Invited paper

PROTECTING SIGNALS FROM EMI - CROSSTALK ANALYSIS

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Abstract: The paper treats a very important aspect of the design of an electromagnetically compatible product - the "crosstalk". In order to understand how to model crosstalk, the fundamental theory and assumptions are presented. A brief discussion on the frequently used analysis methods is given. Throughout the text, a full wave program that uses the method of moments is used to investigate and solve the cross talk problem for a three conductor transmission line.

Keywords: Crosstalk, EMC, EMI, shielding, Wire-Mom

INTRODUCTION

Electromagnetic compatibility is the situation in which electrical and electronic devices and systems work as intended, both within themselves and in their electromagnetic environment.

Electromagnetic interference (EMI) is said to exist when unwanted voltages or currents are present so that they adversely affect the performance of a device or system. Such voltages or currents may reach the victim circuit or device by conduction and/or radiation. In all cases, electromagnetic interference arises because of a combination of three factors: a source, a propagation path, and a response, (Figure 1) at least one of which is unplanned.

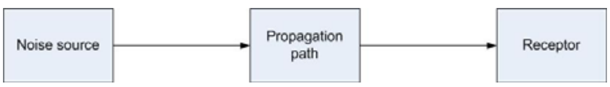

Fig. 1. *Electromagnetic interference*

Radiated noise is noise that arrives at the victim circuit in form of electromagnetic radiation. It causes trouble by inducing extraneous voltages in the victim circuit. Frequently, photoreaction against radiated noise is done by the use of effective shielding techniques.

Conducted noise is noise that arrives at the victim circuit already in the form of an extraneous voltage, typically via the AC or DC power lines.

Different types of EMI, radiated and conducted noise, protection against EMI is extensively described in [1, 2, 3].

A system is electromagnetically compatible with its environment if it satisfies three criteria:

1. It does not cause interference with other systems.
2. It is not susceptible to emissions from other systems.
3. It does not cause interference with itself.

A source (also referred to as an emitter) produces the emission, and a transfer or coupling path transfers the emission energy to a receptor (receiver), where it is processed, resulting in either desired or undesired behavior. Interference occurs if the received energy causes the receptor to behave in an undesired manner. Transfer of electromagnetic energy occurs frequently via unintended coupling modes.

This suggests that there are three ways to prevent interference:

1. Suppress the emission at its source.
2. Make the coupling path as inefficient as possible.
3. Make the receptor less susceptible to the emission.

CROSSTALK

A very important aspect of the design of an electromagnetically compatible product is the "crosstalk". This essentially refers to the unintended electromagnetic coupling between wires and PCB lands that are in close proximity. Crosstalk is distinguished from antenna coupling in that it is a near-field coupling problem. Crosstalk between wires in cables or between lands on PCBs concerns the intrasystem interference performance of the product; that is, the source of the electromagnetic emission and the receptor of this emission are within the same system. Thus this reflects the third concern in EMC: the design of the product such that it does not interfere with itself. With clock speeds and data transfer rates in digital computers steadily increasing, crosstalk between lands on PCBs is becoming a significant mechanism for interference in modern digital systems. There are also cases where crosstalk can affect the radiated and/or conducted emissions of the product. Suppose that a ribbon cable internal to a product is placed in close proximity to wires that connect to a peripheral cable that exits the product. Crosstalk between the two cables can induce signals on the peripheral cable that may radiate externally to the product, causing the product to be out of compliance with the radiated emission regulatory limits. If this internal coupling occurs to the power cord of the product, these coupled signals may cause it to fail the conducted emission regulatory requirements. Crosstalk can also affect the susceptibility of a product to emissions from another

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product. For example, emissions from some other product that are coupled to a peripheral cable of this product may couple, internal to the product, to some other cable internal to it where the susceptibility to this signal may be enhanced.

In order to understand how to model crosstalk, it is important to understand the analysis of two-conductor transmission lines. For a two-conductor transmission line there is no crosstalk. In order to have crosstalk, there must be three or more conductors. However, the notions involved in two-conductor transmission-line theory carry over to a large degree to the case of multiconductor transmission lines and simplify the understanding of the behavior of those lines.

THREE-CONDUCTOR TRANSMISSION LINES AND CROSSTALK

Virtually all of the techniques for the analysis of two conductor transmission lines [1] can be directly extended to the case of coupled transmission lines that consist of any number of parallel conductors. These types of transmission lines are referred to as multiconductor transmission lines (MTLs).

Adding a third conductor to the two-conductor system provides the possibility of generating interference between the circuits attached to the ends of the line conductors resulting from crosstalk. In order to illustrate this important phenomenon, consider the three-conductor line shown in Fig. 2.

