Chapter 4

An efficient methodology for transient simulation of signal integrity problems

The second part of the dissertation deals transmission line effects at RF frequency (1GHz to 300GHz) in MMICs/PCBs. When signal pass through transmission line at gigahertz frequency (1GHz to 300GHz) with a picoseconds rise/fall time, the wire/interconnect/transmission line at both MMIC/Chip and package/housing level are no longer ‘electrical transparent’ to the signals and signal integrity is affected by electromagnetics effects (or called transmission line effects) such as reflection, delay, jitter, attenuation and noise. It is called a signal integrity problem. To take into account these electromagnetics effects in a time domain analysis and to integrate these effects into circuit solver, the macromodel is to be derived. The black box representation that captures electromagnetics behavior at the input/output ports is called a macromodel. Hence, the development of an efficient macromodel and a simple procedure of embedding this macromodel into circuit solvers become necessary for transient simulation of signal integrity in the distributed passive devices in MMIC/PCB.

Any transmission line is best represented by frequency domain, but for transient simulation and to do an analysis of signal integrity problems, it is mandatory to have a time domain model which can be directly connected with transient solver, hence either from measurement or using the full wave simulator, Y parameters of giving under test
network should be extracted. In section 4.2, FIT technique is used to extract the Y parameters of given networks. Once the admittance matrix is available, pole residue form is obtained using the vector fitting technique. In section 4.4.1, we have discussed how to embed pole residue form into circuit solver for transient simulation and analysis. In this chapter, we propose an efficient and simple approach to embed broadband frequency domain responses of passive and linear multiport networks into circuit solver using the vector fitting method for signal integrity analysis.

### 4.1 Introduction

For signal integrity analysis at high frequency, it is mandatory to develop an efficient macromodel of given broadband frequency domain response of any passive and linear multiport network and to club it into circuit solver in a simple and efficient way. There are two major issues in it. One difficulty is the mixed frequency/time domain problem and the other is the computationally expensive [21]. A traditional ordinary differential equation solver such as a SPICE-like circuit simulator cannot efficiently handle this mixed domain problem. The multi port frequency dependent networks equivalent circuit representation is highly desirable for circuit solvers [49]. In spite the CPU expenses and convergence issues of full wave method, the full wave method can be used to extract the network property, such as its Y-parameters, Z-parameters, or S-parameters [50]. Thereafter, the macromodeling techniques [21, 42] can be used to construct macromodel from the given tabulated frequency domain data characterizing the network property of the system. A vector fitting method is robust technique used in this dissertation for constructing
multiport macromodel from tabulated frequency domain data [42].

Y-parameter based macromodel can be used to club with Electromagnetic Transients Program (EMTP)-type circuit solvers via either an equivalent circuit [43] or convolution method [44]. Because of the accuracy problem in equivalent circuit option [45], Norton equivalent and convolution methods may be used for interfacing macro model with electromagnetic transient’s simulators for multiport Y-, Z- and S-parameter-based models [51]. The disadvantage of these techniques is that they require a prior knowledge of source code of the circuit solver. As most solvers do not have a facility to club the user defined models directly, S-parameters, data have to convert into Y-parameters prior to model extraction. This approach is used in this dissertation. Tabulated frequency domain data in the form of the admittance matrix are obtained through Finite Integration Technique (FIT) which is discussed in section 4.2.

In this proposed method, prior knowledge of source code of the circuit solver is not required. Hence the proposed method is simple. Once the pole residue form is created via vector fitting, automatic net list is generated in a foster pole residue form which is then directly embedded into HSPICE or compatible circuit solvers. Equivalent circuit representation or convolution methods are not used here so the proposed method has not suffered from accuracy problem even when the circuit is having complicated interconnecting systems. Hence it is also more efficient than previous work. A complete proposed approach is as shown in Fig. 4.1. A practical example of single ended via connected with two microstrip lines is discussed which shows the validity of the proposed approach.

4.2 Admittance parameters extraction

The tabulated frequency domain data in the form of the admittance matrix are obtained through Finite Integration Technique (FIT). The Finite Integration Technique (FIT) provides a dual grid discretization scheme applicable to various electromagnetic problems. FIT uses following discrete set of Maxwell’s Grid Equations (MGEs):

\[
Ce = -\frac{d}{dt}b, \tilde{C}h = -\frac{d}{dt}d + j, \tilde{S}d = q, Sb = 0,
\]

(4.1)

Where topological matrix \( C \) and \( \tilde{C} \) as the discrete equivalent of the analytical curl operator and discrete divergence operators \( S \) and \( \tilde{S} \).
A method for the calculation of admittance matrix where the admittance matrix $Y(s)$ can alternatively be transformed from its scattering $S$-parameters is efficient and used in this thesis. The $S$-parameters can also be obtained through measurement e.g. vector network analyzer. $Y$-parameter matrix of $N_p$-port subnetwork can be shown as following:

$$
\begin{bmatrix}
Y_{11}(\omega_k) & \cdots & Y_{1n}(\omega_k) \\
\vdots & \ddots & \vdots \\
Y_{n1}(\omega_k) & \cdots & Y_{nn}(\omega_k)
\end{bmatrix}
$$

where $\omega_k = 2\pi f_k$, $k = 1, 2, \ldots, K$ are discrete frequency points within the desired band.

