ELECTROMAGNETIC OPTIMIZATION OF SIGNAL INTEGRITY IN HIGH-SPEED DIGITAL CIRCUITS USING MULTIPLE SPACE MAPPING

J.W. Bandler, R.M. Biernacki,
S.H. Chen and Q.H. Wang

OSA-96-MT-24-R

December 2, 1996

ELECTROMAGNETIC OPTIMIZATION OF SIGNAL INTEGRITY IN HIGH-SPEED DIGITAL CIRCUITS USING MULTIPLE SPACE MAPPING

J.W. Bandler, R.M. Biernacki, S.H. Chen and Q.H. Wang*

Optimization Systems Associates Inc.
P.O. Box 8083, Dundas, Ontario
Canada L9H 5E7

Email osa@osacad.com URL http://www.osacad.com

Abstract

Electromagnetic (EM) optimization of crosstalk and signal delay for design of high-speed digital circuits is, for the first time, handled by Space Mapping. The LC matrices of coupled interconnects, needed for time-domain circuit-level simulations, are determined by employing both an empirical model and EM simulations. A new concept of Multiple Space Mapping is developed to align both models. The mappings are then used for design optimization which is carried out with the speed of the empirical model and the accuracy of the EM model.

This work was supported in part by the Simulation Optimization Systems Research Laboratory and Department of Electrical and Computer Engineering, McMaster University, and in part by the Natural Sciences and Engineering Research Council of Canada under Grants OGP0007239, OGP0042444 and STR0167080 and through the Micronet Network of Centres of Excellence.

* Q.H. Wang is with the Simulation Optimization Systems Research Laboratory and Department of Electrical and Computer Engineering, McMaster University, Hamilton, Ontario, Canada L8S 4L7.

Manuscript submitted December 2, 1996.

SUMMARY

Introduction

The increasing speed of digital circuits and close proximity of circuit components and interconnects at both the printed circuit board (PCB) and multi-chip module (MCM) levels call for accurate analog simulations. Recent advances in field level electromagnetic (EM) simulation, the availability and increasing practicality of EM simulators, and the awareness of problems due to couplings, radiation, etc., are the main reasons why high-speed digital circuit designers are increasingly interested in EM simulation software.

In this paper we present new results for crosstalk and delay minimization for VLSI design. Following recent exciting developments in microwave circuit design [1,2], our goal is to integrate EM simulators for automated interconnect design. To this end we apply Space Mapping [3] which is particularly suitable for designs involving CPU intensive simulators.

A novel concept which we call Multiple Space Mapping is formulated to align empirical models with accurate EM simulations. This concept is applied to the modeling of a two-conductor interconnect by creating one mapping targeting crosstalk and another mapping targeting signal delay. The mappings are then used for design optimization of a test circuit.

Time-Domain Simulations for Signal Integrity

At high operation speeds the interconnect model to consider consists of two coupled transmission lines, as shown in Fig.1. Analog time-domain simulations of digital circuits involving such interconnects are carried out in two steps (e.g., [4-6]). First, the LC matrices (per unit length) of the coupled transmission lines are calculated. Then these matrices together with all other circuit elements are used in a circuit-level time-domain simulator, such as SPICE. In our work we use the AWE time-domain circuit-level simulator COFFEE2 [4,5] to evaluate various circuit responses, including crosstalk. This simulation scheme clearly involves certain intermediate parameters (here, the LC matrices). These intermediate parameters play a special role in the Multiple Space Mapping approach proposed in this paper.

Empirical and EM models of the Interconnects

The high frequency of operation and close proximity of the interconnects call for the modeling of effects such as couplings. We consider two ways of simulating the coupled transmission lines: (1) a set of empirical formulas (see, Walker, [7]), and (2) EM simulations using the ***em*** simulator from Sonnet Software [8].

Sonnet's ***em*** can directly output the LC matrices with the option "-dX". However, to obtain the per-unit-length LC matrices, one needs to divide the ***em*** results by the de-embedded line length, which must be small compared with the wavelength. An alternative way is to postprocess the S parameters calculated by ***em***. The latter approach was adopted in this work.

