Optimal Design of Self-Damped Lossy Transmission Lines for Multichip Modules

Jimmy Shinn-Hwa Wang
Dr. Wayne Wei-Ming Dai

UCSC-CRL-93-43
8 October 1993

Baskin Center for
Computer Engineering & Information Sciences
University of California, Santa Cruz
Santa Cruz, CA 95064 USA

ABSTRACT

This paper presents a simple and robust method of designing the lossy-transmission-line interconnects in a network for multichip modules. This method uses wire-sizing entirely to meet the electrical damping criteria to solve the problems encountered in propagating high-speed signals through unterminated lossy transmission lines on the substrates of multichip modules (MCM). The optimal design method is based on a new improved scattering-parameters (S-parameters) based macro-model of distributed-lumped networks that keeps track of the time-of-flight term in the transfer functions. The wire-sizing optimal design concept is to relate the layout parameter (line width) and the transfer function (damping ratio, and natural undamped frequency) to the signal propagation delay. The optimal design method results in fast and stable signal propagation for single-source multi-receiver networks on multichip modules without using termination resistors. Multi-receiver networks on High Performance MCM process technologies are designed for illustration.

Keywords: Optimal Design, Self-Damped, Lossy Transmission Line, Multichip Module, Perturbation

Contents

1 Introduction

A new revolution in packaging technology called the multichip module (MCM) has several bare chips mounted and interconnected on a substrate. By eliminating the individual chip packages, chips are placed closer together resulting in higher packaging density than the hybrid and Printed Circuit Board (PCB). The metal linewidth on MCM are much wider than those of modern VLSI chip, so the line itself exhibits non-negligible inductance. Compared with the metal lines on PCB, those metal lines on MCM have a larger resistance per-unit-length due to a smaller cross-section. Therefore the metal lines on MCM must be treated as lossy transmission lines.

The interconnection lines on multichip modules exhibit reflections and resonances due to its transmission line characteristics and must therefore be terminated. Different layout parameters and terminations result in output waveforms of different damping conditions. Figure 1.1 shows the underdamped, critical damped, and overdamped output waveforms of a ramp-step input waveform. The overshoot of the underdamped waveform for a 5-volt input is almost one-volt and can be destructive if it is left unterminated.

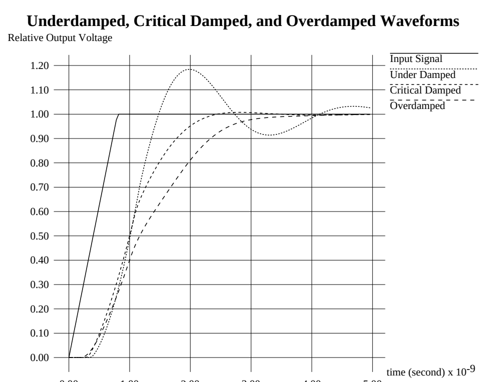

Figure 1.1: **Simulation Waveforms of Different Damping Conditions:** The input waveform is shown in solid line. The under damped output waveform is shown in short dash line. The critical damped output waveform is shown in medium dash line. The overdamped output waveform which has the longest delay is shown in long dash line.

Usually there are two ways to cope with this problem; one is to terminate the lines with clamping diodes and the other is to terminate with resistors. However, when a line on a multichip module is terminated by a pair of voltage-clamping diodes to limit the positive and negative signal swings, the diode when is turned on by signal voltage overshoot conducts a huge amount of current which can cause an increase in power consumption and an increase in power-distribution disturbance. If a line on a multichip module is terminated with a resistor of appropriate value to minimize signal reflections and resonances, the voltage divider formed by the termination resistance and the line characteristic impedance causes substantial and unacceptable attenuation of propagated signals. In addition, each resistor dissipates quiescent power when the line is at non-zero voltage. In modules containing thousands lines, this power dissipation can be excessive, and the problem is compounded by the small packaging size of MCM with limited heat removal capability.

