Modelling of electromagnetic shielding structures for radiated EMI analysis

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Modelling of Electromagnetic Shielding Structures for Radiated EMI Analysis

WANG ZHENG

SCHOOL OF ELECTRICAL & ELECTRONIC ENGINEERING
NANYANG TECHNOLOGICAL UNIVERSITY

2005

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WANG ZHENG

School of Electrical & Electronic Engineering

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Abstract

Electromagnetic (EM) shielding has been used extensively in electronic products to reduce the radiated electromagnetic interference (EMI) from high-speed digital circuits so that the products could comply with the international EMI regulatory limits.

Good estimation of EM shielding for a wide frequency band is an important area of the product design. Unfortunately, estimation of EM shielding is a complex problem because of many factors, such as distance between the shield and the EMI source, nature of the EMI source, location of the source, cavity resonance, frequency and aperture sizes, etc., could influence the shielding effectiveness of a shield. Due to the complexity, in many cases, it is difficult to estimate EM shielding based on analytical or semi-analytical methods.

In this thesis, an electromagnetic code, based on the Method of Moments (MoM) technique, is developed and validated. Its capability to study EMI issues related to enclosure shielding and heat sink has been demonstrated.

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List of Abbreviations

CISPR	International Special Committee on Radio Interference
CONCEPT II	EM solver
dB	decibel
EFIE	electric field integral equation
EM	electromagnetic
EMC	electromagnetic compatibility
EMI	electromagnetic interference
E _{max}	maximum electric field
FCC	Federal Communications Commission, USA
FDTD	finite difference time domain
FEM	finite element method
I/O	input/output
IC	integrated circuits
MoM	method of moments
MFIE	magnetic field integral equation
PEC	perfect electric conductor
PCB	printed circuit board
RF	radio frequency
SE	shielding effectiveness
TLM	transmission line method
TE	transverse electric
TM	transverse magnetic
XFDTD	EM solver
1D	one dimensional
2D	two-dimensional
3D	three-dimensional

List of Symbols

$\mathbf{A}_c$	magnetic vector potential generated by the electric surface current density on the conducting surface
D_n	normal electric flux density
D_t	tangential electric flux density
$\mathbf{E}$	electric field
$\mathbf{E}^i$	incident electric field
$\mathbf{E}^s$	scattered electric field
$\mathbf{E}_c^s$	electric field scattered by the conducting surface
E_t	tangential electric field at the conductor/dielectric boundary
f	frequency
g	the known excitation
f	the unknown function to be determined.
f_N^a	the N-term approximation f
f_n	a set of linearly independent basis functions
α_n	the unknown complex coefficients of the basis functions
h	the testing function
u	either a single or multi-dimensional variable
$G(\mathbf{r};\mathbf{r}')$	free space Green's function
I_α	unknown complex coefficients of the 2D rooftop current basis functions on the conducting surface
$\mathbf{J}$	electric current density
J_α^x	current density coefficients associated with the x-directed basis functions
J_β^y	current density coefficients associated with the y-directed basis functions
$\mathbf{J}_c$	electric conduction current density
k_0	free space wave number
L	linear operator

$\mathbf{r}$	position vectors of the observation point
$\mathbf{r}'$	position vectors of the source points
S_{11}	reflection coefficient
S_c	area occupied by conducting surface
N_x	number of subsections on x-axis
N_y	number of subsections on y-axis
N_z	number of subsections on z-axis
$[V_c^i]$	the generalised voltage matrix that denotes the testing of known incident electric fields on the conductor surface
$[Z_{uu}]$	the generalised impedance matrix that denotes the contributions u-component of the “source” conductor to u-test path of the “test” conductor on the conductor surface by the 2D triangular current basis functions, $u=x, y$ or z .
$[Z_{uv}]$	the generalised impedance matrix that denotes the contributions v-component of the “source” conductor to u-test path of the “test” conductor on the conductor surface by the 2D triangular current basis functions, $u=x, y$ or z .
$[Z_{uc}]$	the generalised impedance matrix that denotes the contributions corner component of the “source” conductor to u-test path of the “test” conductor on the conductor surface by the 2D triangular current basis functions, $u=x, y$ or z .
$[Z_{cc}]$	the generalised impedance matrix that denotes the contributions corner component of the “source” conductor to corner test path of the “test” conductor on the conductor surface by the 2D triangular current basis functions.
Z_L	generalise load impedance
Z_{load}	load impedance

∇	gradient operator
∇'	divergence operator on the source function
ω	angular frequency
ϵ_0	permittivity at free space
ϵ_r	relative permittivity at dielectric
∞	infinity
μ_0	permeability at free space
μ_r	relative permeability at dielectric
ρ	electric conductivity
ρ_s	surface electric charge density
Λ_α^x	x-direction orthogonal triangular basis functions
Λ_β^y	y-direction orthogonal triangular basis functions
Π^x	x-direction pulse function

Chapter 1

Introduction

1.1 Background

Electromagnetic interference and Electromagnetic compatibility (EMI/EMC) first came into people's view in the 1940s and 1950s, problems then were mostly focused on the motor noises that were conducted over power lines and into a sensitive equipment. From then on to 1960s, EMI/EMC was primarily of interest to the military to ensure electronic systems in military services do not interfere with each other. Without proper EMC consideration, accidents such as radar emissions causing inadvertent weapons release, or emission from electronic devices causing navigation systems failure may occur and degrade the defence capability of the weapon systems.

During the 1970s and 1980s, computer industry flourished, interferences from computing devices became a significant problem to broadcast television and radio reception. This made the government decide to regulate the amount of electromagnetic emissions from products in this industry. The Federal Communications Commission (FCC) of USA issued a set of rules to govern the amount of emissions from any type of computing device, and how those emissions were to be measured. Similarly European and other governments began to control

emissions from electronic and computing devices. Nevertheless, during that period, EMI/EMC control was limited to computers, peripherals, and computer communications products.

During the 1990's, the concern over EMI/EMC has been found to broaden dramatically; in fact, many countries have instituted import controls requiring that EMI/EMC regulations be met before products can be imported into the country. The overall compatibility of all devices and equipment must coexist harmoniously in the overall electromagnetic environment. Emissions, susceptibility to emissions from other equipment, susceptibility to electrostatic discharge, all from either radiated or conducted media are controlled. This control now is limited to not only computers, but also any product that may potentially radiate EMI, or that could be susceptible to other emissions, must be carefully tested. Products with no previous need for EMI/EMC control must now comply with the regulations.

EMI can be harmful to the normal operation of a lot of equipment nowadays. The effects of EMI vary from simple annoyance to catastrophe. Some typical examples of the potential effects of EMI are interference to television and radio receptions, loss of data in digital systems, malfunction of medical electronic equipment, malfunction- of automotive microprocessor control systems, malfunction of navigation equipment and inadvertent detonation of explosive devices. Due to ever-increasing speeds of digital circuits, digital circuits on printed circuit boards (PCBs) radiate a significant level of emission that occupies a very wide frequency spectrum. In order to avoid interference from these digital circuits to existing communications services, many countries have imposed mandatory EMI regulations on electronic products. It means that any

electronic device that failed to comply with the EMI regulation is not allowed entry to the respective country. To ensure that radiated emissions from any electronic products are under the control, the international regulatory bodies have enforced mandatory radiated EMI limits for different categories of electronic products [1]. For example, information technology products, such as personal computers, shall meet FCC Part 15 Subpart B Class B limits for USA market and EN55022 Class B limits for European market [2-3]. The two mentioned limits are extracted from the standards and given in Table 1.1 and Table 1.2, respectively.

Table 1-1 FCC part 15 subpart B class B radiated emissions' limits at 3 m

Frequency (MHz)	Test Distance (m)	Field Strength ($\mu\text{V/m}$)
30 – 88	3	100
88 – 216	3	150
216 – 960	3	200
960 – 1000	3	500

Table 1-2 EN55022 class B radiated emissions' limits at 10 m

Frequency (MHz)	Test Distance (m)	Field Strength ($\mu\text{V/m}$)
30 -230	10	31.6
230 -1000	10	70.8

Electromagnetic shielding has been identified as one of the effective methods to control excessive EMI generated by radiation and induction fields from high-speed digital circuits. The shielding of a source of electrical disturbance from a potential receiver is an effective suppression technique in EMC control. A shield is defined as a metallic barrier placed between two regions of space to control the propagation of

electromagnetic fields from one of the regions to the other. Normally, a shield is any enclosure (partial or complete) that is used to lower the electromagnetic field inside (or outside) the space enclosed. The parameter commonly used in shielding designs is called the shielding effectiveness (SE), usually expressed in units of decibel (dB)[4].

Following is a list of some typical shielding structures for EMI control purposes [5]:

- Architectural shielding and shielded rooms;
- Metallic bodies of cars, airplanes, ships, satellites, etc.
- Equipment – cabinets, shelves and product housings;
- Any metallic structures of arbitrary shape, e.g., equipment racks, cable carriers, heat sinks, etc;
- Cable shields;
- Partition shields – over separate elements like IC chips;
- Shielding gaskets – designed to reduce the “imperfections” at the interface of different elements of the shielding system.

Normally, electromagnetic shielding can be classified into system level, board level and IC level according to the actual scale of the special EMI/EMC problems. Board level shields have made a tremendous impact on cost saving in manufacturing in today’s electronic devices. However, higher speed and higher frequency devices generate more heat and making thermal management a greater issue, a combination of a “sealed” board level shield with integral thermal management through the use of a heat sink can provide the solution. Some typical examples of shielding structures for controlling EMI from PCB are shown in Figs 1.1 and 1.2.

Fig. 1.1 Shielding cans and covers for IC

Fig. 1.2 Various configurations of board level shields

1.2 Motivation and Objective

The increasing trend of product miniaturization forces many electronic modules to be housed closely together. This has placed a great challenge for packaging engineers to get the whole circuit working without EMI. Hence, good shielding design becomes crucial for situations that involve equipment of extremely high sensitivity, for examples, mobile phone, radio astronomy, missile guidance, and tracking systems for satellites and space vehicle.

Metal shields around PCBs or ICs remain one of the primary techniques used to control radiated EMI and to provide EMI immunity. Good estimation of the

effectiveness of a shielding solution is important to the packaging engineers. The motivation here is to develop a numerical tool that allows reasonably good prediction of shielding effectiveness (SE) of a shielding structure for a wide frequency range. Also, with this tool, the impacts of various factors, such as cavity resonance, aperture position, EMI source position, cable penetration, aperture size, etc., could be analyzed and studied without the actual shield fabrication and measurement. Once the impacts of these factors are better understood, a set of systematic design guidelines will be established for shielding designs to meet the specific shielding needs. Most current research has been focused on one specific design factor mentioned above and has assumed that other design factors do not affect the design factor under consideration. In reality, all the design factors (cavity resonance, aperture shape, aperture size, cable penetration, etc.) do influence each other to some extent. Hence, a flexible design tool that could provide insight on how all these factors affect each other is important to the designers. The objective here is to develop such a tool.

1.3 Literature Review

A literature survey on the shielding-related research has been carried out here. Most of the early research work mainly focuses on the mechanism of electromagnetic shielding, derived from either theory or experiments.

From the perspective of the methodology, the research on shielding could be summarized in the following methods: numerical method [6-20], analytical method [21-36], approximation method [37-41] and experiment method [42-46].

Among all of these methods, experiment method is the most direct and easiest to understand. It usually requires a proper measuring environment, such as a screened room or an anechoic chamber and the necessary EM field measuring equipments. Of course, this method is expensive due to the heavy investment in the measurement setup. Also, accurate measurements are difficult to achieve due to the measurement setup limitations, for example, cavity resonance in screened rooms, imperfection of absorbers of anechoic chambers and perturbations caused by the measurement antennas or probes. Most of all, the investment of expensive experiment setup will increase the product's development cost.

