Journal of Electron Devices, Vol. 20, 2014, pp. 1755-1760

Journal of Electron Devices
www.jeldev.org

CROSSTALK NOISE REDUCTION USING WIRE SPACING IN VLSI RC GLOBAL INTERCONNECTS

Anushree[1] ,Vikas Maheshwari[2]

Deptt.of ECE, Hindustan College of Science and Technology, Mathura, U.P., INDIA[1]
Deptt. Of ECE,Apeejay Satya University, Sohna Gurgaon[2]
anushree.ece@gmail.com

Received 27-07-2014, online 3-09-2014

ABSTRCAT

This paper presents a closed form 2π crosstalk noise model for on-chip VLSI RC interconnects. It considers a case when unit step input is applied to the aggressor which is adjacent to the victim net and producing crosstalk noise effect over it. In the next section closed form formulae for crosstalk noise pulse width and noise amplitude for RC interconnect have been derived for mathematical implementation of this crosstalk noise. Further, we consider the effect of separation between aggressor and victim lines over crosstalk noise voltage expressions (peak value and width).

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

I. INTRODUCTION

Due to advancement in VLSI technology, feature size is reduced which creates the crosstalk noise problem and also affects the design's timing and functionality goals [1-2].It introduces an effect in VLSI circuit known as crosstalk noise. This crosstalk noise is produced due to the parasitic coupling between interconnects, which degrades the performance and reliability of the circuit. It exhibits a negative impact on the performance of the VLSI circuits. These coupling capacitances may induce unwanted voltage spikes in neighboring nets. By analyzing previous year's trends, the role of this coupling capacitance will be more dominant in the future as feature sizes shrink [3]. In VLSI circuits it is very common that two wires run parallel to each other. One of them known as aggressor and other one is known as victim. The net on which noise is being induced is called the victim net; whereas, the net that induces this noise is known as aggressor net. Victim net and may be modelled as being connected with the help of a distributed coupling capacitance. For crosstalk noise estimation and avoidance problems, various phenomenon, techniques and tools should be incorporated in IC design cycle at the early stages.

With the help of some telegraph equations directly, a set of analytical formulae for peak noise of coactively coupled bus structures were obtained [4,5]. The model presented in [6] considers an Elmore delay like peak noise model for general RC trees but it assumes an infinite ramp input which causes over-estimates the peak-noise for small aggressor slews and large victim nets, which is very likely to occur in today's deep submicron designs. The peak noise obtained in [6] may even be larger than the supply voltage. Devgan's metric has been improved in [7].

The model proposed in [4] gives a closed form formula for RC line with practical boundary conditions. Later on this 2π model is used to find the closed form expression for peak value of crosstalk noise voltage in [8] and further it can be used for calculating crosstalk noise width. This is useful in calculating the values of noise amplitude and noise width. The closed form expression obtained for 2-π model is much better in comparison to the models proposed by Devgan [6] and Vittal [9]. Paper [9] is capable to handling distributed RC network and saturated ramp input. In [10], an improved 2-π model is introduced which considers coupling locations of aggressor and victim nets. Further [11] presents a closed form noise expression when step input is applied at aggressor net. This paper presents a noise reduction technique by varying the spacing between aggressor and victim nets using different sensitivity expressions.

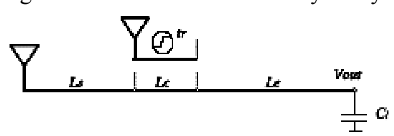
II. PROPOSED CROSSTALK NOISE MODEL

In this section, we first present the 2π model approach and derive its analytical time-domain waveform when unit step input is applied to the aggressor net which is running parallel to the victim net and producing crosstalk noise effect over it. 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.