Instead of terminating with diodes or resistors, the long lines on the thin-film multichip modules can be structured to critically damp the signal to avoid resonances. Doing without the termination eliminates the heat generated by the terminator and solves the heat removal problem. Being unterminated, the long lines in the substrate are structured to exhibit a total resistance that is related to the source resistance of the active devices which drives the lines [6] [1]. Since voltage doubling occurs at the end of the unterminated line due to reflection, a controlled amount of attenuation is tolerable. For performance reason, *slightly* underdamped design gives shorter signal propagation delay with tolerable amount of overshoot [1] [7]. The structured line which is critically damped can transmit input signal frequency components up to the bandwidth of the line without any instability or attenuation. For higher frequencies, attenuation occurs but the line remains stable [6] [3] [1]. These lines are thus called *optimal self-damped lossy transmission lines* [5].

This paper is the first to identify the optimal design concept of optimizing both the maximum path delay and the maximum damping ratio together. This minimization problem is transformed into a least square estimation problem. The least square estimation problem is then solved using the efficient algorithm of Gauss-Marquardt method. The contribution of this paper is to achieve optimal performance completely through wire-sizing. The approaches taken will be presented in the following sections. Section 2 will describe the interpretation of the transfer function. Section 3 will formulate the optimal self-damped design problem. Section 4 will describe the optimization method used. Section 5 will demonstrate the usefulness of the optimization method through examples. Section 6 will present the S-parameter based macro-model with the time-of-flight for a lossy transmission line. Section 7 will describe some previous works done by other researchers. Finally, Section 8 will conclude the findings and contributions of this paper and proposes the future research directions.

2 Transfer Function, Natural Undamped Frequency, Damping Ratio, Line Width, and Signal Propagation delay

The transfer function is defined as:

$$H(s) = \frac{V_{output}(s)}{V_{input}(s)},$$

where s is the complex variable of the Laplace transformation, $V_{output}(s)$ is the Laplace transform of the time domain output waveform $v_{output}(t)$, and $V_{input}(s)$ is the Laplace transform of the time domain input waveform $v_{input}(t)$.

The transfer function $H(s)$ also can be defined as [8]:

$$H(s) = e^{-s\tau} \times \hat{H}(s) \tag{2.1}$$

where τ is the time-of-flight term, and $\hat{H}(s)$ is the part of the transfer function representing the charging curve in the output waveform starting at a delay τ after the input switches at $t = 0$.

The transfer function $H(s)$ can be separated into two terms as in Equation (2.1). Assume a second order approximation is applied to the charging part of the transfer function $\hat{H}(s)$, it can be rewritten as:

$$\hat{H}(s) = \frac{{\omega_n}^2}{s^2 + 2\zeta\omega_n s + {\omega_n}^2}, \tag{2.2}$$

where ω_n is the natural undamped frequency and ζ is the damping ratio of the transfer function

$\hat{H}(s)$ [7]. The natural undamped frequency ω_n is the sinusoidal frequency assuming there is no damping ($\zeta = 0$). ζ is the "ratio of the damping" compared with critical damping. The roots of the denominator polynomial of the transfer function $\hat{H}(s)$ are called the *poles* of $\hat{H}(s)$. Since the poles of $\hat{H}(s)$ are also the poles of $H(s)$, ω_n and ζ are also the natural undamped frequency and damping ratio of the transfer function $H(s)$.

Define the representation of the poles as:

$$s_{1,2} = -\alpha \pm j\omega = -\zeta\omega_n \pm j\omega_n\sqrt{1-\zeta^2}.$$

The damping ratio ζ is:

$$\zeta = \frac{\alpha}{\omega_n} = \frac{\alpha}{\sqrt{\alpha^2+\omega^2}}.$$

For a given input waveform in time domain, the three quantities of the transfer function: the time-of-flight τ, the natural undamped frequency ω_n, and the damping ratio ζ *uniquely* determine the time domain response waveform at the receiver. By the same token, the propagation delay for a signal to reach 50% of its final value at the receiver is also *uniquely* determined by these three quantities: τ, ω_n, and ζ.