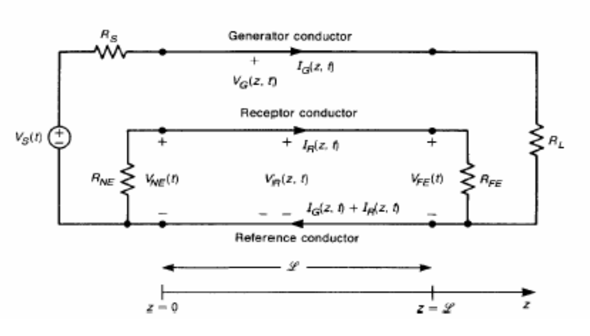

Fig. 2. *The general three-conductor transmission line, illustrating crosstalk*

A source consisting of a source resistance R_S and a source voltage $V_S(t)$ is connected to a load R_L via a generator conductor and a reference conductor. Two other terminations, represented by resistors R_{NE} and R_{FE}, are also connected by a receptor conductor and this reference conductor. These terminations represent the input circuitry looking into the terminals of the terminations. Linear, resistive terminations are shown for illustration, but may include capacitors and/or inductors. The line conductors are assumed to be parallel to the z axis and are of uniform cross section along the line. We will also assume that any surrounding dielectric inhomogeneities also have uniform cross sections along the line axis, so that the lines we will consider will be uniform lines. The generator circuit consists of the generator conductor and the reference conductor and has current $I_G(z, t)$ along the conductors and voltage $V_G(z, t)$ between them. All the voltages are with respect to the reference conductor. The current and voltage associated with the generator circuit will generate electromagnetic fields that interact with the receptor circuit, which consists of the receptor conductor and the reference conductor.

This interaction will induce current $I_R(z, t)$ and voltage $V_R(z, t)$ along the receptor circuit. This induced current and voltage will produce voltages $V_{NE}(t)$ and $V_{FE}(t)$ at the input terminals of the terminations that are attached to the ends of the receptor circuit. The subscripts NE and FE refer to "near end" and "far end," respectively, with reference to the end of the line adjacent to the end of the generator circuit that contains the driving source $V_s(t)$. The line is of total length L and extends from z=0 to z=L.

The objective in a crosstalk analysis is to determine (predict) the near-end and far-end voltages $V_{NE}(t)$ and $V_{FE}(t)$ given the line cross-sectional dimensions, and the termination characteristics $V_S(t)$, R_S, R_S, R_{NE}, and R_{FE}. There are two types of analysis that we might be interested in: time-domain analysis and frequency domain analysis. Time-domain crosstalk analysis is the determination of the time form of the receptor terminal voltages $V_{NE}(t)$ and $V_{FE}(t)$ for some general time form of the source voltage $V_S(t)$. Frequency-domain crosstalk analysis is the determination of the magnitude and phase of the receptor terminal phasor voltages $\hat{\mathbf{V}}_{NE}(j\omega)$ and $\hat{\mathbf{V}}_{FE}(j\omega)$ for a sinusoidal source voltage $V_S(t) = V_S \cos(\omega \cdot t + \phi)$.

Frequency-domain analysis also presumes a steady state, i.e., the sinusoidal source has been attached a sufficient length of time that the transient response has decayed to zero.

Some typical three-conductor lines representing typical configurations to which this analysis applies are shown in Fig. 3. Figure 3 shows wire-type lines consisting of conductors of circular cylindrical cross section (wires). Figure 3a shows a configuration of three wires where one wire serves as the reference conductor for the line voltages. A ribbon cable is typical of this configuration. Figure 3b shows two wires where an infinite, perfectly conducting (ground) plane serves as the reference conductor. The third wire-type configuration shown in Fig. 3c consists of two wires within an overall cylindrical shield that serves as the reference conductor. There are many applications where cables are surrounded by an overall shield as in Fig. 3c in order to prevent unwanted coupling of external electromagnetic fields to the interior wires. All these configurations are shown as being bare wires, i.e., without dielectric insulations. They are then said to be in a homogeneous medium since the surrounding medium has one relative permittivity (that of free space $\varepsilon_{r=}1$). Practical wires (with the exception of high-voltage power transmission lines) have cylindrical dielectric insulations surrounding them for obvious reasons. It is required to develop equations for the per-unit-length capacitances and inductances of these configurations. Determining these per-unit-length parameters in simple, closed-form equations is not possible if dielectric insulations are included, and numerical approximation methods must be used.

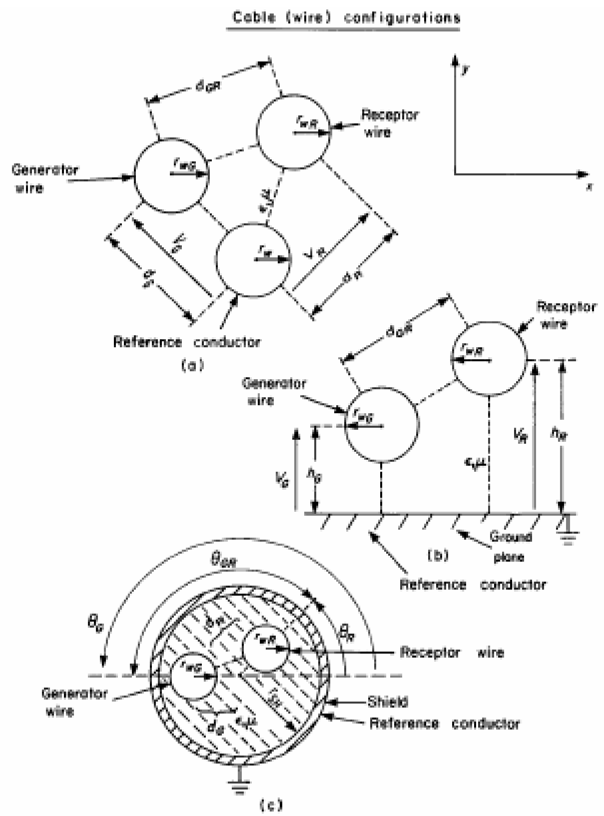

Fig. 3. *Wire-type line cross sections whose reference conductors are (a) another wire, (b) an infinite ground plane, or (c) an overall, cylindrical shield*

