Hybrid S-Parameters for Analysis of Mixed RF-Digital Circuits Subject to External Electromagnetic Interference

Yakup Bayram and John L. Volakis
ElectroScience Laboratory, The Ohio State University
1320 Kinnear Road, Columbus, OH 43212
Email: bayram.2@osu.edu and volakis.1@osu.edu

I. INTRODUCTION

Electromagnetic interference due to on-board power sources (on-board EMI) and signal integrity for mixed signal circuits are important to designers of RF-digital circuits. In addition to on-board EMI, susceptibility analysis of mixed signal circuits subject to external electromagnetic interference (off-board EMI) is also important due to recent security threads. The challenge in the analysis of this class of problems relates to the large size and complexity of the structure and its components. However, for RF-digital systems, an additional challenge is the integrated analysis of circuit elements (the governing equations being the -Kirchhoff’s Voltage (KVL) and Current Law (KCL)) with EM structures governed by Maxwell’s equations. While the former is expressed in terms of voltages and currents, the latter requires the solution of electric and magnetic fields. Due to the inherent non-linearities of the analog and digital circuits having varying input and output impedances, it is impractical to integrate both into a full wave EM solver in the frequency domain. Therefore, EM structure either needs to be transformed into a circuit compatible form via, for instance, Partial Element Equivalent Circuit (PEEC, [1]) method or port analysis via S-Parameter network characterization [2].

Looking at the previous works, we observe that the analysis of off-board EMI on mixed signal circuits is often done using Multiconductor Transmission Line Theory (MTLT, [3]). This is used in conjunction with PEEC with the external field coupling represented via distributed voltage and current sources along the interconnects. However, even though MTLT yields reasonable accuracy for simple Printed Circuit Boards (PCBs). A more robust method is needed to handle complicated non-uniform interconnects, via holes, and multi-layer PCBs. Time domain techniques such as the Finite Difference Time Domain (FDTD) method have been used for concurrent analysis of circuits and EM structures [4]. While time domain techniques also yields accurate results, they suffer from computational inefficiencies due to the meshing of large volumes and simulation of RF-devices with high quality factors. They may also run into convergence problems for circuit elements with stiff differential equations.

In this paper, we extend the port analysis technique to accommodate the external plane wave coupling. In other words, we propose a frequency domain method which overcomes the CPU bottleneck of time domain techniques and concurrently yields increased accuracy as compared to MTLT. As part of this formalism, we introduce additional hybrid S-parameters to establish a link between the existing board ports and external plane waves. Thus, we can handle both on-board and off-board EMI problems simultaneously. This approach can be easily integrated into circuit solvers such as HSPICE and Advanced Design System (ADS, Agilent Technologies). This is achieved by exporting new hybrid S-parameters. The method allows both time domain simulations via broad band network characterization and Harmonic Balance Simulation of non-linear RF components.

II. PROBLEM DEFINITION

Consider the off-board EMI problem shown in Fig. 1(a), displaying a plane wave coupling onto a PCB housing analog and digital components. The frequency of the external plane wave is in the upper range of UHF Band (2-3 GHz) and the electronic circuits consist of both passive and
active components. Further, the PCB is comprised of multi-layer dielectric substrates with typical thicknesses on the order of 0.5-3.0 millimeters.

III. SCATTERING PARAMETERS AND PORT ANALYSIS OF ON-BOARD EMI

We begin the analysis by first revisiting the S-Parameter modeling. We then proceed with the port analysis of on-board EMI.

Scattering parameters are well studied for microwave port analysis. Basically, instead of representing port relations in terms of voltage and currents via Z-Parameters (Impedance) or Y-Parameters (Admittance), S-parameters utilize incident and reflected electromagnetic fields to establish a physics-based mathematical relation among the ports. Impedance and admittance parameters require ports to be terminated with open or short circuits respectively and this makes it difficult in practice. On the other hand, S-parameters are defined with respect to a reference impedance, more prevalently 50Ω, making them more suitable for practical situations.

