Crosstalk Driven Routing Advice

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Abstract

This paper examines the feasibility of generating routing advice for interconnects on printed circuit boards (PCBs) at the design stage, on the basis of crosstalk noise requirements and relevant timing information. A set of analytical expressions have been developed to relate physical design parameters, such as the spacing between a pair of coupled signal lines, to the amount of crosstalk noise generated on these lines. Timing windows have been used to define time intervals during which noise on a specific signal line can affect the functioning of the entire system. An algorithm has been developed, based on the above expressions, to estimate the minimum safe spacing between a pair of coupled signal lines, given the crosstalk noise budget and timing information.

1 Introduction

In modern high-performance digital systems, the containment of noise in the package and signal interconnection sub-system has become an important design consideration. Conventional PCB design methodologies typically attempt to to control noise by using a set of *a priori* high-speed design rules during the initial layout process; simulating the resultant design to check for noise budget violations; and subsequently modifying the design, if necessary, by re-routing potentially noisy nets to reduce noise generation. Several recursions through this iterative simulate – redesign – resimulate process might be required to ensure that all specified signal integrity requirements are met by the current design. The *a priori* design rules enforce the use of interconnect dimensions and topologies which are likely to limit the generation of noise and hence reduce the extent of re-design necessary.

This design approach is inefficient due to several reasons. The *a priori* design rules frequently overconstrain the designer, leading to longer design times and higher manufacturing costs, without ensuring a working design in every case [3]. Simulation generally guarantees accurate characterization of the noise phenomena, but each simulate – redesign iteration requires a significant amount of time, thus increasing design time and costs. Furthermore, simulation tools are typically incapable of providing any routing advice to indicate how a predicted noise problem may be resolved.

A good methodology for avoiding crosstalk noise problems is described in [6], [9], and [11]. The authors indicate that the best solution to managing crosstalk noise is not to specify a restrictive set of spacing and adjacency rules but to check for crosstalk noise violations after routing is completed. Venkatachalam [11] reports that about 10% of the nets require rework after checking. However, this rework is often difficult as the routing channels may be congested.

The long term objective of our work is to automatically generate rules that can be used during design to route congested net-sets with the minimum area while meeting crosstalk requirements [10]. In this method, the *a priori* design rules are replaced by a tool capable of generating noise-driven interconnect routing advice, on a net-by-net basis, during a computer-aided PCB design process. This paper describes initial work towards the development of such a tool for generating wiring advice to avoid a crosstalk problem. In this paper the work is directed towards solving the problem for a coupled pair of wires (which can easily be extended to obtaining spacing rules for a bus). Future work will deal with the more general problem of optimizing the (random) wiring within a congested

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channel.

In the next section, we first summarize the analytic equations that can be used to predict crosstalk and verify these equations through measurement. We then describe how we abstracted the problem, with no loss of accuracy, to make it more amenable for computer operation. In Section 3, we discuss how the concept of timing windows of noise-susceptibility can be used to come up with more aggressive, yet still safe, designs. Finally, we describe the steps in the CAD tool and present conclusions.

2 Crosstalk Noise Estimation

In order to be able to generate crosstalk driven wiring advice during design, it is necessary to develop a computationally efficient method to relate crosstalk noise voltages to physical design parameters. The emphasis is on providing a fast, reasonably accurate engineering approximation of crosstalk noise, as a function of line and signal parameters. For this purpose, we have used a set of simple analytical expressions, based on the work of Feller et al [5].

2.1 Related Work

One of the earliest and best known empirical expressions for transient crosstalk noise was derived by Feller et al in [5] by applying wave equations and the principle of superposition to current and voltage distributions along a pair of coupled transmission lines configured as shown in Figure 1. The authors assume that the lines are lossless, loosely coupled[1], and terminated with matching resistances at both near and far ends, so that $\Gamma_{ne} = \Gamma_{fe} = 0$.