In the next section, the damping ratio ζ and the propagation delay will be used together in the optimization process. The approximate natural undamped frequency ω_n is a byproduct simulation.

3 Formulation of the Optimal Self-Damped Design

To illustrate the basic concept, a single-source multi-receiver network is used. It consists of a set of edges and a set of nodes. The source and the receivers are the nodes of the network graph. This design method finds the optimal width assignment that gives the same damping criteria for each path from source to receiver.

The optimal design problem can be formulated as follows:

Given a network (N) consists of a set of edge E with fixed lengths. The width of each edge belongs to a set of feasible widths $C = \{c_1, c_2, ..., c_{max}\}$ where $(c_i < c_{i+1}, 1 \leq i \leq (max-1))$, the optimal design is to find an optimal width assignment: $W^* \subset C$ and a mapping $M : B \rightarrow W^*$ such that every paths from root to leaves has the shortest delay and the same damping criteria if possible. The objective is to optimize the performance which includes minimizing the signal propagation delay for each path from source to receiver with either no overshoot or controlled amount of overshoot. The width of edge i, w_i, is bound by the minimum and maximum feasible width, $min(w_i) \leq w_i \leq max(w_i)$, where $min(w_i), max(w_i) \in C$. The incremental difference, $\triangle c$, where $\triangle c = (c_{i+1} - c_i), 1 \leq i \leq (max-1)$, between feasible widths is depicted by process technology.

The following notations are defined for the formulation of the optimal self-damped design.

n : is the number of the edges in the network.

m : is the number of the receivers (the output nodes).

w_i : is the width of edge i.

l_i : is the given length of edge i.

τ_j : is the propagation delay for a signal to reach 50% of its final output value transmitted along path from source to receiver j.

ζ_j : is the damping ratio of the lowest conditional frequency of the transfer function from source to receiver j.

ζ_{target} : is the electric damping criteria set forth by the user.

The optimal self-damping design for a general network can be formulated as a general nonlinear programming problem as follows, Let

$$F(w_1, w_2, \ldots, w_n) = Maximum(\tau_j),\ 1 \leq j \leq m. \tag{3.1}$$

$$G(w_1, w_2, \ldots, w_n) = Maximum(|\zeta_j - \zeta_{target}|),\ 1 \leq j \leq m. \tag{3.2}$$

Objective

$$\text{Minimize} \quad \begin{bmatrix} F(w_1, w_2, \ldots, w_n) \\ K \cdot G(w_1, w_2, \ldots, w_n) \end{bmatrix}, \quad \textit{where } K \textit{ is a weight constant.}$$

Constraints

$$\zeta_j(w_1, w_2, \ldots, w_n) \geq \frac{1}{\sqrt{2}}, w_i \in C, 1 \leq i \leq n, 1 \leq j \leq m. \tag{3.3}$$

$$g_i(w_1, w_2, \ldots, w_n) = \{w_i - min(w_i)\} \geq 0, w_i \in C, 1 \leq i \leq n. \tag{3.4}$$

$$h_i(w_1, w_2, \ldots, w_n) = \{max(w_i) - w_i\} \geq 0, w_i \in C, 1 \leq i \leq n. \tag{3.5}$$

The objective of optimal self-damped design is to size the wire width of each edge in order to minimize both the maximum delay shown in Equation (3.1) and the maximum of the damping ratio error shown in Equation (3.2), while maintaining that all output damping ratios satisfy the constrain in Equation (3.3) and all edge widths satisfy the constraints in Equation (3.4) and (3.5).

The way to find the optimal answer is to use the *perturbation method*, which works by perturbing each of the design parameters λ_i a small amount in each direction, and finds the right direction and distance to change λ_i. In this paper, the width of each edge w_i is the chosen design parameter to be changed and a simulation is run at each perturbation to find the propagation delay and the damping ratio. The perturbation results of the propagation delays and the damping ratios are then used to compute the gradient matrix. Details of the optimization method is described in Section 4.