For the homogeneous medium cases, all voltages and current waves will travel down the line with the same velocity

$$v = \frac{v_o}{\sqrt{\varepsilon_r}} \tag{1}$$

where $v_o = 3 \cdot 10^8 \ m/s$ and ε_r is the relative permittivity of the surrounding (assumed homogeneous) medium. If wire insulations are included in the analysis, the surrounding medium will be inhomogeneous and the voltage and current waves will travel down the line with different velocities of propagation further complicating the analysis.

Electrical dimensions and waves

Although Maxwell's equations govern all electrical phenomena, they are quite complicated, mathematically. Hence we use, where possible, simpler approximations to them such as lumped-circuit models and Kirchhoff's laws. The important question here is when we can use the simpler lumped-circuit models and Kirchhoff's laws instead of Maxwell's equations when analyzing a problem. The essence of the answer is when the largest dimension of the circuit is electrically small, for example, much smaller than a wavelength at the excitation frequency of the circuit sources. Typically we might use the criterion that a circuit is electrically small when the largest dimension is smaller than one-tenth of a wavelength. In that case, a physical dimension is said to be electrically small in that the phase shift as a wave propagates across that dimension may be ignored. These concepts give rise to the rule of thumb that lumped-circuit models of circuits are an adequate representation of the physical circuit so long as the largest electrical dimension of the physical circuit is less than, say, one-tenth of a wavelength.

The electrical dimensions of a circuit or other electromagnetic structure need to be calculated to determine whether it is electrically small ($L<1/10\lambda$). One can determine this by first calculating the wavelength at the highest frequency of interest

$$\lambda = \frac{v}{f} = \frac{v_o}{f\sqrt{\varepsilon_r \mu_r}} \tag{2}$$

and then computing k in relation to

$$k = \frac{L}{\lambda} \tag{3}$$

where v_o is the velocity of propagation in free space (air), ε_r relative permittivity and μ_r relative permeability of the surrounding medium.

THE TRANSMISSION-LINE EQUATIONS FOR LOSSLESS LINES

The fundamental assumption involved in the analysis of all multiconductor transmission lines is that the transverse electromagnetic (TEM) mode of propagation is the only field structure present on the line. The TEM field structure assumes that both the electric and the magnetic field vectors lie in the transverse (xy) plane perpendicular to the line (z) axis, i.e., the electric and magnetic fields do not have a component along the line axis. Under the TEM field structure assumption, line voltages $V_G(z, t)$ and $V_R(z, t)$, as well as line currents $I_G(z, t)$ and $I_R(z, t)$, can be uniquely defined for excitation frequencies other than dc. The total current flowing to the right at any line cross section is zero, so that the currents return through the reference conductor. Furthermore, the TEM field structure is identical to static (dc) one. This allows the determination of the per-unit-length parameters of inductance and capacitance strictly from dc methods in the cross-section (xy) plane. The pure TEM field structure cannot exist for (1) imperfect line conductors or (2) an inhomogeneous surrounding medium [4]. Nevertheless, for either cases the deviation from a TEM field structure is typically small for "good conductors," typical line cross-sectional dimensions, and frequencies in the lower GHz range. This is referred to as the quasi-TEM mode assumption.

Hence the per-unit-length parameters of inductance and capacitance can be found using static solution methods, which greatly simplifies their determination. Losses in the line occur via two mechanisms: (1) the losses in the line conductors and (2) losses in the surrounding medium. In order to predict the crosstalk and understand the basic mechanism as well as the parameters affecting it, we will ignore these losses in formulating the multiconductor transmission-line equations in order to simplify their solution. Ignoring losses gives first-order and reasonably accurate predictions of the crosstalk. To obtain the MTL equations that we must solve in order to predict the crosstalk, we construct the per-unit-length equivalent circuit for a Δz section as shown in Fig.4.

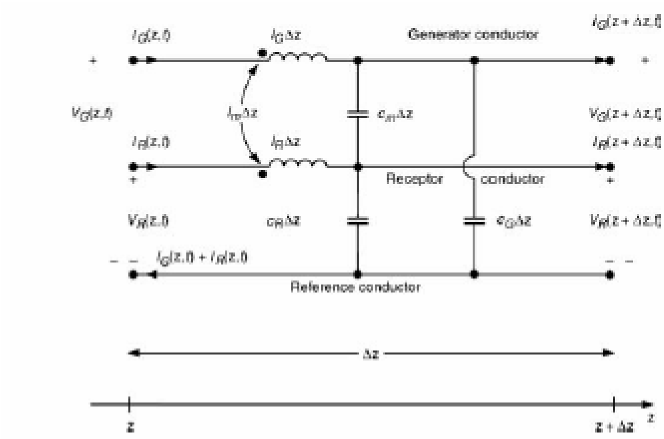

Fig. 4. *The per unit-length equivalent circuit of a three-conductor transmission line*