It is shown that the instantaneous crosstalk voltage $V(x, t)$, induced at a distance x down the passive line, due to any arbitrary signal $V_{in}(t)$ at the near end of the active line, can be given by :

$$V(x, t) = K_f \, x \, \frac{d}{dt}\left[V_{in}\left(t - \frac{T_d \, x}{l}\right)\right] + K_b \left[V_{in}\left(t - \frac{T_d \, x}{l}\right) - V_{in}\left(t - 2T_d + \frac{T_d \, x}{l}\right)\right] \quad (1)$$

where l is the length of the coupled region; L_m and C_m are the inductive and capacitive coupling coefficients; Z_0 is the characteristic impedance of each line; T_d is the time of flight of the signal down the coupled region; and K_f and K_b are the forward and backward coupling coefficients, given by :

$$K_f = -\frac{1}{2}\left(\frac{L_m}{Z_0} - C_m Z_0\right)$$
$$and \ K_b = \frac{l}{4 T_d}\left(\frac{L_m}{Z_0} + C_m Z_0\right)$$

[1] Loose coupling implies that the crosstalk signal on the coupled line is small enough to have no discernible effect on the signal in the active line. This assumption is typically valid for interconnects at the board level.

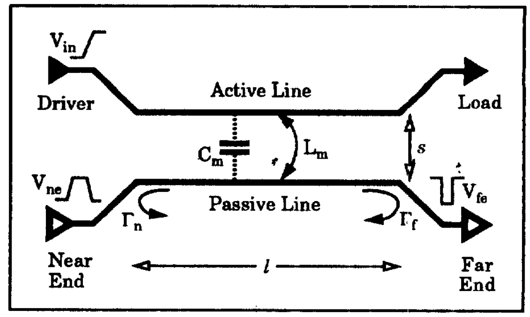

Figure 1: Coupled transmission line pair showing noise parameters and physical dimensions.

Another computationally efficient method for time-domain crosstalk analysis is based on the observation that any signal traveling in a coupled transmission line system can be viewed as a combination of even-mode and odd-mode excitations of the system, such that the effective voltages at the ends of the lines can be determined by superposition of the mode voltages. It is possible, therefore, to determine crosstalk noise voltages in terms of modal parameters of the transmission line system and linearized effective source and load resistances. In [7], Matthaei applies this *mode-superposition* method to obtain expressions for terminal voltages in a pair of coupled symmetric interconnect lines. This method of analysis also assumes loose coupling, and the expressions obtained for the two-line case are similar to those given by Feller et al.

In [1], Canright describes an interesting approach for developing crosstalk equations, based partially on theory and partially on measurement data. In this method, crosstalk equations obtained from theory are modified by adding correction factors to enhance their accuracy. The correction factors are obtained by comparing the noise levels predicted by the equations with actual crosstalk noise measurements on a large number of samples. Curve-fitting and error analysis techniques to determine the type and value of correction factors needed. While the correlation of the corrected equations with actual measurements is very good, this method suffers from the fact that it is not generally applicable to the entire design space. Since the correction factors are obtained from measurements made on a finite set of samples the applicability of the corrected equations is limited to the ranges of design parameters covered by the sample population.

2.2 Crosstalk Equations

It is evident from equation 1, that crosstalk noise generated along the entire length of a coupled quiet line. However, on PCBs, the primary interest is in the cumulative noise signals formed at the ends of the line. From equation 1, putting $x = 0$ for the near end and $x = l$ for the far end, we get the following expressions for near and far end crosstalk voltages on terminated line :

$$V_{NE}(t) = V(0, t) = K_b\ [V_{in}(t) - V_{in}(t - 2T_d)] \quad (2)$$

and

$$V_{FE}(t) = V(l, t) = K_f\, l\, \frac{d}{dt} [V_{in}(t - T_d)] \quad (3)$$

These equations have been used to obtain the near and far end noise pulses for a rising edge on the active line as shown in figure 2.