Following Walker [7], the empirical formulas for the LC parameters are

$$L^{(s)} \approx \frac{\mu_r \mu_0}{K_{L1}} \left(\frac{h}{w} \right) \tag{1}$$

$$L^{(m)} \approx \frac{\mu_r \mu_0}{4\pi} \ln \left[1 + \left(\frac{2h}{d} \right)^2 \right] \tag{2}$$

$$C^{(s)} \approx \varepsilon_r \varepsilon_0 K_{C1} \left(\frac{w}{h} \right) \tag{3}$$

$$C^{(m)} \approx \frac{\epsilon_r \epsilon_0}{4\pi} K_{C1} K_{L1} \left(\frac{w}{h} \right)^2 \ln \left[1 + \left(\frac{2h}{d} \right)^2 \right] \tag{4}$$

$$K_{L1} = \frac{120\pi}{Z_{0(\epsilon_r=1)}} \left(\frac{h}{w} \right) \tag{5}$$

$$K_{C1} = \left[\frac{120\pi}{Z_{0(\epsilon_r=1)}} \sqrt{\frac{\epsilon_{r(eff)}}{K_{L1}\epsilon_r}} \left(\frac{h}{w} \right) \right]^2 \tag{6}$$

$$Z_{0(\epsilon_r=1)} \approx 60 \ln \left(\frac{8h}{w} + \frac{w}{4h} \right) \qquad for \frac{w}{h} \le 1 \tag{7}$$

$$Z_{0(\epsilon_r=1)} \approx \frac{120\pi}{\left(\frac{w}{h} \right) + 2.42 - 0.44 \left(\frac{h}{w} \right) + \left(1 - \frac{h}{w} \right)^6} \qquad for \frac{w}{h} \ge 1 \tag{8}$$

where the superscripts s and m indicate the self and mutual elements, respectively.

The per-unit-length LC matrices of the coupled interconnects (transmission lines), as calculated using (1)-(8) and obtained from simulations by ***em***, do not quite agree with each other [9]. This is illustrated in Fig. 2 in the case of C_{12} ($C_{12} = -C^{(m)}$). Simulations by ***em*** are considered to be accurate and hence are preferred as inputs to the circuit-level simulator. The empirical formulas on the other hand allow for extremely fast calculation of the LC parameters, and are therefore desirable for inclusion in repeated simulations during optimization.

The Concept of Multiple Space Mapping

Space Mapping [3] establishes a mathematical link between models of different complexity and accuracy. In the context of this paper two models are considered: Walker's formulas (a fast, "coarse" model) and simulations by ***em*** (fine model, accurate but slower). The bulk of CPU intensive computations is directed to the coarse model. Then, in order to align the two models, we create and iteratively refine a mapping from the EM parameter space to the empirical model parameters.

We extend this concept to Multiple Space Mapping. While the mathematical details are left out of this summary, the concept can be expressed as follows. Consider the primary parameters $\boldsymbol{\phi}_{em}$ and $\boldsymbol{\phi}_{os}$ of the EM and empirical models, respectively, and a number of responses grouped into N subsets (or vectors) $\boldsymbol{R}^k$, $k = 1, 2, \ldots N$. The responses are not limited to circuit performance functions, but they may also represent some intermediate parameters. Applying Space Mapping to these subsets of responses we establish N different mappings, each targeting one individual $\boldsymbol{R}^k$. This is illustrated by Fig. 3.

The mappings are then incorporated into optimization in the following fashion. For every new point $\boldsymbol{\phi}_{em}$ determined by the optimizer we find N mapped points $\boldsymbol{\phi}_{os}{}^k$ and then N simulations of the coarse model are carried out, each for a specific $\boldsymbol{\phi}_{os}{}^k$ and only to evaluate the corresponding $\boldsymbol{R}^k$. Finally, the overall responses are assembled from those partial responses. In this scheme, it is important to assure larger validity regions for the mappings (quasi-global models).

Results of Multiple Space Mapping Optimization

We consider simultaneous crosstalk and delay (distortion) optimization in the circuit shown in Fig. 4 [10]. Applying a 6 ns trapezoidal voltage shown in Fig. 5 to the circuit input we evaluate the direct output signal V_{out} and the crosstalk signal V_{cross}.