4 Optimization Method for the Optimal Self-Damped Design

Section 4.1 will proof that the original optimization problem can be transformed into a least-squares estimation problem. Section 4.2 will demonstrate the detail of implementation of the least-squares estimation optimization. The proof closely follows those used by Zhu et al. in their technical report [13]. The contribution of this paper is not only to have a different target function but also to identify a new concept where delay has to be include in the optimization objectives in order to find the optimal results with the least maximum path delay.

4.1 Proof of the Correctness of the Least-Squares Estimation Transformation

The optimal wire-sizing problem formulated in Equation (3.1) is a general nonlinear programming problem. The damping ratio error minimization and delay minimization problem can be transformed into a least-squares estimation problem. After the transformation, the resulting least-squares estimation problem can be solved using an efficient optimization method [11] to obtain the optimal width assignment for the original problem. The efficient method is called the Gauss-Marquardt method. It is an improved version of Gauss-Newton method by incorporating a more efficient way of determining the Lagrange multipier to speed up the convergence. The Gauss-Marquardt method combines the best features of the Taylor series methods and the gradient methods. There are two parts in the original optimization problem. The first part is the minimization of the maximum of the 50% signal propagation delay. The second part is the minimization of the maximum of the damping ratio error.

The following is the proof for the first part. Let τ_j be the 50% signal propagation delay of the receiver j $(1 \leq j \leq m)$, where m is the number of receivers. Let column vector $\Theta = \{\tau_1, \tau_2, \ldots, \tau_m\}^T$ represents the delay vector, where T denotes matrix transposition. The summation of all squares of delay errors is:

$$\Phi(w_1, w_2, \ldots, w_n) = \Theta^T \Theta = \sum_{j=1}^{m} \tau_j^2 \tag{4.1}$$

Define the root-mean-square (rms) error of the delay as:

$$\varphi = \sqrt{\frac{\Phi}{m}} = \sqrt{\sum_{j=1}^{m} \frac{\tau_j^2}{m}} \tag{4.2}$$

From Equation (3.1), the maximum delay is $F(w_1, w_2, \ldots, w_n) = Maximum(\tau_j)$. Theorem 1 shows the consistency of minimizing the delay and minimizing the rms delay error when optimization proceeds.

Theorem 1: *Given a single-source multi-receiver network, the root-mean-square error defined in Equation (4.2) and maximum delay defined in Equation (3.1) linearly bound each other.*

Proof: Assume the largest delay is τ_{max}. For all τ_j, $\tau_j \leq \tau_{max}$, $(1 \leq j \leq m)$, where m is the number of receivers and it is a constant for a given network. From Equation (4.2), we have $\varphi = \sqrt{\sum_{j=1}^{m} \tau_j^2/m} \leq \sqrt{\sum_{j=1}^{m} \tau_{max}^2/m} = \tau_{max} = F(w_1, w_2, \ldots, w_n)$. On the other hand, we have $F(w_1, w_2, \ldots, w_n) = Maximum(\tau_j) = \tau_{max} = \sqrt{\tau_{max}{}^2} \leq \sqrt{\sum_{j=1}^{m} \tau_j^2} = \sqrt{m \cdot \sum_{j=1}^{m} \tau_j^2/m} = \sqrt{m} \cdot \varphi$. So φ and $F(w_1, w_2, \ldots, w_n)$ linearly bound each other. □

The following is the proof of the second part. Let column vector $\Omega = \{|\zeta_1 - \zeta_{target}|, |\zeta_2 - \zeta_{target}|, \ldots, |\zeta_m - \zeta_{target}|\}^T$ represents the damping ratio error vector. The summation of all squares of damping ratio errors is:

$$\Psi(w_1, w_2, \ldots, w_n) = \Omega^T \Omega = \sum_{j=1}^{m} (\zeta_j - \zeta_{target})^2 \tag{4.3}$$

Define the root-mean-square (rms) error of the damping ratio as:

$$\psi = \sqrt{\frac{\Psi}{m}} = \sqrt{\sum_{j=1}^{m} \frac{(\zeta_j - \zeta_{target})^2}{m}} \tag{4.4}$$

From Equation (3.2), the maximum damping ratio error is $G(w_1, w_2, \ldots, w_n) = Maximum(|\zeta_j - \zeta_{target}|)$. Theorem 2 shows the consistency of minimizing the damping ratio errors and minimizing the rms damping ratio error when optimization proceeds.

