

**TOWARD DIRECT EM OPTIMIZATION OF
VLSI INTERCONNECTS: VALIDATION
OF COUPLED TRANSMISSION LINE MODELS**

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Toward Direct EM Optimization of VLSI Interconnects: Validation of Coupled Transmission Line Models

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Abstract— With the goal of direct electromagnetic (EM) optimization we present results relevant to crosstalk analysis of VLSI interconnects. We employ EM simulators to extract the LC matrices of coupled interconnects to be used in a circuit level simulator. The results are used to validate a set of empirical formulas. The recent breakthrough concept of Space Mapping optimization linking EM simulations with empirical formulas is outlined.

I. INTRODUCTION

The increasing speed of VLSI 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 for the significant shift in interest of high-speed VLSI circuit designers towards EM simulation software.

The use of EM simulation tools, however, must go beyond the validation of designs obtained by traditional techniques. Following recent exciting developments in microwave circuit design [1-3], our goal here is to integrate EM simulators into a unified optimization environment for automated VLSI interconnect design.

In this paper we present results relevant to crosstalk analysis and minimization for VLSI design. We employ EM simulators to extract the LC matrices of coupled interconnects (transmission lines). These matrices are then used in the circuit level simulator COFFEE2 [4,5] to evaluate various circuit responses, including crosstalk. That simulator has already been interfaced [6] to our optimizers [7].

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II. MODELING THE INTERCONNECTS

High operation speed, thus high frequency, and close proximity of the interconnects call for modeling of such effects as couplings. The model to consider consists of two coupled transmission lines, as shown in Fig.1.

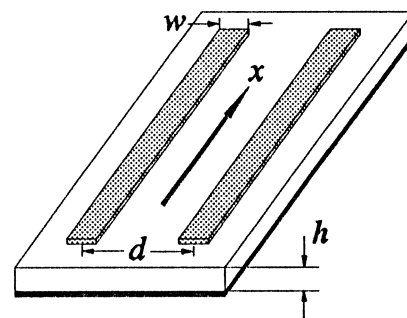


Fig. 1 Two-conductor coupled transmission lines.

In this work we consider three ways of simulating the coupled transmission lines:

- (1) a set of empirical formulas (Walker, [8]),
- (2) EM simulation using the *em* simulator from Sonnet Software, Inc. [9], and
- (3) EM simulation using the IE3D simulator from Zeland Software, Inc. [10].

In all three cases the per-unit-length LC matrices of the coupled interconnects (transmission lines) are evaluated for given values of physical parameters. Such results are suitable for incorporating them into time-domain circuit-level simulators.

III. VALIDATION OF EMPIRICAL FORMULAS

The empirical formulas allow for extremely fast calculation of the per-unit-length LC parameters for the coupled lines, in fact several orders of magnitude faster than EM simulations. However, they have limited validity and accuracy. EM simulations, considered as the most accurate modeling available, are used here to validate the empirical formulas.

With the variety of available EM simulators we had no preference as to the method, or a particular field solver. We have thus cross-checked the results produced by two commercial EM simulators: *em* and IE3D. These results are then compared with those produced by the empirical formulas.

Following Walker [8], the empirical formulas have been directly coded in the OSA90/hope [7] input file using the mathematical expression features of the system. 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.

Zeland's IE3D cannot directly produce the LC matrices. The MGRID subprogram is used first to calculate the S parameters. The S parameters are then postprocessed by the MODUA subprogram to generate an LC equivalent circuit. Finally, the parameters are manually convert into the per-unit-length LC matrices. The resulting LC matrices from both *em* and IE3D were imported into OSA90/hope to generate other results.

Table I summarizes a direct comparison of the per-unit-length LC matrices. Table II provides the percentage differences of the results obtained from EM simulations w.r.t. the empirical formulas.

TABLE I

COMPARISON OF THE CALCULATED LC MATRICES

	Walker	<i>em</i>	IE3D
L_{11} (nH/m)	494.5	510.7	523.4
L_{12} (nH/m)	63.29	58.67	62.70
C_{11} (pF/m)	69.97	57.52	61.06
C_{12} (pF/m)	-7.13	-3.11	-3.06

TABLE II

PERCENTAGE DIFFERENCES

	Walker	<i>em</i>	IE3D
L_{11} (%)	0	3.3	5.8
L_{12} (%)	0	-7.3	-0.9
C_{11} (%)	0	-17.8	-12.7
C_{12} (%)	0	-56.4	-57.1

From Tables II and II we can see that all the results are quite consistent.

It is important to notice that certain parameters may influence the consistency of the results. One such parameter is the conductor metal thickness. It does not appear in the empirical formulas. Also, *em* assumes infinitely thin metallization. Only IE3D allows the user to directly specify the metal thickness and Table III shows its influence on the resulting LC matrices.