The generator and receptor circuits have per-unit-length self inductance l_G and l_R, respectively, associated with them and a per-unit-length mutual inductance l_m between the two circuits. The line currents produce magnetic fluxes penetrating each loop formed by the conductor and the reference conductor, and these inductances represent the effect of those fluxes via Faraday's law. The line voltages (between each conductor and the reference conductor) produce charges on the conductors that generate electric fields between each pair of conductors. This effect is represented by capacitances. The per-unit-length self-capacitances between the generator conductor and the reference conductor and between the receptor conductor and the reference conductor are represented by c_G and c_R, respectively. The per-unit-length mutual capacitance between the generator and receptor conductors is represented by c_M. In a Δz section of the line the total inductance or capacitance is the per-unit-length value multiplied by Δz. The MTL equations can be determined from this per-unit-length equivalent circuit using circuit analysis principles and letting Δ - 0, giving [4]:

$$\begin{aligned}\frac{\partial V_G(z,t)}{\partial z} &= -l_G\frac{\partial I_G(z,t)}{\partial t} - l_m\frac{\partial I_R(z,t)}{\partial t}\\ \frac{\partial V_R(z,t)}{\partial z} &= -l_m\frac{\partial I_G(z,t)}{\partial t} - l_R\frac{\partial I_R(z,t)}{\partial t}\end{aligned} \tag{4a}$$

and

$$\begin{aligned}\frac{\partial I_G(z,t)}{\partial z} &= -(c_G+c_m)\frac{\partial V_G(z,t)}{\partial t} + c_m\frac{\partial V_R(z,t)}{\partial t}\\ \frac{\partial I_R(z,t)}{\partial z} &= c_m\frac{\partial V_G(z,t)}{\partial t} - (c_R+c_m)\frac{\partial V_R(z,t)}{\partial t}\end{aligned} \tag{4b}$$

Writing these equations in matrix form yields [4]:

$$\begin{aligned}\frac{\partial}{\partial z}\mathbf{V}(z,t) &= -\mathbf{L}\frac{\partial}{\partial t}\mathbf{I}(z,t)\\ \frac{\partial}{\partial z}\mathbf{I}(z,t) &= -\mathbf{C}\frac{\partial}{\partial t}\mathbf{V}(z,t)\end{aligned} \tag{5}$$

where

$$\mathbf{V}(z,t) = \begin{bmatrix} V_G(z,t) \\ V_R(z,t)\end{bmatrix} \qquad \mathbf{I}(z,t) = \begin{bmatrix} I_G(z,t) \\ I_R(z,t)\end{bmatrix} \tag{6}$$

and per-unit-length parameter matrices are

$$\mathbf{L} = \begin{bmatrix} l_G & l_m \\ l_m & l_R \end{bmatrix} \tag{7a}$$

$$\mathbf{C} = \begin{bmatrix} (c_G+c_m) & l_m \\ -c_m & (c_R+c_m) \end{bmatrix} \tag{7b}$$

There are different methods of solving these equations and predicting the crosstalk. The crosstalk can either be computed using so called full-wave methods or using some simplified formulation. The full wave methods are based on a rigorous formulation using Maxwell's equations, which then are solved by a numerical method. Some of the more popular numerical methods that are used to essentially solve for the 2D static fields are (1) the method of moments (MoM), (2) the finite-difference method, and (3) the finite-element method (FEM) The advantage of these methods are that all effects are taken into account. The only approximation that is introduced is the one due to the numerical method being used.

One simplified formulation is to use the transmission line equations. This is a simplified formulation since a TEM mode of propagation along the transmission line is assumed. Radiation effects are not taken into account. An approximate method called the inductive-capacitive coupling model, and an exact solution consisting of a SPICE/PSPICE model is thoroughly discussed in [1]. In the inductive-capacitive coupling model, the crosstalk takes place via two distinct coupling mechanisms: magnetic field coupling due to mutual inductance between the two circuits and electric field coupling due to mutual capacitance between the two circuits. There are two key assumptions in this model: (1) weak coupling between the generator and receptor circuits is assumed, that is, the coupling is a one-way effect from the generator circuit to the receptor circuit; and (2) the line is assumed to be electrically short at the frequency of the driving source in the generator circuit.

The per-unit-length parameters

It is of no use to solve the MTL equations if the per-unit-length parameters cannot be determined for the particular line cross-sectional configuration. All the structural information about the line such as type of conductor, wire radii, and wire separation that distinguish one line from another are contained in these two parameters and nowhere else. Corresponding formulas for the PCB type structures are difficult to obtain, and generally must be found using approximate, numerical methods. Again, a TEM mode of propagation along the transmission line is assumed. This means that the per-unit-length parameters for lossless lines, L and C, can be determined by solving static Laplace equation for the cross section of the transmission line. For a homogeneous cross section the L and C matrices are related through:

$$\mathbf{L}=\mu\varepsilon\mathbf{C}^{-1} \quad (8)$$

This relation is used when computing per-unit-length parameters in a way such as only one type of per-unit-length, L or C, has to be computed. Often it is easier to compute the capacitance matrix for a homogeneous cross section, e.g. free space, and determine the inductance matrix through the relation above. When this is done, dielectric materials, if present, is inserted in the capacitance matrix is computed once again. This is the method LC-Calc program [5,6] uses to compute per-unit-length parameters. LC- Calc is a finite difference program for computing the per-unit-length parameters L and C. This program computes the L and C matrices and generates a lumped circuit file that can be used by a circuit analyze program as, for instance, PSPICE. The program LC-Calc [6] is available through the author.

