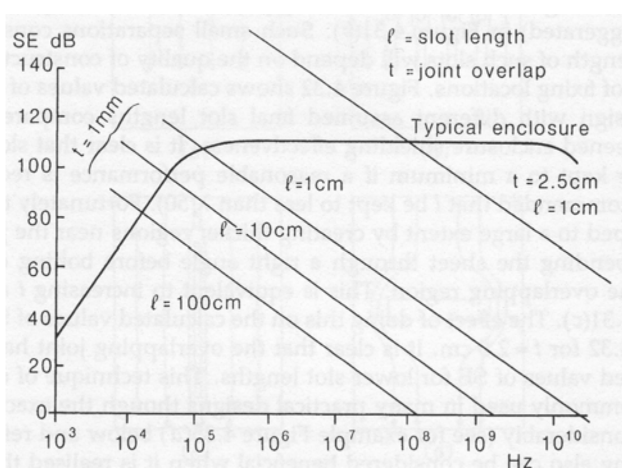

Figure 2.13 Effects of various different length slots of length *l* with overlap *t* on the shield effectiveness as a function of frequency of a typical enclosure.

Besides overlapping a number of other techniques have to be employed. One common solution is to use gaskets. A gasket is a deformable conducting interface between the two materials to be joined. The gasket may be a metallic mesh or a conducting elastomer.

Finally it should be noted that the formulae given in this section are approximate, apply only to plane wave excitation and give results for the far field of the aperture. Therefore their use for other types of excitation or in the near field could result in serious error.

2.6 Alternative ways to describe shield quality.

A possible way to calculate en measure shield quality is the use of the transfer impedance (and admittance) which describes cable shielding reasonably adequately. We shell have a look at this concept for cables and connectors. But it could as well be used for metallic enclosures. In the 1930's Shelkunoff showed that surface transfer impedance (Z_t) was the intrinsic electromagnetic shielding property of cables connectors and back shells. The transfer impedance Z_t for a cable shield that is electrically small in diameter is defined as

$$Z_t = \frac{1}{I_o}\frac{dV}{dz} \tag{2.38}$$

where I_o = the external current on the shield, $dV/dz =$ the open circuit voltage per unit length (x) produced on the internal surface of the shield, and $Z_t =$ the transfer impedance/unit length. If the system is smaller than a wavelength, it is electrically small and all voltages, currents, and fields are the same throughout the system and the system may be analyzed using lumped parameters. This is usually true when the length is less than a tenth wavelength. For connectors, V is a point source

$$Z_t = V_C/I_o \tag{2.39}$$

where V_C is the open circuit voltage on the inside of the shield. Current on one side of the barrier produces voltage on the other side of the barrier due to impedance of the barrier. At low frequencies, the impedance is a resistance due to current diffusion and contact resistance. At high frequencies, the impedance is mutual inductance due to apertures, etc. The effect of the transfer impedance on current inside and outside the shield is shown in Figure 2.14.

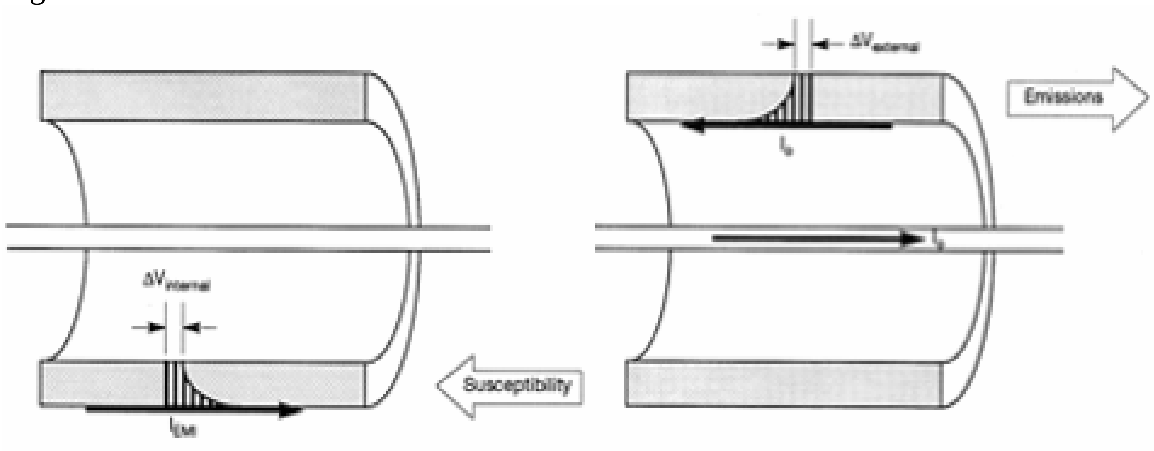

Figure 2.14 The transfer impedance concept: Reciprocity of susceptibility and emission.

The maximum internal shield voltage for a cable of length $L < \lambda/4$ is

$$V = LZ_t I_o \tag{2.40}$$

In general the current and voltage appearing on the signal wire will depend on the cable length and the terminating impedances Z_O at either end. General formula for some cases shows that it is usually best to terminate a long cable at both ends with its characteristic impedance Zo. For a short length similarly terminated the noise current *i* will be

$$i = Z_t L I_o / 2Z_o \tag{2.41}$$

At low frequencies, Z_t is equals to the DC resistance R_o. The DC resistance for a shield with radius a is $R_o = 2/(2\pi a\sigma t)$. The coupling external to internal shield decreases dramatically at a few tens of kHz to be

$$Z_t = R_o(1+j)\left(\frac{t}{\delta}\right) \Big/ \sinh\left((1+j)\left(\frac{t}{\delta}\right)\right) \tag{2.42}$$

where t = wall thickness. The transfer impedance has a square root of frequency (10 dB/decade) dependence above the skin depth cutoff frequency (Cutoff when $R_o \to Z_t$)

For an imperfect shield Z_t is

$$Z_t = R_o(1+j)\left(\frac{t}{\delta}\right) \Big/ \sinh\left((1+j)\left(\frac{t}{\delta}\right)\right) + j\omega M \tag{2.43}$$

where M = shield mutual inductance. Mutual inductance may be due to apertures or porpoising coupling. Porpoising coupling dominates in most braided shields. The effect on the surface impedance of a single hole in a cooper pipe is shown in Figure 2.15.

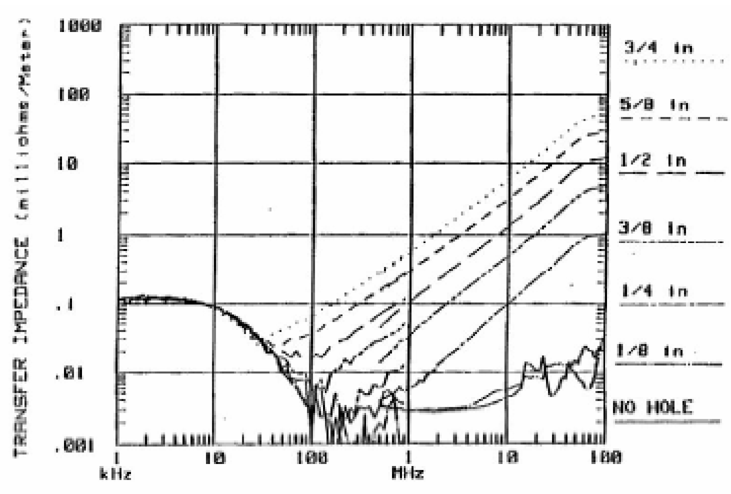

Figure 2.15 Measured surface transfer impedance of a 1-1/4 diameter cooper pipe with a single hole.