First, two mappings are established between the parameters d, w, h and ϵ_r of the empirical model and those used in the simulations by ***em***. The first mapping targets the direct signal while the second mapping targets crosstalk. The responses $\boldsymbol{R}^k$ considered in establishing the mappings are the LC parameters: the first group consists of the self inductance and the self capacitance, while the second group consists of the mutual inductance and the mutual capacitance.

The following design specifications are considered:

$$V_{out}(t) > 0.95 \quad \text{at} \quad t = 1.5 \text{ ns}$$

$$-0.0025 < V_{cross} < 0.0025$$

All lumped component values and the parameters of the coupled lines d, w, h and ϵ_r are considered as the optimization variables (d and w are allowed to be different for each line while h and ϵ_r are kept common for all lines).

Minimax optimization was carried out by OSA90/hope [11] using Walker's formulas and the mappings directly coded in its input file. COFFEE2 was driven by OSA90/hope through a Datapipe [11] connection. Sonnet ***em*** simulations (S parameters) for establishing the mappings as well as for final verification of the results were performed using Empipe [11]. The circuit responses before and after optimization are shown in Figs. 6 and 7.

Conclusions

In the context of combining the respective advantages of EM simulations and empirical formulas for interconnect design we developed a novel Multiple Space Mapping technique. It has been successfully applied to simultaneous crosstalk and delay optimization. The optimization is then carried out with the speed of the empirical model while retaining the accuracy of EM simulations.

References

[1] J.W. Bandler, R.M. Biernacki, S.H. Chen, D.G. Swanson, Jr., and S. Ye, "Microstrip filter design using direct EM field simulation," *IEEE Trans. Microwave Theory Tech.*, vol. 42, 1994, pp. 1353-1359.

[2] Workshop WMFE, *Automated Circuit Design using Electromagnetic Simulators.* 1995 IEEE MTT-S Int. Microwave Symposium (Orlando, FL), 1995.

[3] J.W. Bandler, R.M. Biernacki, S.H. Chen, P.A. Grobelny and R.H. Hemmers, "Space mapping technique for electromagnetic optimization," *IEEE Trans. Microwave Theory Tech.*, vol. 42, 1994, pp. 2536-2544.

[4] R. Griffith, E. Chiprout, Q.J. Zhang and M.S. Nakhla, "A CAD framework for simulation and optimization of high-speed VLSI interconnections," *IEEE Trans. Circuits Syst.*, vol. 39, 1992, pp. 893-906.

[5] E. Chiprout and M.S. Nakhla, *Asymptotic Waveform Evaluation and Moment Matching for Interconnect Analysis.* Boston, MA: Kluwer, 1994.

[6] Q.J. Zhang and M.S. Nakhla, "Statistical simulation and optimization of VLSI interconnects," *Int. J. Analog Integrated Circuits and Signal Processing*, Special Issue on High-Speed Interconnects, vol. 5, 1994, pp. 95-106.

[7] C.S. Walker, *Capacitance, Inductance, and Crosstalk Analysis.* Norwood, MA: Artech House, 1990.

[8] ***em***™ and ***xgeom***™, Sonnet Software, Inc., 1020 Seventh North Street, Suite 210, Liverpool, NY 13088.

[9] J.W. Bandler, R.M. Biernacki and S.H. Chen, "Toward direct EM optimization of VLSI interconnects: validation of coupled transmission line models," *Proc. 1995 Canadian Conf. on Electrical and Computer Engineering* (Montreal, Canada), 1995, pp. 377-380.

[10] S. Lum, M.S. Nakhla and Q.J. Zhang, "Sensitivity analysis of lossy coupled transmission lines," *IEEE Trans. Microwave Theory Tech.*, vol. 39, 1991, pp. 2089-2099.

[11] *OSA90/hope*™ and *Empipe*™, Optimization Systems Associates Inc., P.O. Box 8083, Dundas, Ontario, Canada L9H 5E7.

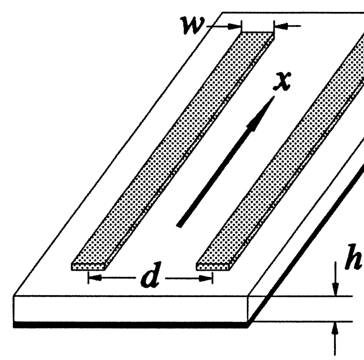

Fig. 1. Two-conductor coupled transmission lines.