NUMERICAL EXAMPLE OF CROSSTALK PROBLEM FOR THREE-CONDUCTOR TRANSMISSION LINE

Structure definition

The crosstalk has been investigated for three-conductor transmission line structure that is terminated to a common ground plane. The structure that has been investigated is presented in Figure 5. In the Figure, $V_s(t)=V_o\sin(2\pi ft)$ V_o=1V. Geometrical data is L=15cm, H=2mm, the distance between the generator and the receptor is 2mm and the radius of the wires is 0,1mm.

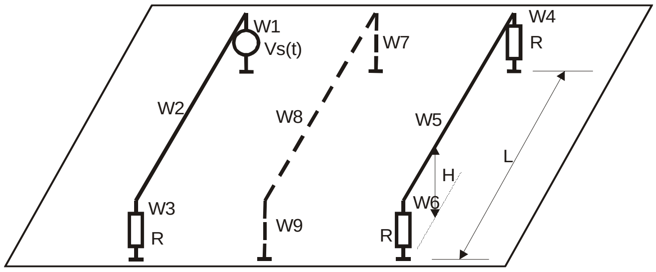

Fig. 5. *Investigated structure*

The considered line length is a wavelength at f=2000MHz. Hence the line is electrically short for frequencies bellow approximately 200MHz.

Three different cases were subject of analysis. First a generator-receptor structure has been analyzed. One of the traces (W1-W2-W3) was excited at one end and terminated with resistor R at the other end. The second trace (W4-W5-W6) was terminated with resistors R at both ends.

For the computation the conductors were divided in 31 current elements along the length and 5 in the transverse direction.

The total current in the conductors for the frequency range 10MHz-1000MHZ was calculated. The total current was inspected through the resistor in conductor W4 and this response is further referred as near-end crosstalk. The total current inspected through the resistor in conductor W6 gives far-end crosstalk response.

The knowledge of which type of coupling dominates is very important when steps to reduce coupling are taken. Generally, the primary means of noise coupling can be determined by changing the signal-source impedance. When a circuit or termination has high impedance the capacitive coupling is predominant. The noise voltage created at the receiver is in fact the noise current multiplied by the receiver impedance.

Correspondingly, when the circuit or termination impedance is low, the primary means of coupling is inductive. The reason for this is that low impedances encourage the flow of appreciable currents, resulting in large amount of flux and hence large inductive coupling.

Therefore, two different load cases were considered:

Low circuit impedance R=50Ω (load case 1, in the text);

High circuit impedance R=1000Ω (load case 2, in the text).

Computational method

The total currents in the conductors were computed and visualized using the EMC frequency domain software Wire-MoM [7]. Wire-MoM is a method of moments program for the analysis of wire structures. The wire structure can be placed in free space or over an infinite perfectly conducting ground plane. Wires can also be connected to the ground plane. The program computes the current in the wires in the frequency domain due to either voltage sources on the wires or an incident plane wave. Also the near field, far field and S-parameters can be computed. Impedance elements that serve as models for discrete components (R, L and C) can be placed at arbitrary positions along the wires. Several tools are available for creating and modifying complex wire structures, e.g. tools for creating sphere, box, cylinder etc. Computed results can be visualized directly in the program, radiation pattern can be plotted as a color-coded 3D-plot. The program is based on the EFIE (Electric Field Integral Equation) formulated in the frequency domain.

The electric field integral equation (EFIE) for the current distribution on three-dimensional conducting wire structures in free space is solved in the frequency domain using the method of moments (MoM).

$$\overline{E} = f(\overline{J}) \tag{9}$$

where $\overline{E}$ is incident field and $\overline{J}$ is the induced current. The electric field integral equation can be derived from the Maxwell's equations.

The method of moments is a technique for solving complex integral equations by reducing them to a system of simpler linear equations. The induced current $\overline{J}$ is expanded as a finite sum of known basis functions, $\overline{b_i}$

$$\overline{J} = \sum_{i=1}^{M} J_i \overline{b_i} \tag{10}$$

where coefficients J_i are to be determined. The unknown coefficients of the induced current are the terms of the $\overline{J}$ vector, and are related to the conductors (wires).

As a result of a Wire-MoM simulation both near end and far field points can be inspected to give the EM-field intensities. The induced currents on the wires are computed as functions of coordinates and frequency.

The accuracy of the program has been validated with measurements [5] as well as other commercial field computational programs. EMC Software Wire-Mom [7] is a free-ware and can be obtained with permission of the author. Although there is no help file, the program is user friendly and also [8] is very useful for start.

Near end and far end crosstalk

In the first case that was considered, the circuit impedance was set to be low i.e. R=50Ω (load case 1) simulating dominant inductive coupling.

Figure 6 shows the frequency distribution of the current i.e. the near and far end crosstalk for load case 1.