Approximation or semi-empirical approximation method provides quick solutions that are easily interpreted, but accurate quantitative results may be difficult to obtain. It can only be used under certain specific conditions related to shield geometry and the frequency. In the low-frequency range, a quasi-static approach allows an accurate analytic description of the field penetration inside the shielded region. When apertures are electrically small, they can be replaced by in terms of equivalent electric and magnetic dipoles on an infinite surface and EM fields external to shield can be estimated.

Analytical methods provide an efficient means of calculating shielding effectiveness, enabling the effect of certain design parameters to be investigated. More recent efforts include an analytical formulation of the shielding effectiveness of an empty enclosure with apertures by using an equivalent circuit for the shorted waveguide and aperture impedance, but only up to the fundamental cavity mode resonance. Normally, a

power-balanced method, Bethe's small-hole theory will be used. The theory of modal expansions developed by Mendez in the seventies' is also a popular theory. It is often used in an ideal cavity, in the presence of low-frequency current and charge distributions, and is developed using the scalar and dyadic Green's functions. The usage of the analytical methods requires the users to have a strong EM background knowledge and they are also applicable to some very simple shielding structures.

Numerical methods permit treatment of complicated problems not solvable by analytical methods. Although they only provide approximate solutions, the solutions are sufficiently accurate for engineering design purposes. We can almost deal with EM shielding problems with almost different shapes and under different conditions by using the appropriate numerical methods and powerful computer resources. In recent years, the common numerical methods, which have been used by researchers to solve the EM shielding problems includes: Method of Moments (MoM) [6,8,11-12,14-16,18], Finite Element Method (FEM) [10], Finite Difference Time Domain (FDTD) [12], Transmission Line Method (TLM) [20], and some hybrid methods that combine the various numerical methods just mentioned. FEM can provide an effective means of characterizing the internal fields to first order. Both FDTD and TLM methods have proven to be powerful tools for analyzing electromagnetic problems in the time domain. In order to apply the method, a certain number of cells each containing a symmetrical condensed node, discretize the analyzed space. However, to deal with radiated fields external to shielding structure, additional numerical algorithm to handle far-field boundary condition is needed. For solving electromagnetic fields external to shielding structure, the MoM is the most suitable method due to its ease of modelling far-field EM problems.

If fine details of an EM problem are to be modelled, all numerical methods suffer a similar problem, the excessive computational overhead. The FDTD and TLM methods require lengthy computational times to handle late-time oscillations due to the resonant nature of the enclosure. The MoM has a difficult time with narrow slots and the fine discretization requirements that create large matrices to be solved by the computer. Hence, the selection of the numerical method is very much problem-dependent. For open boundary problem, such as estimation of radiated EM field in far-field region, MoM becomes an attractive method. Also, MoM is an ideal choice in the frequency domain calculation since its formulation is frequency domain in nature. Hence, for estimation of radiated EM fields with and without the presence of shields, MoM has become a natural choice.

From the perspective of applications, most of the interest is focus on electric field shielding problems. Only limited research involves the magnetic field shielding such as the problem of TM transmissions by a metallic conducting shield at extremely low frequencies [6]. This is mainly because most of EMI/EMC regulations specify the radiation limits in terms of the electric field at higher frequencies from 30 MHz till several GHz. The classic EM shielding topics often include: the study of shielding effectiveness of empty enclosures with apertures, it involves the number, shape and dimension of the apertures [7-9,17,21-22,37]; the calculation and prediction of resonance frequency [20]; EM coupling to an enclosure via a wire penetration or with attached cable [7,28-30]; EM shielding behaviours of wire-meshed or perforated screens [12-13,38-39]; Shielding effectiveness of bent slot [31]; Metallic planes in the shielded enclosures which produce the multi-section enclosures[8,11,14,22,32], and

some research is carried out on the field which is focused on the EM shielding effectiveness of ICs on PCBs [40,43-44,48].

Most of the research mentioned above has been focused on a shielding design against external plane wave EM fields. Very little research has been carried out to study shielding design with an EMI source inside the shielding structure, for example, a noisy PCB housed in a product casing. The research in this thesis will focus on the modelling of shielding structures with an internal source excitation.

1.4 Contributions of the Thesis

The major contributions of this thesis are thought to be:

- Successful development and implementation of a full-wave MoM electromagnetic code, written in C++ programming language, that is capable of modelling any metallic shielding structure made up of rectangular units;
- Comprehensive validation of the MoM electromagnetic code through numerical convergence test and comparison with experimental results and simulation results in published literature;
- Detailed study of the influences of slots, wire penetration through the slot and different source locations to the shielding effectiveness of a rectangular shielded box;

- Detailed study of EMI impact caused by different grounding configurations of a typical heat sink structure.

1.5 Organization of the Thesis

The thesis is organized as follows:

Chapter 1 has introduced the historical development of EMI/EMC and the importance of shielding study in EMI control. A thorough literature review on shielding design has also been carried out. The motivation, objective and the contributions of the thesis are finally discussed.

Chapter 2 begins with a brief introduction of MoM in solving EM problems. With the application of the field equivalence theorem, the conducting surface of the shield is replaced by the two-dimensional (2D) equivalent current density. By applying the necessary boundary conditions, the electric field integral equation (EFIE) is formulated and finally the MoM technique is applied to solve the unknown current densities on the shield numerically.

Chapter 3 focuses on the validation of the MoM electromagnetic code developed by the author. Firstly, a numerical convergence test is carried out to demonstrate the convergence of numerical solution with increasing mesh sizes. Next, the simulation results of some typical shield structures are compared with the experimental results and the simulation results obtained by another established EM solver (CONCEPT II)

in the literature. Good agreement with both experimental results and other well-established simulated results are demonstrated.

Chapter 4 shows the capability of the developed MoM code in studying shielding effectiveness of a rectangular shielded box with different slots and different source locations inside the box. The effect of shielding effectiveness caused by a wire penetration through the aperture on the shielding box is also investigated. Finally, the developed MoM code is employed to study impact on heat sink emission caused by different arrangement of grounding schemes of the heatsink.

Chapter 5 concludes the thesis and discusses possible future work for shielding-related research.

Chapter 2

Numerical Modelling Method

The Method of Moments (MoM) is a general concept that can be applied to any field types. It is based on the formulation of integral equations for any specific problems with well-defined boundary conditions. It has been established to be one of the powerful tools for solving electromagnetic field problems numerically. For the solution of integral equations, the MoM offers a lot of advantages and is frequently used for the analysis of radiation and scattering problems in frequency domain. Harrington first systematically investigated the application of the MoM for solving EM field problems, in his published work in 1968 [49].

2.1 Introduction

The Method of Moments is a general technique, which can apply to fields of any types, not necessarily electromagnetic fields. It reduces a functional equation to a matrix equation, and then solves the matrix equation by known techniques. Consider a linear functional equation as follows:

$$L(f) = g \tag{2.1}$$

where L is a linear operator;
 g is the known excitation;
 f is the unknown function to be determined.

The first step in solving equation (2.1) is to approximate the unknown function f by a linear combination of functions called the basis functions, in the domain of L , given by:

$$f = f_N^a = \sum_{n=1}^N \alpha_n f_n \quad (2.2)$$

where f_N^a is the N-term approximation of f ;
 α_n are the unknown complex coefficients of the basis functions;
 f_n is a set of linearly independent basis functions.

For an exact solution, N should be infinite. Therefore, equation (2.2) is an approximation for the unknown function f . Substituting equation (2.2) into equation (2.1), and making use of the linear property of the operator L , equation (2.1) can be approximated as follows:

$$\sum_{n=1}^N \alpha_n L(f_n) \approx g \quad (2.3)$$

Equation (2.3) is, strictly speaking, an ill-defined statement. The approximately equal sign is the problem, but is necessary because the left and right sides cannot be exactly equal. Now a set of weighting functions, or testing functions is defined in the domain of L , and then taking a suitable inner product on equation (2.3) with each of the

testing functions, resulting in N precisely defined linear equations. Finally, the N unknown complex coefficients α_n are solved by using matrix techniques.

An inner product $\langle f, h \rangle$ between two functions, f and h is defined as

$$\langle f, h \rangle = \int_D f^*(u) h(u) du \quad (2.4)$$

where D is the domain of integration;

h is the testing function;

u is either a single or multi-dimensional variable;

superscript $*$ denotes the complex conjugate.

If f is real, equation (2.4) becomes

$$\langle f, h \rangle = \int_D f(u) h(u) du \quad (2.5)$$

Now, define a set of N testing functions $w_1, w_2, w_3, \dots, w_N$ in the spatial domain of L , and take the inner product of equation (2.3) with each of the testing functions. The result is

$$\sum_{n=1}^N \alpha_n \langle w_m, L(f_n) \rangle = \langle w_m, g \rangle \quad (2.6)$$

$m=1, 2, 3, 4, \dots, N$

The equation (2.3) is now transformed into a set of N equations. This set of equations can be written in the matrix format as

$$[A_{mn}] [\alpha_n] = [g_m] \quad (2.7)$$

where

$$[A_{mn}] = \begin{bmatrix} \langle w_1, L(f_1) \rangle & \langle w_1, L(f_2) \rangle & \cdots & \langle w_1, L(f_N) \rangle \\ \langle w_2, L(f_1) \rangle & \langle w_2, L(f_2) \rangle & \cdots & \langle w_2, L(f_N) \rangle \\ \cdots & \cdots & \cdots & \cdots \\ \cdots & \cdots & \cdots & \cdots \\ \langle w_N, L(f_1) \rangle & \langle w_N, L(f_2) \rangle & \cdots & \langle w_N, L(f_N) \rangle \end{bmatrix} \quad (2.8)$$

$$[\alpha_n] = \begin{bmatrix} \alpha_1 \\ \alpha_2 \\ \cdot \\ \cdot \\ \cdot \\ \alpha_N \end{bmatrix} \quad [g_m] = \begin{bmatrix} \langle w_1, g \rangle \\ \langle w_2, g \rangle \\ \cdot \\ \cdot \\ \cdot \\ \langle w_N, g \rangle \end{bmatrix} \quad (2.9)$$

The unknown complex coefficients of the basis functions are determined by:

$$[\alpha_n] = [A_{mn}]^{-1} [g_m] \quad (2.10)$$

Where $[A_{mn}]^{-1}$ is the inverse matrix of $[A_{mn}]$.

The discretization of a continuous equation by the MoM involves the projection of the continuous linear operator onto finite-dimensional subspaces defined by the basis and

testing functions. The choice of basis and testing functions is the principal issue arising within a MoM implementation. Practical factors affecting the selection of basis functions include the desired accuracy of the approximate solution, the relative complexity of the resulting matrix entries, and computational constraints that place an upper limit on the matrix size. While one would expect that the desired goal would always be to obtain the best accuracy with the fewest basis functions, the need to apply the approach to a wide variety of different problems may motivate a less than optimal formulation. The basis and testing functions should be linearly independent and able to accurately approximate f and $L(f)$, respectively. It cannot be overemphasized that for good results the basis and testing functions must be chosen with the particular operator L in mind. Although the domain of L may be restricted to functions that satisfy certain differentiability requirements, in many cases the differentiability, or “smoothness”, requirements of the basis functions can be shifted to the testing functions. The simplest basis and test functions are the pulse and impulse functions, respectively. However, the pulse basis function with impulse testing function requires much finer grid to provide reasonably good accuracy, which usually places a huge burden on the computational effort. On the other hand, the entire domain basis and test functions produce the least number of unknowns to be solved by the computer. However, the mathematical implementation can be rather complex. In this thesis, a compromise between complexity and computational effort is reached by using the triangular function as the basis function and the pulse function as the test function.