### 4.3 Vector fitting technique

In the Vector Fitting Method, the objective is to identify a rational approximation of the each element of 4.2 that approximates a given broadband frequency response $Y_{ij}(s)$ as closely as possible.

$$Y_{ij}(s) = \sum_{n=1}^{N} \frac{r_{n_{ij}}}{s - q_n} + d_{ij} + s e_{ij}$$

(4.3)

denote, $r_{n_{ij}}$ and $q_n$ respectively, the residues and poles, both of which are complex, in general and while the terms $d_{ij}$ and $e_{ij}$ are real constants, $i, j = 1, 2, \ldots, N_p$.

Vector fitting is an iterative procedure for solving 4.3. Each iteration consists of a pole identification step followed by a residue identification step with known pole. An iterative process is used to enhance accuracy, and $Y_{ij}(s)$ converges to an optimized rational function fit.

The pole identification step requires in each iteration to solve in the least squares (LS) sense equation

$$W(s) = \sigma(s) \cdot Y_{ij}(s) = \sum_{n=1}^{N} \frac{\hat{r}_{n_{ij}}}{s - \hat{q}_n} + \hat{d}_{ij} + s \hat{e}_{ij}$$

(4.4)

where

$$\sigma(s) = 1 + \sum_{n=1}^{N} \frac{\tilde{r}_{n_{ij}}}{s - \tilde{q}_n}$$

(4.5)

Substituting 4.5 into 4.4 and for a given sample points $M_f$, following linear equation in
compact form can be derived:

\[ A_f \cdot u = b_f \]  \hspace{1cm} (4.6)

where

\[ A_f = \begin{bmatrix}
\Omega_f & \Omega_{f,1} \\
\Omega_f & \Omega_{f,2} \\
\vdots & \vdots \\
\Omega_f & \Omega_{f,K}
\end{bmatrix}, \]

\[ \Omega_f = \begin{bmatrix}
\frac{1}{\sigma_1} & \cdots & \frac{1}{\sigma_n} \\
\vdots & \ddots & \vdots \\
\frac{1}{\sigma_M} & \cdots & \frac{1}{\sigma_n}
\end{bmatrix} + \begin{bmatrix}
S_1 \\
S_2 \\
\vdots \\
S_M
\end{bmatrix}, \]

\[ \Omega_{f,k} = \begin{bmatrix}
\frac{-H_k(S_1)}{\sigma_1} & \cdots & \frac{-H_k(S_1)}{\sigma_n} \\
\vdots & \ddots & \vdots \\
\frac{-H_k(S_M)}{\sigma_1} & \cdots & \frac{-H_k(S_M)}{\sigma_n}
\end{bmatrix} \\
\]

\[ u = \begin{bmatrix}
u_1^T \\
\vdots \\
u_k^T \\
\tilde{u}^T
\end{bmatrix}^T, \]

\[ u_k = \begin{bmatrix}
\hat{r}_{1,k} \\
\vdots \\
\hat{r}_{N,k} \\
d_k \\
e_k
\end{bmatrix}^T, \]

\[ \tilde{u} = \begin{bmatrix}
\hat{r}_1 \\
\vdots \\
\hat{r}_N
\end{bmatrix}^T, \]

\[ \tilde{b}_f = \begin{bmatrix}
\tilde{b}_{f,1}^T \\
\vdots \\
\tilde{b}_{f,k}^T
\end{bmatrix}^T, \]

\[ \tilde{b}_{f,k} = \begin{bmatrix}
H_k(S_1) \\
\vdots \\
H_k(S_M)
\end{bmatrix}^T, \]

where \( k = 1, 2, \ldots, K \).

Solving the least squares problem of (4.6), the coefficients of \( \sigma(s) \) and product of \( \sigma(s) \cdot Y_{ij}(s) \) are obtained. \( Y_{ij}(s) \) can be expressed as \( Y_{ij}(s) = \frac{W(s)}{\sigma(s)} \), which shows that poles of \( Y_{ij}(s) \) coincides with the zeros of \( \sigma(s) \). The zeros of \( \sigma(s) \) are calculated from \( u \), which are calculated as the eigenvalues of the matrix

\[ H = A - b \tilde{u}^T \]  \hspace{1cm} (4.7)
where \( A \) is a diagonal \( \sigma(s) \) matrix containing the starting poles, \( b \) is a column vector of ones and \( \tilde{u}^T \) is a row-vector containing the residues for \( \sigma(s) \).

The second step is the calculation of the residues of \( Y_{ij}(s) \). In principle, it can be directly obtained from 4.6; instead, equation 4.3 is solved with the obtained zeros of \( \sigma(s) \), which again constitutes a least squares problem to improve accuracy, and the residues of \( Y_{ij}(s) \) are computed. An iterative process is used to enhance accuracy, and \( Y_{ij}(s) \) converges to an optimized rational function fit.