At the far end of the quiet line a square noise pulse whose amplitude and duration both depend on the rise time of the active line signal, is predicted. At the near end, the type of noise pulse generated depends on the length of the coupled lines. For the long line case, $(t_r < 2t_d)$, the near end crosstalk signal reaches a saturation value of $K_b V_0$. For short line, where $2t_{d'} < t_r$, the peak noise voltage is given by $2t_{d'}K_b V_0/t_r$. In either case, the duration of the noise pulse is $2T_d + t_r$, and magnitude of the slopes of the rising and falling edges of the noise pulse are equal to the slope if the edge of the active line signal.

These near and far end noise pulses have been called the near and far end *primitive pulses* in the rest of this paper. This is done in order to easily distinguish them from the more complex cumulative noise waveforms which are obtained as a result of the superposition of these primitive pulses and their reflections.

2.3 Validation through Measurement

Measurements were carried out on a pair of symmetric coupled lines on a test board, with signals of different rise times and amplitudes. The peak observed noise voltages at the far and near ends is tabulated in table 1 and table 2 respectively. The measurements obtained for a 0.6ns rise time were used

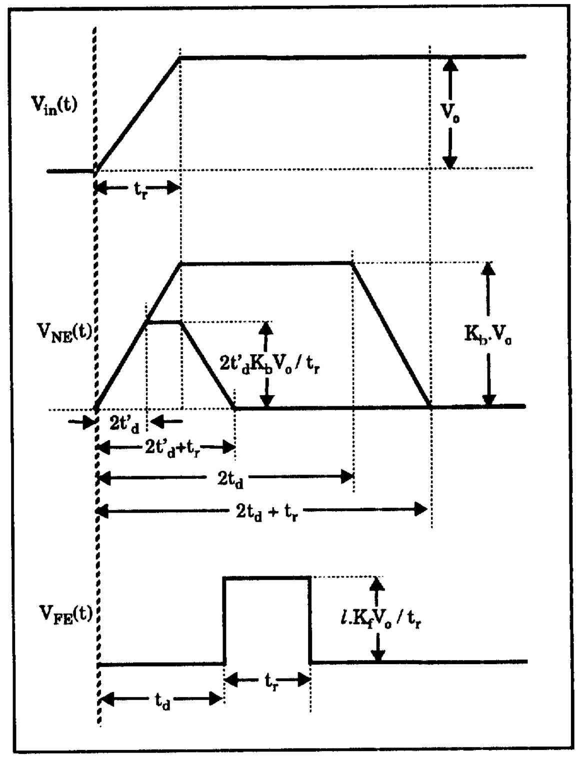

Figure 2: Near and Far End Crosstalk Primitive Pulses

to calculate the coupling coefficients for each case. Equations 2 and 3 were used to calculate predicted peak noise in each case, and these calculated values are compared graphically to the measurement data in figure 3.

V_0 (in mV)	t_r (ns)	Measured V_{FE} (mV)	Calculated V_{FE} (mV)	Error Percent
504	0.12	78	87	11
171	0.25	13.7	14.3	4
125	0.46	6.2	5.7	8
175	0.6	6.1	6.1	0 [2]

Table 1: Measured & Calculated Far End Crosstalk.

The accuracy, especially of the near-end predictions, appears to be within our acceptable range (±10%), though error appears to increase with smaller rise times. The relatively larger errors in calculated far

[2]These measurements were used to calculate the coupling coefficients, so they provide a perfect match with calculated values.

V_0 (in mV)	t_r (ns)	Measured V_{NE} (mV)	Calculated V_{NE} (mV)	Error Percent
504	0.12	45	43.2	4
171	0.25	14.8	14.7	1
125	0.46	·11	10.7	3
175	0.6	15	15	0 [2]

Table 2: Measured & Calculated Near End Crosstalk.

end crosstalk noise may be partially due to inaccuracies in the determination of the rise-time of the signal driving the active line. For these measurements, the rise times were actually determined by observing the signal traces on an oscilloscope. For the faster rising signals, this method is not very accurate.