TABLE III

PERCENTAGE DIFFERENCES BETWEEN RESULTS BY WALKER'S FORMULAS AND IE3D

	Walker	IE3D metal thickness	
		0.1 mm	0.58 mm
L_{11} (%)	0	5.8	-30.0
L_{12} (%)	0	-0.9	30.7
C_{11} (%)	0	-12.7	1.1
C_{12} (%)	0	-57.1	-16.8

We have also compared the models over a range of parameter values. For example, Fig. 2 shows the mutual inductance L_{12} and the mutual capacitance C_{12} versus the line spacing.

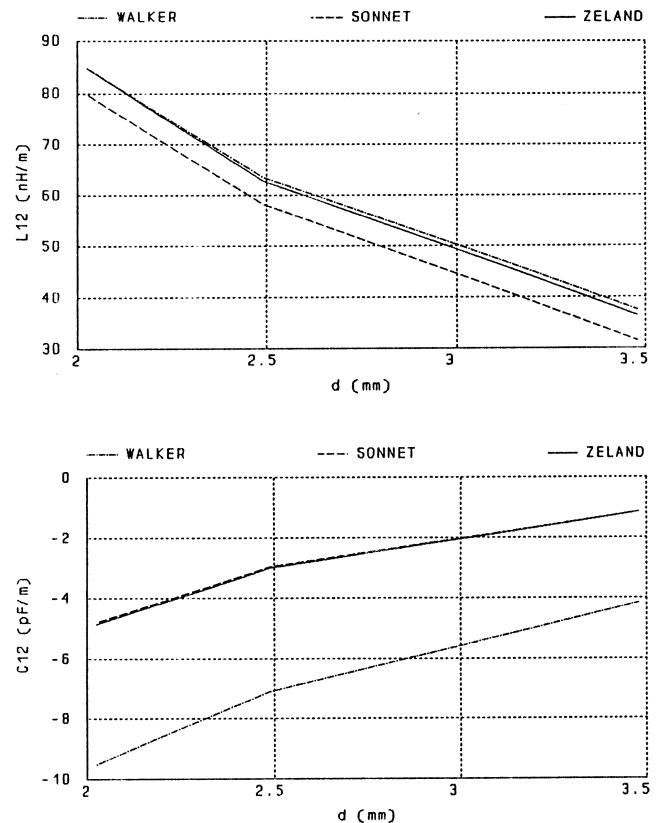


Fig. 2 L_{12} and C_{12} vs. the line spacing.

Similar results were obtained for the line width and the substrate thickness. The largest discrepancy appears when varying the line width (see Fig. 3).

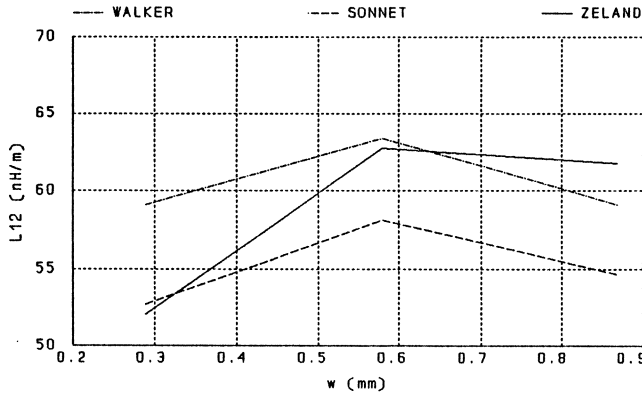


Fig. 3 L_{12} vs. the line width.

IV. CROSSTALK ANALYSIS

The differences between the LC matrices obtained by the three approaches considered can be better understood by comparing the responses of a more complex circuit embedding the coupled transmission lines as subcircuits.

We consider crosstalk analysis of the circuit shown in Fig. 4 (Lum, Nakhla and Zhang [11]).

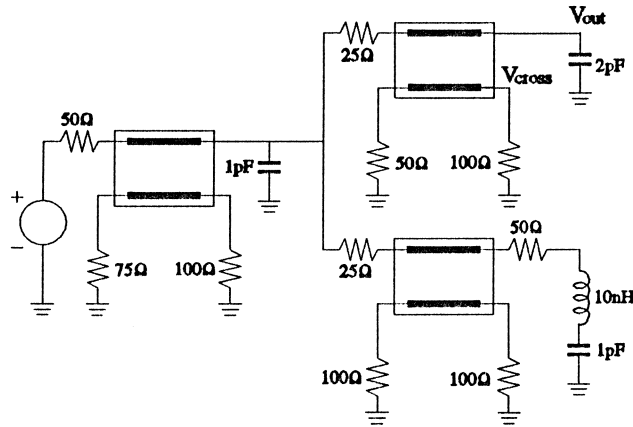


Fig. 4 The interconnect circuit for model comparison.

The LC matrices determined from the empirical formulas and from EM simulations are incorporated into the AWE time-domain circuit-level simulator COFFEE2 [5] developed at Carleton University. The simulator is driven by OSA90/hope through a Datapipe [7] connection. In this manner it is ready for optimization controlled by OSA90/hope.