Theorem 2: *Given a single-source multi-receiver network, the root-mean-square damping ratio error defined in Equation (4.4) and maximum damping ratio error defined in Equation (3.2) linearly bound each other.*

Proof: Assume the largest damping ratio is ζ_{max} and it deviates from ζ_{target} the most, the maximum damping ratio error is $|\zeta_{max}-\zeta_{target}|$. (The proof can be applied to the case where the smallest damping ratio ζ_{min} deviates from ζ_{target} the most. We only have to replace all ζ_{max} with ζ_{min} in the proof.) For all ζ_j, $|\zeta_j-\zeta_{target}| \leq |\zeta_{max}-\zeta_{target}|$, $(1 \leq j \leq m)$, where m is the number of receivers and it is a constant for a given network. From Equation (4.4), we have $\psi = \sqrt{\sum_{j=1}^{m}(\zeta_j - \zeta_{target})^2/m} \leq \sqrt{\sum_{j=1}^{m}(\zeta_{max} - \zeta_{target})^2/m} = |\zeta_{max} - \zeta_{target}| = Maximum(|\zeta_j - \zeta_{target}|) = G(w_1, w_2, \ldots, w_n)$. On the other hand, we have $G(w_1, w_2, \ldots, w_n) = Maximum(|\zeta_j - \zeta_{target}|) = |\zeta_{max} - \zeta_{target}| = \sqrt{(\zeta_{max} - \zeta_{target})^2} \leq \sqrt{\sum_{j=1}^{m}(\zeta_j - \zeta_{target})^2} = \sqrt{m \cdot \sum_{j=1}^{m}(\zeta_j - \zeta_{target})^2/m} = \sqrt{m} \cdot \psi$. So ψ and $G(w_1, w_2, \ldots, w_n)$ linearly bound each other. □

From the above two theorems, we have:

Theorem 3: *Given a single-source multi-receiver network, minimizing the two least-squares estimation problem minimizes the maximum path delay $F(w_1, w_2, \ldots, w_n)$ defined in Equation (3.1) and maximum damping ratio error $G(w_1, w_2, \ldots, w_n)$ defined in Equation (3.2) of the original nonlinear programming problem.*

4.2 Implementation of the Least-Squares Estimation Optimization

The Gauss-Marquardt method is used to solve the least-square estimation problem. Theorem 3 shows the consistency between the minimizing of the original problem and the minimizing of the transformed root-mean-square estimation problem. Starting with a arbitrary initial solution of width assignment $W^{(0)} = \{w_1{}^{(0)}, w_2{}^{(0)}, \ldots, w_n{}^{(0)}\}^T$, based on the Gauss-Marquardt method, the width assignment W is optimized according to the following formula:

$$W^{(k+1)} = W^{(k)} - (J^T J + \lambda\Lambda)^{-1} J^T \begin{bmatrix} \Theta|_{W^{(k)}} \\ \Omega|_{W^{(k)}} \end{bmatrix} \quad (4.5)$$

where k is the number of iteration, $\Theta|_{W^{(k)}}$ is the column vector of delays at all the receivers at the $k-th$ iteration, and $\Omega|_{W^{(k)}}$ is the column vector of damping ratio errors at all the receivers at the $k-th$ iteration. J is the $2m \times n$ sensitivity matrix, J^T is the transposition matrix of J where the