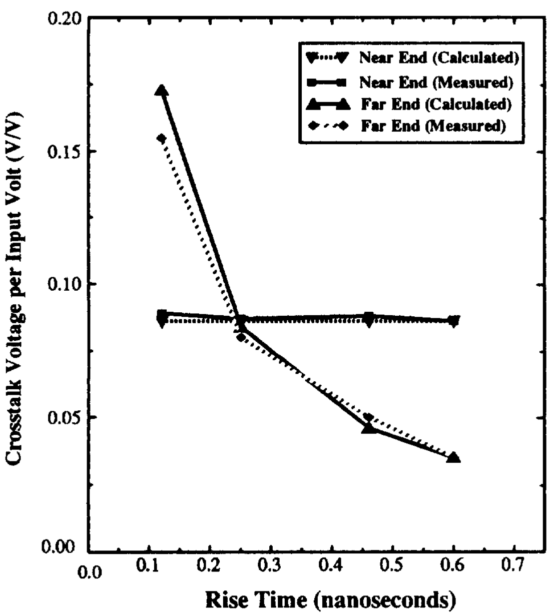

Figure 3: Far and Near End Crosstalk Noise Voltages

2.4 Reflected Noise Pulses

In the analysis presented in [5], Feller et al have considered only terminated signal lines in which all noise pulses, upon reaching either end of a line, are absorbed in matching terminations. While this simplifies the formulation of crosstalk equations, it severely limits the applicability of these equations by excluding reflected noise pulses from the analysis.

We have used a simple, lattice diagram based approach in order to characterize the secondary noise

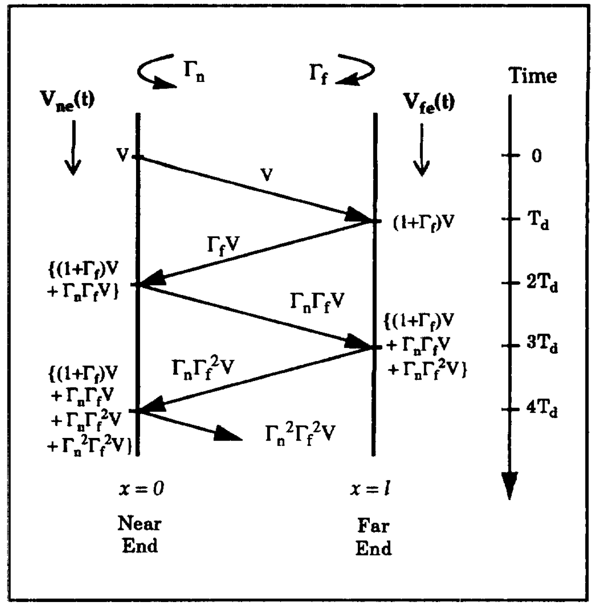

Figure 4: Lattice Diagram Analysis of Reflected Waveforms

pulses which arise due to reflections at the ends of the line, as shown in figure 4. Reflected components of both the near and far end primitives (and their reflections), will contribute to the the total noise at both ends of of an interconnection line. A sketch of a typical cumulative noise waveform (observed at the near end) arising due to reflected crosstalk noise pulses, is shown in 5.