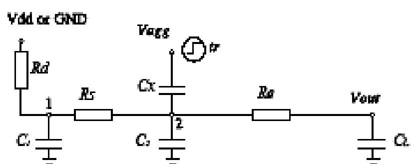
II.1 2-π Model and its analytical waveform

For explaining the crosstalk noise model, the 2-π 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 location. Similarly, L_c and L_e represent the interconnect length of victim net at the coupling and after the coupling location, respectively. 2-π model is reduced to a 2-π type RC model as shown in Figure 1(b). This reduction helps so much while calculating the value of crosstalk noise at the receiver's 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. Modelling of victim net is done by an effective resistance R_d and other RC parameters such as C_x, C_1, R_s, C_2, R_e and C_L. Figure 1(b) helps in the finding the values of these parameters 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)

Fig.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,

$$\frac{1}{Z_1} = \frac{1}{R_d} + sC_1 . \quad (1)$$

Then at node 2, we have

$$\frac{1}{Z_2} = \frac{1}{(Z_1 + R_s)} + sC_2 + \frac{1}{R_e + \frac{1}{sC_L}} . \quad (2)$$

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

$$V_2(s) = \frac{Z_2}{Z_2 + \frac{1}{sC_L}} V_{agg}(s) . \quad (3)$$

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

$$V_{out}(s) = V_2(s) \frac{\frac{1}{sC_L}}{R_e + \frac{1}{sC_L}} . \quad (4)$$

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

$$V_{out}(s) = \frac{a_2 s^2 + a_1 s}{s^3 + b_2 s^2 + b_1 s + b_0} V_{agg}(s) \quad (5)$$

where the co-efficients are,

$$a_2 = \frac{K_1}{K_2} \quad (6)$$

$$a_1 = \frac{(R_d + R_s)C_x}{K_2} \quad (7)$$

$$b_2 = \frac{(C_2+C_x)(R_e C_L (R_d+R_s) + R_d R_s C_1) + R_d R_e C_1 C_L + C_L R_d R_s C_1}{K2} \quad (8)$$

$$b_1 = \frac{(R_d + R_s)(C_x + C_2 + C_L) + (R_e C_L + R_d C_1)}{K_2} \quad (9)$$

$$b_0 = \frac{1}{K_2} \quad (10)$$

$$K_1 = C_x R_d R_s C_1 \quad (11)$$

$$K_2 = R_d R_s C_1 C_L R_e (C_x + C_2) . \quad (12)$$

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

$$H(s) = \frac{a_2 s^2 + a_1 s}{s^3 + b_2 s^2 + b_1 s + b_0} \equiv \frac{k_1}{s - s_1} + \frac{k_2}{s - s_2} + \frac{k_3}{s - s_3} . \quad (13)$$

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 $f_i(t) = k_i e^{s_i t}$ with (i=1, 2, 3).

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

$$V_{agg} = \begin{cases} 1 & t \geq 0 \\ 0 & \text{otherwise} \end{cases} . \quad (14)$$

If we find out its laplace transform it will be,

$$V_{agg}(s)=\frac{1}{s} \quad . \tag{15}$$

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

$$V_{out}(s)=\frac{k_i}{(s-s_i)}V_{agg}(s) \quad . \tag{16}$$

From (15) and (16),

$$V_{out}(s)=-\frac{k_i}{s_i}\frac{1}{s}+\frac{k_i}{s_i}\frac{1}{(s-s_i)} \quad . \tag{17}$$

So, the time domain expression of the output voltage is

$$v_{out}(t)=-\frac{k_i}{s_i}(1-e^{s_i t}) \text{ (for i=1, 2,3)} . \tag{18}$$

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

$$v_{out}(t)=v_{out1}(t)+v_{out2}(t)+v_{out3}(t) \quad . \tag{19}$$

The 2-π model has been tested extensively and its waveform can be shown as explained in next sections.

II.2 Closed form noise amplitude and width

Equation (19) 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 (5) using dominant-pole approximation method,

$$V_{out}(s)\approx\frac{a_1 s}{b_1 s+b_0}.V_{agg}(s) \quad . \tag{20}$$

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

So,

$$V_{out}(s)\approx\frac{t_x}{t_v\left(s+\frac{1}{t_v}\right)} \tag{21}$$

where the co-efficient are,

$$t_x=(R_d+R_s)C_x \tag{22}$$

$$t_v=(R_d+R_s)(C_x+C_2+C_L)+(R_eC_L+R_dC_1) \quad . \tag{23}$$

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 expression shown in (21) can be expressed in time domain and given in (24).

$$v_{out}(t)\approx\frac{t_x}{t_v}e^{-t/t_v} \tag{24}$$

where $t \geq 0$.

From the noise expression shown in (24), 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 (24):

$$v_{\max}\approx\frac{t_x}{t_v} \quad . \tag{25}$$

Equation (25) represents the maximum amplitude of noise which is obtained at t=0.