$(i,j)th$ element $J^T(i,j) = J(j,i)$, Λ is a diagonal matrix in which the values of its diagonal elements are the same as the diagonal elements of $J^T J$, and λ is the *Lagrange Multiplier* properly selected to speed up the convergence of the optimization process [11]. Round-off occurs when computing $w_i^{(k)}$ so that Equation (3.4) and (3.5) are always satisfied. The physical meaning of $J^T \begin{bmatrix} \Theta|_{W^{(k)}} \\ \Omega|_{W^{(k)}} \end{bmatrix}$ is that it represents the gradient around the current width assignment $W^{(k)}$. To obtain the sensitivity matrix J, the $(i,j)th$ element is defined as:

$$J(i,j) = \begin{cases} \frac{\partial \Theta[i]}{\partial w_j}, & \text{if } 1 \leq i \leq m \\ \frac{\partial \Omega[i-m]}{\partial w_j}, & \text{if } m+1 \leq i \leq 2m \end{cases} . \tag{4.6}$$

The partial derivatives are computed using central difference method. The optimization continues until the maximum damping ratio error is less than a prescribed value, the maximum damping ratio error cannot be further improved, the maximum delay cannot be further improved, or the iteration number exceeds a preset limit. The convergence to the optimal values of Gauss-Marquardt method is proved in [11].

5 Experiment Results

The examples tested are constructed with High Performance MCM process technologies published by Frye [4]. The important parameters of the MCM process are listed in Table 5.1. In the uniform width case, all the widths are equal to $25\mu m$ for all the examples tested. All the drivers are modeled with a step input voltage source in series with the parallel combination of a 12Ω resistor and a $4.3pF$ capacitor. All the receiver are modeled using a $2.5pF$ capacitor. The damping ratio target is chosen to be 0.8 or $\frac{1}{\sqrt{2}}$ for shorter propagation delay with a controlled amount of overshoot [1] [7].

6 S-Parameter Based Macro-Model with the Time-of-Flight Extraction

In order to precisely analyze the lossy transmission lines on MCM substrates, the scattering parameter (S-parameter) based macro-model [9] [10] is used to find the approximated transfer function $H(s)$. S-parameters represent the interrelationship of a set of incoming and outgoing waves of a multiport component. Some of the advantages of using S-parameters are:

1. They are easier to measure and to work with at high frequencies compared to using other types of parameters.

2. All lumped circuit elements including short and open circuits have unique analytical S-parameter descriptions.

3. They provide a very convenient means for describing distributed elements such as lossy transmission lines.

The S-parameter based macro-model simulator can handle both lumped circuit elements and lossy/lossless transmission lines including loops. With the description of elements by S-parameters, an efficient network reduction algorithm is employed to reduce the large interconnect network into one multi-port component together with the source and receiver circuit elements. Pade approximation [10] or Exponential Decaying Polynomial Function approximation [9] is used to derive the macro-model.

However, the delay associated with transmission line networks consists of the exponentially charging time and a pure propagation delay representing the finite propagating speed of electro-magnetic signals in the dielectric medium. This propagation delay, so called time-of-flight delay and denoted by τ, is impossible to model perfectly by a finite order of approximation. So, the time-of-flight τ, more precisely the factor $e^{-s\tau}$, must be extracted from the transfer function of the circuit.

The new and important improvement in the S-parameter based macro-model simulator is the extraction of the exact time-of-flight term in transfer functions [8]. By extracting the time-of-flight of scattering parameters for basic components, an effective network reduction is developed

Later Zhou et al. presents a distributed-RLC model and a second order approximation in their performance-driven-layout paper [12] which extends the wire-sizing algorithm to cover transmission lines on MCM. It is possible that the second order approximation to be used in order to keep the optimization manageable. Zhou et al. has published a paper [12] using the second order without the extraction of the time-of-flight to formulate the performance-driven-layout.

There are three problems with the performance-driven-layout described in Zhou et al. paper.