The expression for the cumulative noise waveform at the near end of a line with reflection coefficients Γ_n and Γ_f at the near and far ends, is found to be [10]:

$$\begin{aligned} V_{NE}^T(t) &= (1+\Gamma_n)V_{NE_0}(t) \\ &+ (1+\Gamma_n)\Gamma_n\Gamma_f V_{NE_0}(t-2T_d) + \cdots \\ &+ (1+\Gamma_n)(\Gamma_n^{i-1}\Gamma_f^{i-1})V_{NE_0}(t-2(i-1)T_d) \\ &+ \cdots \\ &+ (1+\Gamma_n)\Gamma_f V_{FE_0}(t-T_d) \\ &+ (1+\Gamma_n)\Gamma_n\Gamma_f^2 V_{FE_0}(t-3T_d) + \cdots \\ &+ (1+\Gamma_n)(\Gamma_n^{i-1}\Gamma_f^{i})V_{FE_0}(t-(2i-1)T_d) \\ &+ \cdots \qquad (4) \end{aligned}$$

where $V_{NE_0}(t)$ and $V_{FE_0}(t)$ refer to the near and far end primitive pulses which were derived in section 2.2. The corresponding expression for the cumulative far end noise voltage is :

$$V_{FE}^T(t) = (1+\Gamma_f)V_{FE_0}(t)$$

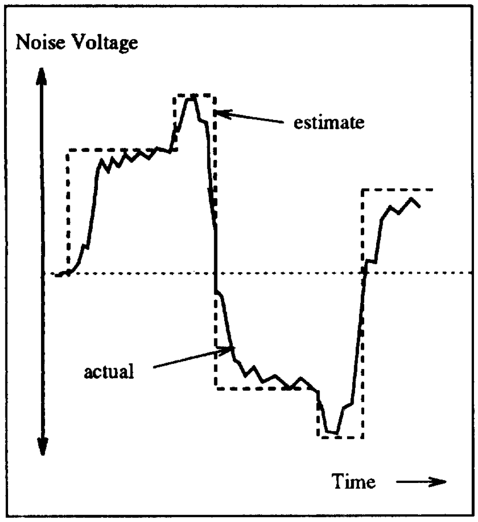

Figure 5: Cumulative Noise Waveform due to Reflection

$$
\begin{aligned}
&+ \quad (1+\Gamma_f)\Gamma_n\Gamma_f V_{FE_0}(t-2T_d) + \cdots \\
&+ \quad (1+\Gamma_f)(\Gamma_n^{i-1}\Gamma_f^{i-1})V_{FE_0}(t-2(i-1)T_d) \\
&+ \quad \cdots \\
&+ \quad (1+\Gamma_f)\Gamma_n V_{NE_0}(t-T_d) \\
&+ \quad (1+\Gamma_f)\Gamma_f\Gamma_n^2 V_{NE_0}(t-3T_d) + \cdots \\
&+ \quad (1+\Gamma_f)(\Gamma_f^{i-1}\Gamma_n^{i})V_{NE_0}(t-(2i-1)T_d) \\
&+ \quad \cdots \qquad (5)
\end{aligned}
$$

The number of reflection terms to be summed in estimating the net noise can be determined by defining a threshold level. When the magnitude of a reflected pulse falls below this threshold, further reflections will be ignored. In our implementation, we also provide an automatic mechanism to discard reflections which occur after the current system clock cycle has expired.

In the process of using these expressions for generating routing advice (see Section 4), we will need to symbolically manipulate the terms involved. This process becomes impractical if the actual shape of the near end primitive pulse is retained. Hence, in this work, this trapezoidal pulse is abstracted as a rectangular pulse which can be represented by magnitude and width equal to that of the primitive pulse. Since the far-end primitive pulse is rectangular to begin with, no modifications are necessary there. These rectangular pulses must be abstracted and superposed, using expressions 4 and 5, in a manner which will provide a conservative (upper bound) estimate of the actual noise signals, as shown in figure 5.

3 Timing Windows

Typically, in digital systems, there are many signals whose values are capable of affecting the functioning of the system during only certain specific parts of a clock cycle. Hence, these signals are susceptible to external disturbances during certain time intervals, and relatively insensitive during other intervals [8]. In this section, we illustrate how timing windows, based on device and system timing information, can be used to define the period when a given net is sensitive to noise. A simple example is shown in Fig. 6, where a signal line drives the input, D, of a clocked, positive edge-triggered flip-flop. We can define a timing window around the positive edge of the clock, such that any noise at D, which falls outside this window, will not affect the output of the flip-flop.