II.2.2 Noise 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 (24),

$$v_{out}(t)\approx\frac{t_x}{t_v}e^{-\frac{t}{t_v}} \tag{26}$$

and

$$t=t_v\ln\frac{1}{v_{out}}.\frac{t_x}{t_v} \quad . \tag{27}$$

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

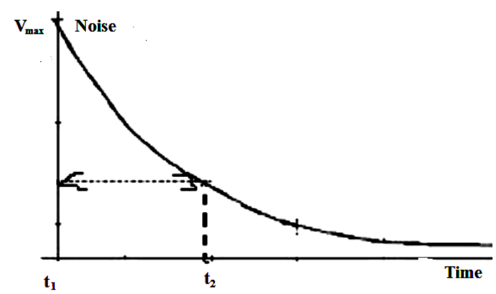


Fig.2: Illustration of the noise width

The value of t_2 can be calculated using (27). Hence, t_2 can be derived as

$$t_2=t_v\ln\frac{1}{v_{out}}\frac{t_x}{t_v} \quad . \tag{28}$$

At t_2, the noise voltage is v_t.

and $$t_2=t_v\ln\frac{1}{v_t}\frac{t_x}{t_v} \quad . \tag{29}$$

The noise width is given by

$$t_{width} = t_2 - t_1 \tag{30}$$

$$t_{width} = t_v \ln \frac{1}{v_t} \frac{t_x}{t_v} \tag{31}$$

This paper assume the value of threshold voltage v_t to be half of the value of the peak noise voltage v_{max}.

$$v_t = \frac{v_{max}}{2} \tag{32}$$

then (25), (31) and (32),

$$t_{width} = t_v \ln 2 \tag{33}$$

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 statement can be expressed clearly by some noise amplitude versus noise width plots.

III. NOISE AVOIDANCE TECHNIQUE

In this section, we investigate the noise avoidance with the help of this model. This paper presents wire spacing technique for reducing the crosstalk noise effect on victim net produced by aggressor net and evaluate various sensitivity expressions representing it.

Wire Spacing- Wire spacing is a method which can be helpful for the reduction in value of coupling capacitance. This further helps in the maximum noise and noise width reduction with an area penalty. As spacing between two wires increase some of the field lines contributing to coupling capacitance fail to reach the neighbour wire (victim net) and start contributing to ground capacitance. If we consider a wire of fixed length, its coupling capacitance decreases while its ground capacitance increases, as its spacing to a neighbor wire increases. On the other hand, coupling capacitance of the wire increases while ground capacitance decreases, when its spacing to a neighbour wire decreases. From our model

$$\frac{\partial v_{max}}{\partial C_x} = (R_d + R_s) \frac{t_v - t_x}{t_{v^2}} \tag{34}$$

$$\frac{\partial v_{max}}{\partial C_1} = -\frac{t_x}{t_{v^2}} . R_d \tag{35}$$

$$\frac{\partial v_{max}}{\partial C_2} = -\frac{t_x}{t_{v^2}} .(R_d + R_S) \tag{36}$$

$$\frac{\partial v_{max}}{\partial C_L} = -\frac{t_x}{t_{v^2}} .(R_d + R_s + R_e) \tag{37}$$

$$\frac{\partial t_{width}}{\partial C_x} = (R_d + R_s) \tag{38}$$

$$\frac{\partial t_{width}}{\partial C_1} = R_d \tag{39}$$

$$\frac{\partial t_{width}}{\partial C_2} = (R_d + R_s) \tag{40}$$

$$\frac{\partial t_{width}}{\partial C_L} = (R_d + R_s + R_e) \tag{41}$$

By considering the signs of sensitivity expressions, effect of varing wire spacing (aggressor net and victim net) on crosstalk noise amplitude and width can be examined.

IV. SIMULATION RESULTS AND DISCUSSIONS

Equations (25) and (33) represents the analytical formulae for peak noise and noise width for 2π model. These expressions have been tested extensively and the efficacy of the proposed models are justified by comparing it with HSPICE simulations. 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 to obtain high fidelity and to detect the corner scenarios,. For the test circuits, the driver resistance Rd is from 20 to 2000 Ω·, the loading capacitance CL is from 4 to 50fF, the length parameters Ls, Lc, and Le 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.

Our proposed model is compared with Vittal [3] named as V-pay and SPICE.