Pertinent to our case, let us consider the $k^{th}$ incident and reflected mode at the $i^{th}$ port. Typically, the mode amplitudes are denoted by $a^k_i$ (incident) and $b^k_i$ (reflected) which can be related to the voltages and currents at the $i^{th}$ port via the relations

$$a^k_i = \frac{V^k_i + Z_i I^k_i}{2\sqrt{|\Re(Z_{ref_i})|}}$$
$$b^k_i = \frac{V^k_i - Z_i I^k_i}{2\sqrt{|\Re(Z_{ref_i})|}}$$

As usual, $Z_{ref}$ is the chosen reference impedance for the corresponding $i^{th}$ port. Accordingly, the corresponding voltage and current for the propagating $k^{th}$ mode at the terminals are given via the relations,

$$V^k = -\int_C \mathbf{E}^k \cdot d\mathbf{l}$$
$$I^k = -\oint_C \mathbf{H}^k \cdot d\mathbf{l}$$

It must be noted that there is no clear definition of voltage and current for non-TEM modes because the integrals (2) for time-harmonic case are not path independent. However, the ambiguity with voltage and current definition disappears for the quasi-Transverse Electromagnetic Waves (quasi-TEM).

Since the dominant propagating mode between the interconnects and the reference plane is quasi-TEM mode, we can carry out our analysis based on this definition for the voltage and currents. Henceforth, we will remove the superscript $k$ for the incident and reflected waves at ports. When we consider all possible ports within a RF-digital system, the relation between the incident and reflected waves at ports takes the form,

$$[\bar{b}] = [\bar{S}] [\bar{a}]$$

where $S_{ij} = \frac{b_j}{a_i}$ with all ports are terminated with their corresponding reference impedance $Z_{ref}$.

For on-board EMI analysis, microwave ports are defined at the intersections of circuit components, and interconnects (see Fig. 1(b)). Full wave EM analysis is performed to compute the corresponding N-Port S-parameters and finally the resulting scattering matrix is exported into circuit solvers such as HSPICE or ADS for the simulation of the overall geometry including RF-digital circuit elements and power sources.
IV. HYBRID S-PARAMETER APPROACH TO ACCOUNT FOR EXTERNAL EMI

A key aspect of our formulation for external EMI analysis of RF-digital boards is the modification of the N-port $[S]$ matrix with the addition of a non-physical $(N+1)^{th}$ port (see Fig. 2). This additional port can be thought as representing the incident and scattered far zone plane waves from the entire board. Thus, the scattering matrix for the mixed excitation analysis takes the form

$$
\begin{pmatrix}
  b_1 \\
  \vdots \\
  b_N \\
  b_{N+1}
\end{pmatrix} = 
\begin{pmatrix}
  S_{1,1} & \cdots & S_{1,N} & HS_{1,N+1} \\
  \vdots & \ddots & \vdots & \vdots \\
  S_{N,1} & \cdots & S_{N,N} & HS_{N,N+1} \\
  S_{N+1,1} & \cdots & S_{N+1,N} & S_{N+1,N+1}
\end{pmatrix}
\begin{pmatrix}
  a_1 \\
  \vdots \\
  a_N \\
  a_{N+1}
\end{pmatrix}
$$

\begin{align*}
  \begin{pmatrix}
    S^{PCB-PCB} & HS^{EF-PCB} \\
    HS^{PCB-EF} & S^{EF-EF}
  \end{pmatrix}
  \begin{pmatrix}
    [a]^{PCB} \\
    a_{N+1}
  \end{pmatrix}
  &+ \begin{pmatrix}
    a_1 \\
    \vdots \\
    a_N
  \end{pmatrix}
\end{align*}

(4)

where $HS_{i,N+1}$ refers to the hybrid scattering parameter contributing to the $i^{th}$ port due to the external field coupling to the PCB ports and $[a]^{PCB}$ contains the incident waves at the physical PCB ports. Since we are only interested in coupling to the ports on PCB, reflected wave on the EMI port can be neglected and later in this section we explain how $S_{N+1,i}$ is treated. Hence, (4) becomes

$$
\begin{pmatrix}
  b_1 \\
  \vdots \\
  b_N
\end{pmatrix} = 
\begin{pmatrix}
  S^{PCB-PCB} & [a]^{PCB} + \begin{pmatrix}
    HS^{EF-PCB} \\
    a_{N+1}
  \end{pmatrix}
\end{pmatrix}
\begin{pmatrix}
  a_1 \\
  \vdots \\
  a_N
\end{pmatrix}
$$

(5)