2.2 The Formulation of EFIE

In some cases, the magnetic field integral equation (MFIE) is used to analyze the electromagnetic shielding issues, but they are mostly focused on the scattered fields. This method, however, is only applicable when the domain of interest is completely closed. For practical shielding structures, they are seldom completely closed. Hence the electric field integral equation (EFIE) is chosen to solve the problems.

For typical shielding structures in PCB and IC applications, the conductor thickness is usually much smaller compared to the wavelength of the frequency concerned. Hence, the conductor could be approximated by 2D conductor with zero thickness. For any 2D conductor carrying time-harmonic current, it radiates time-harmonic electric field as follows:

$$\mathbf{E}(\mathbf{r}) = -j\omega\mathbf{A}(\mathbf{r}) - \nabla\Phi(\mathbf{r}) \quad (2.11)$$

The magnetic vector potential $\mathbf{A}(\mathbf{r})$ is defined as:

$$\mathbf{A}(\mathbf{r}) = \mu_0 \iint_{S_c} \mathbf{J}_c(\mathbf{r}') G(\mathbf{r}; \mathbf{r}') ds' \quad (2.12)$$

The electric scalar potential $\Phi(\mathbf{r})$ is defined as:

$$\Phi(\mathbf{r}) = \frac{1}{\epsilon_0} \iint_{S_c} \rho_c(\mathbf{r}') G(\mathbf{r}; \mathbf{r}') ds' \quad (2.13)$$

where $J_c(\mathbf{r}')$ is the surface current density on the conductor in A/m;

$\rho_c(\mathbf{r}')$ is the surface charge density on the conductor in C/m²;

S_c is the surface area of the conductor in m²;

$\mu_0 = 4\pi \times 10^{-7}$ H/m is the free-space permeability;

$\epsilon_0 = 8.854 \times 10^{-12}$ F/m is the free-space permittivity.

$G(\mathbf{r}; \mathbf{r}')$ is the free-space Green's function and is defined as:

$$G(\mathbf{r}; \mathbf{r}') = \frac{e^{-jk_0|\mathbf{r}-\mathbf{r}'|}}{4\pi|\mathbf{r}-\mathbf{r}'|} \quad (2.14)$$

Where $\mathbf{r}$ is the position vector of the observation point

$\mathbf{r}'$ is the position vector of the source point

$k_0 = \omega\sqrt{\mu_0\epsilon_0}$ is the free-space wave number

Substituting equations (2.12) and (2.13) into equation (2.11) leads to:

$$\mathbf{E}(\mathbf{r}) = -j\omega\mu_0 \iint_{S_c} \mathbf{J}_c(\mathbf{r}')G(\mathbf{r}; \mathbf{r}')ds' - \frac{\nabla}{\epsilon_0} \iint_{S_c} \rho_c(\mathbf{r}')G(\mathbf{r}; \mathbf{r}')ds' \quad (2.15)$$

$J_c(\mathbf{r}')$ and $\rho_c(\mathbf{r}')$ are related by the equation of continuity as follows:

$$\begin{aligned} \nabla \cdot \mathbf{J}_c(\mathbf{r}') &= -j\omega\rho_c(\mathbf{r}') \\ \therefore \rho_c(\mathbf{r}') &= \frac{\nabla \cdot \mathbf{J}_c(\mathbf{r}')}{-j\omega} \end{aligned} \quad (2.16)$$

Substituting equation (2.16) into equation (2.15) leads to:

$$\mathbf{E}(\mathbf{r}) = -j\omega\mu_0 \iint_{S_c} \mathbf{J}_c(\mathbf{r}')G(\mathbf{r}; \mathbf{r}')ds' + \frac{\nabla}{j\omega\epsilon_0} \iint_{S_c} \nabla \cdot \mathbf{J}_c(\mathbf{r}')G(\mathbf{r}; \mathbf{r}')ds' \quad (2.17)$$

Equation (2.17) is called the electric field integral equation (EFIE), which forms the basis of modelling metallic shielding structure for any given geometry.

2.3 Numerical Solution of EFIE

The surface current densities in the EFIE are unknown functions to be determined, and the integro-differential operator in the EFIE is a linear operator, and the incident electric fields are the known excitations. Based on the numerical procedure described in section 2.3, the unknown current densities in the EFIE are solved as follows.

To apply the MoM on the EFIE for solving the unknown currents, the first step is to discretize the conductor. Any arbitrary conductor can be considered as a combination of appropriately interconnected flat conducting patches. Typical shielding structures are usually rectangular in shapes, hence, it is convenient to subdivide a rectangular shielding structure, as shown in Fig. 2.1, into Δx , Δy and Δz sub-sections along the x , y and z Cartesian coordinate axes, respectively. The conducting surface is composed of rectangular flat patches of size $\Delta x \times \Delta y$, $\Delta y \times \Delta z$ or $\Delta z \times \Delta x$, depending on the orientation of the conducting surface. For more complex structures, where their boundaries do not align with the Cartesian coordinates, it would have to be stepwise approximation.

Fig. 2.1 Sub-dividing conductor

Once the conductor has been subdivided into smaller patches, pre-determined basis functions must be selected to represent the current distribution on the conductor. To handle the current continuity between any two adjacent subdivided patches, the 2D triangular functions are chosen as the sub-sectional current density functions for the current distribution on the 2D conductor. To fully describe the current density $J_c(\mathbf{r})$ on the conductor, two orthogonal vector components are needed. For example, using the Cartesian coordinate system, two orthogonal triangular basis functions, $\Lambda_\alpha^x(\mathbf{r})$ and $\Lambda_\beta^y(\mathbf{r})$, are needed to represent the current distribution on the conductor that lies in the x-y plane. The mathematical definitions of $\Lambda_\alpha^x(\mathbf{r})$ and $\Lambda_\beta^y(\mathbf{r})$ are given as follows, where the superscript indicates the direction of the current flow and the subscript represents the respective two patches that the current basis function is situated.

$$\Lambda_{\alpha}^x(\mathbf{r}_{\alpha}') = \begin{cases} 1 - \frac{|x' - x_{\alpha}'|}{\Delta x}, & -\Delta x \leq x' - x_{\alpha}' \leq \Delta x, -\frac{\Delta y}{2} \leq y' - y_{\alpha}' \leq \frac{\Delta y}{2}, z' = z_{\alpha}' \\ 0, & \text{otherwise} \end{cases}$$

$$\Lambda_{\beta}^y(\mathbf{r}_{\beta}') = \begin{cases} 1 - \frac{|y' - y_{\beta}'|}{\Delta y}, & -\Delta y \leq y' - y_{\beta}' \leq \Delta y, -\frac{\Delta x}{2} \leq x' - x_{\beta}' \leq \frac{\Delta x}{2}, z' = z_{\beta}' \\ 0, & \text{otherwise} \end{cases} \quad (2.18)$$

The current density distribution $J_c(\mathbf{r}')$ on the conductor that lies in the x-y plane is given as:

$$\mathbf{J}_C(\mathbf{r}') = \sum_{\alpha=1}^{N_x} J_{\alpha}^x(\mathbf{r}_{\alpha}') \Lambda_{\alpha}^x(\mathbf{r}_{\alpha}') \hat{\mathbf{x}} + \sum_{\beta=1}^{N_y} J_{\beta}^y(\mathbf{r}_{\beta}') \Lambda_{\beta}^y(\mathbf{r}_{\beta}') \hat{\mathbf{y}} \quad (2.19)$$

where N_x and N_y are the total numbers of x-directed and y-directed basis functions on the conductor, respectively; J_{α}^x and J_{β}^y are the current density coefficients associated with the x-directed and y-directed basis functions, respectively.

Figs. 2.2 and 2.3 show the x-directed and y-directed current density basis functions on the conductor, respectively.

Fig. 2.2 X-directed 2D triangular functions lie on x-y plane

Fig. 2.3 Y-directed 2D triangular functions lie on x-y plane

To ensure continuous current flow between two joining conducting surfaces of different orientations, 2D corner triangular basis functions are needed around the right-angle bends by combining two 2D half-triangular basis functions together. Each

2D corner triangular basis function consists of a rising 2D half-triangular basis function on one conducting surface and a falling 2D half-triangular basis function on another conducting surface of different orientation. The 2D half-triangular basis functions share the same unknown coefficient. Fig. 2.4 shows an example of 2D half-triangular basis function.

Fig. 2.4 A 2D corner triangular basis function at the bend that joins the x-y and y-z conducting surfaces

Assuming that the conductor for the shield is a perfect electric conductor (PEC), the tangential electric field on the surface of the PEC is zero. It means that the resultant tangential electric field, which is equal to the tangential incident electric field plus the tangential scattered electric field due to the conductor, is equal to zero. This boundary condition leads to equation (2.20).

$$E_x^i(r'_a)_{\tan} = -E_x^s(r'_a)_{\tan} \approx -L[J(r'_a)]_{\tan} \quad (2.20)$$

where

$E_x^i(r'_a)$ is tangential electric field component of the incident electric field;

$E_x^s(r'_a)$ is tangential electric field component of the scattered electric field;

$$L[J(r'_a)] = -j\omega\mu_0 \iint_{\Omega} J_c(r'_a)G(r;r')ds' + \frac{\nabla}{j\omega\epsilon_0} \iint_{\Omega} \nabla' \cdot J_c(r'_a)G(r;r')ds'$$

To have a set of well-defined linear equations, the next step of the MoM is to select a suitable testing procedure. Here “testing procedure” refers to a numerical technique whose solutions satisfy the boundary conditions (e.g. zero tangential electric field on the conducting surface) at selected spatial domains of a conductor. The point matching method, which uses Dirac delta functions as the test functions, is the most simple test procedure as solutions are satisfied at discrete points, but convergence is slow. Galerkin method, which uses the same basis functions as the test functions, satisfied the solutions on spaces occupied by the basis functions, gives fast convergence but at the expense of intensive computational effort involved in the numerical integration. As a compromise between computational efficiency and convergence rate, 1D pulse functions are selected as the test functions for the electric field. The testing procedure satisfies the solutions at discrete line paths all over the conductor. The general testing path is shown in Fig. 2.5.

The x-directed pulse function $\Pi^x(r_m)$ is as follows

$$\Pi^x(r_m) = \begin{cases} 1, & -\frac{\Delta x}{2} \leq x - x_m \leq \frac{\Delta x}{2}, y = y_m, z = z_m \\ 0, & otherwise \end{cases} \quad (2.21)$$

Where $\mathbf{r}_m = (x_m, y_m, z_m)$ is the reference point of the x-directed 1D pulse function.

