JAVA APPLET-BASED VIRTUAL LABORATORY FOR EMI/EMC TRAINING

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The objectives of this Tutorial are to explain some of the basic EMI/EMC interactions through the application of numerical experimentation based on the use of applets. The advantages of this approach are twofold: First, complex mathematics are avoided thus focusing on physical principles and making this material accessible to a wider audience. Second, sophisticated computer codes and numerical techniques are not employed giving the user an easy to drive interface which resembles the simplicity and immediacy of a physical experiment. The emphasis of the Tutorial is on fundamentals and no attempt is made to tackle complex problems.

The Tutorial is based around a Powerpoint presentation describing the strengths and limitations of the models employed. These models are then implemented as Java Applets and are embedded in the presentation. Thus, an interactive simulation environment is provided that enables engineers to explore how each parameter affects EMC and thus help them to devise effective approaches to mitigation. The Tutorial focuses on three fundamental aspects of EMI/EMC namely electromagnetic shielding, electromagnetic emissions, and electromagnetic immunity. Internet access to selected Java Applets for personal use after completion of the Tutorial will be given to all registered Tutorial participants.

Keywords: Electromagnetic Compatibility, EMC/EMI Training, Electromagnetic Shielding, Emission, Immunity.

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Contents:

- **Introduction**
- **EM Shielding Effectiveness**
- **EM Emissions**
- **EM Immunity**
- **Concluding Remarks**

INTRODUCTION

EMC/EMI Training is challenging due to the

- **complexity of EM interactions**
- **abstract nature of mathematical formulations**
- **complexity of practical systems**
- **very wide frequency range**
- **large differences in physical scale**
- ……..

We need flexible educational tools that suit the complexities of practical problems, the background of our engineers and their style of living and learning

JAVA Applet-based training offers a virtual laboratory with many advantages including:

- **delivery close to the normal place of work (online)**
- **can be tailored to the application area of the trainee**
- **allows rapid experimentation for illustrating concepts/techniques**
- **can be easily revisited by the trainee**
- **can be easily updated with new developments/requirements**
- **allows rapid access to the instructor**
- **can provide a framework for study at different depths**

...We show below three examples under development...

EM Shielding Effectiveness (SE)

- •Introduction and Aims
- •General Objectives
- •Basic Concepts
- •Detailed Model Development
- •Model Extensions
- •Applet-based Experimentation for SE
- •Appendix and Further Reading

•Introduction and Aims

A basic electromagnetic interaction (EM) affecting EMC/SQ is when an external EM field (e.g. due to a radio transmitter) penetrates through apertures (e.g. cooling holes, access openings) to establish EM fields inside enclosures (e.g. equipment cabinets).

Depending on the magnitude and spectral content of these fields, signals may be induced on circuits inside the enclosure which may cause malfunction and/or permanent damage (a susceptibility problem). Similarly, fields generated by circuits inside enclosure may 'leak' through apertures to propagate in the external environment potentially causing EMI to other users (emission problem).

In both these cases of paramount importance is the establishment of the shielding effectiveness (SE) of the cabinet.

In practical systems a certain amount of shielding is required to minimise emissions and immunity problems. The following give some idea of what is expected:

- SE of the order of 20dB is the minimum worthwhile value
- A SE of 50 to 60 dB is a typical average to cope with most problems
- For some test equipment and transmitters a SE of the order of 100dB may be necessary
- SE in excess of 120dB is very difficult to get in practice (state of the art)

For further details see: *"Principles and Techniques of Electromagnetic Compatibility", C. Christopoulos, CRC Press 1995*

Penetration of EM energy through the equipment cabinet may be due to one of more of the following mechanisms:

- EM energy penetration through the walls of the cabinet

  This is done by diffusion in cases where the electrical conductivity of the wall material is not high. Also, low-frequency magnetic fields can penetrate even through high-conductivity walls.

- EM energy penetration along wire interconnects

  This typically happens when EM energy is guided by wire penetrations such as signal and mains cables

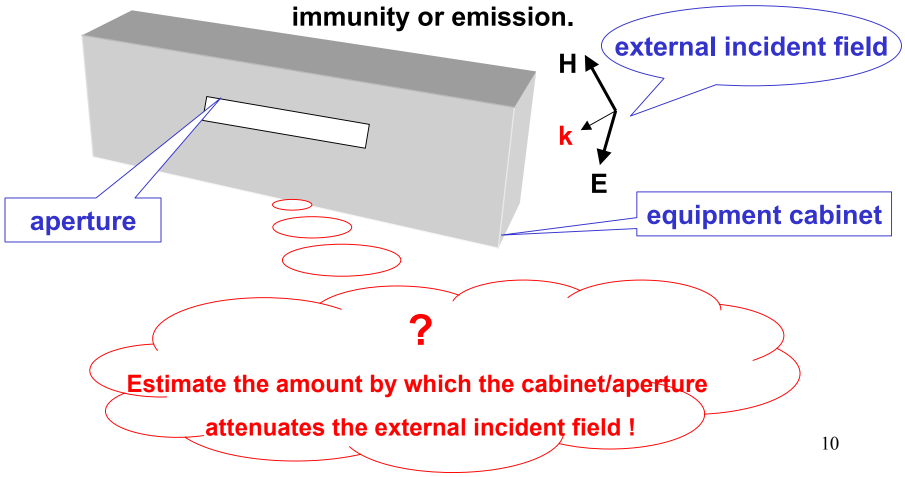
- EM energy penetration through apertures

  Apertures may be, access holes, ventilation openings, doors, poorly joined panels etc.

In this unit we focus on the calculation of the shielding effectiveness of cabinets with apertures (assuming perfectly conducting walls).

The central aim of this unit is to:

Understand the penetration of external incident EM fields through apertures on equipment cabinets and calculate the Shielding Effectiveness (SE). SE is the same whether one considers immunity or emission.

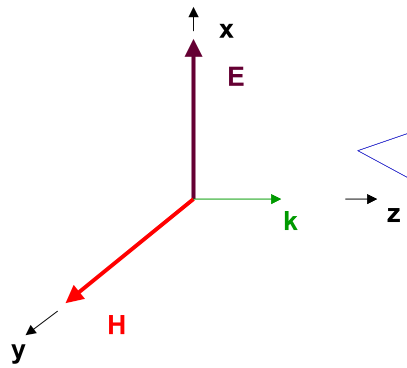


•General Objectives:

- establish the basic modelling methodology for understanding field penetration through apertures
- study the impact of cabinet dimensions on SE.
- study the impact of aperture dimensions and number of apertures on SE.
- quantify the influence of the cabinet contents on SE
- Establish design rules for SE

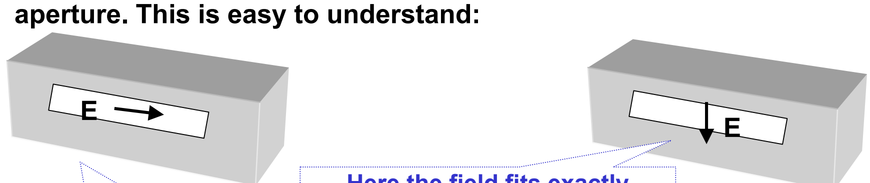
•Basic Concepts:

- A simple type of EM wave which can help us understand the behaviour of more complex waves is the **uniform plane wave**. For this type of wave the **electric E** and **magnetic H** field components are perpendicular to each other and to the **direction of propagation k**.



In this example the electric field has only an x-component, the magnetic field a y-component and the propagation direction is along z.

- **More complex waves** may be often analysed in terms of plane waves. All the essential concepts can be illustrated by focusing on plane waves.
  - The **worst case** as far as penetration is concerned is when the electric field is perpendicular to the longest dimension of the aperture. This is easy to understand:



Here the field fits exactly the boundary conditions-hence it penetrates easily

Here the electric field is parallel to the walls of the slot, but the tangential component of the electric field should be zero on a conducting surface-hence the field finds it difficult to penetrate

- The aperture (the slot in this case) will allow certain frequencies to penetrate relatively easily while energy at other frequencies will penetrate with difficulty. Intuition indicates that frequencies with wavelengths larger that the largest physical dimension of the slot will not penetrate easily.

- As stated above, penetration through the slot is frequency-dependent. In addition, since the cabinet resonates at certain frequencies, the EM energy which has penetrated past the slot will distribute unevenly at different frequencies according to the resonant properties of the cabinet. The presence of PCBs and other components inside the cabinet complicates matters still further. A formula for resonances in an empty box is given next.

Resonance frequencies of an empty rectangular box with conducting walls and without apertures:

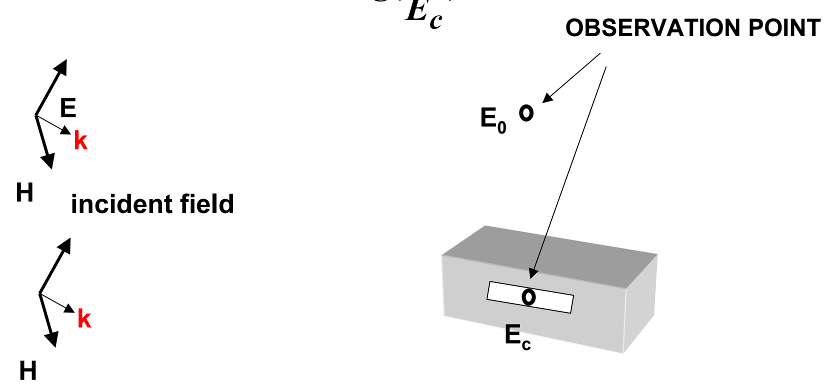
$$f_{MHz} = 150\sqrt{\left(\frac{m}{a}\right)^2 + \left(\frac{n}{b}\right)^2 + \left(\frac{p}{d}\right)^2}$$

Where, the frequencies are in MHz, (a,b,d) are the internal dimensions of the box in meters, and (m,n,p) are integers which specify the particular mode or resonance. No more than one of these integers can be zero. Each integers gives the number of half wavelengths fitting in each direction.

