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NOISE MODELLING FOR RC INTERCONNECTS IN DEEP SUBMICRON VLSI CIRCUIT FOR UNIT STEP INPUT

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Abstract*:* This paper presents a closed form crosstalk noise modelling for on-chip VLSI RC interconnects using 2π model. For low frequency of operation, the interconnect can be modelled as distributed RC segments with sufficient accuracy. This crosstalk noise modelling is carried out for the case when step input is applied to the aggressor which is adjacent to the victim net. The proposed model represents the noise voltage waveform. In this paper, the original 2π model is further simplified and the closed form formulae for noise pulse width and noise amplitude for RC interconnect have been derived. This model considers various parameters, such as coupling location (near driver and near receiver) and course distributed RC characteristics for victim net. The proposed crosstalk model results in an error of less than 6% when compared to that of the SPICE simulation.

Keywords: Noise Modelling, Crosstalk, On-Chip RC Interconnect, Step Input, VLSI

I. INTRODUCTION

Interconnect of integrated circuits in the deep submicron (DSM) technology plays an important role in determining the performance and signal integrity [1-2]. Due to the rapid advances in Very Large Scale Integration (VLSI) technology, minimum feature size is reduced to sub-quarter micron and switching time to tenths of pico seconds or even lesser. Decreasing feature size increases crosstalk noise which is produced due to the parasitic coupling between interconnects. The amount of this crosstalk noise can be calculated by using circuit and layout techniques [3]. This reduction in the feature size also affects the design's timing and functionality goals [4]. The model proposed in [5] provides a closed form formula for the waveform in the RC line with practical boundary conditions. Later on 2π model is used to find the closed form expression for crosstalk noise in [6]. This is useful in calculating the values of noise amplitude and noise width. Compared with previous crosstalk noise models the 2π-model is proven to be more accurate. There are various ways of calculating the crosstalk noise amplitude and crosstalk noise width. One crosstalk noise estimation method using Devgan's metric is explained in [9], but this method is not valid for fast slew rate. Also the magnitude of the induced noise is not depending upon the values of capacitances and resistances. Various circuits such as latches and dynamic circuits are very sensitive to crosstalk noise at both the input and output ends. Due to this crosstalk noise the performance of these high speed circuits degrades. So this makes coupling an important issue for interconnects. There are different models proposed to capture the coupling aware noise modelling for different frequency range of applications [8-17]. But the drawback of these models is that they deal with fully coupled structures and hence are not suitable for partially coupled lines. The model proposed in this paper considers a 2π RC interconnect crosstalk noise model and can accurately deal with partially coupled lines. This model considers various parameters such as aggressor's slew at the coupling location, the coupling location at the victim net (near driver or near receiver) and the coarse distributed RC characteristics. For low frequency of operation, the interconnect can be modelled as distributed RC segments with sufficient accuracy. This model is very accurate, with less than 6% error on average compared with HSPICE simulations.

The contribution of this paper is as follows: a novel and accurate model has been proposed for crosstalk aware noise for on-chip VLSI RC interconnects.

This paper is summarized as follows: Section II presents the proposed crosstalk noise model for step input excitation. Section III shows and describes the simulation results. Finally Section IV concludes the paper.

II. PROPOSED CROSSTALK NOISE MODEL

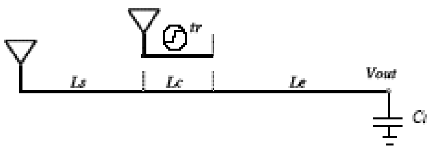
In this section, we first present the 2π model approach and derive its analytical time-domain waveform. Then we will focus on two key metrics for the 2π model, i.e., peak noise (amplitude) and noise width. The closed form expressions are then derived for both peak amplitude of the noise and the noise width. We then extend the 2π model to handle general RC trees, followed by extensive validation of the model.