$S^{PCB-PCB}$ is already known from full wave port analysis of the board without external field excitation and the matrix in (5) with the last row and the column removed. The second term, $HS^{EF-PCB}$ contains the last column of (5) and refers to the quasi-TEM port coupling from the incident plane wave and is computed with all ports terminated with $Z_{ref}$. However, the induced fields due to the plane wave excitation is not necessarily quasi-TEM. Therefore, we proceed to extract quasi-TEM contributions by exploiting the orthogonality of the propagating modes between the interconnects and the reference plane. To do so, we start with the definition of total voltage between any interconnect and the reference plane,

$$V^{total} = -\int_C \frac{E^{total}}{c} \cdot dl = -\int_C \left( \frac{E^{inc}}{c} + \frac{E^{scat}}{c} \right) \cdot dl = V^{inc} + V^{scat}
$$

(6)

We further decompose the scattered voltage $V^{scat}$ into contributions from the active sources on the PCB board ($V^{scat}_{PCB-Sources}$) and that the induced voltage from the external plane wave ($V^{scat}_{EMI}$). That is, we write $V^{total}$ as

$$V^{total} = V^{inc} + V^{scat}_{PCB-Sources} + V^{scat}_{EMI}
$$

(7)

The contribution of $V^{scat}_{PCB-Sources}$ is already accounted for via $S^{PCB-PCB}$ and $V^{inc}$ is the voltage from the incident field in the absence of the whole structure. Therefore, it is constant at each port regardless of the port termination. It implies that $V^{scat}_{EMI}$ is the only propagating term that contributes to the hybrid S-parameters in $H^{EF-PCB}$.

Fig. 2. Modeling of external field as additional quasi-TEM port
To find the hybrid S-parameters, we terminate each port on the PCB with their corresponding reference impedances. Thus, $a_{PCB}^P$ becomes zero in (5) and from (1) induced scattering voltage yield

$$H_{SEF-PCB} = \frac{1}{|E_0|} \begin{pmatrix} V_{\text{scat EMI}_1} \\ \vdots \\ V_{\text{scat EMI}_N} \end{pmatrix}, \quad V_{\text{EMI}} = -\int_C (\overline{E_{\text{scat EMI}_1}} \cdot e^{TEM}(x, y, z)) \cdot dl \quad (8)$$

where $a_{N+1}$ is set to $|E_0|$ the magnitude of the incident plane wave. Also, $e^{TEM}$ is the quasi-TEM field stamp at the ports and can be easily obtained from N-port full wave S-parameter simulation or analytic expressions for simple structures. Since $b_{N+1}$ is of no interest, the last row in (5) is padded with zeros, thus suppressing reflections from EMI port if EMI port is not matched. This can be readily proved using Mason’s rule. Having obtained the hybrid S-parameters, we can then proceed to export $[N+1 \times N+1]$ S-parameter matrix to any choice of circuit solvers with the corresponding EMI port and voltage source as shown in Fig. 2. In addition to obtaining induced scattered voltage using the aforementioned method, $V^{inc}$ should be also added to the port voltage to account for the total voltage induced at the ports for linear loads. However, it must be noted that non-linear circuit elements require different treatment than linear elements depending on the simulation algorithm used such as SPICE or HBM and this issue will be addressed in more detail in the presentation.

V. VALIDATION EXAMPLE

To validate the proposed method for a confined internal and external port excitation, we consider the two layer PCB shown in Fig. 3 (a) subject to a plane wave with $E_x = 500V/m$ and $E_y = 750V/m$. The wave direction of incidence is along ($\theta = 42^0$ and $\phi = 326^0$) and operating at 2 GHz.

Each port is first terminated with a 50Ω reference load and a full wave analysis is performed to find the $[4 \times 4] S^{PCB-PCB}$ in the absence of the plane wave excitation. Subsequently, the same geometry is solved with a full wave solver to find the hybrid S-parameters using the relation in (8) under plane wave excitation. The resulting $[5 \times 5]$ S-parameter matrix was exported to ADS and each port was terminated with the corresponding loads as shown under configuration 2 in Fig. 3 (a). The ADS results are compared with the full wave solver used to characterize the entire problem with the corresponding port terminations. It must be noted that incident excitation does not have z-component. Therefore, there is no contribution of $V^{inc}$ to the total induced voltage at the ports. As seen in Fig. 3 (b), plots of port voltages due to the external EMI plane wave excitation, the proposed method and the full wave solver results are in good agreement validating the proposed methodology.

![Validation Geometry: Plane Wave Coupling to Two Layer PCB](a)

![Comparison of Induced Voltages at Each Port](b)

Fig. 3. (a) Validation Geometry: Plane Wave Coupling to Two Layer PCB - (b) Comparison of Induced Voltages at Each Port
REFERENCES