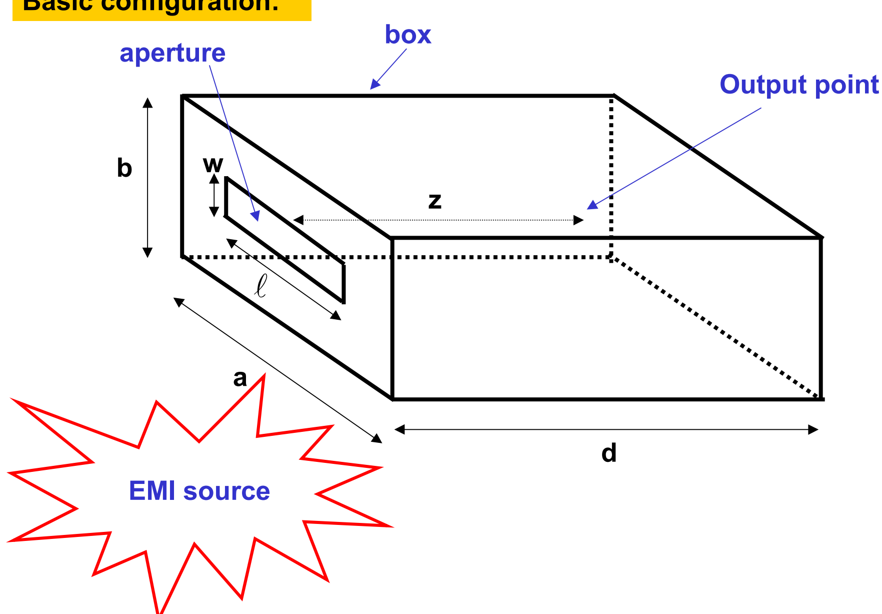
- All these factors must taken into account in any model used to predict SE

- The **Shielding Effectiveness (SE)** is defined as the ratio in decibels of the incident electric field E_0 without the cabinet, to the field with the cabinet present E_c. In both cases the field is calculated or measured at the same point.

$$SE = 20\log\left(\frac{E_0}{E_c}\right)$$



Basic configuration:



•Detailed Model Development

The basic strategy is to derive simplified models of the slot (slotline) and of the cabinet (short-circuited waveguide) and combine them to study penetration and coupling .

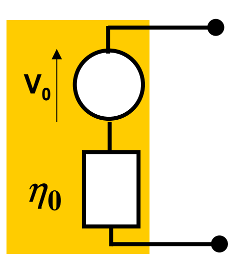
Each part in this interaction is modelled in the simplest possible way. The following parts need to be modelled:

- The incident field
- The aperture/slot
- The cabinet

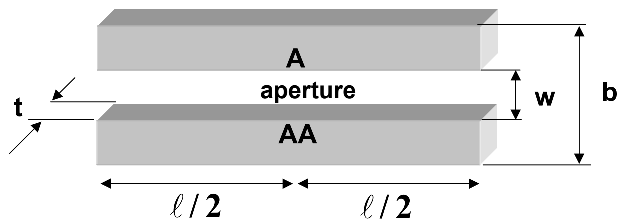
We develop models for each of the above part in turn.

Model for the Incident Field:

As far as the cabinet is concerned, the incident field can be described by a Thevenin equivalent circuit where the voltage source is V_0, and the impedance is the intrinsic impedance of the medium (377 Ohm in air). The exact value of V_0 is not important for SE calculation as it cancels out when the ratio of the two fields is formed.



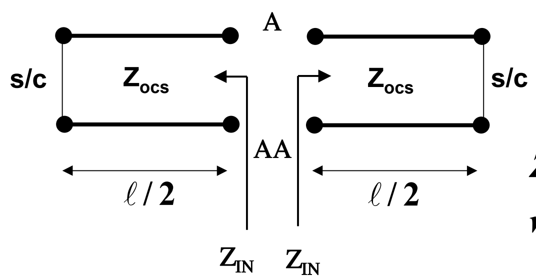
Model for the Aperture:



We will construct an equivalent circuit for the aperture as seen across its centre (points A-AA).

Propagation along the aperture is similar to propagation along a coplanar stripline. There are in fact two such lines each of length l/2, which are, to the left and to the right, terminated by short-circuits (the front walls of the cabinet). The characteristic impedance of this line Z_{ocs} may obtained from the literature.

Hence, the equivalent circuit across points A-AA is:

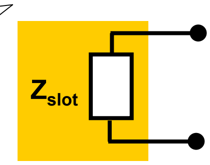


$$Z_{IN} = jZ_{ocs}\tan\left(\beta_0 \ell / 2\right)$$
$$where\ \beta_0 = 2\pi / \lambda$$

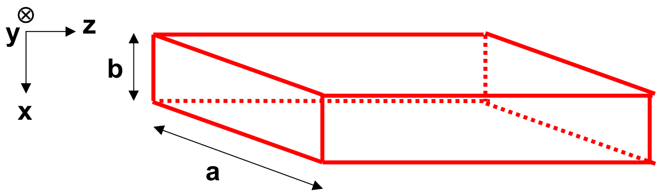
Hence, the equivalent slot impedance is equal to the input impedances looking left and right in parallel:

$$Z_{slot} = \frac{1}{2}\frac{\ell}{a} jZ_{ocs}\tan\left(\beta_0 \ell / 2\right)$$

Scaling factor l/a accounts for the different length of slot and cabinet. Formulae for Z_{ocs} are given in the appendix at the end of this presentation.



Model for the Cabinet:



We consider the cabinet as a rectangular waveguide with propagation along z. The cutoff wavelength is

$$\lambda_c = {2a}/{n}$$

n is and integer and we choose n=1 to get the lowest possible cutoff frequency

The guide wavelength is then: $$\lambda_g = \frac{\lambda}{\sqrt{1-\left({\lambda}/{\lambda_c}\right)}}$$

This corresponds to waves bouncing between the sidewalls with no variation in x and n number of half sinusoidal variations in the y-direction ($TE_{0,n}$ mode). The guide characteristic impedance is:

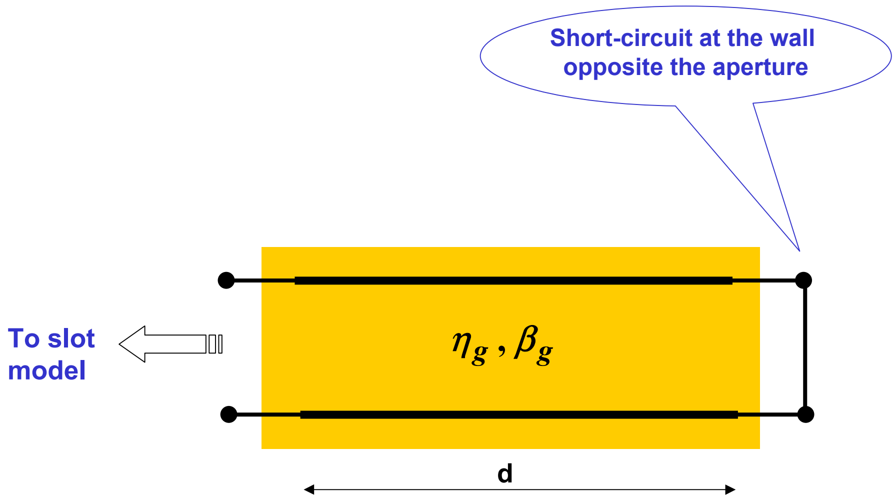
$$\eta_g = {\sqrt{\frac{\mu}{\varepsilon}}} \Big/ {\sqrt{1-\left(\frac{\lambda}{\lambda_c}\right)^2}}$$

Choosing n=1 (lowest frequency mode) and expressing the guide quantities in terms of the dimensions, we get:

$$\eta_g = \frac{\eta_0}{\sqrt{1-\left({\lambda}/{2a}\right)^2}} \qquad \beta_g = \beta_0\sqrt{1-\left({\lambda}/{2a}\right)^2}$$

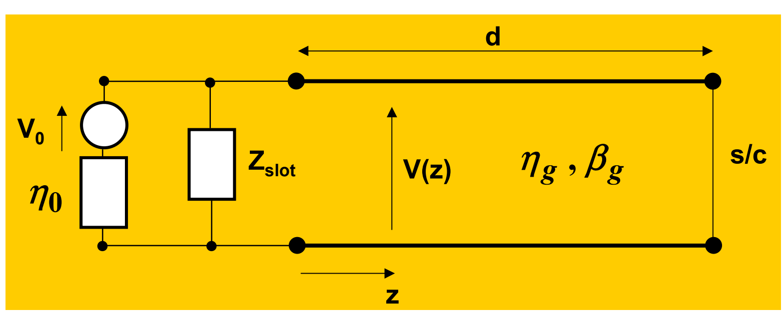
$$where, \quad \eta_0 = 377\Omega, \ \beta_0 = {2\pi}/{\lambda}$$

Hence, the model for the cabinet looks as follows:



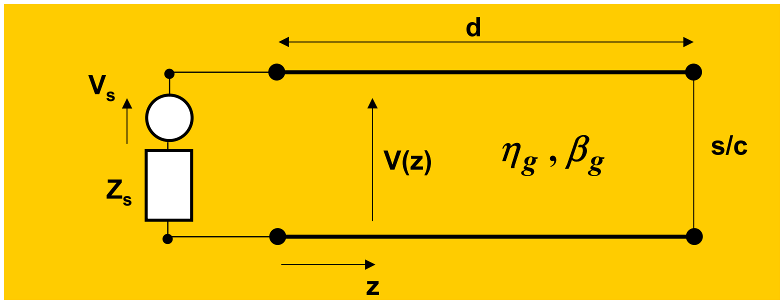
Complete model of the System:

We now put together all the various elements to form the complete model of the system:



In this model, the field inside the cabinet and at a distance z from the side with the slot is represented by the voltage V(z).