II.1 Model and its analytical waveform

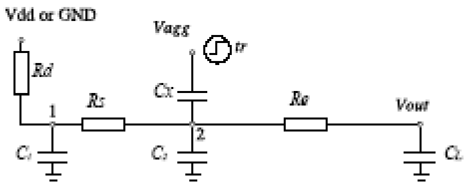
For explaining the crosstalk noise model, the 2π model is explained first. In this explanation, victim net is a RC line. An aggressor line is placed near the victim net, as shown in Figure 1(a). Let the aggressor voltage pulse at the coupling location be a unit step input. In Figure 1(a), L_s represents the interconnect length of victim net before the coupling. Similarly, L_c and L_e represent the interconnect length of victim net at the coupling and after the coupling, respectively. 2-π model is reduced to a 2-π type RC model as shown in Figure 1(b). This reduction of 2-π model is very useful while calculating the value of crosstalk noise at the receiver end. This model contains two π type RC circuits, known as 2-π model. One RC circuit is located before the coupling and the other is after the coupling. The victim driver is modelled by an effective resistance R_d and other RC parameters are C_x, C_1, R_s, C_2, R_e and C_L. The values of these parameters are computed from the geometric information from Figure 1(b) in the following manner:

The node 2 is known as coupling node; this is set to be the centre of the coupling portion of the victim net i.e. $L_s+L_c/2$ from the source. Let the upstream and downstream interconnect resistance/capacitance at node 2 be R_s/C_s and R_e/C_e, respectively. Then capacitance values are set to be $C_1 = C_s/2$, $C_2 = (C_s+C_e)/2$ and $C_L = C_e/2+C_1$.

In some cases one lumped RC for the victim net can be used, but 2π model can model the coarse distributed RC characteristics.The resulting 2π model can be solved analytically.



(a)



(b)

Figure 1. (a) Layout of a victim net and aggressor above it; (b) 2-π model crosstalk noise model

From Figure 1(b), the impedance at node 1, Z_1 satisfying the following,

$$\frac{1}{Z_1} = \frac{1}{R_d} + sC_1$$

Then at node 2, we have

$$\frac{1}{Z_2} = \frac{1}{(Z_1 + R_s)} + sC_2 + \frac{1}{R_e + \frac{1}{sC_L}}$$

$V_2(s)$ denotes the s-domain voltage at node 2, then

$$V_2(s) = \frac{Z_2}{Z_2 + \frac{1}{sC_L}} V_{agg}(s)$$

The output voltage V_{out} in the s-domain is

$$V_{out}(s) = V_2(s)\ \frac{\frac{1}{sC_L}}{R_e + \frac{1}{sC_L}} \tag{1}$$

Substituting Z_1, Z_2, and V_2 into $V_{out}(s)$ yields,

$$V_{out}(s) = \frac{a_2 s^2 + a_1 s}{s^3 + b_2 s^2 + b_1 s + b_0}\ V_{agg}(s) \tag{2}$$

where the co efficient are

$$a_2 = \frac{K_1}{K_2}$$

$$a_1 = \frac{(R_d + R_s)C_x}{K_2}$$

$$b_2 = \frac{((C_2+C_x)(R_e C_L(R_d+R_s)+R_d R_s C_1)+R_d R_e C_1 C_L+C_L R_d R_s C_1)}{K2}$$

$$b_1 = \frac{((R_d + R_s)(C_x + C_2 + C_L) + (R_e C_L + R_d C_1))}{K_2}$$

$$b_0 = \frac{1}{K_2}$$

$$K_1 = C_x R_d R_s C_1$$

$$K_2 = R_d R_s C_1 C_L R_e (C_x + C_2)$$

The transfer function H(s) can be expressed into the poles/residues form as,

$$H(s) = \frac{a_2 s^2 + a_1 s}{s^3 + b_2 s^2 + b_1 s + b_0} \equiv \frac{k_1}{s - s_1} + \frac{k_2}{s - s_2} + \frac{k_3}{s - s_3}$$

The three poles s_1, s_2 and s_3 are the three roots of $s^3+b_2s^2+b_1s+b_0=0$, which can be obtained analytically using standard mathematical techniques. The time domain function of each pole/residue is $f_i(t) = k_i e^{S_i t}$ where (i=1, 2, 3).