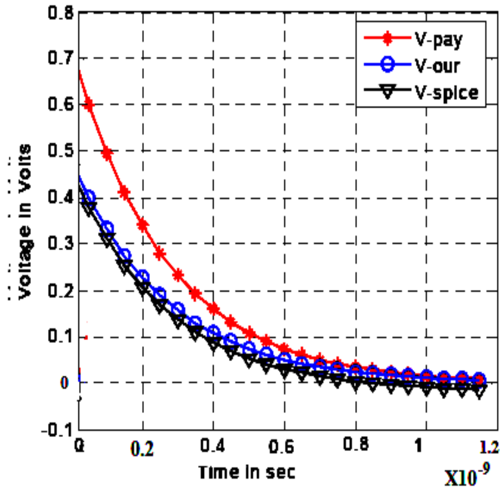


Fig..3: .Crosstalk Noise waveform for two coupled transmission lines

Our 2π model is also tested on a set of randomly generated multiple-pin nets with general RC tree structures. Experimental results show that 2-π model still works better for general RC trees.Figure 3 represents Crosstalk Noise waveform for two coupled transmission lines Also Figure 4(a) and Figure 4(b) shows the scatter diagram comparing the 2-π model (y-axis) with HSPICE (x-axis) simulations for 25 randomly generated four-pin nets for peak noise and noise width respectively.

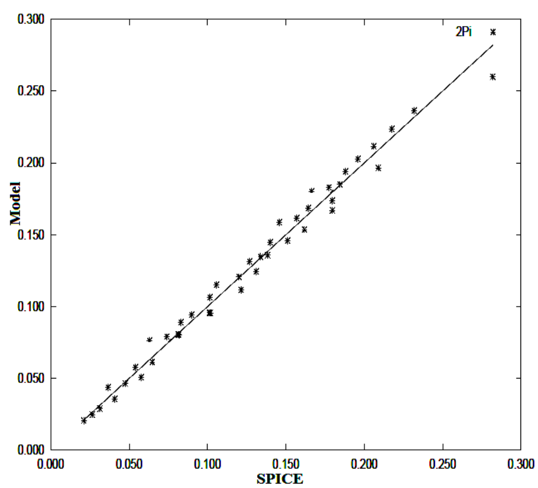


Fig. 4(a): Comparison of 2-π model versus HSPICE simulation for 25 randomly generated RC trees forpeak noise

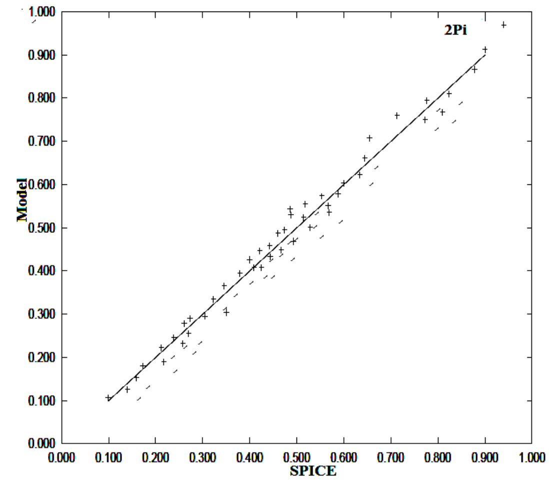


Fig.4(b):. Comparison of 2-π model versus HSPICE simulation for 25 randomly generated RC trees for noise width

The branching wire length ranges from 1 to 1500 μm considering branching location anywhere from driver to receiver. HSPICE simulations are performed on distributed RC networks (long wire) which is divided into wires each segment is of 10μm. In case of all test circuits, the 2-πmodel gives average errors for peak noiseand noise width are just 2.2% and 4.53%, respectively.

V. CONCLUSIONS

This paper presents a much improved 2-π crosstalk noise model and noise avoidance by varying the wire (aggressor and victim nets) spacing for RC on-chip VLSI interconnects. Error presented by this model is less than 5% on average compared with SPICE simulation, for both peak noise voltage and noise width. Along this, the effect of wire spacing variation can be explained by different sensitivity expressions. In this paper, we considered the unit step input for aggressor which is placed near the victim net. The 2-π model presented in this paper will be useful in other applications at various levels to help noise aware DSM circuit.

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