AN ACCURATE, COMPUTATIONALLY EFFICIENT CROSSTALK MODEL FOR ROUTING HIGH-SPEED MCMS

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ABSTRACT

A crosstalk model that is both accurate and easily incorporated into an MCM router has been developed. This model accounts for coupling from multiple layers out to three wires in either direction and accounts for the shielding involved with those wires. An algorithm that calculates the effect of reflections from terminals with mismatched impedances is also included as part of the model. Finally, it is shown to be impractical to use a more complex crosstalk model without including some sort of timing simulator.

I. INTRODUCTION

With the ever increasing clock speeds and decreasing feature sizes on integrated circuits, noise considerations when routing multichip modules are becoming increasingly more important. To compound the noise problem, lower voltages are being used in newer circuits resulting in significantly smaller noise margins.

Most of the existing MCM routers are modifications of printed circuit board layout tools. These routers work fine when noise is not a consideration; however, nearly all modern designs need to account for induced noise.

The existing MCM routers that do account for noise use greatly oversimplified crosstalk models. These models are slightly better than totally ignoring crosstalk. The typical model used in existing routers either blindly limits the parallel path length of one net with another net, or imposes a conservative spacing rule on the design.

An accurate crosstalk model is important for two reasons. The first and most important reason is that a crosstalk violation can render a design useless and can require a significant amount of time and money to correct. The second reason is that an overly conservative model can add layers to a design, resulting in increased cost and decreased performance.

II. CROSSTALK CALCULATIONS

When a voltage is induced in a line, two voltage pulses (traveling in opposite directions) are created. The voltage pulse that travels towards the driver end of the driven line is called backward or near-end crosstalk. The pulse that travels away from the driver end is called forward or far-end crosstalk. The two pulses are different in both shape and magnitude so separate equations are needed for each pulse. The following equations were developed by Feller et. al. in [3] and are also discussed in [6].

The near-end crosstalk is given by:

$$V_{NE}(t) = K_{NE}\left[V_{in}(t) - V_{in}(t - 2t_d)\right] \qquad (1)$$

where $V_{in}(t)$ is the input voltage, t_d is the transit time for the signal to cross the coupled region, and K_{NE} is the near-end coupling coefficient. K_{NE} is given by:

$$K_{NE} = \frac{1}{4t_d}\left(\frac{L_m}{Z_o} + C_m Z_o\right) \qquad (2)$$

where L_m is the mutual inductance, C_m is the mutual capacitance, and Z_o is the characteristic impedance of the lines.

The above equations are valid for both long and short lines. However, the peak value of the crosstalk depends on the length of the coupled region. The peak value is:

$$V_{peak} = \begin{Bmatrix} K_{NE}V_o, & for\ t_r \le 2t_d \\ 2t_d K_{NE}\dfrac{V_o}{t_r}, & for\ t_r \ge 2t_d \end{Bmatrix} \qquad (3)$$

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where t_r is the rise time of the input pulse and V_o is the peak input voltage. The coupled length where $t_r = 2t_d$ is called the saturation length. The near-end crosstalk does not increase after the saturation length has been reach between two nets. Since V_{NE} has a saturation value, it is important to monitor the total crosstalk between the two nets. The total near-end crosstalk between the two nets cannot exceed $K_{NE}V_o$.

The far-end crosstalk is given by:

$$V_{FE}(t) = K_{FE}\ell\frac{d}{dt}\left[V_{in}(t - t_d)\right] \qquad (4)$$

where K_{FE} is the far-end coupling coefficient. K_{FE} is given by:

$$K_{FE} = -\frac{1}{2}\left(\frac{L_m}{Z_o} - C_mZ_o\right) \qquad (5)$$

The peak crosstalk voltage is given by:

$$V_{peak} = K_{FE}\ell\frac{V_o}{t_r} \qquad (6)$$

where ℓ is the length of the coupled region.

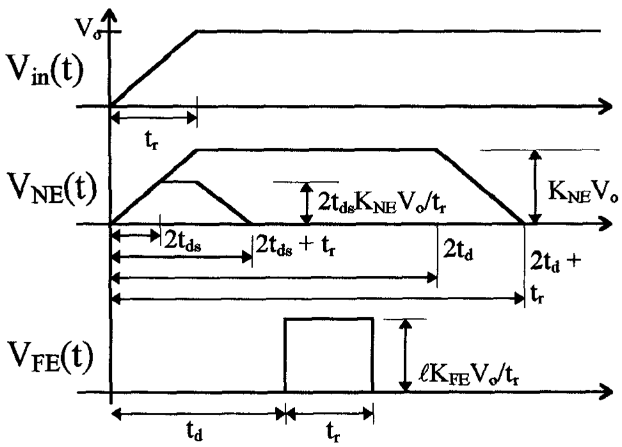


Figure 1. Primitive Crosstalk Pulses [6].