For the aggressor with unit step input with normalized $V_{dd}=1$, i.e.

$$V_{agg} = \begin{cases} 1 & t \geq 0 \\ 0 & \text{otherwise} \end{cases}$$

its Laplace transform is,

$$V_{agg}(s) = \frac{1}{s} \tag{3}$$

Then for each pole/residue pair, the s-domain output is given by,

$$V_{out}(s) = \frac{k_i}{(s - s_i)} V_{agg}(s) \tag{4}$$

From (3) and (4),

$$V_{out}(s) = -\frac{k_i}{s_i}\frac{1}{s} + \frac{k_i}{s_i}\frac{1}{(s - s_i)}$$

So, the time domain expression of the output voltage is

$$v_{out}(t) = -\frac{k_i}{s_i}\left(1 - e^{S_i t}\right) \quad \text{(for i=1, 2, 3)}$$

Therefore, the final noise voltage is simply the summation of the voltage waveform from each pole/residue pair.

$$v_{out}(t) = v_{out1}(t) + v_{out2}(t) + v_{out3}(t) \tag{5}$$

The 2π model has been tested extensively and its waveform from (5) can be shown to be almost identical compared to HSPICE simulations.

II.2 Closed form noise amplitude and width

Equation (5) gives the final noise voltage waveform. In this subsection, formulae for noise amplitude and noise width are obtained. This is achieved when we will further simplify the original 2π model.

II.2.1 Noise Amplitude Calculation

On simplifying (2) using dominant-pole approximation method,

$$V_{out}(s) \approx \frac{a_1 s}{b_1 s + b_0} . V_{agg}(s)$$

where $V_{agg}(s) = \frac{1}{s}$

So,

$$V_{out}(s) \approx \frac{t_x}{t_v\left(s + \frac{1}{t_v}\right)} \tag{6}$$

where the co-efficient are,

$$t_x = (R_d + R_s)C_x$$

$$t_v = (R_d + R_s)(C_x + C_2 + C_L) + (R_e C_L + R_d C_1)$$

The term t_x represents the RC delay term from the upstream resistance of the coupling element times the coupling capacitance. The term t_v represents distributed Elmore delay of victim net.

The output voltage shown in (6) can be expressed in time domain and given in (7).

$$v_{out}(t) \approx \frac{t_x}{t_v} e^{-t/t_v} \tag{7}$$

where $t \geq 0$.

From the noise expression shown in (7), it is evident that noise monotonically decreases as $t \geq 0$. The value of noise will be maximum at t=0. This maximum value of noise can be calculated by putting t=0 in (7).

So,

$$v_{\max} \approx \frac{t_x}{t_v} \tag{8}$$

Equation (8) represents the maximum amplitude of noise which is obtained at t=0.

II.2.2Noise Width Calculation

The noise width for a noise pulse is defined to be the length of time interval so that the noise spike voltage v is larger than or equal to v_t where v_t represents the threshold voltage.

From (7),

$$v_{out}(t) \approx \frac{t_x}{t_v} e^{-\frac{t}{t_v}}$$

So, $$t = t_v \ln \frac{1}{v_{out}} . \frac{t_x}{t_v} \tag{9}$$

Noise width is the width of time interval between t_1 and t_2.