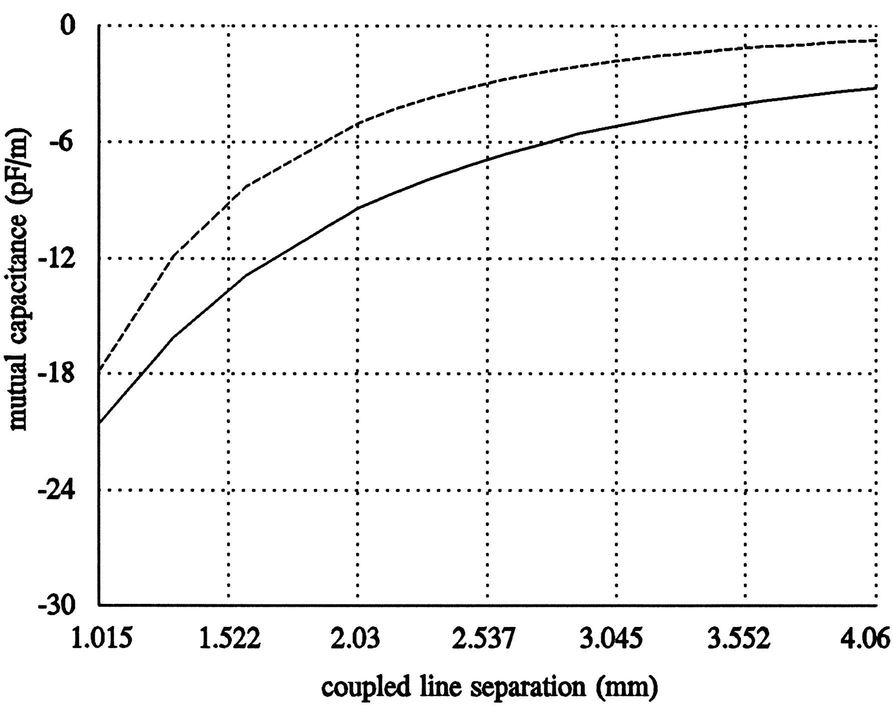

Fig. 2. C_{12} calculated from Sonnet *em* simulations (----) and the empirical formulas (3)-(8) (——).

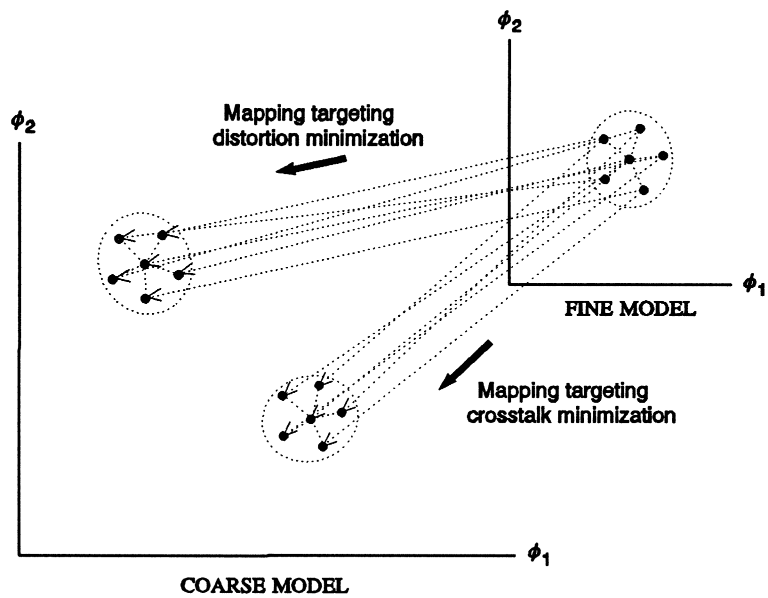

Fig. 3. Multiple mappings between the fine and coarse models .