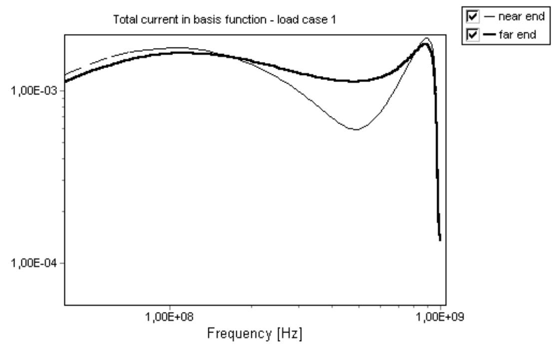

Fig. 6. *Near end and far end crosstalk for load case 1*

Figure 7 shows the near and far end crosstalk for load case 2, modeled in the program when high circuit impedance R=1000Ω was set. When the circuit impedance is high, the primary means of coupling is capacitive.

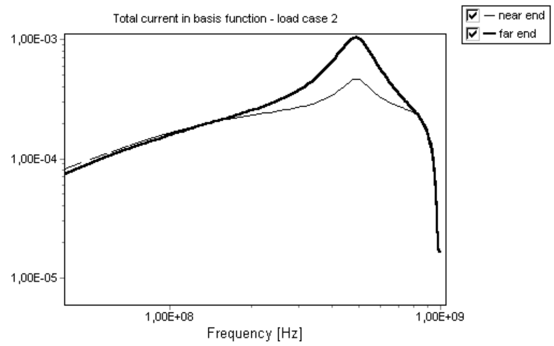

Fig. 7. *Near end and far end crosstalk for load case 2*

From Figure 6 and Figure 7 it can be seen that current flows in the receptor i.e. the victim trace as a result of the mutual coupling between the conductors. The behavior of the investigated structure is different, depending on the circuit impedance. From the results it can be observed that inductive coupling favors lower frequency components whereas capacitive coupling favors transfer of the higher-frequency components of a signal. Capacitive transfer becomes prevalent as frequency increases because the capacitive reactance coupling the two circuits decreases with frequency.

METHODS FOR REDUCING CROSSTALK

We now consider methods for reducing the crosstalk in a three-conductor line. Suppose the near- or far-end crosstalk in the previous three-conductor line exceeds desired levels, causing interference with the terminations at the ends of the receptor circuit. For wire-type lines, there are two common methods for reducing the crosstalk; replace the generator and/or receptor wire with a shielded wire or a twisted pair. Consider replacing the receptor wire with a shielded wire, as shown in Fig. 8. In order to illustrate the effect of a shielded wire on crosstalk, we will use an infinite ground plane as the reference conductor. Other choices of reference conductors such as another wire will give the same conclusions.

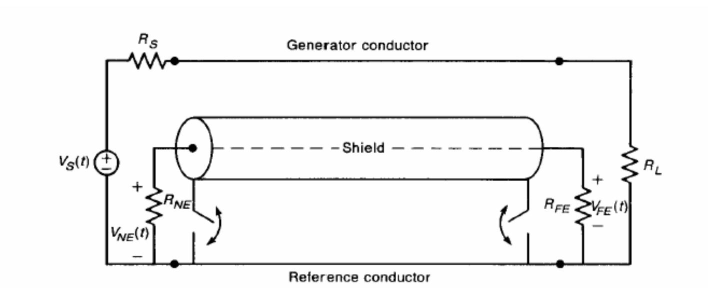

Fig. 8. *Shield around the receptor circuit wire to reduce crosstalk*

Typically, the shield is connected to the reference conductor ("grounded") at one end or at both ends. If the shield is connected to the reference conductor at either end, the shield voltage is reduced to zero and the capacitive coupling contribution is removed.

This is the origin of the notion that a shielded wire eliminates electric field or capacitive coupling wherein the electric field lines from the generator circuit terminate on the shield and not on the receptor wire. In order for the shield to eliminate capacitive coupling, the shield voltage must be zero. For an electrically short line, grounding the shield at either end will cause the voltage all along the shield to be approximately zero. As the line length increases, electrically, the shield must be grounded at multiple points spaced some 1/10 λ along it in order to approximate this.

Next we consider inductive coupling. A shield must be grounded at both ends in order to eliminate inductive coupling. In order to show this, consider the magnetic fields generated by the generator wire current as shown in Fig. 9.

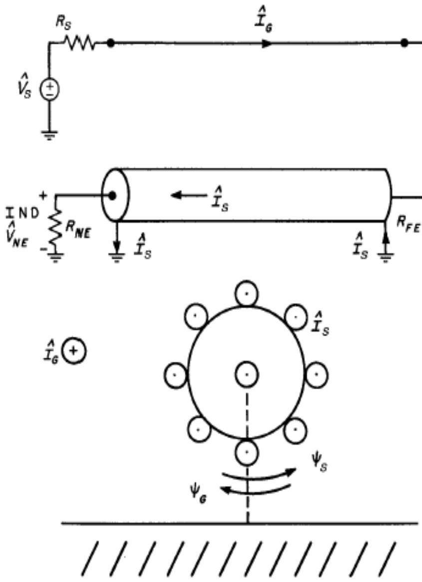

Fig. 9. *Illustration of the effect of placing a shield around a receptor wire on inductive coupling*

The generator wire current I_G produces a magnetic flux ψ_G in the shield–ground plane circuit. This induces by Faraday's law an emf in the shield circuit that produces a secondary current I_S flowing back along the shield. The flux of this induced shield current tends to cancel that of the generator wire current. It is this process that allows a shielded wire to eliminate inductive or magnetic field coupling. Observe that if the shield is not grounded at both ends, then there is no path for allowing a current, I_S, to flow back along the shield, thereby generating a magnetic flux through the shield–ground plane loop that counteracts the flux due to the generator wire loop. Hence if the shield is not grounded at both ends, inductive coupling will not be eliminated.