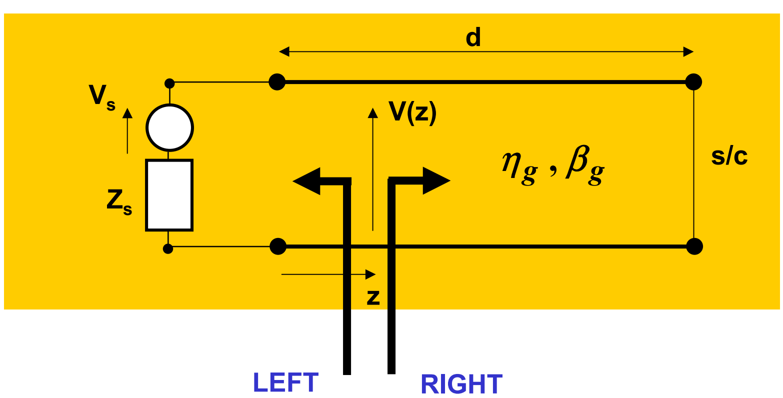
The incident field and the slot models may be combined to form a Thevenin equivalent as shown below;



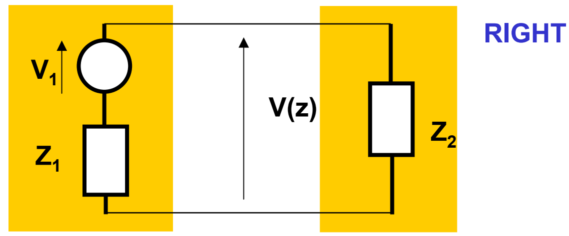
$$V_s = V_0 \frac{Z_{slot}}{Z_{slot} + \eta_0} , \quad Z_s = \frac{Z_{slot}\eta_0}{Z_{slot} + \eta_0}$$

This is now a standard TL circuit which can be solved for V(z).

To simplify calculations, we replace the circuit to the LEFT and to the RIGHT of the output point at z by their Thevenin equivalents:



Thevenin equivalent at a distance z from the slot:



$$V_1 = \frac{V_s}{\cos(\beta_g z) + j\dfrac{Z_s}{\eta_g}\sin(\beta_g z)}$$

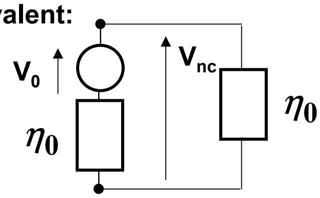
$$Z_2 = j\eta_g \tan\left[\beta_g (d - z)\right]$$

$$Z_1 = \frac{Z_s + j\eta_g \tan(\beta_g z)}{1 + j\dfrac{Z_s}{\eta_g}\tan(\beta_g z)}$$

The voltage at z representing the electric field inside the box is then:

$$V(z) = V_1 \frac{Z_2}{Z_1 + Z_2}$$

In the absence of the box the field is given by the following equivalent:



$$V_{nc} = V_0 / 2$$

Hence, the shielding effectiveness of the cabinet at point z is:

$$SE = 20\log\left(\frac{E_0}{E_c}\right) = 20\log\left(\frac{V_0}{2V(z)}\right)$$

•Model Extensions

The procedure outlined shows how to obtain the electric field shielding effectiveness for an empty (unloaded) cabinet. The magnetic field shielding effectiveness may also simply be obtained by calculating the ratio of currents rather than voltages at point z.

We now investigate two additional aspects of shielding effectiveness which affect the behaviour of practical systems:

- **loaded cabinets**
- **multiple apertures**

These two extensions to the theory are dealt with by modifying the standard TL model described for the unloaded cabinet.

Loaded cabinet:

A simple, phenomenological, way to take account of the contents of the cabinet and their impact on SE is to assume that the waveguide representing the cabinet is lossy. In its simplest form this model assumes distributed losses and a correction term ζ thus replacing the guide characteristic impedance Z_g and propagation constant β_g by

$$\eta'_g = (1+\zeta - j\zeta)\eta_g \ , \quad \beta'_g = (1+\zeta - j\zeta)\beta_g$$

The correction factor is of the order of the inverse of the loaded cabinet Q-factor

$$\zeta \simeq Q^{-1} \ll 1$$

The use of the correction factor is explained in "Immittance transformation using precision air-dielectric coaxial lines and connectors" D. Woods, Proc. IEE, Vol. 118, 1971, pp. 1667-1674

Multiple apertures:

If there are n identical apertures then the equivalent slot impedance is:

$$Z_{slot} = n \frac{1}{2} \frac{\ell}{a} jZ_{ocs} \tan\left(\beta_0 \ell / 2\right)$$

Circular apertures:

If the aperture is circular of diameter d_h then to a good approximation the previous formulae may be used if we set

$$\ell = w = \frac{\sqrt{\pi}}{2} d_h$$

•Applet-based Experimentation for SE

- Show how the size and shape of the cabinet affects SE
- Show how the size and shape of the slot affects SE
- Learn about the particular behaviour of SE near cabinet resonances
- Understand the merits of using a larger number of smaller slots to increase SE
- Assess how the contents of a cabinet affect SE especially near resonances

Semi-empirical formula for SE[*]:

A simple approximate formula used often to estimate the SE of a cabinet with a slot is

$$SE = 20\log\left(\frac{\lambda}{2\ell}\right)$$

This formula does not take into account the width of the slot or the dimensions of the cabinet and therefore gives at best very approximate results. SE curves obtained using this formula are plotted for comparison with the results of the more sophisticated model shown here (see numerical experiments that follow).

[*] *H.W. Ott, "Noise Reduction Techniques in Electronic Systems", 2nd edition, Wiley 1988*

Using the Applet you can estimate the shielding effectiveness of a cabinet:

You will need to specify the following-

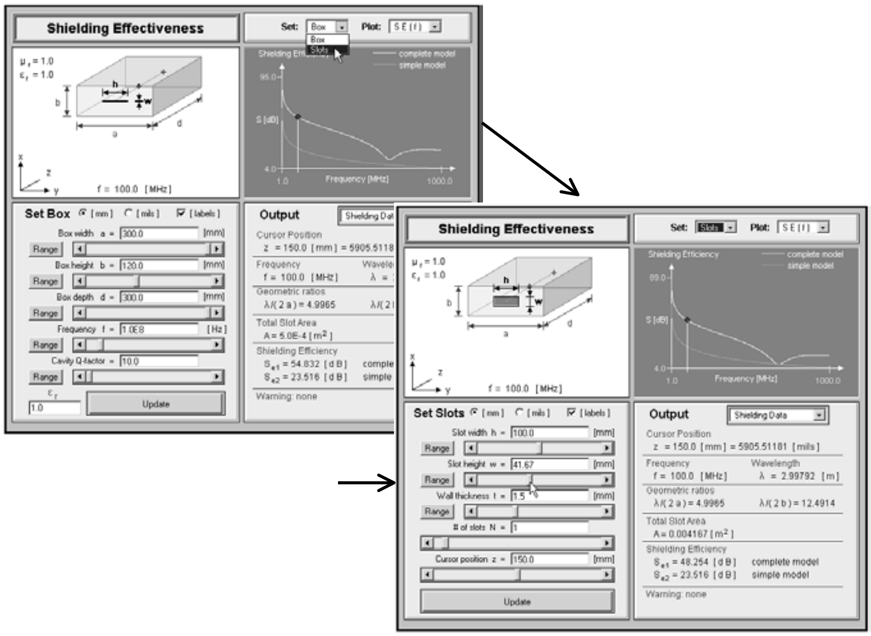
•dimensions of the cabinet

•dimensions of the slot

•position of the output point z (SE will be different depending on where the field is measured inside the box

See an example of this applet

in the next slide!



Completion of this exercise should have given you an insight into the following:

- the way in which cabinet size and shape affect SE
- the way in which slot size and shape affect SE
- the frequencies at which SE is at its lowest
- the configuration of slots that gives maximum SE
- effect of loading inside the cabinet on SE
- practically achievable values of SE at different frequencies
- distribution of field inside the cabinet

APPENDIX: Formulae for Z_{ocs}

Formulae for the characteristic impedance of a coplanar microstrip line may be found in , *"Microstrip Lines and Slotlines", K.C. Gupta, R. Garg, I. J. Bahl, Artech House,1979 (chapter 7).*

They are summarised below: $Z_{ocs} = 120\pi K\left(w_e / b\right) / K'\left(w_e / b\right)$

Where, K and K' are elliptic integrals and w_e is the effective width

$$w_e = w - \frac{5t}{4\pi}\left[1 + \ln\left(\frac{4\pi w}{t}\right)\right]$$

Approximate formula:

$$Z_{ocs} = 120\pi^2\left[\ln\left(2\frac{1+\sqrt[4]{1-(w_e/b)^2}}{1-\sqrt[4]{1-(w_e/b)^2}}\right)\right]^{-1}$$

$$for \quad w_e < b/\sqrt{2}$$

FURTHER READING:

1. "Shielding effectivenessof a rectangular enclosure with a rectangular aperture", M.P. Robinson et al, Electronics Letters, 15 Aug. 1996, Vol.32, No 17, pp.1559-1560
2. "Analytical evaluation of the shielding effectiveness of enclosures with apertures", M.P. Robinson et al, IEEE Trans. on EMC, Vol. 40, No 3, 1998, pp.240-248
3. "Assessment of the shielding effectiveness of a real enclosure", R. De Smedt et al, Int. Symp. On EMC, Sept. 14-18, Rome, Italy, pp. 248-253
4. "Numerical and experimental study of the SE of a metallic enclosure", F. Olyslager et al, IEEE Trans. On EMC, Vol. 41, No 3, 1999, pp. 202-213

ELECTROMAGNETIC EMISSIONS

•Introduction and Aims
•General Objectives
•Basic Concepts
•Wire Interconnects as Radiating Antennas
•Applet-based Experimentation for Emissions
•Further Reading

•Introduction and Aims

A basic electromagnetic interaction (EM) affecting EMC/SQ is when signals on wire interconnects or other components generate fields which couple to adjacent circuits or which propagate by radiation over large distances to couple to other circuits.

Depending on the magnitude and spectral content of these radiated fields circuits may malfunction and/or be permanently damaged.

Investigations into the level of radiated fields from circuits are described under the term emission studies.

The reverse problem whereby a circuit is the victim to EM energy is referred to as a susceptibility/immunity study.