Figure 1 shows the wave forms of both the backward and forward crosstalk pulses. $V_{NE}(t)$ has two possible pulses for the reasons discussed above. The smaller pulse is for short lines. The term t_{ds} represents t_d of the short line. The polarity of $V_{FE}(t)$ is assumed to be positive in the chart but is not necessarily the case.

Usually, only the nearest neighbor of any particular conductor needs to be considered when calculating crosstalk, because at low frequencies the effects of other conductors tend to be negligible. However, for very high speed MCM systems this may not be true. Consider the bus in Figure 2. Line 3 is quiet while the

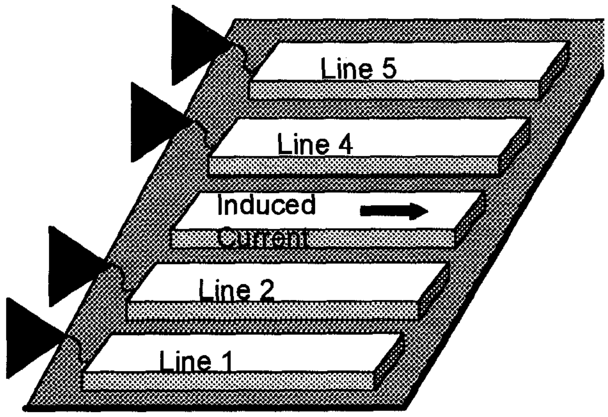


Figure 2. Five Conductor Bus.

other four lines are driven. Other crosstalk-driven auto-routers simply ignore the effects from *next-nearest-neighbors* (lines 1 and 5).

In order to make the crosstalk calculations as accurate as possible, it was necessary to include not only the next-nearest-neighbors like lines 1 and 5, but also the *next-next-nearest-neighbors* like lines 0 and 6 (if they were shown) in the crosstalk calculations of this router. In many circuits, the coupling of lines that far removed is not necessary, and the loss in router performance would be greater than the value of the increased accuracy of the crosstalk calculations. For this reason, the coupling limits within the router described here, are *dynamically* determined at runtime based on the crosstalk coefficients, the input voltages, the size of the design, and the noise margins. The coupling limits for lines that are parallel, but on different layers, are also dynamically determined. Dynamically determining the coupling limits helps balance the trade-offs between accuracy and speed.

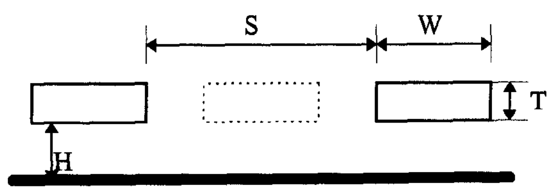


Figure 3. Wiring Geometry.

One problem with trying to calculate the crosstalk between two lines that have another line in between them is that the line in the middle partially shields them from each other (as in Figure 3). In order to determine the amount of shielding, we start with the relationship $C_m \propto L_m$ for a given frequency. From this relationship and the equations above, it can be shown that $V_{NE} \propto C_m$ and $V_{FE} \propto C_m$. Using the mutual capacitance approximation by Sakurai in [5] and assuming that the $0.07(T/H)^{0.222}$ term is negligible, the following ratio can be derived:

$$\frac{C_{mwo}}{C_{mw}} = \frac{\varepsilon_{wo}}{\varepsilon_w}\left[1 + 27.27\frac{T}{H}\right] \tag{7}$$

where C_{mw} is the mutual capacitance with the intermediate line present, C_{mwo} is the mutual capacitance without the intermediate line present, and ε_w and ε_{wo} are the dielectric constants with and without the intermediate line present, respectively.

Notice that the equation does not directly depend on distance. This means that the calculated crosstalk between any two wires that have another wire in between must be divided by the factor above regardless of the relative position of the shielding wire.

III. REFLECTIONS

Reflections are due to changes in impedance in a signal path and can be a large source of noise in high speed systems. In many cases, the terminals of a net have a different impedance than the net itself. If the terminals are not impedance matched with the signal wire, then large reflections can occur and must be taken into account.