The width of the window, t_{win}, will be determined by factors such as gate setup (t_{su}) and hold (t_h) times, parametric variations in device timing behavior, and clock skew (t_{skew}). The position of the window, relative to the clock edges, will be determined by the logic and interconnect propagation delays involved, and other system timing considerations. These dependencies have been dealt with in greater detail in [10]

In [8], Purks uses timing windows primarily to detect adjacent line pairs where crosstalk analysis is not necessary due to particular timing constraints. In this work, we have used a more general approach, where, given a cumulative crosstalk noise waveform as shown in figure 5, only that portion of the waveform which overlaps with noise-sensitive timing windows of the corresponding net is considered while calculating total noise on the net. This is, of course, an aggressive noise-budgeting technique, not suitable for critical nets. However, in designs where routing density is of primary importance, this timing window based noise analysis may provide significant advantages.

4 Generation of Routing Advice

A CAD tool has been built using the system of equations and techniques described above. It comes up with a spacing rule for each net pair by using the following sequence:

1. Generate a symbolic expression for the primitive pulses at each end of the line, in terms of the driving signal parameters, and the physical design variables.

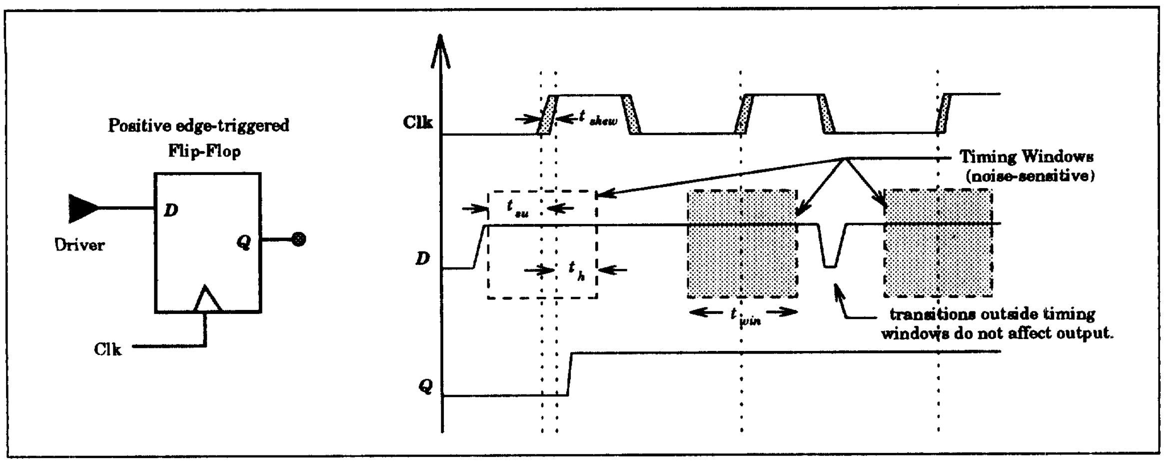

Figure 6: Timing specifications for a positive edge-triggered flip-flop, showing noise-sensitive timing windows at input.

2. Using these primitives as building blocks, derive an expression for the entire cumulative waveform resulting from reflection. Define and use a threshold voltage level, so that reflected pulses may be ignored when they fall below that level.

3. Extract the timing information (position and duration) of each primitive and reflected pulse, with respect to the system clock. Using this information, compare the the timing of each noise pulse with the noise sensitive windows for that particular net. Discard pulses which lie entirely outside the timing window.

4. Obtain an expression for the peak voltage level resulting from the superposition of the remaining noise pulses. Equate this to the available noise budget. Solve to obtain the acceptable values of coupling parameters.