The x-component of tangential electric field on the x-y plane of the conductor is tested by taking the inner product between $\Pi^x(\mathbf{r}_m)$ and equation (2.20). Substituting the x-directed component of equation (2.19) into equation (2.20), lead to

$$\langle \Pi^x(\mathbf{r}_m), E_x^i(\mathbf{r}_a')_{\tan} \rangle = - \sum_{\alpha=1}^{N_\alpha} J_\alpha^x(\mathbf{r}_a') \langle \Pi^x(\mathbf{r}_m), L[\Lambda_\alpha^x(\mathbf{r}_a')]_{\tan} \rangle \quad (2.22)$$

The physically meaning of equation (2.22) is that the solutions satisfy the boundary conditions on the conducting surface, along the x-directed unit test paths all over the conductor. Equation (2.22) can be rewritten as

$$V_c^i = \sum_{\alpha=1}^{N_\alpha} I_\alpha Z_{cc}(\mathbf{r}_m; \mathbf{r}_a') \quad (2.23)$$

$$m=1,2,3,\dots, N_\alpha$$

where

$$\begin{aligned} V_c^i &= \langle \Pi^x(\mathbf{r}_m), E_x^i(\mathbf{r}_a')_{\tan} \rangle ; \\ I_\alpha &= \sum_{\alpha=1}^{N_\alpha} J_\alpha^x(\mathbf{r}_a') ; \\ Z_{cc}(\mathbf{r}_m; \mathbf{r}_a') &= - \langle \Pi^x(\mathbf{r}_m), L[\Lambda_\alpha^x(\mathbf{r}_a')]_{\tan} \rangle \end{aligned}$$

Equation (2.23) is the result of testing the x-component electric fields on the x-y conducting surface. To form the complete set of equations, similar testing procedures are carried out on the x and y components, y and z components and z and x components of the tangential electric fields on conducting surfaces lie in the x-y, y-z and z-x planes, respectively.

Finally, a full set of linear equations may be expressed in the matrix format given as

$$[V] = [Z][I] \quad (2.24)$$

Fig. 2.5 Test paths of the electric field on the conducting surface

2.4 Creation of Geometry

According to theory of the MoM described above, discretization of the shield conductor is the first step and also the essential pre-process part of this EM code. The dimensions of the shielding structure and the dimensions of the subsections must be given as the input quantities to the simulation code. Normally, to achieve reasonable accuracy, the dimension of each subsection is no more than one tenth of the

wavelength (in the air) of the highest frequency to be simulated. Basically, a shielding structure composed of several 2D planes, by giving both the coordinates of the points which are at the two end points of the diagonal line of every plane and the length, width and height of the shielding structure, the simulator will discretize the conducting surfaces automatically. In this thesis, a flexible mesh generator is developed to pre-process the discretization of the shielding structures. By using this mesh generator and according to the actual requirement of the simulations, the shielding structures to be simulated can be divided into uniform space and non-uniform space subsections whose basic unit is a rectangular patch with various size, so that some tiny and complicated structures can be modelled without excessive computational effort so that normal PCs could handle shielding structures of moderate size. Also, the finer mesh near some tiny or sensitive parts of the whole shielding structure provides better accuracy for the simulation results.

2.5 Evaluation and Organization of Impedance Matrix $[Z]$

Accurate evaluation of the $[Z]$ matrix is important as it decides the accuracy of the solution. As described earlier, the physical meaning of the matrix elements of $[Z]$ is the contributions to the testing of the electric fields on the conducting surfaces due to the current basis functions. The electric field produced by any current density is the sum of the magnetic vector potential term ($j\omega\mathbf{A}$) and electric scalar potential term ($\nabla\Phi$). Due to the divergence operation of the current density in the continuity equation, each triangular current basis function would be associated with a charge pulse doublet as shown in Fig. 2.6 (Taking x-directed triangular current basis function as the example). Using the x-directed 1D pulse functions as test functions gives raise

to physical unit test paths along x , with intervals $x - \frac{\Delta x'_a}{2}$ and $x + \frac{\Delta x'_a}{2}$, as shown in Fig.

2.6. Hence, the evaluation of the matrix elements of $[Z]$ involves line integrals of $j\omega\mathbf{A}$ and $\nabla\Phi$ along the unit test paths.

Line integral of $j\omega\mathbf{A}$ is determined by a numerical integration method. The line path between $x - \frac{\Delta x'_a}{2}$ and $x + \frac{\Delta x'_a}{2}$ is divided into smaller equal-space intervals, and the method of mid-point integration [50] is applied. The advantage of this integration scheme is the avoidance of evaluating the integral at the endpoints of the line paths to exclude singularity at the endpoint. But this integration method often converges very slowly. To improve this situation, Richardson's extrapolation technique [49,50] is applied, which produces a higher order estimate with fewer intervals.

Because of the gradient operator, line integral of $\nabla\Phi$ results in the difference of electric scalar potentials at the endpoints of each unit test path, which is equal to $\Phi(x + \Delta x'_a / 2) - \Phi(x - \Delta x'_a / 2)$. The magnetic vector potential $j\omega\mathbf{A}$ and electric scalar potential $\nabla\Phi$ are integrals of current density and charge density, respectively. Evaluation of $j\omega\mathbf{A}$ and $\nabla\Phi$ requires accurate estimates of the following integrals: surface integral of a 2D triangular current basis function with respect to an observation point; surface integral of a 2D pulse charge function with respect to an observation point.

Fig. 2.6 (a) A x-directed triangular current basis function
(b) An associated charge pulse doublet

All the above integrals involve integration of the free space Green's function with either triangular current basis function or pulse charge function as source function.

The free space Green's function is

$$G(\mathbf{r}; \mathbf{r}') = \frac{e^{-jk_0 R}}{4\pi R}, \text{ Where } R = |\mathbf{r} - \mathbf{r}'| \text{ and } k_0 = \omega \sqrt{\mu_0 \epsilon_0}$$

Due to the highly singular nature of Green's function when R approaches zero, it may lead to significant errors when the observation point is very close to the sub-sectional elements that contain the source functions by using conventional numerical surface integration techniques to calculate the above integrals. To overcome this problem, the Taylor series expansion is applied on $e^{-jk_0 R}$. The Green's function can now be expressed in terms of the series as follows

$$\frac{e^{-jk_0 R}}{4\pi R} = \frac{1}{4\pi} \left(\frac{1}{R} - jk_0 - \frac{k_0^2 R}{2} + j \frac{k_0^3 R^2}{6} + \dots \right) \quad (2.25)$$

For R approaching zero, taking only the first two terms of equation (2.25) is accurate enough to represent the Green's function. The singular term $1/R$ is extracted and integrated analytically, and the integral of the second term may be determined either analytically or numerically.

A shielding structure can be constructed with six planes in Cartesian coordinate. Six directions, as shown in Fig. 2.7, are defined in order to facilitate the coding of MoM simulation.

Fig. 2.7 Definitions of six directions

For the corners formed by two planes of different directions, the following twelve possible current directions for the corner basis functions are given in Fig. 2.8.

Fig. 2.8 Definition of corners

After discretization, based on MoM technique, the matrix elements of $[Z]$ are evaluated and organized as follows.

$$Z = \begin{bmatrix} Z_{xx} & Z_{xy} & Z_{xz} & Z_{xc} \\ Z_{yx} & Z_{yy} & Z_{yz} & Z_{yc} \\ Z_{zx} & Z_{zy} & Z_{zz} & Z_{zc} \\ Z_{cx} & Z_{cy} & Z_{cz} & Z_{cc} \end{bmatrix} \quad (2.26)$$

The descriptions of the sub-matrices are given as follows:

- $[Z_{uu}]$ – the generalised impedance matrix that denotes the contributions u -component of the “source” conductor to u -test path of the “test” conductor on the conductor surface by the 2D triangular current basis functions, $u=x, y$ or z .
- $[Z_{uv}]$ – the generalised impedance matrix that denotes the contributions v -component of the “source” conductor to u -test path of the “test” conductor on the conductor surface by the 2D triangular current basis functions, $u=x, y$ or z .
- $[Z_{uc}]$ – the generalised impedance matrix that denotes the contributions corner component of the “source” conductor to u -test path of the “test” conductor on the conductor surface by the 2D triangular current basis functions, $u=x, y$ or z .

- $[Z_{cc}]$ – the generalised impedance matrix that denotes the contributions corner component of the “source” conductor to corner test path of the “test” conductor on the conductor surface by the 2D triangular current basis functions.

2.6 Current Distribution

After the matrix $[Z]$ is accurately evaluated, the unknown complex coefficients of the current basis functions can be determined by equation (2.27). Once the current distribution is found, other quantities such as radiated electric field and shielding effectiveness can be calculated.

$$[I] = [Z]^{-1} [V_c'] \quad (2.27)$$

Equation (2.27) can be used for both radiation and scattering problems. The distinction between a radiator and a scatter is the location of the incident electric field source. If the source is on the structure, the structure is considered a radiator. If the source is away from the structure, the structure is viewed as a scatter. For a conductor excited by a voltage source V_j at the j^{th} patch of the conducting surface, the voltage column matrices $[V_c']$ is given as

$$[V_c^i] = \begin{bmatrix} 0 \\ \cdot \\ \cdot \\ 0 \\ V_j \\ 0 \\ \cdot \\ \cdot \\ 0 \end{bmatrix} = V_j \quad (2.28)$$

The unknown complex current coefficients can be determined by

$$[I] = [Z]^{-1} V_j' = [Y] V_j' \quad (2.29)$$

Hence, the j^{th} column of the impedance matrix gives the current distribution of the conductor for a unit voltage source applied to the j^{th} patch of the conducting surface of the conductor.

2.7 Calculation of Radiated Electric Field

The radiated electric field of the conductor is obtained by treating the conductor as a combination of N current carrying patches. For radiated EMI analysis, since the radiated electric field is of interest and therefore only the $j\omega A$ term will be used in the calculation of the radiated electric field because the contribution from $\nabla\phi$ will be rather small as R is in the far-field zone. Hence, taking x direction as an example, the x-directed magnetic vector potential is given as follows

$$A_x(\mathbf{r}) = \mu_0 \left(\sum_{\alpha=1}^N I_x \iint_{\Delta S_\alpha} \Lambda_\alpha^x(\mathbf{r}') G(\mathbf{r}; \mathbf{r}') ds' \right) \quad (2.30)$$

For a field point sufficiently far away from the current carrying elements, the free space Green's function $G(\mathbf{r}; \mathbf{r}')$ is invariant with respect to the changes of the position over the source elements. Therefore instead of integrating all the 2D triangular current densities over the patches, the far-field approximation is employed to speed up the calculation of radiated electric field. With the far field approximation, the x-directed magnetic vector potential is given as

$$A_x(\mathbf{r}) = 0.5 \mu_0 \left(\sum_{\alpha=1}^N I_x \Delta S_\alpha G(\mathbf{r}; \mathbf{r}') \right) \quad (2.31)$$

Where, ΔS_α are the unit surface patches occupied by the 2D triangular current basis functions on the conducting surfaces.

From equation (2.31), $A_x(\mathbf{r})$, $A_y(\mathbf{r})$, $A_z(\mathbf{r})$ can all be determined. In a conventional spherical coordinate system, the electric far-field components are calculated as follows

$$E_\theta(r) = -j\omega A_\phi(\mathbf{r}) \quad (2.32)$$

$$E_\phi(r) = -j\omega A_\theta(\mathbf{r}) \quad (2.33)$$

where

$$A_{\theta}(\mathbf{r}) = A_x(\mathbf{r})\cos\theta\cos\phi + A_y(\mathbf{r})\cos\theta\sin\phi - A_z(\mathbf{r})\sin\theta \quad (2.34)$$

$$A_{\phi}(\mathbf{r}) = -A_x(\mathbf{r})\sin\phi + A_y(\mathbf{r})\cos\phi \quad (2.35)$$

Chapter 3

Validation of the EM Code

A two-step approach is adopted to validate the MoM electromagnetic code developed based on the EFIE formulation given in Chapter 2. First, a numerical convergence test will be carried out to ensure that the numerical solution converges with finer mesh size. Next, the simulated results will be validated with other simulation results obtained from other EM solvers in published literature. Once the EM code is validated, a comprehensive study of the shielding effectiveness of some typical shielding structures will be carried out in chapter 4.