### 4.4 Inclusion of macromodel into circuit solver

There are basically two ways to incorporate the macromodel of distributed interconnect structures into circuit simulators (1) synthesize the equivalent circuit using basic circuit elements of the circuit simulator, or (2) use recursive convolution. In this chapter, we present the application of (1) techniques for interfacing rational function approximations of the Y parameter with SPICE.

#### 4.4.1 Foster pole residue representation

Once the macro model of the given network is created as described in the preceding section, for embedding macromodel into HSPICE or compatible circuit solvers, following syntaxes in Foster Pole-Residue form is used [52]. Transconductance \( G(s) \) form:

\[
G_{\text{xxx}} n + n - \text{FOSTER} \text{in} + \text{in} - k_0 k_1 + (ReA_1, ImA_1)/(Rep_1, Imp_1)
+ (ReA_2, ImA_2)/(Rep_2, Imp_2) + (ReA_3, ImA_3)/(Rep_3, Imp_3) \ldots \ldots (4.8)
\]

A pole-residue pair is represented by four numbers - real and imaginary part of the residue, then real and imaginary part of the pole. We must make sure that \( Re[\pi]<0; \) otherwise, the simulations will certainly diverge. For example, to represent \( G(s) \) in the
4.4 Inclusion of macromodel into circuit solver

Figure 4.2: (a) Model of single ended via connecting with two microstrip lines (b) Schematic circuit diagram of the test model: Two-port network having source and load forms of equation 4.9,

\[
G(s) = 0.001 + 1 \times 10^{-12}s + \frac{0.0008}{s + 1 \times 10^{10}} + \frac{0.001 - j0.006}{s - (-1 \times 10^8 + j1.8 \times 10^{10})} + \frac{0.001 + j0.006}{s - (-1 \times 10^8 - j1.8 \times 10^{10})}
\] (4.9)

We will input in the net list file as: G1 1 0 FOSTER 2 0 0.001 1e-12+ (0.0004, 0)/ (-1e10, 0) (0.001, -0.006)/ (-1e8, 1.8e10).

The netlist file can be prepared as per above form for each elements of 4.2. This netlist file can be well inserted into HSPICE or its compatible circuit solvers for transient simulation.
4.5 Numerical results

The Printed Circuit Boards (PCBs) may hold large numbers of discontinuities, and their presence causes signal distortion problems. For this reason, an accurate modeling of discontinuities at PCB level is essential for signal integrity (SI) analysis. A suggested methodology is validated for a simple via hole connecting with microstripes in multilayer PCB. A single ended via connected with two microstrip lines is shown in Fig. 4.2. Y-parameter matrix of this network is extracted by Finite Integration Technique (FIT) using CST Microwave Studio (MWS). The example network is having input and output port, port 1 and port 2 respectively. In FIT, for transient simulation, Gaussian pulse source and ABC boundary condition are applied. Each element of the $Y$-parameter matrix is approximated with the vector fitting method. 50 poles, including 10 real and 40 complex conjugate poles are drawn out to meet the parameters of the test network up to 6 GHz. Good agreements can be observed in Fig. 4.3 between the FIT simulated $Y$-parameters and the macromodel $Y$-parameters based on the vector fitting method. For embedding the macromodel into circuit solvers, netlist file which is the equivalent sub circuit in time domain, is automatically created and inserted into HSPICE or equivalent circuit simulator by the proposed approach in this thesis. Obtained time domain model is employed to perform the transient analysis of simple via hole in multilayer PCBs. The conventional circuit is having 50 ohm source and 50 ohm load impedance at port 1 and 2 respectively as shown in Fig. 4.2. The circuit is excited at port 1 by a pulse having a rise/fall time of 0.02 ns and a pulse width of 4 ns and transient simulation is performed. Transient simulation results are shown in Fig.4.4. The result indicates that the line is ringing with overshoots.

4.6 Conclusion and discussion

The second piece of dissertation proposes an efficient and uncomplicated access to embed broadband frequency domain responses of passive and linear networks into circuit solver using the vector fitting method. In this thesis, systematic methodology is discussed and details of embedding the macromodel into circuit solver are described. Single ended via connected with two microstrip lines test case is presented which validate that the
Figure 4.3: Y parameters of the single ended via simulated by FIT in CST MWS and Vector Fitting method. (a) Magnitude. (b) Phase
4.6 Conclusion and discussion

Figure 4.4: Transient voltage waveforms: (a) at port 1 denoted as $V_{in}$ and at port 2 denoted as $V_{out}$.

The suggested approach is desirable for a transmission line, high speed interconnects as well on board discontinuities such as via holes. A practical test case indicates that the analysis of signal integrity for complicated interconnecting systems can be done efficiently. Ideally, any passive and linear microstrip circuit can be studied and transient simulation can be executed utilizing the proposed approach before its real execution, so the design cost and time will be concentrated. In future, the proposed method can be used to perform the transient simulation of variety of EMC-EMI, Signal Integrity and Power Integrity problems in the high-speed complex hybrid interconnect and lumped circuit systems.