NUMERICAL EXAMPLE FOR SHIELDING EFFECTIVENESS

In order to reduce crosstalk, a shield conductor (W7-W8-W9) was added in parallel both to the generator and receptor conductors at equal distance from both conductors, as shown in Figure 5. The shield is connected to the reference conductor ("grounded") at one end or at both ends. In the text, the shielding technique with two ends grounded is considered as "shield A". In the next case a grounding wire (W9) from the shield conductor was removed. Further in the text, the shielding technique with single end grounded is referred as "shield B".

Shielding Effectiveness For Inductive Coupling

With inductive coupling, the physical mechanism involved is a magnetic flux density B from some interference source that links with a current loop in the victim circuit. According to Lenz's law this will induce voltage

$$v = -NA(\frac{dB}{dt}) \qquad (11)$$

where in this case N=1 and A is the area of the current loop in the victim trace.

Shielding against inductive coupling means controlling the dimensions of the current loops in the circuit.

The effectiveness of the shield in terms of grounding technique is investigated. Although only the near end crosstalk will be presented, computations showed that the far end crosstalk confirms the conclusions that will be derived.

A. Shield A – two ends grounded

Figure 10 compares the near end crosstalk in the receptor without shield and with two ends grounded shield.

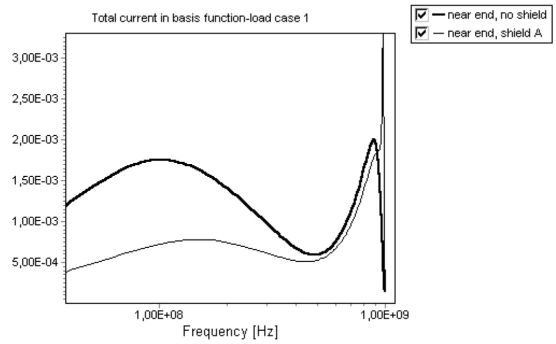

Fig. 10. *Near end crosstalk with and without grounded shield for load case 1*

From the results in Figure 10 it can be seen that for load case 1, when inductive coupling dominates, the shield trace works as intended in the frequency range considered. However it can be seen that the shield is most effective for frequencies up to 350MHz.

B. Shield B – single end grounded

Many engineers blindly apply single-end grounding for all product designs, without realizing that additional and more complex problems are created using this ground methodology. The effect of the single end grounding (shield B) was investigated and the results are presented in Figure 11 and Figure 12.

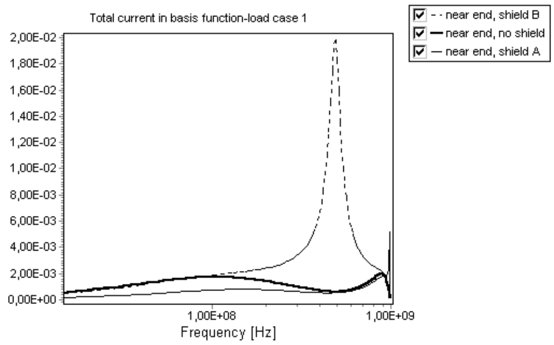

Fig. 11. *Near end crosstalk - shield A versus shield B, load case 1, f=(10-1000)MHz*

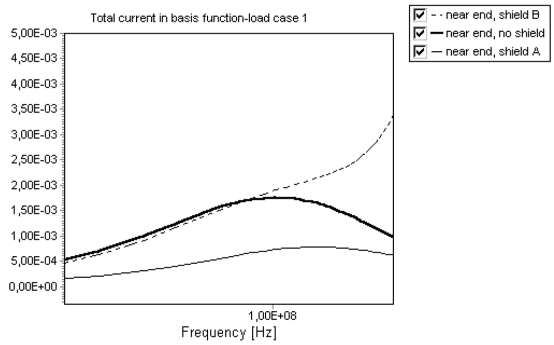

Fig. 12. *Near end crosstalk - shield A versus shield B, load case 1, f=(10-350)MHz*

For inductive coupling single end grounding is less effective compared to the previous case when the shield was grounded in two ends. For higher frequencies the single end grounded shield trace in fact can increase the crosstalk instead of decrease it (Figure 11).

In conclusion, when inductive coupling dominates the shield trace works as intended because current flows in the shield trace and therefore the shield has to be grounded in two ends.

Shielding Effectiveness For Capacitive Coupling

Capacitive coupling involves the passage of interfering signals through mutual or stray capacitance. Using shielding cables may reduce noise due to capacitive coupling. The shield is a Gaussian or equi-potential surface on which can electric fields discharge and return to ground without affecting the receptor conductor.