Applying a 6 ns trapezoidal voltage shown in Fig. 5 to the circuit input we evaluate various responses.

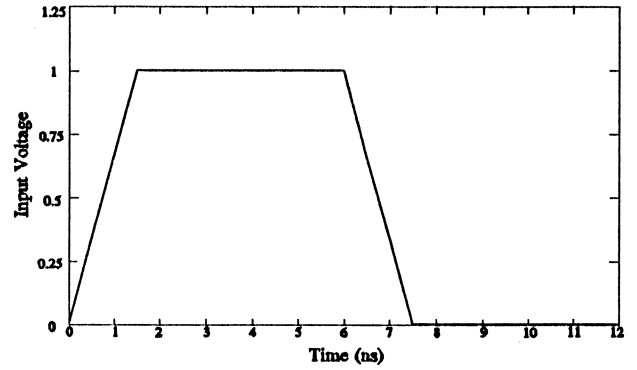


Fig. 5 Trapezoidal input signal.

Three voltage waveforms at the first output port corresponding to the three coupled-line models are shown in Fig. 6. They appear to be almost identical.

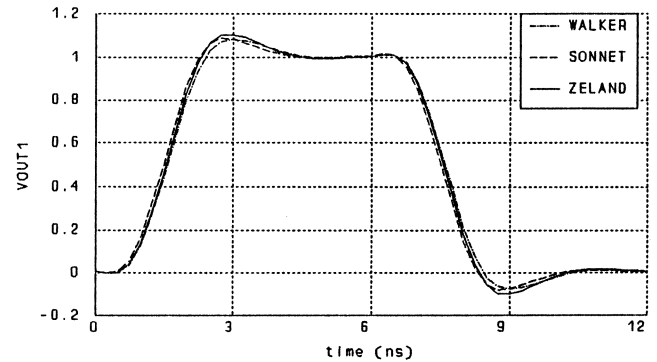


Fig. 6 Voltage waveforms at the first output port.

The corresponding crosstalk voltage waveforms at the first output port are shown in Fig. 7. In this case we can notice a larger difference between the results. The responses corresponding to both EM simulations are fairly consistent while the response generated by the empirical formulas differs from the other two by up to 50%.

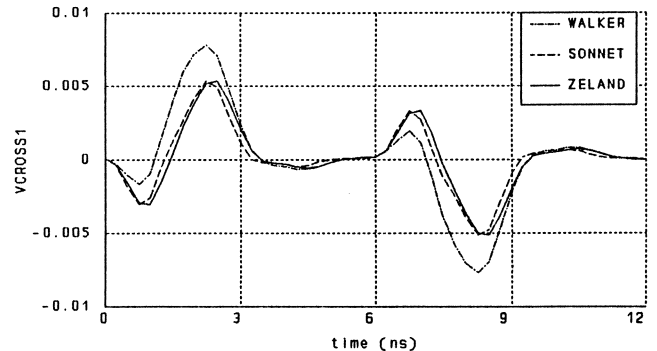


Fig. 7 Crosstalk waveforms at the first output port.

V. SPACE MAPPING OPTIMIZATION

Direct EM optimization has already been well advanced in the area of microwave and millimeter-wave circuits [1-3] addressing problems such as response interpolation due to fixed grids, databases, gradient evaluation, etc. The recent breakthrough technique of Space Mapping optimization [12,13] is particularly suitable for designs involving CPU intensive simulators. Space Mapping establishes a mathematical link between models of different complexity and accuracy. In the context of this paper two models can be considered: the empirical formulas (a faster model) and EM simulations (an accurate but slower model).

In SM optimization we carry out conventional design optimization with the empirical model. The bulk of CPU intensive computations is thus directed to the faster model. Then, in order to align the two models, we create and iteratively refine a mapping from the EM simulator parameter space to the empirical model parameters. Once the mapping is established the inverse mapping is used to find the EM solution as the image of the optimal solution. In other words, we map the optimal empirical model back into the EM parameters.

VI. CONCLUSIONS

EM simulations have been used to validate a set of empirical formulas for calculating the per-unit-length LC parameters of coupled interconnects. The resulting LC matrices are incorporated into a time-domain circuit-level simulator to carry out crosstalk analysis of a circuit embedding as subcircuits such coupled interconnects.

Comparison of the results shows a very good agreement between EM simulations obtained using two commercially available software packages: *em* and IE3D. The results based on the empirical formulas were fairly consistent with EM simulations. They were closer for the voltages at the output ports directly linked to the input. In the case of crosstalk the difference was larger.

In the context of combining the respective advantages of EM simulations and the empirical formulas for interconnect design we described the recent breakthrough concept of Space Mapping optimization. It has been successfully applied to microwave circuit design problems. In the follow-up to this work it will be expanded to link the EM simulations and the empirical formulas considered. The goal is to use the empirical formulas in the design optimization process as much as possible while retaining the accuracy of EM simulations.

VII. ACKNOWLEDGEMENTS

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