5. Obtain acceptable values of physical design variable from empirically-derived look-up tables containing coupling parameters as functions of conductor spacing. In principal, a field solver could be used as well.

This procedure is illustrated with the help of a simple design problem in the following section.

A Design Example

In this section, we consider the problem of obtaining the minimum spacing, s, required between a pair of interconnection lines, in order to meet a specified crosstalk noise budget. The problem configuration and timing specifications are shown in figure 7.

The problem involves microstrip lines on an alumina substrate with $\epsilon_r = 10$. The coupled 50 ohm lines are 10 centimeters long, and are series terminated at the near end by matching 50 ohm resistors. The lines are not terminated at the far end, and drive gate inputs which can be assumed to behave as open circuits. The signal driving the active line has a rise time $t_r = 0.1$ns, and a steady state amplitude $V_0 = 5$V. The noise budget available for crosstalk noise is assumed to be 0.2 Volts. We have assumed that the lines are part of a synchronous circuit, with a two phase clock, and shown in figure 7. The drivers are assumed to switch during ϕ_1, while the signal at the gate inputs at the receiving end, is latched during ϕ_2. Accordingly, we have defined, for the sake of illustration, timing windows at the near and far ends, as shown in the figure.

In order to solve this problem, we first calculate the necessary intermediate variables :

- Line propagation delay, $T_d = (l\sqrt{\epsilon_r})/3 \times 10^8$ For $l = 0.1$m and $\epsilon_r = 10$, we have $\mathbf{T_d = 1.05ns}$.

- The far end reflection coefficient, for this case, is given by $\Gamma_f = (\infty - 50)/(\infty + 50)$. Hence we get $\mathbf{\Gamma_f = 1}$.

- The near end reflection coefficient, Γ_n, is zero, since there is a matching 50 ohm resistor terminating each line at the near end.

- Since $2T_d = 2.1$ns $< t_r = 0.1$ns, we can treat this

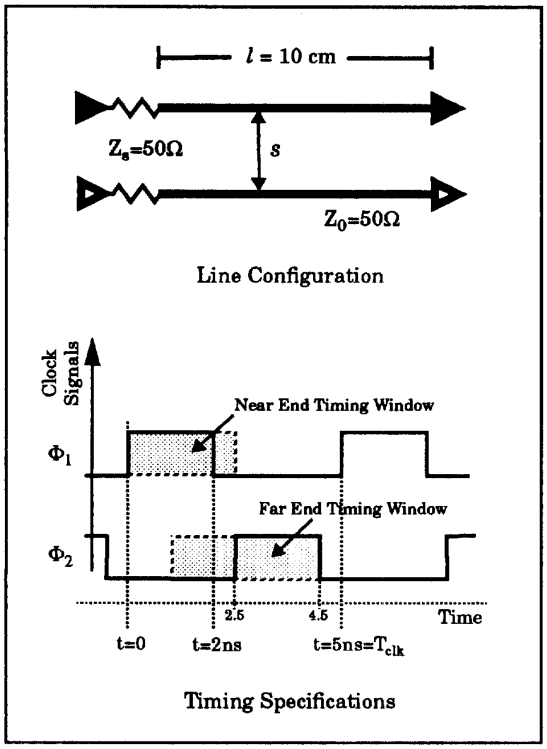

Figure 7: A Two-Line Design Example

as a long-line case, and assume saturated near-end crosstalk.

We can now apply the steps outlined in the previous section, to this problem:

1. **Generation of Primitives**: From the general forms of the near and far end primitive pulses shown in figure 2, applying the specific values for the given design problem, we get:

   - *Near End Primitive* : This is a trapezoidal pulse, with height $K_b V_0$, and total width of $2T_d + t_r$, and rise and fall times equal to t_r. As explained earlier, we abstract this as a rectangular pulse with the parameters of magnitude $= 5K_b$, and duration = 2.2 ns, placed at the beginning of the clock cycle ϕ_1
   - *Far End Primitive* : This is a rectangular pulse, defined, in this case, but a magnitude of $5K_f \times 10^9$, and a duration of 0.1 ns, placed 2.1 ns after the beginning of the ϕ_1 clock cycle.