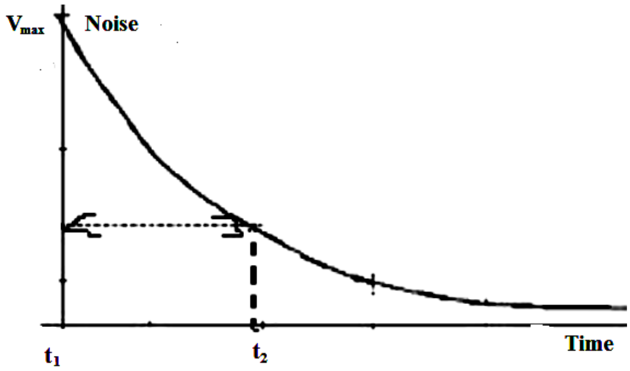


Fig-2 Illustration of the noise width

The value of t_2 can be calculated using (9). Hence, t_2 can be derived as,

$$t_2 = t_v \ln \frac{1}{v_{out}} \frac{t_x}{t_v}$$

At t_2, the noise voltage is v_t.

So, $$t_2 = t_v \ln \frac{1}{v_t} \frac{t_x}{t_v}$$

Noise width is given by,

$$t_{width} = t_2 - t_1$$

$$t_{width} = t_v \ln \frac{1}{v_t} \frac{t_x}{t_v} \tag{10}$$

In this paper, we assume the value of threshold voltage v_t to be half of the value of the peak noise voltage v_{max}.

i.e. $$v_t = \frac{v_{\max}}{2} \tag{11}$$

From (8), (10) and (11),

$$t_{width} = t_v \ln 2 \tag{12}$$

This expression represents the width of the noise voltage waveform.

Note that when the time increases beyond t_2, the noise voltage becomes very less. In the above calculation we ignored the effect of that noise. In some conditions peak noise exceeds certain threshold voltage but remains immune to the noise. This can be expressed clearly by some noise amplitude versus noise width plots.

III. SIMULATION RESULTS AND DISCUSSIONS

The 2π model and its analytical formulae for peak noise and noise width as shown in (8) and (12), respectively, have been tested extensively and the efficacy of the proposed models are justified when compared to HSPICE simulations. To obtain high fidelity and to detect the corner scenarios, we run our 2π model, Devgan model [9], Vittal model [3], and HSPICE simulations on 1500 randomly generated circuits with realistic parameters in a 0.18 μm technology. For the test circuits, the driver resistance R_d is from 20 to 2000 Ω·, the loading capacitance C_L is from 4 to 50fF , the length parameters L_s, L_c, and L_e are from 1 to 2000 μm, the wire width/spacing is either 1x or 2x minimum width/spacing, and the aggressor slew is from 10 to 500 ps. The simulation results show that the average errors for peak noise estimation using Vittal and the proposed 2-π model are 9%, and less than 2%, respectively.

Table 1 summarizes the percentage of nets that fall into certain error ranges using the 2-π model with closed-form peak noise and noise width expressions from (8) and (12) compared to those from HSPICE simulations. It can be found that using the proposed model both peak noise and noise width calculation results in an error of as less as when compared with those of SPICE simulations, and almost 97% nets have less than 6% errors.

Table 1: The percentage of nets that fall into the error ranges for peak noise (V_{max}) and noise width (t_{width}) from the 2-π model

Error Range	V_{max}	t_{width}
within +/- 20%	98.9%	99.78%
within +/- 20%	94.19%	97.98%
within +/- 20%	91.89%	95.78%
within +/- 20%	89.32%	88.98%
Average Error	1.98	2.01

Figure 3 shows the changes in crosstalk when the input rise time varies and all of geometric parameters are fixed. Our proposed model is compared with Vittal [3] named as V-pay and SPICE.

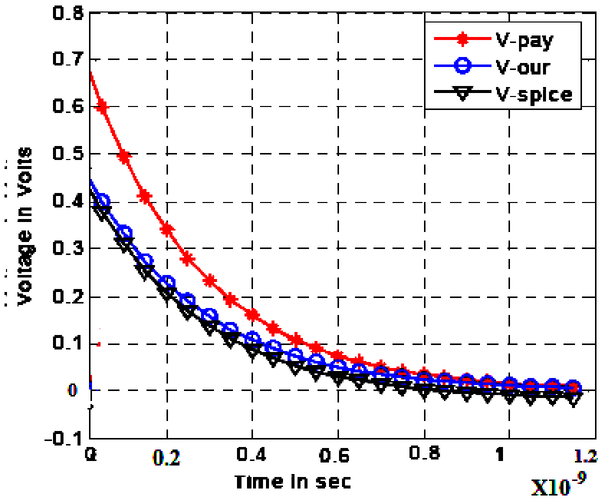


Fig.3.Crosstalk Noise waveform for two coupled transmission lines.