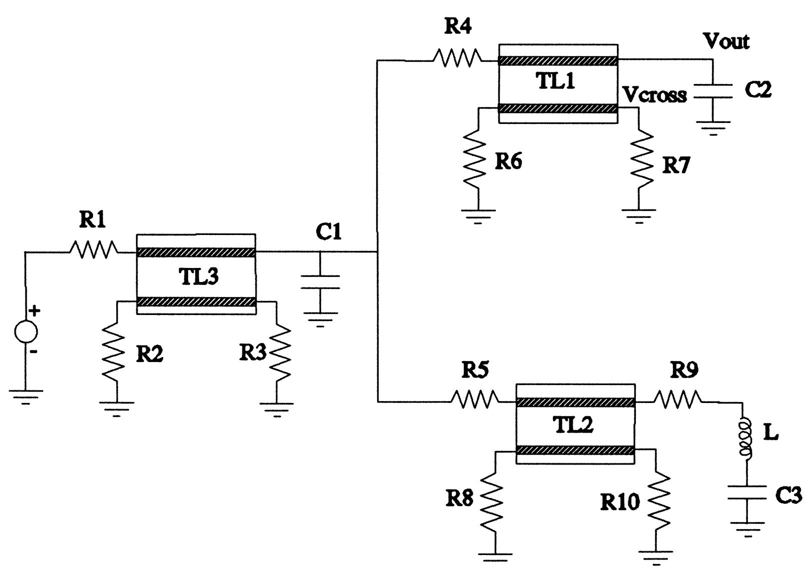

Fig. 4. The interconnect circuit for crosstalk and delay optimization.

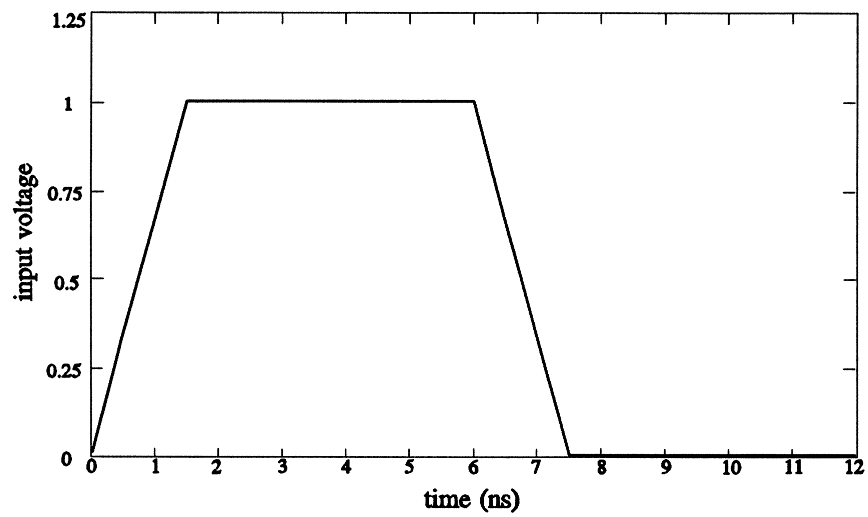

Fig. 5. Trapezoidal input signal.