3.1 Numerical Convergence Test

A simple conducting loop with zero thickness, as shown in Fig. 3.1, will be modelled and simulated to test for the numerical convergence. We keep N_y (number of subsections on y-axis) constant and make N_x (number of subsections on x-axis) and N_z (number of subsections on z-axis) equal. For the numerical convergence test, N_x and N_z are increased gradually from 2 to 16 by doubling (i.e. 2, 4, 8 and 16) with N_y fixed at 1. In the simulation, the loop is excited with a voltage source of 1V across the infinitesimal gap (delta gap) at point “a”. The simulated frequencies are from 50 MHz to 1 GHz with 20 MHz steps. The real and imaginary parts of the input

impedance versus frequency at point “a” are plotted in Fig. 3.2 and Fig. 3.3, respectively. The blown-up versions of Fig. 3.2 and Fig. 3.3 are plot in Fig. 3.4 and Fig. 3.5, respectively. From the results, the half-wavelength resonant frequency moves closer to the theoretical value of 750 MHz with increasing of N_x and N_z . It is a clear indication of numerical convergence.

Fig. 3.1 Square conducting loop along the x-z axis

Fig. 3.2 Real part of the input impedance of the loop shown in Fig. 3.1

Fig. 3.3 Imaginary part of the input impedance of the loop shown in Fig. 3.1

Fig. 3.4 Real part of the input impedance of the loop shown in Fig. 3.1

Fig. 3.5 Imaginary part of the input impedance of the loop shown in Fig. 3.1

Since typical shielding structures consist of six conducting surfaces in three different planes (x-y, y-z and z-x planes), it is necessary to ensure that the geometry handling feature of the MoM code functions well in the three different planes. Hence, the same square loop is simulated again by rotating it in two other orientations, as shown in Fig. 3.6 and Fig. 3.7. For the loop shown in Fig. 3.6, N_z is fixed at 1, and N_x and N_y are equal to 16; for the loop shown in Fig. 3.7, N_x is fixed to 1, while N_y and N_z are equal to 16. For both loops, voltage source of 1V across the infinitesimal gap (delta gap) at point “a” is applied. Again, the simulated frequencies are from 50 MHz to 1 GHz with 20 MHz steps. The real and imaginary parts of the input impedance versus frequency for the loop in Fig. 3.6 are plotted in Fig. 3.8 and Fig. 3.9, respectively. The real and imaginary parts of the input impedance versus frequency for the loop in Fig. 3.7 are plotted in Fig. 3.10 and Fig. 3.11 respectively.

As illustrated in Fig. 3.2, Fig. 3.3, Fig. 3.8, Fig. 3.9, Fig. 3.10 and Fig. 3.11, the real and imaginary parts of the input impedance of the three loops in three different orientations are identical. Hence, the geometry handling feature of the MoM code is proven to work properly. With this flexible geometry handling feature, we could model any shielding structures with six different conducting planes of various sizes.

Fig. 3.6 Square conducting loop along the x-y axis

Fig. 3.7 Square conducting loop along the y-z axis

Fig. 3.8 Real part of the input impedance of the loop shown in Fig. 3.6

Fig. 3.9 Imaginary part of the input impedance of the loop shown in Fig. 3.6

Fig. 3.10 Real part of the input impedance of the loop shown in Fig. 3.7

Fig. 3.11 Imaginary part of the input impedance of the loop shown in Fig. 3.7

The next simulated structure is a fully enclosed rectangular box without apertures or slots as shown in Fig. 3.12. The dimensions of the shielded rectangular box are given in Table 3-1. A dipole is placed at the centre of the box as a radiating source. For the simulation, the dipole is driven at its centre point. The simulated frequencies are from 100 MHz to 3 GHz in steps of 50 MHz. The mesh size for all the six surfaces of the shielded box is 1 cm.

Fig. 3.12 A dipole antenna totally enclosed in the shielded box

Table 3-1 Shielded box dimensions

Dimension along x-axis (cm)	Dimension along y-axis (cm)	Dimension along z-axis (cm)	Source size (cm)
12	12	10	1×4 (dipole)

As most international radiated EMI regulatory limits are specified at a test distance of 3 m, the radiated electric field at 3 m from the radiating source will be determined. The radiated electric field is first calculated without the shielded box and then re-calculated again with the shielded box. The difference of the two calculated results is called the shielding effectiveness (SE) of the shielded box.

Cavities of any shape made from conducting materials can act as containers for electromagnetic waves, confining them until their energy is dissipated by currents flowing in the resistive walls of the container. Even if a structure has holes in it, at certain frequencies there will be resonant behaviour that will cause field amplitudes to become very large in some locations within the cavity.

One way to explain the behaviour of a cavity is to imagine a waveguide with the open ends being closed by metal plates. These will act as good reflectors, resulting in the allowed modes within the waveguide being reflected back and forth within the enclosure. Feynman once has given a different picture, which illustrates the relationship between the fields around a resonant circuit and those within a cavity. According to his illustration, a resonant cavity has transformed from the resonant circuit through evolution of R/C circuits. In fact, for a rectangular cavity with dimensions, a , b , d metres containing a lossless dielectric the resonant frequencies for the TE and TM modes are calculated by using equation (3.1) [4].

$$f_{mnl} = \frac{c}{2\pi\sqrt{\mu_r\epsilon_r}} \sqrt{\left(\frac{m\pi}{a}\right)^2 + \left(\frac{n\pi}{b}\right)^2 + \left(\frac{l\pi}{d}\right)^2} \quad (3.1)$$

where f_{mnl} are the cavity resonance frequencies

C is the velocity of light $= 3 \times 10^8$ m/sec

μ_r is the relative permeability of the medium within the enclosure

ϵ_r is the relative permittivity of the medium within the enclosure

m, n, l are the number of variations in the standing wave pattern in the x, y, z directions, respectively.

a, b, d are the lengths of the cavity along the x, y, z directions, respectively.

It is well known that both transverse electric (TE) and transverse magnetic (TM) modes can exist in a rectangular cavity resonator. The resonant frequencies of TM and TE modes in a resonator depend on the direction of propagation, which could be either in the x, y or z-direction. The subscript “mnl” denotes a specific TM or TE standing wave pattern in a cavity. For the given shielded enclosure to be simulated, the dimension along the x-axis is equal to that of the y-axis. All possible resonant modes and the associated resonant frequencies below 3 GHz are listed in Table 3-2 with z-axis as the wave propagation direction. All possible resonant modes and the associated resonant frequencies below 3 GHz are listed in Table 3-3 with either x or y axes as the direction of wave propagation [51].

For the fully enclosed shielded box with no aperture, to ensure that all possible cavity resonant modes are detected, simulations will be carried out with the radiating dipole positioned along the x, y and z-axis. The simulated SE in dB for the dipole in x, y and z axes are shown in Figs 3.14, 3.15 and 3.16, respectively. Figs 3.14 through 3.16 show that all the cavity resonant frequencies listed in Tables 3-2 and 3-3 have been

clearly detected except for 2.32 GHz. To detect this cavity resonant frequency, two dipoles with the same dimension as the one mentioned above are placed simultaneously along x and z-axes, as shown in Fig. 3.13, Fig. 3.17 shows that all the possible cavity resonant frequencies can now be detected with this arrangement of source excitation.

Table 3-2 Possible modes and resonant frequencies with z-axis as the propagation direction

Mode	Frequency (GHz)
TM_{110}^z	1.77
TM_{111}^z	2.32
TM_{120}^z	2.8
TM_{210}^z	2.8
TE_{101}^z	1.95
TE_{011}^z	1.95
TE_{111}^z	2.32
TE_{201}^z	2.92
TE_{021}^z	2.92

Table 3-3 Possible modes and resonant frequencies with x (y) axis as the propagation direction

Mode	Frequency (GHz)
$TM^{x(y)}_{110}$	1.95
$TM^{x(y)}_{111}$	2.32
$TM^{x(y)}_{210}$	2.92
$TE^{x(y)}_{101}$	1.77
$TE^{x(y)}_{011}$	1.95
$TE^{x(y)}_{012}$	2.92
$TE^{x(y)}_{111}$	2.32
$TE^{x(y)}_{201}$	2.8

Fig. 3.13 Two dipoles along x and z axes simultaneously

Fig. 3.14 SE of the shielded box with the radiating dipole placed along x axis

Fig. 3.15 SE of the shielded box with the radiating dipole placed along y-axis

Fig. 3.16 SE of the shielded box with the radiating dipole placed along z-axis

Fig. 3.17 SE of the shielded structure with two radiating dipoles along x and z axes simultaneously

3.2 Comparison with Experiment Results and Other Softwares

In the following part of validation, a cubic shielding box of dimensions $10\text{ cm} \times 10\text{ cm} \times 10\text{ cm}$ with a $2\text{ cm} \times 8\text{ cm}$ slot in the front plane will be simulated. In the simulation, a one-volt voltage source will be applied across the middle of the slot, as shown in Fig. 3.18. The S_{11} parameter seen by the voltage source from 100 MHz to 3 GHz will be calculated. Fig. 3.19 shows comparison of the simulated S_{11} from the developed MoM code and XFDTD, which is based on the FDTD numerical method. In both of these two simulations, the grid sizes are the same 1 cm.

According to equation (3.1), for the given cubic shielded box, the possible cavity resonant frequencies below 3 GHz are 2.1 GHz and 2.6 GHz. Both cavity resonant frequencies are clearly observed in both the MoM and XFDTD simulated results. The theoretical slot resonant frequency is 1.9 GHz, which is very close to the cavity resonant frequency 2.1 GHz, so it is not visible in the plot due to the dominant nature of the cavity resonance.

Through the comparison, the results obtained by the MoM code developed in the thesis shows reasonable agreement with results obtained by the XFDTD. Two resonant frequencies are observed, the first resonant frequency of both simulations is closely matched; the second resonant frequency is slightly differed by about 150 MHz.

Fig. 3.18 10 cm³ cubic box with a 2 cm × 8 cm slot at centre of the front plane

Fig. 3.19 Comparison of S_{11} simulated by MoM and XFDTD

Another rectangular shielded box with a slot shown in Fig. 3.20 will be used as another example for validation purposes. The SE simulated using the developed MoM code will be compared with published measured and simulated results in the literature. The published simulated SE was obtained by another commercial EM simulation software, called CONCEPT II, which is also a full-wave frequency-domain EM code developed by the Technical University of Hamburg-Harburg, Germany. It is also based on MoM but using different types of basis functions.

Fig. 3.20 Geometry of the enclosure used in the simulations

Fig. 3.21 Comparison of SE between the developed Mom code and CONCEPT II [52] – a vertical slot (5 cm $\times$ 20 cm) at centre of front panel and electric dipole antenna inside

For the given shielded box, the length, width and height (a, b and d) all have the same dimension, which is 50 cm. For the first simulation, a vertical slot (w=20 cm and l=5 cm) is cut at the centre of the front panel. A z-directed dipole is placed within the centre of the shielded box inside. The radiated electric field at 3 metres away from the dipole without and with the shielded box are calculated from 100 MHz to 1 GHz. The difference of the two fields is the SE of this shielded box with the slot. The comparison of the simulated result using CONCEPT II is given in Fig. 3.21. According to equation (3.1), the possible cavity resonance frequencies of this cubical

box up to 1 GHz are 424.3, 519.6, 670.8, 734.8, 848.5, 900.0, 948.7 and 995.0 MHz. From Fig. 3.21, it can be observed that the results obtained by the developed MoM are in close agreement with the results simulated using CONCEPT II, both simulation results detect the cavity resonance frequencies near 424.3, 734.8 and 948.7 MHz. The possible slot resonance frequency are 375 MHz and 750 MHz, since they are close to the cavity resonance frequency 424.3 MHz and 734.8 MHz, to most degree, they will be overshadowed by the cavity resonance frequency, so it is reasonable that our simulation results do not show these resonance frequencies clearly.