In this unit we focus on emission studies.

Radiation of EM energy from a wire interconnect may be studied in connection with :

•impact on adjacent wires/circuits

•general pattern of radiated fields

The first case is normally studied under the heading of cross-talk (implying near-filed coupling described by mutual capacitance/inductance). It is relevant to intra-system EMC.

The second case implies radiation some times over large distances and therefore near-field and far-field radiation from circuits acting very much like antennas. It is relevant to inter-system EMC.

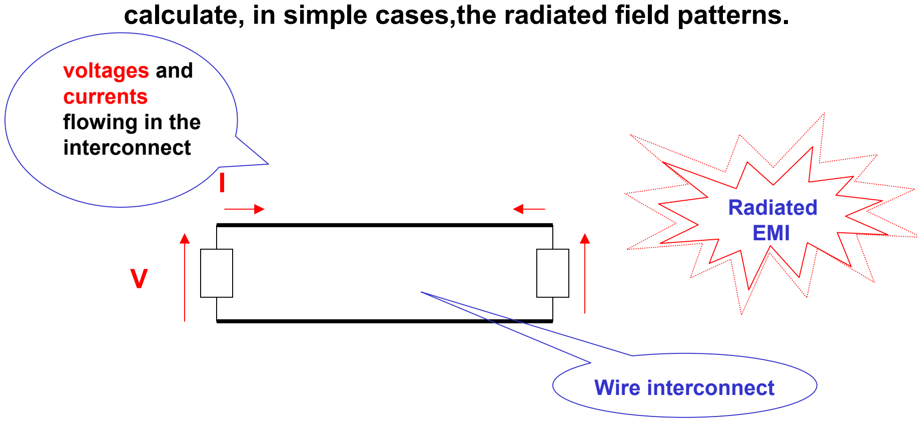
Both processes are important and often the boundaries between them are not easy to establish.

However, the methodology of dealing with these two cases is different and hence it is profitable to tackle each case separately.

In this unit we deal with the second case of general radiated field patterns.

The central aim of this unit is to:

Understand the radiation of EM fields from wire interconnects and calculate, in simple cases,the radiated field patterns.

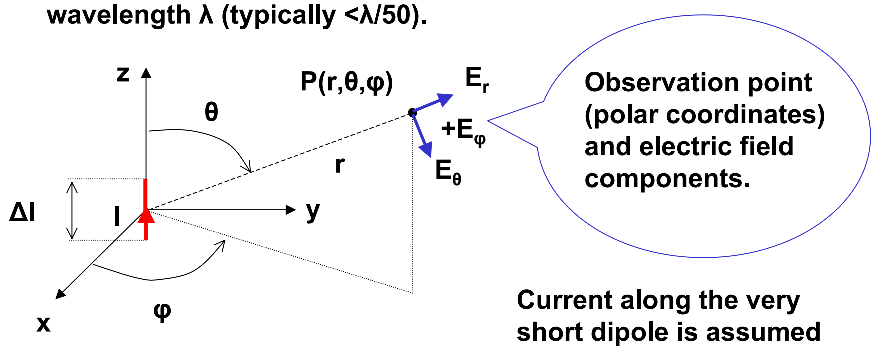


•General Objectives

- establish the basics for understanding radiated emissions from simple wire interconnects
- study the patterns of radiation from interconnects
- study the impact of line length, configuration, presence of conducting planes and of terminations
- Discuss emissions from more complex configurations

•Basic Concepts

•A simple wire radiator which can help us understand the behaviour of more complex radiators is the very short dipole. This radiator is often referred to also as a Hertzian dipole. The length of this radiator is Δl and must be much smaller than the wavelength λ (typically <λ/50).



Current along the very short dipole is assumed constant and equal to I.

The field components at a point a distance r away from a very short dipole are given by the formulae:

$$E_\vartheta = \frac{j\omega\mu}{4\pi}(I\Delta l)\frac{e^{-j\beta r}}{r}\sin\vartheta\left[1+\frac{1}{j\beta r}+\frac{1}{(j\beta r)^2}\right]$$

$$E_r = \frac{j\omega\mu}{2\pi}(I\Delta l)\frac{e^{-j\beta r}}{r}\cos\vartheta\left[\frac{1}{j\beta r}+\frac{1}{(j\beta r)^2}\right]$$

$$E_\varphi = 0$$

$$H_r = H_\vartheta = 0$$

$$H_\varphi = \frac{(I\Delta l)}{4\pi}\frac{e^{-j\beta r}}{r}\sin\vartheta\left[j\beta+\frac{1}{r}\right]$$

$$\boxed{where,\quad \beta = 2\pi/\lambda}$$

We see that emissions, even from this simple structure display complex behaviour, depending on r and θ. Also several terms appear which vary as $1/r$, $1/r^2$, $1/r^3$.

At large distances (several wavelengths) only the 1/r terms remains and the emitted field has practically only two non-zero components E_θ and H_φ. We are then at the far-field region.

$$E_\vartheta \simeq j\eta\frac{\beta(I\Delta l)}{4\pi}\frac{e^{-j\beta r}}{r}\sin\vartheta$$

$$H_\varphi \simeq j\frac{\beta(I\Delta l)}{4\pi}\frac{e^{-j\beta r}}{r}\sin\vartheta$$

$$where,\quad \eta \equiv 377\Omega$$

Intrinsic impedance of free space

In the far-field the radiated power density is:

$$W_r = \frac{|E_\vartheta(r,\vartheta,\varphi)|^2}{2\eta} = \frac{\eta}{2}\left|\frac{\beta(I\Delta l)}{4\pi}\right|^2 \frac{(\sin\vartheta)^2}{r^2} \quad , W/m^2$$

The radiation intensity U defined below is:

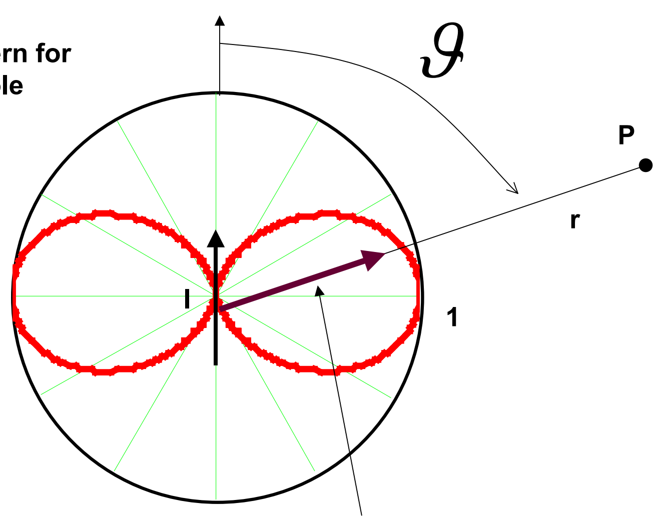
$$U \equiv r^2 W_r = \left[\frac{\eta}{2}\left|\frac{\beta(I\Delta l)}{4\pi}\right|^2\right](\sin\vartheta)^2 \quad , W/\textit{unit solid angle}$$

$$= U_{pk}(\sin\vartheta)^2$$

A plot of the U/U_{pk} is shown next valid for any angle φ.

Radiation pattern for very short dipole

φ=const.



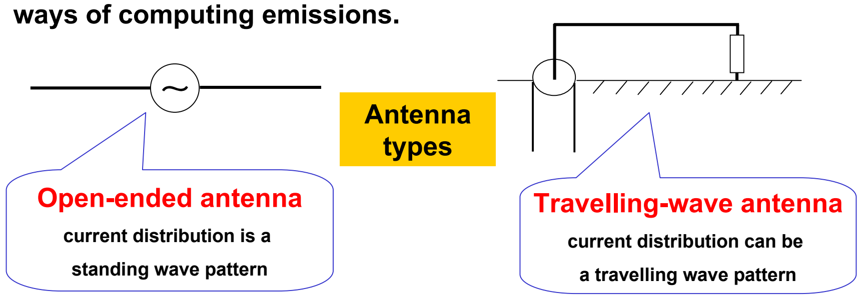
Normalised radiation intensity at point P, independent of φ.

$$U(\vartheta)/U_{pk} = (\sin\vartheta)^2$$

•Wire Interconnects as Radiating Antennas

The voltage and current distribution on an open-circuit transmission line is similar to that found on open-ended antennas. It therefore follows that we can learn a lot about emissions from wire interconnects by studying the properties on antennas.

We will do this first to gain a broad appreciation of the emission properties of wires before looking at systematic computer-based ways of computing emissions.



•A standing wave pattern can be constructed as the superposition of travelling waves propagating in opposite directions.

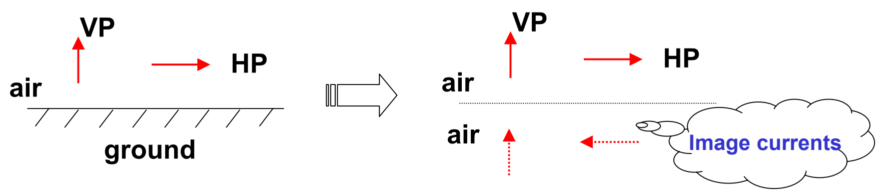
•In a travelling wave antenna where there is a single travelling wave, the amplitude of the wave is constant and the phase is linearly varying.

•In an open-ended (resonant) antenna the amplitude varies but the phase is constant.

•The pattern of emission depends on the mix of travelling waves on the radiating structure.

The proximity of current carrying conductors (interconnects) to other objects also affects the emission patterns. As an example we examine the simple but practical case of a conductor above a ground plane. A good insight may be gained by applying image theory to account for the impact of the ground plane.

We can distinguish two cases: A conductor perpendicular to the ground plane (Vertical Polarisation-VP), and a conductor parallel to the ground plane (Horizontal Polarisation-HP). Imaging in its case is different



We can see that the field at any point is made out of TWO components: one from the original current and the other from its image.The two component may combine constructively (max emission) or destructively (min emission). It depends on:

•direction of current at the image relative to the conductor current). We note that for VP currents are in the same direction, while for HP currents are in opposite directions.