2. **Building the Cumulative Noise Waveforms**: Substituting the variables from this sample design in expressions 4 and 5, we get the following expressions for total noise at each end of the quiet line :

$$V_{NE}^T(t) = V_{NE_0}(t) + \Gamma_f \, V_{FE_0}(t - T_d) \quad (6)$$

   And

$$V_{FE}^T(t) = (1 + \Gamma_f) V_{FE_0}(t) \quad (7)$$

3. **Timing Analysis** : In figure 7, the timing window for the near end is shown to extend from the positive edge of clock ϕ_1 till 2.5 ns after this edge; while that at the far end stretches from 1 ns after the positive ϕ_1 edge, to 4.5 ns after this edge. From the previous step, we can see that both components of $V_{NE}^T(t)$ overlap with the near end timing window, and the entire $V_{FE}^T(t)$ pulse lies within the far end timing window. Hence, we can use equations 6 and 7 directly in further computations.

4. **Estimation of Coupling Parameters**: From the earlier steps, and equations 6 and 7, we can write the estimated peak values of crosstalk noise at either end of the line as follows :

$$V_{NE_{peak}} = 5K_b + (5K_f \times 10^9)$$

   And

$$V_{FE_{peak}} = 2 \times (5K_f \times 10^9) = K_f \times 10^{10}$$

   These expressions may now be equated to the noise budget of 0.2 V, giving $\mathbf{K_b = 0.02}$ and $\mathbf{K_f = 2 \times 10^{-11}}$.

5. **Generation of Required Spacing**: The K-coefficients calculated above may be used to obtain the values of the coupling capacitance, C_m, and inductance, L_m, between the given lines. From this information, it is possible to generate the required spacing between the two lines, from a look-up table or, as discussed earlier, from a field solver.

5 Conclusions

In this work, we have attempted to develop an analysis based technique for generation of noise-driven routing advice during design. Our results indicate that for the simple case of pairwise coupled lines, it is possible to implement such a technique as a better design-time alternative to *a priori* high-speed design rules. In either case, in order to guarantee proper functioning, the resulting design must be simulated and tested. However, this technique should enable more aggressive designs, while reducing the need for subsequent re-design.

In order to formalize the techniques presented in this paper, into a pragmatic design tool, several outstanding issues will require further investigation. Some of these issues are outlined below :

- It should be possible to extend this technique to the more general case of transmission line pairs of unequal length. This would, however, introduce an additional symbolic term in our formulation, rendering the computations more complex, and increasing the possibility of error in the manner in which overlaps of reflected waveforms are calculated.

- The effect of discrete discontinuities such as vias and bends in the interconnection lines have to be considered. It is not possible to characterize the effects of these discontinuities in any simple manner, using the techniques outlined in this paper. In [4], it is suggested that the magnitude of the reflected pulses from such discontinuities is small enough to be neglected, unless the number of discontinuities is very large. However, this issue merits further investigation. A related issue is the effect of these discontinuities on timing. Since the methods presented in this paper depend strongly on the timing of each noise pulse in order to calculate the total noise, any significant variation in timing would also affect accuracy.

- Frequency dependent effects such as the skin effect are clearly not computable using our methods. In [2], it is suggested that the crosstalk noise levels obtained through simulation, after accounting for skin effect, are actually lower than the estimates which do not consider such effects.

Acknowledgements

The authors would like to thank members of the Signal Integrity Department at Tandem Computers, and George Katopis of IBM for their valuable insights. Thanks are also due to Slobodan Simovich, Sharad Mehrotra and Jeff Byrd at NCSU for their assistance.

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