For the next simulation, the cubical shielded box with the same dimension mentioned above is used again. Instead of a vertical slot, a horizontal slot ($w=5$ cm, $l=20$ cm) is now located at centre of the front panel. A square loop antenna whose four sides are all 4 cm is placed at the centre of the shielded box. The SE is estimated from 100 MHz to 500 MHz. For the measurement setup in [8], in order to minimize parasitic effects of the surrounding due to imperfections in the anechoic rooms and in order to load the internal antenna as much as possible, the enclosure without front panel instead of the total absence of the enclosure was used as the reference to evaluate the shielding effectiveness. So in our simulation, we follow the same instruction like those in the literature. Fig. 3.22 shows the comparison of simulated result obtained by the MoM code with the result obtained by CONCEPT II as well as the measured result published in [8]. In the simulations, the mesh size near the slot and the mesh size for the antenna sources is less than $1/20$ of the wavelength of the maximum simulated frequency. The different mesh sizes are needed in order to speed up the computational time and still obtaining acceptable accuracy. Fig. 3.22 shows that all of the three results detect one cavity resonance frequency below 500 MHz. In this

simulation, the possible slot resonant frequencies are all beyond 1 GHz, so they cannot be observed in all of our results since the frequency range of interest in the simulations are below 1 GHz here. Also, the simulated MoM results show better agreement with the measured result, as compared to the CONCEPT II result. The general trends of the three results match quite well until 450 MHz, where the CONCEPT II result show a resonant frequency at around 500 MHz. Both the simulated MoM result and the measured result do not have resonant at that frequency.

Fig. 3.22 Comparison of SE between the developed MoM code and published data in [8] – a horizontal slot (20 cm × 5 cm) at centre of the front panel

In the third simulation, the same shielded box with the same loop antenna mentioned in the second simulation will be used again but this time a thinner slot ($w=2$ mm, $l=400$ mm) is located at the centre of the front panel. The simulated SE is calculated from 100 MHz to 1 GHz. The same problem has also been studied in [8] through

measurement and simulation by CONCEPT II. In this case, the reference configuration for the calculation of the shielding effectiveness was the configuration with no enclosure at all. Fig. 3.23 shows the comparison of the two simulated results and measured results. It is clearly observed that the MoM code developed manages to detect the theoretical cavity resonance frequencies more accurately than CONCEPT II. The cavity resonance frequencies near 424.3, 519.6, 670.8, 734.8, 900.0 and 948.7 MHz can be clearly seen. Also, at these resonant frequencies, our simulated results show closer agreements with the measured results than those from CONCEPT II, especially at resonant frequencies near 424.3, 519.6 and 734.8 MHz. Here the possible slot resonance frequency is 375 MHz, which is close to the cavity resonance frequency 424.3 MHz, the cavity resonance frequency is the dominant one.

Fig. 3.23 Comparison of SE between the developed MoM code and published data in [8] – a horizontal thin slot (400 mm × 2 mm) at centre of the front panel

There are still some difference between the numerical results and the measurement results, especially between the frequency range 420MHz-700MHz, according to the published paper's author's opinion, the discrepancy at low frequency is due to the fact that the test are close to the noise level; the source of the difference at high frequency is partly due to the fact that a ferrite tiled chamber is less accurate at higher frequency. Actually this is partly because in the numerical method, it is very difficult to simulate the real experiment setup 100%, there are always some details can not be simulated by using the current code, it makes the numerical experimental environmental much ideal compared with real setup, but we can still find the mainly trend of the SE of the same issue, the better agreement with the other code-CONCEPT II can also illustrate this point.

Chapter 4

Modelling of Electromagnetic Radiation Structures

In this chapter, two kinds of typical electromagnetic radiation structures – shielded enclosure and heat sink will be introduced, some relative items about the EMI issues on these two structures will also be investigated comprehensively.

4.1 Enclosure Shielding Study

Due to stringent electromagnetic compatibility (EMC) regulatory requirements in Europe and USA, designing a product to meet these requirements becomes very important. The inherent shielding properties offered by the metallic enclosures for packaging purposes can be used to suppress radiated EMI emitted by the electronics housed within the enclosures. In order to achieve optimum shielding performance for a given piece of equipment and its metallic enclosure, the location of noisy circuits with respect to the shielding apertures, the mechanical designs such as size, location and amount of the required apertures of the enclosure must all be properly analyzed and studied [54]. The actual enclosures always need holes for ventilation, input and output connections, control panels or displays. The design of these openings and holes to achieve good shielding effectiveness can be a tough task because of the complicated electromagnetic couplings from the radiating sources to these apertures.

This chapter demonstrates the application aspect of the developed MoM EM code in shielding design study of typical electronics enclosures.

4.1.1 Shielding Structure with Different Apertures

In the simulations that follow, the rectangular metallic enclosure is fully sealed with six conducting surfaces. There is only a slot (or aperture) in one of the surfaces. The size of the aperture is much smaller than the wavelength of the field. The electromagnetic fields 3 metres away from a known source (electric or magnetic dipole) will be calculated first and then the field is recalculated with source placed within the enclosure. The difference between the two calculated fields, expressed in decibel (dB), will be the shielding effectiveness (SE) of the given enclosure.

Fig. 4.1 shows the coordinate system that will be adopted for all the simulations to be carried out in this chapter.

Fig. 4.1 Geometry of the shielded enclosure used in the simulations

The cubic metallic box to be modelled (Fig. 4.1) whose dimensions are: 500 mm $\times$ 500 mm $\times$ 500 mm. This box consists of six conducting surfaces, which are assumed to be perfect electric conductor (PEC). One of the surfaces is a reconfigurable front panel. The source here will be an electric dipole or a magnetic loop antenna at the centre of the box. According to equation (3.1), the cavity resonance frequencies below 1 GHz of this metallic box are calculated as: 424 MHz, 519 MHz, 670 MHz, 734 MHz, 848 MHz, 900 MHz, 948 MHz and 995 MHz.

4.1.1.a Effect of Vertical Slot and Horizontal Slot

The impacts on SE caused by a horizontal slot and a vertical slot (electric dipole along z -axis, located at the centre of the box), as shown in Fig. 4.2, are investigated. The

simulated frequency range is from 100 MHz to 1 GHz with steps of 20 MHz. The simulated results are shown in Fig. 4.3.

From Fig. 4.3, it can be seen that the orientation of the slot with respect to the radiating source has a great influence in the shielding performance, even the slot dimension is identical. For an electric dipole along z-axis, a horizontal slot causes discontinuity of the electric field and it leads to field leaks through the slot. For the case of a vertical slot, the discontinuity of the electric field is less and it results in better shielding performance. Hence, the source polarization has a strong influence in shielding performance. The shielding performance is the worst when the radiated electric field coincides with a slot whose longer dimension is perpendicular to the propagation direction of the electric field. For this case study, the shielding performance can vary as much as 10 dB.

Fig. 4.2 Front panel with either horizontal slot or vertical slot

Fig. 4.3 Comparison of SE of between with vertical slot and horizontal slot at the centre of the front panel

4.1.1.b Effect of Different Width and Length of a Slot

In the next part of study, the influence of different width and length of a slot on SE of a shielded enclosure will be investigated. The shielded box shown in Fig. 4.1 will be used again for the study. A slot is located at centre of the front panel and a z-directed electric dipole is placed at the centre of the box. Firstly, a slot with a fixed 5 cm width, and a variable length is simulated. The SE is investigated from 100 MHz to 1 GHz with steps of 20 MHz. To study the impact of different length of a slot on SE of the shielded box, the SE with slot length of 10 cm, 30 cm and 40 cm are simulated

and plotted in Fig. 4.4. From the simulated results, it is obvious that the SE degrades with increasing length of slot.

The same shielded box is simulated again with a fixed slot length but with variable width. For the simulations, the slot's length is kept constant at 40 cm and the slot's width varies at 5 cm and 10 cm respectively. The SE is illustrated in Fig. 4.5. Compared to the SE shown in Fig. 4.4, it is observed that the influence to SE of slot with different width is not as strong as that of slot with different length. It has confirmed our previous observation that with a z-directed dipole, SE is more sensitive to the length of the slot, which is perpendicular to the propagating direction of the electric field.

Fig. 4.4 SE of the shielded box with fixed slot's width and variable slot's lengths

Fig. 4.5 SE of the shielded box with fixed slot's length and variable slot's width

4.1.1.c Effect of Different Number of Apertures

In Fig. 4.6, three different configurations of the front panel with different number square apertures are considered. Fig. 4.7 shows the influence on SE due to different number equal-sized square apertures on the front panel of the shielded box. When the number of the square apertures increases from one to two, there is about 5 dB decrease in SE, when the number of the square apertures increase from two to four, the SE further reduces to about 5-6 dB again. The results provide us a rough

estimation of shielding degradation for the given shielded box due to increasing number of apertures.

In Fig. 4.8, we consider different configurations of the front panel with different number thin slots. Each configuration consists of one, two or three thin equal-sized slots ($400\text{ mm} \times 5\text{ mm}$). For configurations with more than one slot, the slots are separated by 5 cm and positioned centrally on the front panel. The SE is calculated from 100 MHz to 1 GHz . Fig. 4.9 shows the comparison of SE for these three configurations. It is clearly observed that more slots degrade the SE of the shielded box. The simulation results allow us to estimate the degradation of SE with increasing number of slots.

Fig. 4.6 Geometry of the front panel with one, two and four square apertures

Fig. 4.7 SE of cubic box with different number square apertures on the front panel

Fig. 4.8 Geometry of the front panel with one, two and three thin slots

Fig. 4.9 SE of cubic box with different number thin slots on the front panel

4.1.2 Shielding Structure with Wire Penetration through a Slot

Normally, wires or wire bundles will be placed through a slot or an aperture of a shielded enclosure in order to transfer the signals between the devices inside and outside the enclosure. These wires will degrade the shielding effectiveness of the shielded enclosures. In this section, we demonstrate the flexibility of the MoM code to investigate the possible influence that caused by a wire placed through an aperture of a shielded enclosure. To provide a better understanding of the effects of wire on shielding effectiveness, two different wire configurations will be modelled and simulated. Using the shielded box shown in Fig. 4.1 as an example, where, a , b and d

are all 500 mm, $w=100$ mm and $l=100$ mm. An electric dipole is placed inside the box, simulation of a straight wire and a bent wire are carried out respectively. The straight wire penetrated into the shielded box through the square aperture, whereas the bent wire has an additional vertical arm, which extends to the bottom of the shielded box. For both cases, the wire is extended 300 mm (half a wavelength at 500 MHz) exterior to the cavity. Fig. 4.10 shows the SE of the above two wire configurations.

It is rather obvious that both wire configurations deteriorate the SE across the frequency range of 100 MHz to 3 GHz. The bent wire configuration deteriorates the SE further.

Fig. 4.10 Effect on SE due to two wire configurations with length half wavelength at 0.5 GHz interior and exterior to the cavity shown in Fig. 4.1.

4.1.3 Shielding Structure with Source at Different Locations Inside

Again the box shown in Fig. 4.1 with a horizontal aperture of $5 \text{ mm} \times 20 \text{ mm}$ in the front panel is used for simulation. In this case, a small square loop antenna of side $h=40 \text{ mm}$ will be positioned vertically at different positions in the vertical symmetry plane ($y=250 \text{ mm}$) of the enclosure as shown in Fig. 4.11. The source will be placed at different positions: centre excitation, backward excitation (position 1), excitation shifted backwards and up (position 2) and an excitation shifted up (position 3). The effect of the distance of the source to the opening in the enclosure is given in Fig. 4.12, where the results of the simulations up to 500 MHz with 20 MHz steps are shown.