In this paper 2π model is also tested on a set of randomly generated multiple-pin nets with general RC tree structures. As a result experimental results show that 2-π model still works better for general RC trees. Figure 4 shows the scatter diagram comparing the 2-π model (y-axis) with HSPICE (x-axis) simulations for 25 randomly generated four-pin nets (i.e., with two branches). The experimental setting is the same as those for 2-pin nets. The branching wire length ranges from 1 to 1500 µm. The branching location can be anywhere from driver to receiver. HSPICE simulations are performed on distributed RC networks by dividing each long wire into every 10µm segment. Again, for all test circuits, the 2-π model gives very good estimation. The average errors for peak noise and noise width are just 2.2% and 4.53%, respectively.

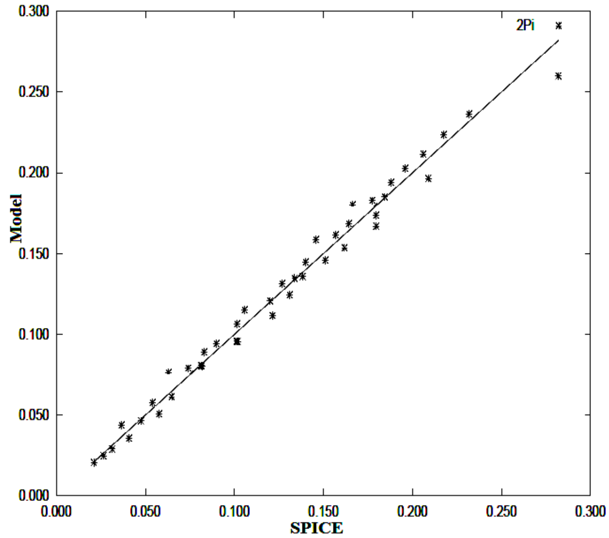


Fig. 4: Comparison of 2-π model versus HSPICE simulation for 25 randomly generated RC trees for (a) peak noise

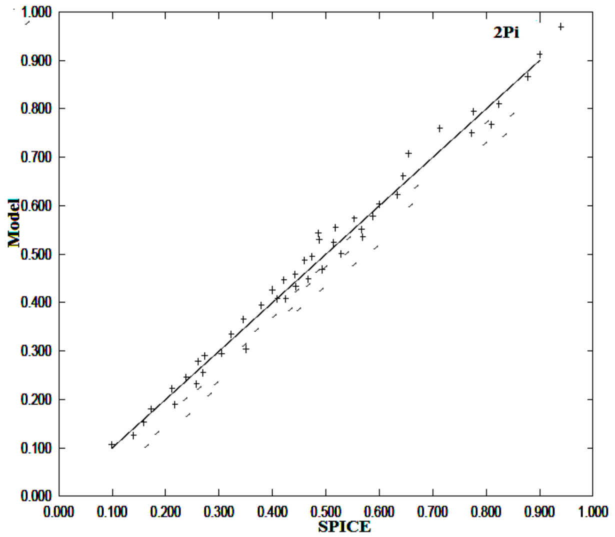


Fig. 5: Comparison of 2-π model versus HSPICE simulation for 25 randomly generated RC trees for noise width

IV. CONCLUSIONS

This paper presents a much improved 2-π crosstalk noise model for RC interconnects. Error presented by this model is less than 6% on average compared with HSPICE simulation, for both peak noise voltage and noise width estimation. In this paper unit step input is used for aggressor which is present near the victim net. The 2-π model explained above will be useful in much other application at various levels to guide noise aware DSM circuit.

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