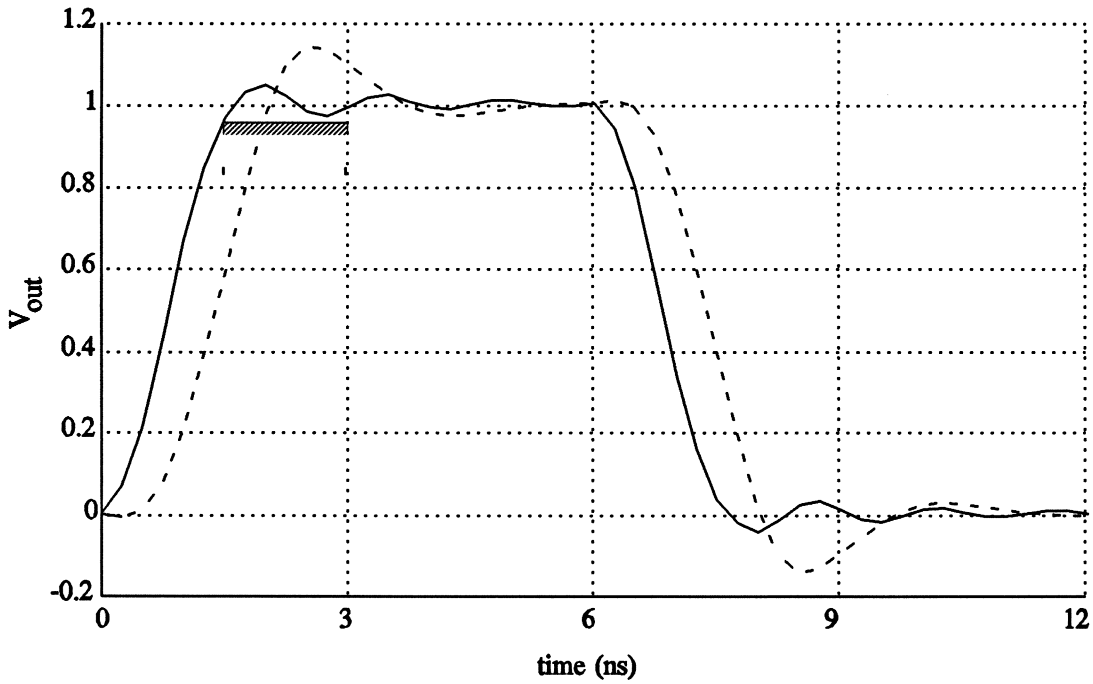

Fig. 6 Direct signal output V_{out} before (----) and after (——) optimization.

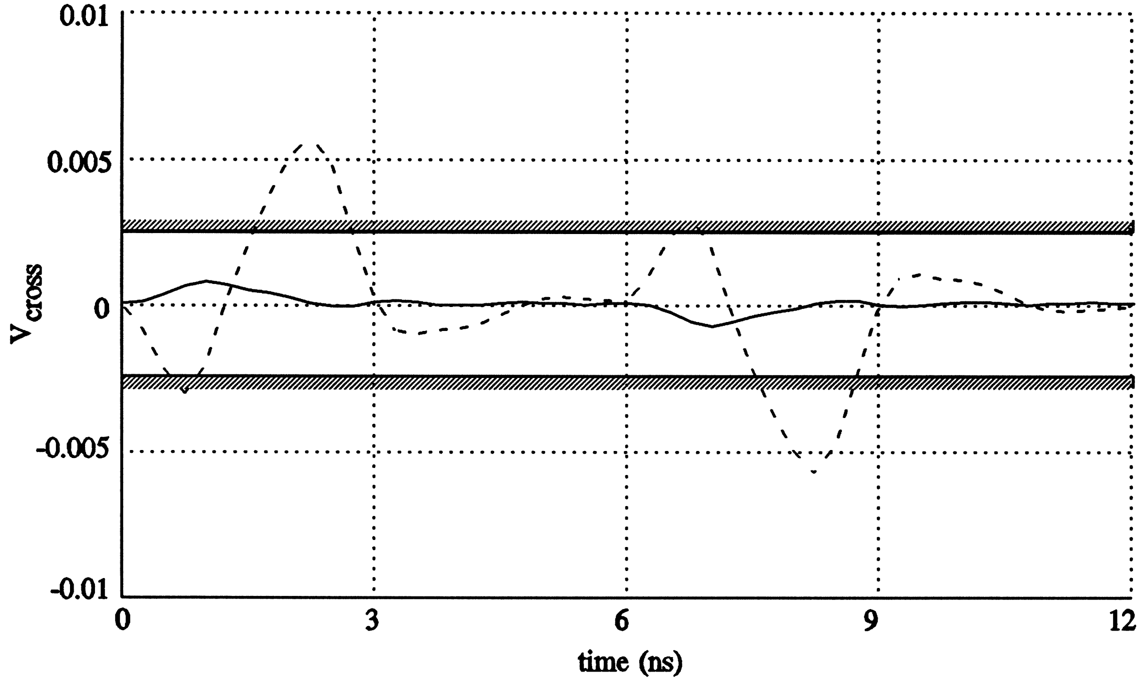

Fig. 7. Crosstalk waveforms V_{cross} before (----) and after (——) optimization.