The lattice diagram method was used to account for reflections from the terminals. This accurate method of calculating reflections is used by North Carolina State University [6][1]. It is based on the superposition of the primitive pulses (in Figure 1) of the induced noise and the time of flight between terminals.

The total noise at each end of the net is the sum of the contributions by both the near-end and far-end voltages and their reflections. This makes the near-end total voltage to be [6]:

$$\begin{aligned} V_{NE}^T(t) &= (1 + \Gamma_{NE})V_{NE_o}(t) \\ &+ (1 + \Gamma_{NE})\Gamma_{NE}\Gamma_{FE}V_{NE_o}(t - 2T_d) + \ldots \\ &+ (1 + \Gamma_{NE})(\Gamma_{NE}^{i-1}\Gamma_{FE}^{i-1})V_{NE_o}(t - 2(i - 1)T_d) \\ &+ \ldots \\ &+ (1 + \Gamma_{NE})\Gamma_{FE}V_{FE_o}(t - T_d) \\ &+ (1 + \Gamma_{NE})\Gamma_{NE}\Gamma_{FE}^2V_{FE_o}(t - 3T_d) + \ldots \\ &+ (1+\Gamma_{NE})(\Gamma_{NE}^{i-1}\,\Gamma_{FE}^{i})V_{FE_o}(t-(2i-1)T_d) \\ &+ \ldots \end{aligned} \tag{8}$$

where $V_{NEo}(t)$ and $V_{FEo}(t)$ are the near-end and far-end primitive pulses.

Obviously, these equations become quite cumbersome as the number of terminations increase. Therefore, an algorithm was developed that calculates the crosstalk from reflections on a net with N terminations.

The algorithm works by calculating the initial crosstalk pulse shape, size, and time arrived at each termination. These pulses are then reflected to each of the other terminations with the new peak value and arrival time calculated. If the new peak value is below a cutoff value, that reflection is assumed to be negligible and is ignored. The cutoff value is determined heuristically. The reflection process continues until all additional reflections are below the cutoff value.

The result of the above process is a list of pulses on each termination. Each pulse has a different start time and different peak value. Pulses also vary in shape as shown in Figure 1. The list of pulses is then combined by superposition and the maximum value is calculated.

IV. CALCULATION VS. SIMULATION

When using equations to calculate the crosstalk values, the worst case must be assumed. The worst case is that all induced crosstalk pulses overlap. However, this is not always the case. Figure 4 shows one common situation where the crosstalk pulses may not overlap. The uncoupled region of net 1 creates a temporal space between the two pulses associated with the two coupled regions. A crosstalk pulse induced on net 1 must travel through the uncoupled region while the crosstalk pulse in the second region is being induced. The gap between the two pulses increases linearly with the length of the uncoupled region. Therefore, the crosstalk induced between net 1 and net 4 may actually be less than the equations predict.

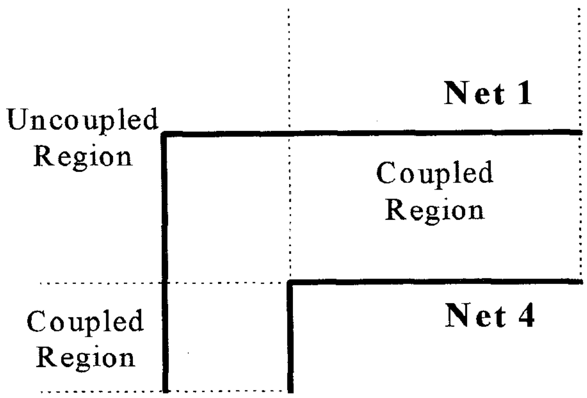


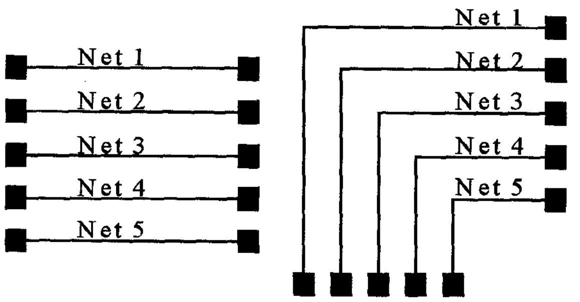
Figure 4. Uncoupled Portion of Nets.