•the difference in electrical distance between the observation point and the conductor and its image. If the difference in physical distance between these two paths is d then the electrical distance is

$$\frac{2\pi}{\lambda}d \quad ,\lambda \text{ is the wavelength}$$

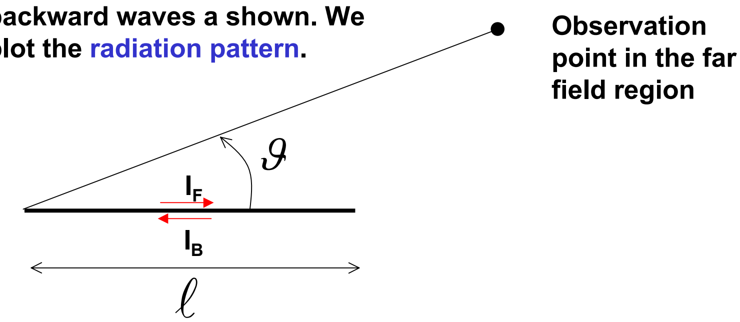
The strength of the emitted field at the observation point will therefore depend on whether we have VP or HP and on the frequency as it affects directly electrical distance and therefore phase. This explains why in EMC measurements we do height scans at each frequency etc.

In complex practical systems it can be difficult to predict emissions without resort to computational models. However, in order to get a better understanding, we will present here three different models of increasing levels of complexity

- emissions from single wires where there is a mix of forward and backward travelling current waves
- emission from wires driven by asymmetrically placed sources
- emissions from complete circuits where the current is known from numerical or other studies

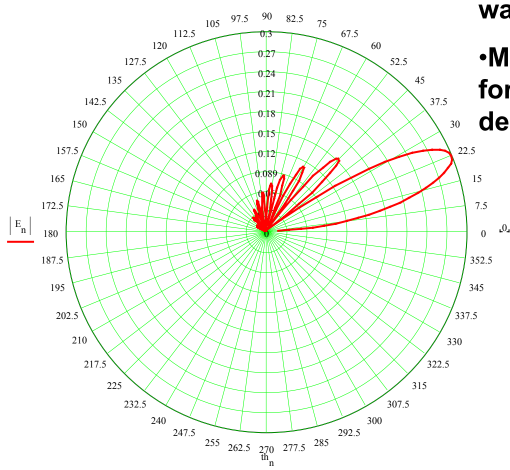
Emissions from single wires where there is a mix of forward and backward travelling current waves

The configuration we examine is a wire with forward and backward waves a shown. We plot the radiation pattern.



Forward propagating current wave I_F

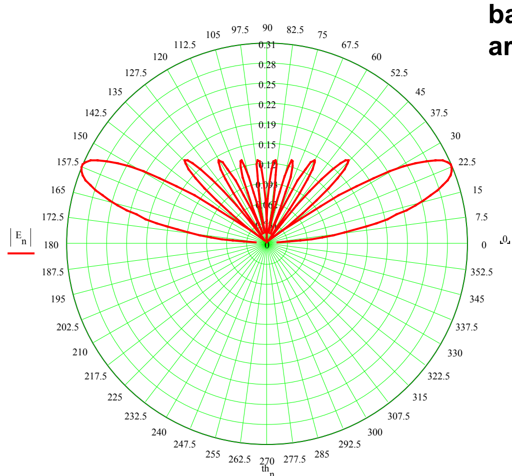
Backward propagating current wave I_B



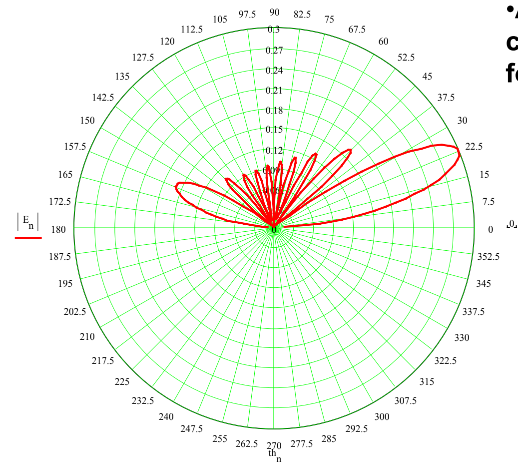
•Only forward wave (I_B=0)

•Ratio of wire length to wavelength is 5

•Maximum radiation in the forward direction at about 22 degrees



•As before,but forward and backward current waves are equal



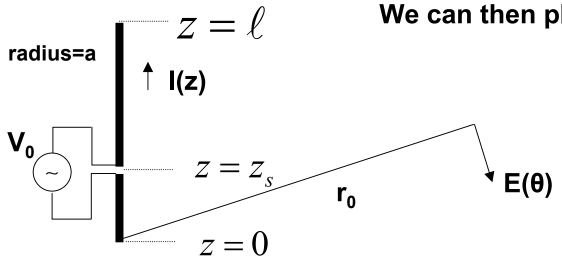
•As before, but backward current is 50% of the forward current

Emission from wires driven by asymmetrically placed sources

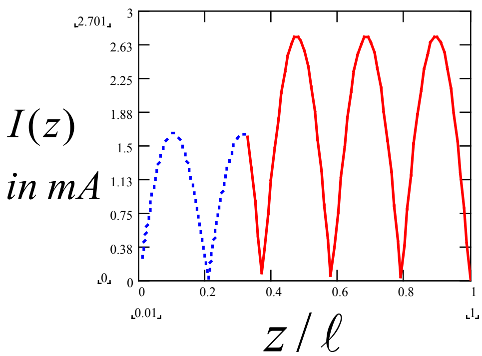
The problem we tackle here has two aspects (see ref. [3]):

• first, we work out the current distribution along a wire driven by a sinusoidal source anywhere along its length

•second, based on this current distribution, we work out the emitted field



We can then plot I(z) and E(θ)



$a / \ell = 10^{-4}$

$z_s = 0.33\ell$

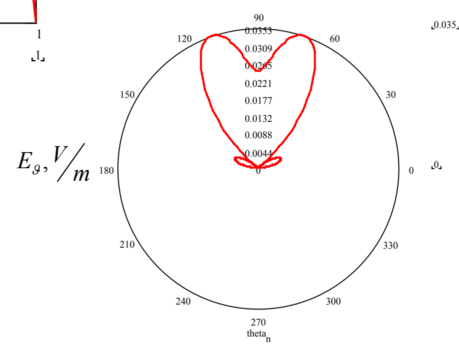
$k\ell = 15$

$V_0 = 1.0\,Volts$

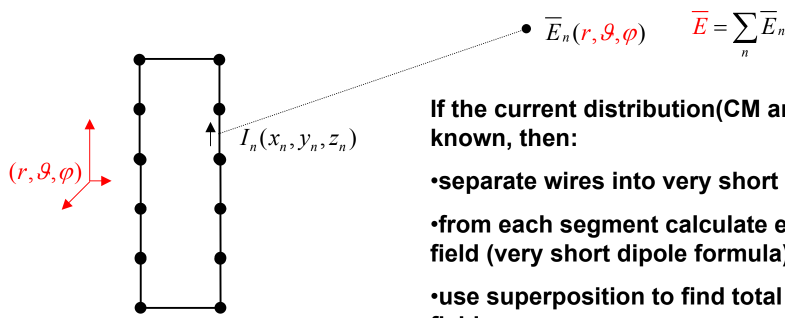
$k\ell = 3.14$

$\ell = 0.6m$

$r_0 = 3m$



Emissions from complete circuits where the current is known from numerical or other studies

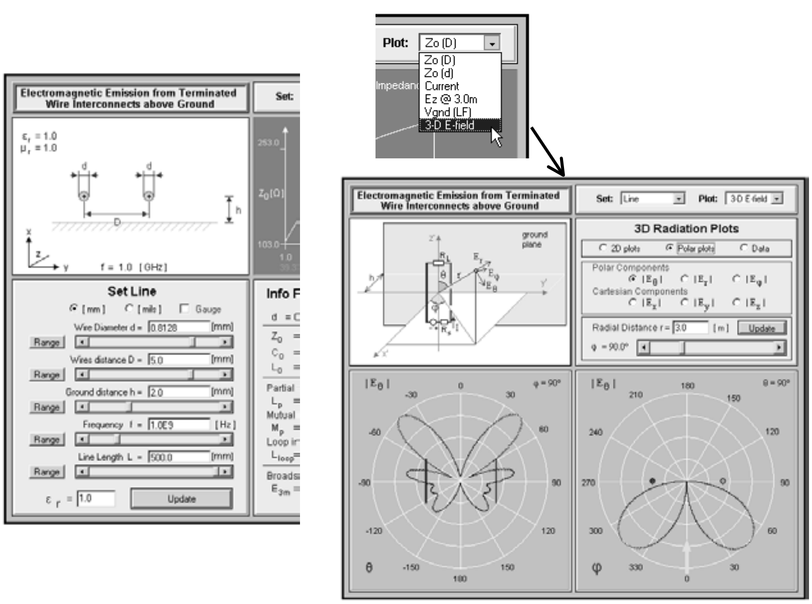


If the current distribution(CM and DM) is known, then:

- **separate wires into very short segments**
- **from each segment calculate emitted field (very short dipole formula)**
- **use superposition to find total emitted field**

•Applet-based Experimentation for Emissions

See an example of this applet in the next slide!



Completion of this exercise should have given you an insight into the following:

- **the way in which terminations affect emission**
- **the way in which line length and spacing affect emissions**
- **the directional patterns of emission from interconnects**
- **the impact of ground planes on emissions**

•FURTHER READING

1. **"Analysis of Multiconductor Transmission Lines", C. R. Paul, Wiley 1994, (chapters 4,5)**
2. **"Principles and Techniques of Electromagnetic Compatibility", C. Christopoulos, CRC Press 1995, (chapters 7,8)**
3. **"EMC Analysis Methods and Computational Models", F. M. Tesche, M. V. Ianoz and T. Karlsson, Wiley 1997, (chapter 4)**
4. **"Antenna Theory: Analysis and Design", C A Balanis, Wiley 1997 2nd Edition (Chapter 10)**
5. **"Control and Measurement of Unintentional EM Radiation", W Scott Bennett, Wiley 1997**

Electromagnetic Immunity

•Introduction and Aims

•General Objectives

•Basic Concepts

•Model Formulation

•Models and Applet-based experimentation for Immunity

•Further Reading

•Introduction and Aims

A basic electromagnetic interaction (EM) affecting EMC/SQ is when an EM field couples to wires (i.e. induces voltages/currents on wire interconnects between circuits).