Fig. 4.12 shows that, the distance between source and the slot does influence the SE of the box, the further the source is from the slot, the better is the SE. In general, the difference in SE is not so obvious for frequency below 250 MHz . When the frequencies go beyond 250 MHz , the difference becomes noticeable.

Fig. 4.11 Different positions of the square loop antenna in the box in the vertical symmetry plane $y=250$ mm

Fig. 4.12 Influence of the source positions shown in Fig. 4.9 with a horizontal slot on front panel and loop antenna inside

4.2 Heat Sink Emission Study

Ventilation is commonly used to remove heat generated by PCB. However, if ventilation cannot provide efficient heat removal, heat sinks will be mounted on PCB to improve the heat removal process. From the EMC design point of view, heat sink can become an unintentional antenna that radiate significant amount of emission. Circuits designers need to know the effective ways to reduce radiated emission from a heat sink whenever possible. Grounding of a heat sink has been used by many designers for EMI suppression purposes. However, how the heat sink should be grounded and how many points to be grounded to achieve lowest possible radiation can be a difficult decision [1,55]. In this section, the developed MoM code is employed to investigate the effectiveness of EMI suppression by a variety of grounding configurations for a standard sized heat sink.

4.2.1 Heat Sink Emission with No Grounding Points

Fig. 4.13 shows the heat sink under investigation. It is a metallic block measuring 88.9 mm (3.5 in.) by 63.5 mm (2.5 in.) by 38.1 mm (1.5 in.). The metallic block is located at 6.0 mm (0.2361 in.) above a conducting ground plane. The size of this grounding plane is: 160 mm $\times$ 120 mm (6.290 in. $\times$ 4.740 in.). The heat sink is excited by a vertical source extending from the ground plane to the base of the heat sink and offset 12.7 mm (0.5 in.) in the x and y axes from the centre of the heat sink in order to excite both the even and odd modes. The excitation voltage source is kept

at 1V across the full frequency range. Fig. 4.14 gives the maximum radiated electric field 3 metres away from the given heat sink structure. In the next few simulations, the heat sink will be grounded to the ground plane at different points through grounding posts, which are 6 mm $\times$ 6 mm (0.2362 in.) square legs. To check the effectiveness of the

Fig. 4.13 Geometry of the heat sink

various configurations of grounding on controlling radiated emissions from the heat sink, the electric field plot in Fig. 4.14 is used as a reference for subsequent comparison.

The EMI suppression effectiveness (SE) here is defined as the difference between the reference radiated electric field (no grounding point) and radiated electric field from grounded heat sink, expressed in decibels.

Fig. 4.14 The reference electric field without grounding points

4.2.2 Heat Sink Emission with Different Grounding Points Located

The grounding points are located at different positions of the heat sink. The different configurations of the grounding points are shown in Fig. 4.15. The effects of the groundings of heat sink are shown in Fig. 4.15.

According to the simulation results shown in Fig. 4.16, it is obvious that suppression effectiveness increase with higher number of grounding points. However, an increased number of grounding points also increase the resonant frequency of the structure. The resonances observed are believed to be caused by the capacitance between the heat sink and the ground reference plane and the inductance contributed by the grounding posts. As more posts are added, the resultant inductance decreases, which results in higher resonant frequencies.

Fig. 4.15 Different configurations of the grounding posts

Fig. 4.16 SE of different grounding points configurations

4.3 Conclusion Remarks

This chapter has demonstrated the application aspects of the electromagnetic code in studying and analyzing two typical electromagnetic radiated structures: shielded enclosure and heat sink.

Firstly, the impacts of the aperture's size, number of apertures, wire penetration and source locations on the shielding effectiveness of a shielded box were investigated. The simulation results allow estimation of SE for a typical shielded enclosure of an electronics product.

With the help of the SE plots, the characteristics of any practical shielded box with ventilation holes and wire penetration can be understood. As the final SE of an enclosure is dependent on a number of influencing factors, such simulation and studies will provide the necessary inputs for the design engineers for EMC compliance purposes.

In the next section, several configurations of grounding scheme of a typical heat sink structure have been investigated. From the simulation results, it is obvious that different grounding configurations have direct impact on the EMI suppression effectiveness of the heat sink. For the given heat sink, more grounding points result in better suppression. However, more grounding points also push the resonant frequencies higher. Therefore, the number of grounding points depends very much on the required EMI suppression for the specific frequency range.

Heat sink structures are widely used in electronic designs for heat dissipation or partial shield purposes. They can also become potential EMI radiator if no attention is paid on the heat sink grounding. How to grounding the heat sink to achieve the required EMI suppression is important to the designers. The MoM code developed in this thesis has proven to be handy for such an investigation.

Chapter 5

Conclusion and Future Work

5.1 Conclusion

Based on the MoM, an electromagnetic simulation code capable of modelling shielding structures was successfully developed and implemented in C++ programming language. A comprehensive validation of the electromagnetic code was also carried out. A numerical convergence test was performed to ensure that the numerical solution converged with finer grid. Very good numerical convergence was observed. Next, the shielding effectiveness of a typical shielding structure was simulated using the developed code and another established code (CONCEPT II) and close agreement between the two codes was demonstrated.

The flexibility of the developed code is fully exploited to study the slot size, number of slot and slot positioning on the shielding performance of a shielded box. This kind of study is important to designers of electronics designers, as shielding is a complex subject that could only be solved by a full-wave solver. Also the code can be extended to investigate the impact of different methods of heat sink grounding on radiated emission from the heat sink. Of course, the code can be extended further to many EMC design study purposes for high-speed electronics products.

5.2 Recommended Future Work

In the existing MoM numerical implementation, the metal shield is assumed to be perfect electric conductor (PEC) with zero thickness. Hence, the loss due to skin effect of finite thickness conductor has not been accounted for. The skin effect of the metal can be modelled as a surface resistance and will be included in the MoM formulation.

As the aperture size reduces, modelling grid size has to be reduced to yield accurate simulation results. This imposes heavy memory storage requirement on the computing resources and slows down the computational efficiency. A more flexible pre-process mesh generation algorithm can be developed to handle both fine and coarse grid sizes for a shielding structure, based not only on current rectangular subsection, but also the subsections of other shape, without significant loss of accuracy. The new algorithm will reduce the storage requirements further and hence, opens the door for more challenging and realistic shielding problems to be modelled and simulated.

A hybrid algorithm that combines the FDTD with the current MoM can be considered later, since FDTD is good at handling with the near field problems, if FDTD can be added to deal with the field inside the shielding structure, viz., near the source, meanwhile the MoM can deal with the far field outside the shielding structures, viz., away from the source, the current EM code will be more flexible and powerful.

The apertures with other shapes except for the regular rectangular shape should be considered later such as the circular apertures, elliptical apertures or diamond

apertures by using more flexible basis functions to represent the expansion of the surface current density together with the more flexible mesh generation method. Meanwhile the effect of the substrate can be added into the current MoM code, so that many actual shielding structures will be modelled accurately since a lot of shielding problems in reality have the influence of various substrates.

After accomplishing the work mentioned above, a comprehensive set of design guidelines that could allow designers to design a shield without going through the trial-and-error process can be built up, which can save the product development time and cost greatly by finding and eliminating potential EMI problems in shielding designs at the early design stage of product packaging.

References

- [1] **B. Archambeault, C. Brench, and O. M. Ramahi**, “EMI/EMC Computational Modeling Handbook”, second edition, Kluwer Academic, 2001.
- [2] “FCC Title 47 part 15, 18 and 68 of the US Code of Federal Regulations”, Federal Communications Commission, Washington DC.
- [3] “Limits and Methods of Measurement of Radio Interference Characteristics of Information Technology Equipment”, CISPR Publication 22, 1985.
- [4] **P. A. Chatteron, and M. A. Houlden**, “EMC: Electromagnetic Theory to Practical Design”, John Wiley & Sons, 1992
- [5] **A. Tsaliovich**, “Electromagnetic shielding handbook for wired and wireless EMC applications”, Kluwer Academic, 1999.
- [6] **E. H. Newman, and M. Kragalott**, “ Moment method analysis of the electric shielding factor of a conducting TM shielding at ELF”, IEEE Trans. Electromagnetic Compatibility, vol. 37, no. 3, Aug. 1995, pp. 400-408.
- [7] **Eng Swee Siah, K. Sertel, J. L. Volakis, V. V. Liepa and R. Wiese**, “Coupling studies and shielding techniques for electromagnetic penetration through apertures on complex cavities and vehicular platforms”, IEEE Trans. Electromagnetic Compatibility, vol. 45, no. 2, May 2003, pp. 245-257.
- [8] **F. Olyslager, E. Laermans, D. De Zutter, S. Criel, R. De Smedt, N. Lietaert, and A. De Clercq**, “Numerical and experimental study of the shielding effectiveness of a metallic enclosure”, IEEE Trans. Electromagnetic Compatibility, vol. 41, no. 3, Aug. 1999, pp. 202-213.

- [9] **Min Li, J. L. Drewniak, S. Radu, J. Nuebel, T. H. Hubing, R. E. DuBroff, and T. P. Van Doren**, “An EMI estimate for shielding-enclosure evaluation”, IEEE Trans. Electromagnetic Compatibility, vol. 43, no. 3, Aug. 2001, pp. 295-304.
- [10] **C. F. Bunting**, “Shielding effectiveness in a two-dimensional reverberation chamber using finite-element techniques”, IEEE Trans. Electromagnetic Compatibility, vol. 45, no. 3, Aug. 2003, pp. 548-552.
- [11] **W. Wallyn, D. De Zutter, and H. Rogier**, “Prediction of the shielding and resonant behavior of multisection enclosures based on magnetic current modelling”, IEEE Trans. Electromagnetic Compatibility, vol. 44, no. 1, Feb. 2002, pp. 130-138.
- [12] **Min Li, J. Nuebel, J. L. Drewniak, R. E. DuBroff, T. H. Hubing, and T. P. Van Doren**, “EMI from airflow aperture arrays in shielding enclosures-experiments, FDTD, and MOM modelling”, IEEE Trans. Electromagnetic Compatibility, vol. 42, no. 3, Aug. 2000, pp. 265-275.
- [13] **R. De Smedt, J. De Moerloose, S. Criel, D. De Zutter, F. Olyslager, E. Laermans, W. Wallyn, and N. Lietaert**, “Approximation simulation of the shielding effectiveness of a rectangular enclosure with a grid wall”, IEEE International Symposium on Electromagnetic Compatibility, vol. 2, Aug.1998, pp. 1030-1034.
- [14] **W. Wallyn, and D. De Zutter**, “Modeling the shielding effectiveness and resonance of metallic shielding enclosures loaded with PCBs”, IEEE International Symposium on Electromagnetic Compatibility, vol. 2, Aug. 2001, pp. 691-696.
- [15] **H. Singer, H.-D. Bruns, and G. Burger**, “State-of-the-art in the method of moments”, IEEE International Symposium on Electromagnetic Compatibility, Aug.1996, pp. 122-127.