Simulation accounts for these temporal spaces caused by uncoupled regions. This is why it is important to simulate designs. Without accounting for these temporal spaces, it is impractical to use a more complex crosstalk model than the one described here.

V. RESULTS

In order for the internal crosstalk calculations to be useful, they need to be accurate. ContecSPICE was used as the standard with which to compare the router's internal crosstalk calculations.

The first test was to compare the router's crosstalk values with ContecSPICE's values for small test cases. These test cases were created manually to represent net configurations commonly seen in typical designs. Two of these test cases are shown in Figure 5.



Example 1 Example 2

Figure 5. Examples for Crosstalk Calculations.

A small sample of the results for example 1 and example 2 are in Table 1 and Table 2 (values are in volts). Each example was run many times varying which nets were present and the lengths. These tables show that the rough crosstalk calculations by the router are quite accurate. They are certainly accurate enough to use as a first estimate until the design is simulated, and are far more accurate than the parallel-path-length method that other routers use [2][4][7]. In many cases, these calculations are accurate enough to stand alone. The data in these tables also show that the router generally misses on the conservative side of the true crosstalk value. Notice that the larger errors occur in example 2 on net 1 when net 2 is not present. This is caused by the temporal spaces discussed earlier.

Table 1. Results from Example 1.

	ContecSPICE	Router	% Difference
Net 1	0.04426	0.04558	3.0
Net 2	0.08323	0.08635	3.7
Net 3	0.08538	0.08892	4.1
Net 4	0.08323	0.08635	3.7
Net 5	0.04426	0.04558	3.0
Net 1	0.00804	0.00849	5.6
Net 3	0.05007	0.05184	3.5
Net 4	0.08102	0.08266	2.0
Net 5	0.04381	0.04446	1.5
Net 1	0.04199	0.04189	-0.2
Net 2	0.04814	0.04927	2.3
Net 4	0.04814	0.04927	2.3
Net 5	0.04199	0.04189	-0.2
Net 1	0.00217	0.00223	2.8
Net 4	0.04241	0.04300	1.4
Net 5	0.0411	0.04077	-0.8

Table 2. Results from Example 2.

	ContecSPICE	Router	% Difference
Net 1	0.04574	0.04649	1.6
Net 2	0.08251	0.08380	1.6
Net 3	0.07679	0.07769	1.2
Net 4	0.06626	0.06708	1.2
Net 5	0.03265	0.03297	1.0
Net 1	0.00738	0.00801	8.5
Net 3	0.04299	0.04341	1.0
Net 4	0.06437	0.06394	-0.7
Net 5	0.03232	0.03214	-0.6
Net 1	0.04347	0.04297	-1.2
Net 2	0.04833	0.04913	1.7
Net 4	0.03637	0.03662	0.7
Net 5	0.03100	0.03022	-2.5
Net 1	0.00172	0.00190	10.5
Net 4	0.03159	0.03129	-0.9
Net 5	0.03040	0.02939	-3.3

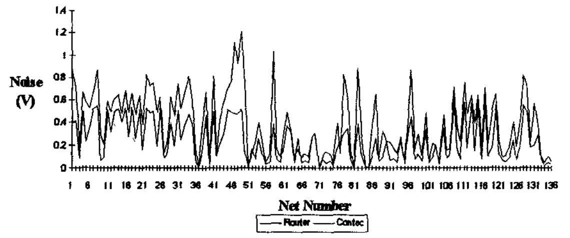


Figure 6. Crosstalk Model VS. ContecSPICE.

The next step in determining the accuracy of the crosstalk prediction code was to test the code on an actual design. Figure 6 shows the crosstalk estimates for a 4-chip design (provided by the Mayo Foundation) versus the values from ContecSPICE. The chart shows that the estimations by the router are conservative. On some nets, the router estimates the crosstalk to be approximately double that of the ContecSPICE value. For the most part, this error is due to the irregular shapes of the nets and the effects of the temporal spaces. The potential for overestimating crosstalk increases with the irregularity of the shape of the net. This is due to the timing effects discussed above.

VI. CONCLUSIONS

This crosstalk model appears to be as accurate as practical without accounting for the temporal spaces created from the bends in nets. MCM design examples provided to date have validated the accuracy of the model and demonstrated the need of including coupling from non-neighboring wires. The data indicates that any significant increase in the accuracy of the noise calculations will be in the form of a simulator, because the actual timing of the noise pulses plays a crucial part in the actual noise induced in a net.

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