- The formulation of the optimization problem with the critical damping constrains is changed to an optimization without any constrain in implementation.
- The distributed-RLC model is used in the formulation but distributed-RC model is actually used in the optimization.
- The approach used by Zhou et al. is a two-step approach, first step is to perform the wire-sizing to achieve shorest delay, the second step is to critically damp the output signals.

The Zhou's critical damping design method consists of adding resistors in series at the receiving terminals. This adding of serial resistor at the receiver is like adding the matching serial termination resistor at the driver, both consume power when the line is at non-zero voltage and take extra space to accommodate the resistors.

8 Concluding Remarks

This paper is the first to identify the performance-driven-layout optimal objective is to minimize both the maximum path delay and the maximum damping ratio together. The optimal performance is achieved *entirely* through wire-sizing on any general network which may contains *loops*. The use the new S-parameter based macro-model which keeps track of the exact time-of-flight helps in producing accurate path delays. The results in Section 5 shows significant improvements over the uniform typical width assignment designs.

Due to the limited space and heat dissipation problem, the interconnections on multichip modules are better left unterminated. The metal and dielectric film thickness commonly used in the fabrication of a multichip module is well suited for the design of optimal, self-damped interconnections. This optimal design method uses wire-sizing entirely which eliminating the excessive heat dissipation that accompanies with resistive terminators.

Future research includes taking cross coupling, frequency dependent models, and 3-D discontinuity (vias and bends) into consideration, and verification of the results through measurements.

9 Acknowledgment

This work is supported in part by the National Science Foundation Presidential Young Investigator Award under the Grant MIP-9009945, , and in part by IBM Corp.

References

[1] John. R. Brews. Overshoot-controlled RLC interconnections. *IEEE Trans. on Electron Device*, ED-38:76–87, January 1991.

[2] Jason Cong, Kwok-Shing Leung, and Dian Zhou. Performance-driven interconnect design based on distributed RC delay model. Technical Report CSD-920043, University of Clifornia, Los Angeles, Oct. 1992.

[3] R. C. Frye and K. L. Tai. Interconnection lines for wafer-scale-integrated assemblies. *U. S. Patent 4,703,288*, 1987.

[4] Robert C. Frye. Physical scaling and interconnection delays in multichip module. To Be Published, 1993.

[5] Robert C. Frye and Howard Z. Chen. High speed interconnection using self-damped lossy transmission lines. In *Symposium on High Density Integration in Comunication and Computer Systems*, pages 36–37, 1991.

[6] C. W. Ho, D. A. Chance, C. H. Bajorek, and R. E. Acosta. The thin-film module as a high performance semiconductor package. *IBM Journal Res. Develop*, 26(286), 1982.

[7] Benjamin C. Kuo. *Automatic Control Systems.* Prentice-Hall, Inc., third edition, 1975.

[8] Haifang Liao and Wayne Dai. Extracting time-of-flight from scatering-parameter based macro-model. Technical Report 93-35, University of Clifornia, Santa Cruz, 1993.

[9] Haifang Liao, Wayne Dai, Rui Wang, and F. Y. Chang. Scatering-parameter based macro model of distributed-lumped networks using expnontially decayed polynomial function. In *Proc. of 30th ACM/IEEE Design Automation Conference*, pages 726–731, 1993.

[10] Haifang Liao, Rui Wang, Wayne Dai, and R. Chandra. Scatering-parameter based macro model of distributed-lumped networks using pade approximation. In *Proc. of International Symposium on Circuits and Systems*, 1993.

[11] D. W. Marquardt. An algorithm for least-squares estimation of nonlinear parameters. *Journal of Society of Indust. Appl. Math.*, 11:431–441, 1963.

[12] D. Zhou, F. Tsui, J. S. Cong, and D. S. Gao. A distributed-RLC model for MCM layout. In *Proc. of IEEE Multi-Chip Module Conference MCMC-93*, pages 191–197, 1993.

[13] Qing Zhu and Wayne Dai. Optimal sizing of high speed clock networks based on distributed RC and lossy transmission line models. In *IEEE/ACM International Conference on Computer-Aided Design*, 1993.