Depending on the magnitude and spectral content of these induced signals circuits may malfunction and/or be permanently damaged.

Investigations into the level of induced signals and the likely behaviour of circuits are variously described as susceptibility or immunity studies.

The reverse problem whereby the circuit emits EM energy is referred to as emission study.

In this unit we focus on susceptibility/immunity studies.

Coupling of EM energy into a wire interconnect may be caused by:

•adjacent wires/circuits

•EM fields from distant wires/circuits

The first case is normally studied under the heading of cross-talk (implying near-filed coupling described by mutual capacitance/inductance). It is relevant to intra-system EMC.

The second case implies interaction of circuits with an incident field (often a plane wave) representing far-field radiation from distant circuits acting very much like antennas. It is relevant to inter-system EMC.

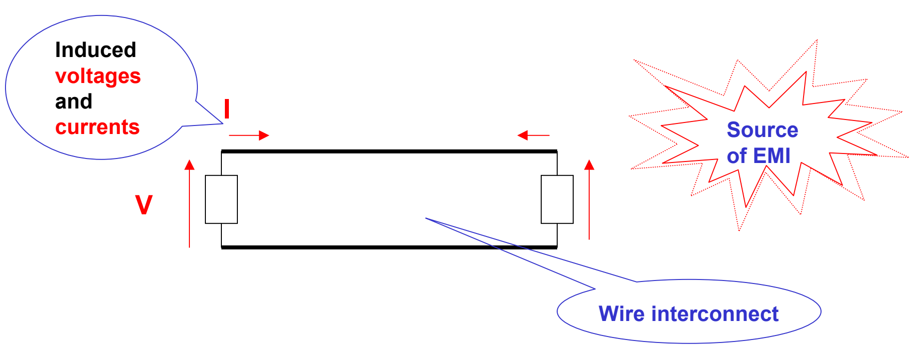
Both processes are important and often the boundaries between them are not easy to establish.

However, the methodology of dealing with these two cases is different and hence it is profitable to tackle each case separately.

In this unit we deal with the second, far-field case.

The central aim of this unit is to:

Understand the coupling of external incident EM fields into wire interconnects and calculate, in simple cases, induced voltages and currents at circuit terminals.



•General Objectives

•establish the basic modelling methodology for understanding field-to-wire coupling

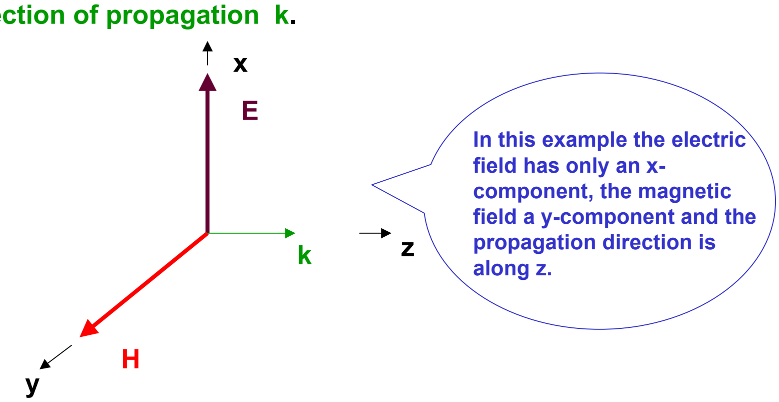
•study the impact of field polarisation and propagation direction on coupling

•study the impact of line length, configuration, presence of conducting planes and of terminations

•establish basic guidelines to reduce field-to-wire coupling

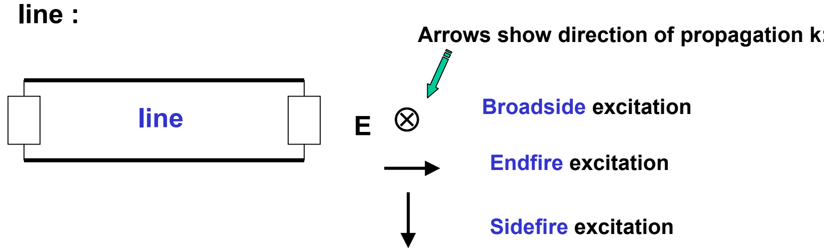
•Basic Concepts

•A simple type of EM wave which can help us understand the behaviour of more complex waves is the uniform plane wave. For this type of wave the electric E and magnetic H field components are perpendicular to each other and to the direction of propagation k.



- **Polarisation** of the electric field refers to the direction of the electric field relative to the configuration of the line (e.g. electric field vector parallel to the line, transverse to the line, or, in some other direction). Physical intuition suggests that polarisation affects the strength of coupling.

- **Direction of propagation** refers to the direction in which EM energy is transported in space by the wave. Clearly, the direction of incidence of EM energy on the line will affect coupling. See below some simple wave propagation directions for exciting a line :



- **Line terminations** refers to the loads at the two ends of a wire interconnect. As an EM waves impinges on the line the induced currents and voltages and wave reflections will depend on the terminations

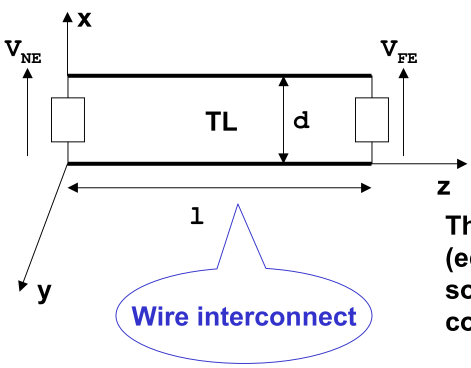
- **TEM conditions** in wire interconnects refer to the situation where the electric and magnetic field between the wires have only transverse components to the direction of propagation (strictly true only for lossless lines)

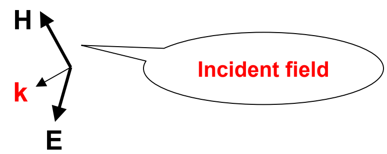
- **Conducting planes** in the vicinity of the interconnect, or as a part of it, may affect coupling e.g. by causing reflections to incident waves.

•Model Formulation

Formulation of a mathematical model of field-to-wire coupling:

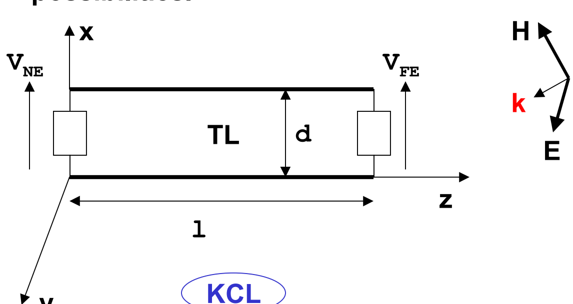
The basic strategy is to supplement the well known transmission line (TL) equations with additional source terms which represent coupling to the external field.





There are at least three possible (equivalent) formulations of the source terms to account for coupling to the field!

In an often used formulation, the total voltage V and current I are described, as shown, with source terms dependent on the incident electric field in x-direction and incident magnetic field in y-direction. There are however other, equivalent, modelling possibilities.



KCL

$$\frac{dI}{dz} + j\omega C_d V = -j\omega C_d \int_0^d E_x^i dx$$

KVL

$$\frac{dV}{dz} + j\omega L_d I = j\omega \int_0^d B_y^i dx$$

The main alternative model formulations are in terms of source terms depending on the incident, electric field only, or, the magnetic field only.

For a critical discussion of the various models and their equivalence see:

"On the contribution of the EM field components in field-to transmission line interaction"

C.A. Nucci and F. Rachidi, IEEE Trans. On EMC, 37, No. 4, Nov. 1995, pp. 505-508.

The model as shown is formulated in the frequency-domain (time is not the independent variable, harmonic excitations of frequency f are assumed). If a pulsed field is applied it must first be analysed into its frequency components, the line response to each component found by solving the previous equations, and the results combined to obtain the total response (only possible for linear systems).

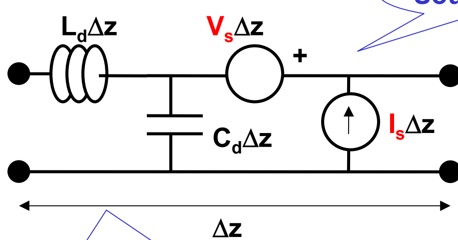
Alternatively, the model may be formulated in the time-domain to obtain a more direct solution for the effects of pulsed fields on the line which is also valid for general non-linear systems.

Both, frequency- and time-domain approaches are used in formulating models which can be solved either analytically (for simple cases) ,or, numerically.

Analytical solutions are possible for many cases especially at low-frequencies where neglecting transmission line effects simplifies the equations. Although low-frequency results cannot be used at high frequencies they nevertheless provide an insight into basic coupling mechanisms which is invaluable to the designer.

Numerical solutions are obtained by dividing the line into many small segments, each smaller than a wavelength, and inserting in each segment voltage and current sources which represent the coupling to the electric and magnetic field components respectively, as explained earlier. This process is illustrated below:

Source terms



$$V_s(z) = j\omega \int_0^d B_y^i(x,z)dx$$

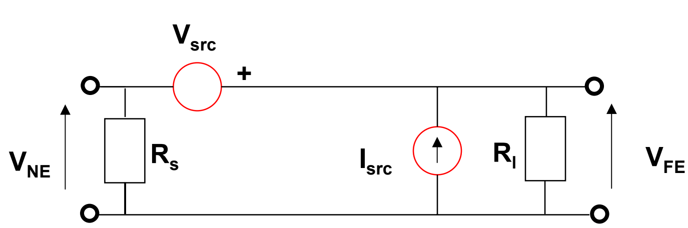
$$I_s(z) = -j\omega C_d \int_0^d E_x^i(x,z)dx$$

Parameters of the TL segment

There are THREE types of solutions to the coupling models:

Simple model

This is a low-frequency model where all excitation sources are lumped together, the inductance and capacitance of the line is neglected and the near-end and far-end impedances are assumed to be approximately equal to the characteristic impedance of the line. This model is sketched out below.