A. Shield A – two ends grounded

When capacitive coupling dominates, two ends grounded shielding works as desired for all frequencies considered (10-1000) MHz. This can be clearly seen in Figure 13.

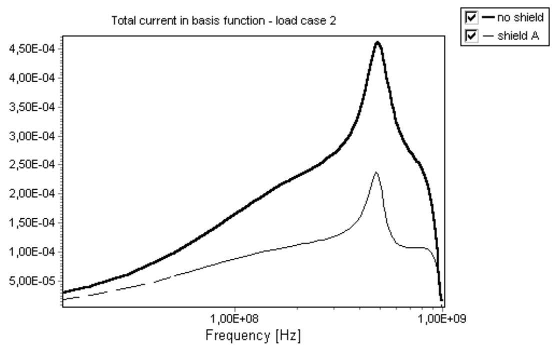

Fig. 13. *Near end crosstalk with and without grounded shield for load case 2*

B. Shield B – single end grounded

Figure 14 and Figure 15 show computational results for the effectiveness of the single end grounding technique, when capacitive coupling dominates between the conductors. Figure 15 differs from Figure 14 in the frequency range considered.

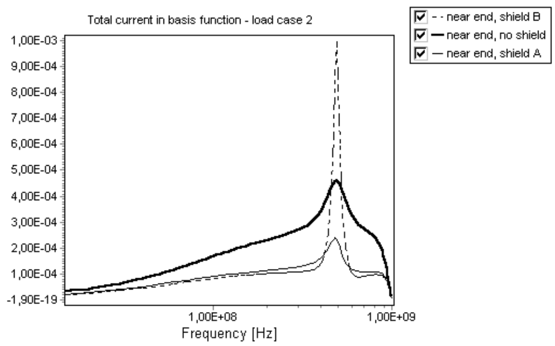

Fig. 14. *Near end crosstalk - shield A versus shield B, load case 2, f=(10-1000)MHz*

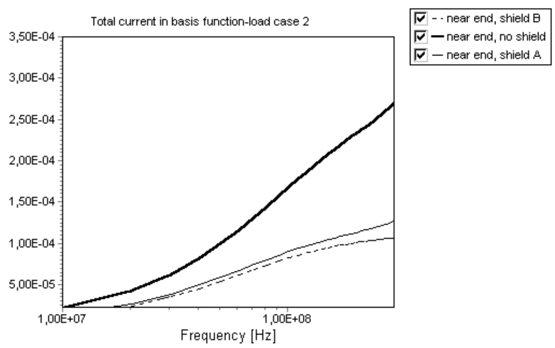

Fig. 15. *Near end crosstalk - shield A versus shield B, load case 2, f=(10-350)MHz*

From these results, one can conclude that for capacitive coupling the use of single end grounded shield reduces the crosstalk more effectively compared to shield with two ends grounded. This conclusion is valid for frequencies up to 350MHZ (see Figure 15) i.e. where the line is still electrically short. As the line length increases, electrically, the shield must be grounded at multiple points along it in order to reduce crosstalk for higher frequencies.

For the considered wire structure, for higher frequencies the shield trace in fact can increase the crosstalk instead of decrease it (see Figure14). For frequencies higher than 350MHz additional and more complex problems are created using one end grounded shield due to the fact that the line has become electrically long and resonance are observed above 700MHz.

CONCLUSIONS

From the computations, it was found that for low circuit impedance, the primary means of coupling is inductive and inversely, for high circuit impedance the primary means of coupling is capacitive. In other words, if a coupling situation is known to be primarily via the electric field, decreasing the circuit impedance may reduce the amount of interference coupled. Correspondingly, coupling that occurs primarily via the magnetic field (current loops) can be reduced by increasing the circuit impedance.

For the investigated structure and the considered load cases, the computations of the crosstalk showed that shielding trace could be used to decrease the crosstalk between the generator and the receptor. To fully utilize the shielding effectiveness, the shield has to be adequately grounded. If the shield is grounded at least one end, the capacitive coupling contribution is zero, and the inductive coupling is affected by the shield only if the shield is grounded at both ends and the frequency is greater than the break frequency, shield $f > f_{sh}=R_{sH}=2/\pi L_{sh}$.

Adequately grounded shield trace is more effective when the circuit impedance is high (capacitive coupling).

When inductive coupling dominates the shield has to be grounded in two ends. For the investigated wire structure the shield works most effectively if it is grounded in two ends for a frequency range (10-350) MHz where the line is electrically short.

When capacitive coupling dominates the shield trace could be grounded in two ends, too. However, we have seen that the shield inherently eliminates capacitive coupling so long as it is "grounded" at either end. For the considered wire structure, this conclusion is valid for frequencies up to 350MHZ. For frequencies higher than 350MHz additional and more complex problems are created using one end grounded shield.

The above conclusions indicate that a major consideration when performing EMC analysis on a product design is to realize what is the impedance of both source and receptor units, where in the frequency spectrum the problem is observed, how strong is the source energy level and what is the potential harmful interference etc.

From further computations, which are not shown, it was seen that it is useful to place the shield trace close to the generator trace instead of in the middle of the generator and victim trace.

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