- [16] **C. F. Bunting, and Shih-pin Yu**, “Statistical shielding effectiveness-an examination of the field penetration in a rectangular box using modal MOM”, IEEE International Symposium on Electromagnetic Compatibility, vol. 1, Aug. 2002, pp. 210-215.
- [17] **D. Florean**, “Computer aided predictions and experimental validation of radiated emission from a shielding box with different size apertures”, IEEE International Symposium on Electromagnetic Compatibility, vol. 1, Aug. 2001, pp. 195-198.
- [18] **M. Kimmel, and H. Singer**, “Numerical computation of anisotropic shielding materials based on the method of moments”, IEEE International Symposium on Electromagnetic Compatibility, Aug. 1996, pp. 87-91.
- [19] **B. Archambeault, and K. Hatashita**, “Modeling and measurements of EMI shielding performance”, IEEE International Symposium on Electromagnetic Compatibility, Aug. 1995, pp. 347-348.
- [20] **J. Hesselbarth, and R. Vahldieck**, “Resonance frequencies calculated efficiently with the frequency-domain TLM method”, IEEE Microwave and Wireless Components Letters, vol. 13, no. 5, May 2003, pp. 190-192.
- [21] **G. Cerri, R. De Leo, and V. M. Primiani**, “Theoretical and experimental evaluation of the electromagnetic radiation from apertures in shielded enclosures”, IEEE Trans. Electromagnetic Compatibility, vol. 34, no. 4, Nov. 1992, pp. 423-432.
- [22] **M. P. Robinson, T. M. Benson, C. Christopoulos, J. F. Dawson, M. D. Ganley, A. C. Marvin, S. J. Potter, and D. W. P. Thomas**, “Analytical formulation for the shielding effectiveness of enclosures with apertures”, IEEE Trans. Electromagnetic Compatibility, vol. 40, no. 3, Aug. 1998, pp. 240-248.

- [23] **H. A. Mendez**, “On the theory of low-frequency excitation of cavity resonators”, IEEE Trans. Microwave Theory and Techniques, vol. 18, no. 8, Aug. 1970, pp. 444-448.
- [24] **F. De Medulenaere, and J. Van Bladel**, “Polarizability of some apertures”, IEEE Trans. Antennas and Propagation, vol. 25, no. 2, Mar. 1977, pp. 198-204.
- [25] **N. A. McDonald**, “Polynomial approximations for the electric polarizabilities of some small apertures”, IEEE Trans. Microwave Theory and Techniques, vol. 33, no. 11, Nov. 1985, pp. 1146-1149.
- [26] **N. A. McDonald**, “Polynomial approximations for the transverse magnetic polarizabilities of some small apertures”, IEEE Trans. Microwave Theory and Techniques, vol. 35, no. 1, Jan. 1987, pp. 20-23.
- [27] **N. A. McDonald**, “Simple approximations for the longitudinal magnetic polarizabilities of some small apertures”, IEEE Trans. Microwave Theory and Techniques, vol. 36, no. 7, Jul. 1988, pp. 1141-1144.
- [28] **M. C. Render, and A. C. Marvin**, “A multiple mode equivalent transmission line model of crosstalk between conductors in shielded environments”, IEEE Trans. Electromagnetic Compatibility, vol. 37, no. 3, Aug. 1995, pp. 488-494.
- [29] **D. W. P. Thomas, A. Denton, T. M. Benson, T. Konefal, J. F. Dawson, A. C. Marvin, S. J. Porter, C. Christopoulos, and J. Paul**, “Electromagnetic coupling to an enclosure via a wire penetration”, IEEE International Symposium on Electromagnetic Compatibility, vol. 1, Aug. 2001, pp. 183-188.
- [30] **M. Feliziani, and F. Maradei**, “EMI prediction inside conductive enclosures with attached cables”, IEEE International Symposium on Electromagnetic Compatibility, vol.1, Aug.2001, pp. 167-172.

- [31] **Jean-Fu Kiang, Shau-Hua Chen, Chung-I G. Hsu, and Ching-Her Lee**, “Shielding effectiveness of bent slots”, IEEE International Symposium on Antennas and Propagation Society, vol. 4, Jul.1999, pp. 2680-2683.
- [32] **D. W. P. Thomas, A. C. Denton, T. Konefal, T. Benson, C. Christopoulos, J. F. Dawson, A. Marvin, S. J. Porter, and P. Sewell**, “Model of the electromagnetic fields inside a cuboidal enclosure populated with conducting planes or printed circuits boards”, IEEE Trans. Electromagnetic Compatibility, vol. 43, no. 2, May 2001, pp. 161-169.
- [33] **G. P. Caixeta, and F. J. Pissolato**, “ Analysis of the electromagnetic field transient inside a shielded structure”, IEEE International Symposium on Electromagnetic Compatibility, vol. 1, Aug. 2001, pp. 239-242.
- [34] **C. F. M. Carobbi, L. M. Millanta, and L. Chiosi**, “The high-frequency behaviour of the shield in the magnetic-field probes”, IEEE International Symposium on Electromagnetic Compatibility, vol. 1, Aug. 2000, pp. 35-40.
- [35] **T. Konefal, J. F. Dawson, A. Denton, T. M. Benson, C. Christopoulos, A. C. Marvin, S. J. Porter, and D. W. P. Thomas**, “Electromagnetic field predictions inside screened enclosures containing”, International Conference and Exhibition on Electromagnetic Compatibility, Jul. 1999, pp. 95-100.
- [36] **J. R. James, A. J. Race and L. A. Scott**, “Electromagnetic shielding degradation effects in composite material enclosures”, Electronics letters, vol. 35, no. 3, Feb. 1999, pp. 209-211.
- [37] **T. Martin, M. Backstrom, and J. Loren**, “Semi-empirical modelling of apertures for shielding effectiveness simulations”, IEEE Trans. Electromagnetic Compatibility, vol. 45, no. 2, May 2003, pp. 229-237.
- [38] **K. F. Casey**, “ Electromagnetic shielding behavior of wire-mesh screens”, IEEE Trans. Electromagnetic Compatibility, vol. 30, no. 3, Aug. 1988, pp. 298-306.

- [39] **F. B. Beck, C. R. Cockrell, and M. D. Deshpande**, “Electromagnetic shielding effectiveness of rectangular enclosure with perforated walls”, Proceedings of the Digital Avionics Systems Conference, vol. 1, Oct. 1999, pp. 3.D.3-1 - 3.D.3-6.

- [40] **N. Sudarshan, S. K. Chatterji, K. Suryanarayana, and G. V. Rao**, “EM shielding effectiveness of metal sandwiched composites”, Proceedings of the International Conference on Electromagnetic Interference and Compatibility, Feb. 2002, pp. 99-103.

- [41] **K. F. Casey**, “Low-frequency electromagnetic penetration of loaded apertures”, IEEE Trans. Electromagnetic Compatibility, vol. 20, no. 2, May 1978, pp. 296-305.

- [42] **S. Criel, L. Martens, and D. De Zutter**, “Near- and far- field characterization of perforated screens: theoretical and experimental study of proposed definitions for the shielding performance”, IEEE International Symposium on Electromagnetic Compatibility, Aug. 1994, pp. 298-302.

- [43] **J. J. Rollin, L. Roy, and G. Arcari**, “A novel technique for measuring IC package shielding effectiveness”, Proc. of IEEE Instrumentation and Measurement Technology Conference, May 1999, pp. 1927-1929.

- [44] **J. P. Spratt, B. C. Passenheim, R. E. Leadon, S. Clark, and D. J. Strobel**, “Effectiveness of IC shielded packages against space radiation”, IEEE Trans. Nuclear Science, vol. 44, no. 6, Dec. 1997, pp. 2018-2025.

- [45] **L. O. Hdeft, B. Bergsrud, and T. Young**, “Measured electromagnetic shielding of small composite enclosure/junction boxes”, IEEE International Symposium on Electromagnetic Compatibility, vol. 2, Aug. 2001, pp. 697-701.

- [46] **J. D. Nordgaard, and R. M. Sega**, “Measured and predicted coupling of electromagnetic radiation into a cylindrical cavity through a small aperture,” Proc. 8th Int. Zurich Symp. Electromagnet Compatibility, Mar. 1989, pp. 93-98.
- [47] **T. Konefal, J. F. Dawson, A. Denton, T. M. Benson, C. Christopoulos, A. C. Marvin, S. J. Porter, and D. W. P. Thomas**, “Electromagnetic field predictions inside screened enclosures containing”, International Conference and Exhibition on Electromagnetic Compatibility, Jul. 1999, pp. 95-100.
- [48] **M. S. Sarto, S. Di Michele, and P. Leerkamp**, “Electromagnetic performance of innovative lightweight shields to reduce radiated emissions from PCBs”, IEEE Trans. on Electromagnetic Compatibility, vol. 44, no. 2, May 2002, pp. 353-363.
- [49] **R. F. Harrington**, “Field Computation by Moment Methods”, second edition, IEEE Press, 1993.
- [50] **R. Bancroft**, “Understanding Electromagnetic scattering using the moment method”, Artech House, 1996.
- [51] **C. A. Balanis**, “Advanced Engineering Electromagnetic”, John Wiley & Sons, 1989.
- [52] **W. Wallyn, F. Olyslager, E. Laermans, D. De Zutter, R. De Smedt, and N. Litaert**, “Fast evaluation of the shielding efficiency of rectangular shielding enclosure”, IEEE International Symposium on Electromagnetic Compatibility, vol. 1, Aug. 1999, pp. 311-316.
- [53] **D. M. Pozar**, “Microwave Engineering”, John Wiley & Sons, 2nd edition, 1998.
- [54] **B. Archambeault, and O. Ramahi**, “Evaluating tools which predict the shielding effectiveness of metal enclosures using a set of proposed standard EMI

modelling problems”, IEEE International Symposium on Electromagnetic Compatibility, Aug. 1998, pp. 517-521.

[55] <http://www.ewh.ieee.org/cmte/tc9/>

Annex

All simulations are run by using a Pentium IV 2.4 GHz PC.

Structure	No. of frequencies	Mesh Size	Runtime
Fig 3.14	60	1626	609 mins
Fig 3.15	60	1626	606 mins
Fig 3.16	60	1626	624 mins
Fig 3.17	60	1630	612 mins
Fig 3.19	60	1367	611 mins
Fig 3.21	46	4132	11694 mins
Fig 3.22	41	4410	14723 mins
Fig 3.23	91	4565	47101 mins
Fig 4.3 (horizontal slot)	46	4243	11816 mins
Fig 4.3 (vertical slot)	46	3973	8367 mins
Fig 4.4 (L = 10 cm)	46	4279	13753 mins
Fig 4.4 (L = 30 cm)	46	4225	13165 mins
Fig 4.4 (L = 10 cm) & Fig. 4.5 (W = 5 cm)	46	4189	13128 mins
Fig 4.5 (W = 10 cm)	46	3893	9388 mins
Fig 4.7 (1 aperture)	21	3983	5126 mins
Fig 4.7 (2 apertures)	21	4230	5175 mins
Fig 4.7 (4 apertures)	21	4472	5221 mins
Fig 4.9 (1 slot)	46	4189	13095 mins
Fig 4.9 (2 slots)	46	4174	14160 mins
Fig 4.9 (3 slots)	46	4331	13099 mins
Fig 4.10 (straight wire)	60	4002	30523 mins
Fig 4.10 (bent wire)	60	4012	30594 mins
Fig 4.12 (centre)	21	3980	4569 mins
Fig 4.12 (position 1)	21	3980	6995 mins

Fig 4.12 (position 2)	21	3987	10811 imins
Fig 4.12 (position 3)	21	3987	6910 mins
Fig 4.14 (no ground point)	60	1060	1226 mins
Fig 4.16(1# ground point)	60	1062	1228 mins
Fig 4.16(1-2# ground points)	60	1065	1230 mins
Fig 4.16(1-4# ground points)	60	1070	1234 mins
Fig 4.16(5-8# ground points)	60	1070	1234 mins
Fig 4.16(1-8# ground points)	60	1080	1242 mins