TEM model

This a model valid up to medium frequencies and it is based on solving the TL equations assuming TEM approximation (line length much larger than separation, no losses). This solution can be done in either the frequency or time domains. Here, the frequency-domain solution is used.

High-frequency model

A high-frequency field model valid for all frequencies including lines with substantial losses and arbitrary separation. Here a solution based on the Method of Moments (MoM) must be used. *This model is beyond the scope of this presentation.*

We now look at some specific examples to get some intuitive understanding and quantitative information regarding coupling to wires.

There are three basic interconnect configurations:

- **A parallel wire interconnect**
- **A wire-above-ground interconnect**
- **A parallel wire interconnect above ground**

For each configuration we apply one or more of the three different models described (Simple, TEM and High-frequency models).

Examples of sidefire, broadside and endfire excitations are studied.

•Models and Applet-based Experimentation for Immunity

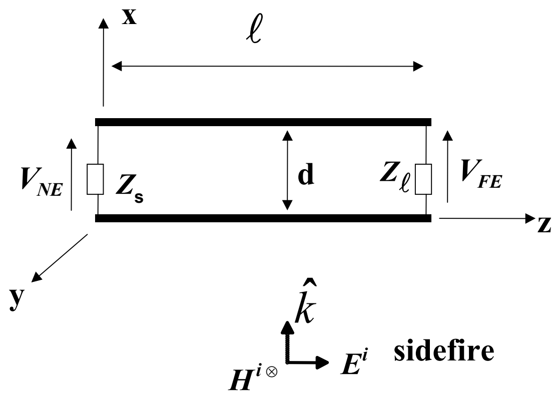
- Show how the length, spacing,and line terminations affect induced voltages
- Establish the range of validity of the different models
- Learn about the inaccuracies of coupling predictions based on the assumption of electrically very short lines
- Study the behaviour of an electrically long line that radiates substantially (requires high-frequency model)
- Compare the severity of coupling due to different excitation modes (sidefire, broadside and endfire). Establish worst case and suggest possible remedies.

objectives (cont.):

- Compare coupling for parallel wire, wire-above-conducting plane, and parallel wires above conducting plane lines
- Establish the circumstances when a full-field solution for the interconnect configuration is necessary
- Based on the numerical experiments you have done, suggest broad design guidelines to minimise coupling from external fields to wire interconnects

SIDEFIRE Excitation:

The configuration studied is shown below. The line length is ℓ and the spacing is d. The near-end (NE) and far-end (FE) quantities are shown, together with the corresponding terminations Z_s and Z_ℓ.



For this excitation the electric filed has only a z-component and the magnetic field a y-component. Using these field components the equivalent sources are calculated and inserted into the field-to-wire coupling model described earlier.

Basic transmission line theory is then used to obtain solutions for the terminal voltages at the line terminations.

The following information is required:

- **magnitude of the electric field component E_0 .The magnetic field need not be explicitly supplied as for a plane wave**

$$H = E/\eta$$

where η is the intrinsic impedance of the medium. In air,

$$\eta = \eta_0 = \sqrt{\mu_0 / \varepsilon_0} = 377\Omega$$

- the frequency of the excitation f. This permits the calculation of the line phase constant,

$$\beta = 2\pi / \lambda = \omega / v_p$$

- v_p is the velocity of propagation on the line. We assume a lossless line.
- the geometrical and material properties of the line. These allow the calculation of the per unit length capacitance and inductance of the line, its characteristic impedance, velocity of propagation etc.
- the impedance of line terminations. For simplicity, we assume that the impedances at terminations are resistive.

In the next slides we summarise the formulae used and in the WEB-based exercises for sidefire coupling

Simple model for sidefire coupling on parallel wire interconnect:

$$V_{NE} = -\frac{Z_s}{Z_s + Z_\ell}\left[-j(d\ell)\beta E_0 \frac{\sin(\beta d/2)}{(\beta d/2)} e^{-j(\beta d/2)}\right]$$

$$V_{FE} = \frac{Z_\ell}{Z_s + Z_\ell}[as\ above]$$

In these formulae :

$$j = \sqrt{-1}$$

$$\beta = 2\pi / \lambda$$

$$V_{NE}, V_{FE}\ are\ phasors\ (E = E_0 + j0)$$

Advantages of the simple model:

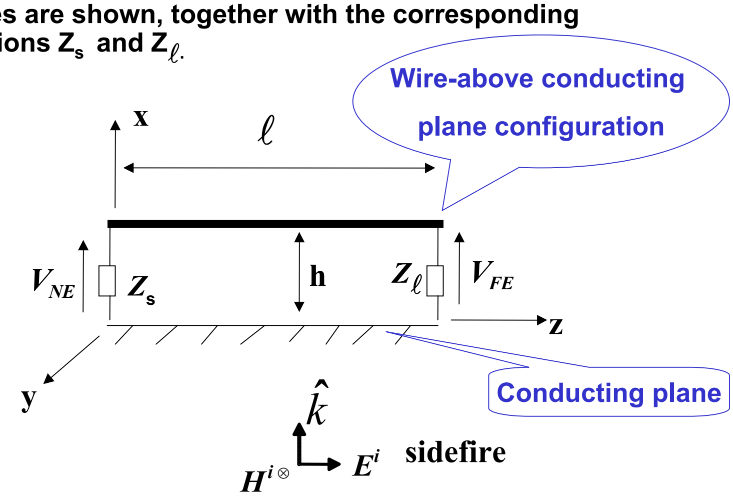
- very easy to use, results obtained quickly
- clear physical understanding of parameters affecting coupling

Limitations of the simple model:

- line must be electrically very short (line length<<λ)
- excitation sources lumped together
- line L and C are neglected
- impedance at line terminations must be of the same order as the line characteristic impedance
- re-radiation from the line is neglected

SIDEFIRE Excitation:

The configuration studied is shown below. The line length is ℓ and the height is h. The near-end (NE) and far-end (FE) quantities are shown, together with the corresponding terminations Z_s and Z_ℓ.



Simple model for sidefire excitation in an wire-above-conducting-plane configuration:

$$V_{NE} = -\frac{Z_s}{Z_s + Z_\ell}\left[j2(h\ell)\, E_0 \beta \frac{\sin(\beta h)}{(\beta h)} \right]$$

$$V_{FE} = \frac{Z_\ell}{Z_s + Z_\ell}\left[as\ above \right]$$

TEM model for the sidefire excitation of a parallel-wire interconnect :

$$V_{NE} = Z_s \frac{E_0}{D} d e^{-j\beta d/2} \frac{\sin(\beta d/2)}{(\beta d/2)}\left[\frac{Z_\ell}{Z_c}\left[\cos(\beta\ell) - 1\right] + j\sin(\beta\ell) \right]$$

$$V_{FE} = -Z_\ell \frac{E_0}{D} d e^{-j\beta d/2} \frac{\sin(\beta d/2)}{(\beta d/2)}\left[\frac{Z_s}{Z_c}\left[\cos(\beta\ell) - 1\right] + j\sin(\beta\ell) \right]$$

where,

$$D = \cos(\beta\ell)(Z_s + Z_\ell) + j\sin(\beta\ell)\left(Z_c + \frac{Z_s Z_\ell}{Z_c} \right)$$

TEM model for the sidefire excitation of a wire-above-conducting-plane interconnect :

$$V_{NE} = -Z_s 2h \frac{E_0}{D} \frac{\sin(\beta h)}{(\beta h)} \left[\frac{Z_\ell}{Z_c} [\cos(\beta \ell) - 1] + j \sin(\beta \ell) \right]$$

$$V_{FE} = Z_\ell 2h \frac{E_0}{D} \frac{\sin(\beta h)}{(\beta h)} \left[\frac{Z_s}{Z_c} [\cos(\beta \ell) - 1] + j \sin(\beta \ell) \right]$$

where,

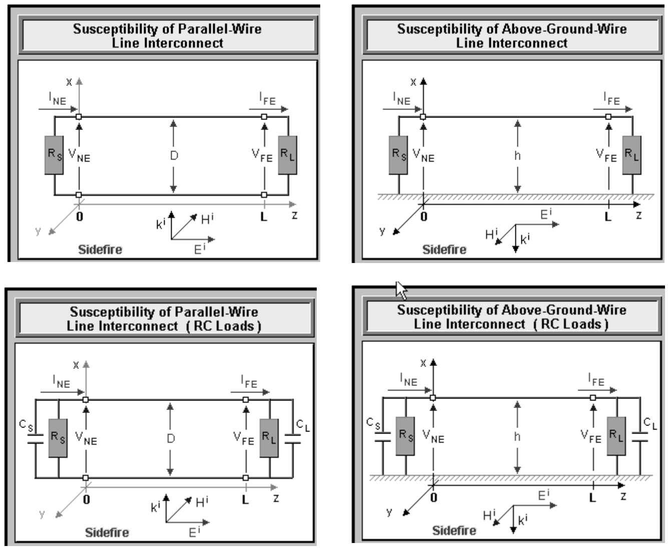
$$D = \cos(\beta \ell)(Z_s + Z_\ell) + j \sin(\beta \ell)(Z_c + \frac{Z_s Z_\ell}{Z_c})$$

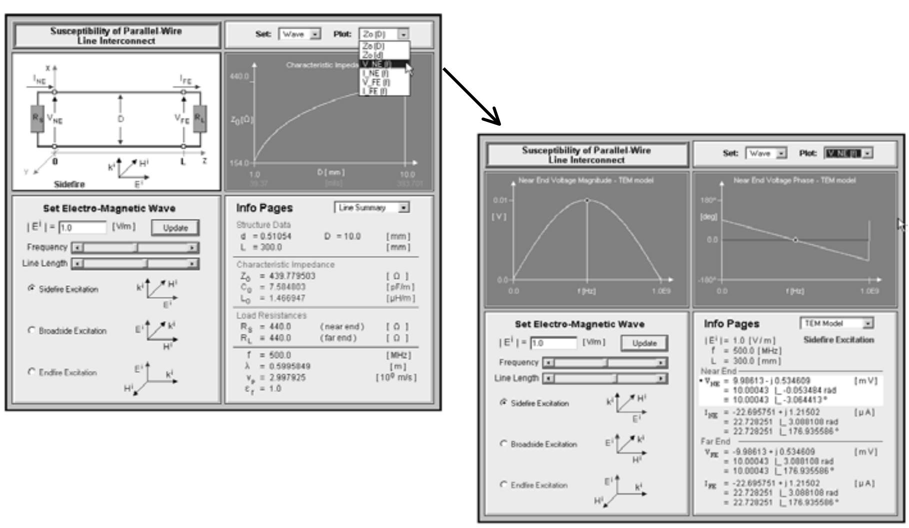
See an example of this applet in the next slide!

Sidefire excitation of a parallel-wire and wire-above-conducting-plane interconnects.

Find out how the length, spacing and line terminations affect the near- and far-end voltages.

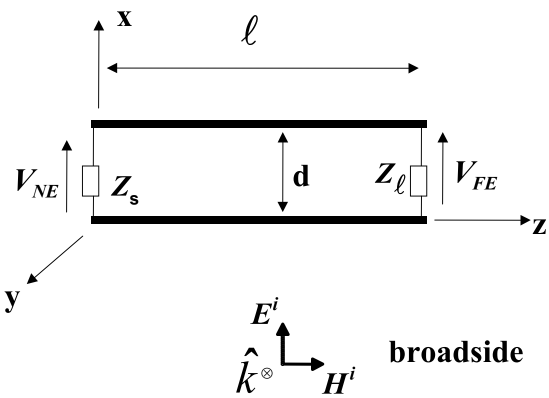
How do the simple and TEM models compare?


Susceptibility of Parallel-Wire Line Interconnect
Susceptibility of Above-Ground-Wire Line Interconnect
Susceptibility of Parallel-Wire Line Interconnect (RC Loads)
Susceptibility of Above-Ground-Wire Line Interconnect (RC Loads)
Sidefire


Susceptibility of Parallel-Wire Line Interconnect
Set Electro-Magnetic Wave
Info Pages
Sidefire Excitation
Broadside Excitation
Endfire Excitation
Near End Voltage Magnitude - TEM model
Near End Voltage Phase - TEM model

Broadside Excitation:

The configuration studied is shown below. The line length is ℓ and the spacing is d. The near-end (NE) and far-end (FE) quantities are shown, together with the corresponding terminations Z_s and Z_ℓ.



For this excitation, the electric field has only an x-component and the magnetic field a z-component. The same approach and notation as for the sidefire excitation is adopted.

Three models are employed and each has the same advantages and limitations as for the sidefire case.

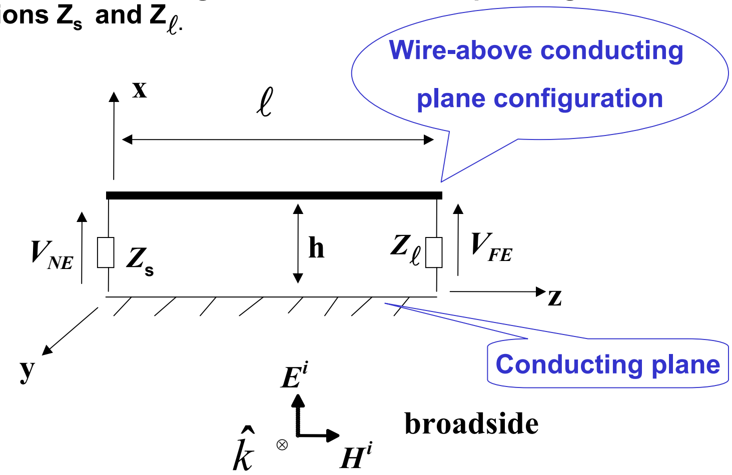
You should now look at the predictions of these models and obtain results to compare with the sidefire excitation.

Simple model for the broadside excitation of a parallel-wire line :

$$V_{NE} = V_{FE} = \frac{Z_s Z_\ell}{Z_s + Z_\ell}\left[-j\omega C_d (d\ell) E_0\right]$$

BROADSIDE Excitation:

The configuration studied is shown below. The line length is ℓ and the height is h. The near-end (NE) and far-end (FE) quantities are shown, together with the corresponding terminations Z_s and Z_ℓ.



Simple model of the broadside excitation of a wire-above-conducting-plane interconnect :

$$V_{NE} = V_{FE} = \frac{Z_s Z_\ell}{Z_s + Z_\ell}\left[-j\omega C_d \, 2(h\ell)E_0\right]$$

TEM model for the broadside excitation of a parallel-wire interconnect :

$$V_{NE} = -Z_s \frac{dE_0}{D}\left[\cos(\beta\ell) - 1 + j\frac{Z_\ell}{Z_c}\sin(\beta\ell)\right]$$

$$V_{FE} = Z_\ell \frac{dE_0}{D}\left[1 - \cos(\beta\ell) - j\frac{Z_s}{Z_c}\sin(\beta\ell)\right]$$

where,

$$D = \cos(\beta\ell)(Z_s + Z_\ell) + j\sin(\beta\ell)(Z_c + \frac{Z_s Z_\ell}{Z_c})$$

TEM model for the broadside excitation of a wire-above-conducting-plane interconnect :

$$V_{NE} = -Z_s \frac{2hE_0}{D}\left[\cos(\beta\ell) - 1 + j\frac{Z_\ell}{Z_c}\sin(\beta\ell)\right]$$

$$V_{FE} = Z_\ell \frac{2hE_0}{D}\left[1 - \cos(\beta\ell) - j\frac{Z_s}{Z_c}\sin(\beta\ell)\right]$$

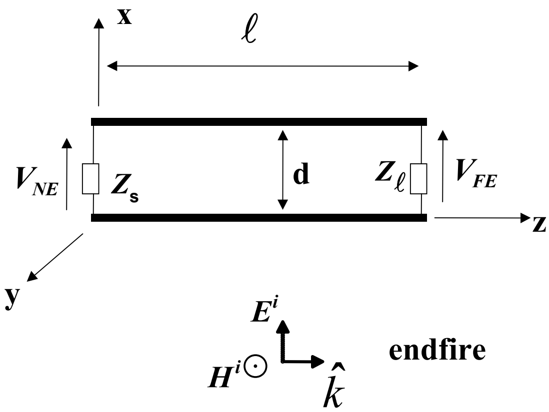
where,

$$D = \cos(\beta\ell)(Z_s + Z_\ell) + j\sin(\beta\ell)(Z_c + \frac{Z_s Z_\ell}{Z_c})$$

Now try the applet!

ENDFIRE Excitation:

The configuration studied is shown below. The line length is ℓ and the spacing is d. The near-end (NE) and far-end (FE) quantities are shown, together with the corresponding terminations Z_s and Z_ℓ.



For this type of excitation the electric field has an x-component only and the magnetic field a y-component. The relevant models are given below:

<u>Simple</u> model of an endfire excitation of an interconnect:

$$V_{NE} = -\frac{Z_s}{Z_s + Z_l} V_{src} + \frac{Z_s Z_\ell}{Z_s + Z_\ell} I_{src}$$

$$V_{FE} = \frac{Z_\ell}{Z_s + Z_\ell} V_{src} + \frac{Z_s Z_\ell}{Z_s + Z_\ell} I_{src}$$

where,

$$V_{src} = j(d\ell)\beta E_0 e^{-j\beta\ell/2}$$

$$I_{src} = -j\omega C_d (d\ell) E_0 e^{-j\beta\ell/2}$$

For a parallel-wire interconnect

For a wire-above-conducting-plane interconnect replace d by 2h in the formulae for V_{src} and I_{src}.

TEM model of endfire excitation of wire interconnect:

For parallel-wire interconnect

$$V_{NE} = -jZ_s \frac{dE_0}{D} \sin(\beta\ell)\left[1 + \frac{Z_\ell}{Z_c}\right]$$

$$V_{FE} = Z_\ell \frac{dE_0}{2D}\left[1 - \frac{Z_s}{Z_c}\right]\left[1 - \cos(2\beta\ell) + j\sin(2\beta\ell)\right]$$

where,

$$D = \cos(\beta\ell)(Z_s + Z_\ell) + j\sin(\beta\ell)\left(Z_c + \frac{Z_s Z_\ell}{Z_c}\right)$$

For wire-above-conducting-plane interconnect replace d by 2h in the above formulae.

Completion of these exercises should have given you an insight into the following:

- **the way in which line length affects coupling (i.e. resonances)**
- **the way in which the loop area of the line (dℓ) affects coupling**
- **in what way the three excitations are different and whether there is any merit in configuring the line in any particular way if the primary EMI excitation threat is known**
- **the influence of terminations on severity of coupling for the different excitations**
- **worst case coupling and best configuration/excitation for maximum immunity (design guidelines)**
- **range of applicability of the three models**

Simple model (low frequency), TEM model (medium frequency), High-frequency model

•Further Reading

1. "Analysis of Multiconductor Transmission Lines", C. R. Paul, Wiley 1994, (chapter 7)
2. "Principles and Techniques of Electromagnetic Compatibility", C. Christopoulos, CRC Press 1995, (chapter 7)
3. "EMC Analysis Methods and Computational Models", F. M. Tesche, M. V. Ianoz and T. Karlsson, Wiley 1997, (chapter 7)

CONCLUDING REMARKS

•Many important EMC phenomena can be studied by simple models

•Applets can be used to relieve the student from excessive calculation and to illustrate trends in design

•Applet based material offers a virtual laboratory accessible from anywhere via the internet

•We plan to enhance and extend this work to provide a sophisticated and effective learning environment