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EXPLICIT MODELS FOR RLC TRANSIENT DECOUPLED INTERCONNECT CROSSTALK GLITCHES AND DELAY NOISE EFFECTS USING 90NM PROCESS TECHNOLOGY

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ABSTRACT
Performance of DSM chips has suffered due to the resultant noise effects. Noise effects are result of couplings between adjacent neighboring wires. The need is to calculate the effect of all the produced noises. It is done by carefully analyzing the single wire simultaneously with all the couplings. This paper unveils a decoupled RLC transient equivalent model for victim net which results in accurate and flexible calculations. The proposed model can also compute the delay and noise effects for different slew rates and time delays of aggressor's signal. Computation of maximum crosstalk glitch effect on victim net under different combinations of input signals can also be performed. PSPICE simulation results have raised a strong case. It has ensured the accuracy and flexibility of equivalent victim's interconnect model with higher simulation speed.
Keywords: noise, delay effects, wires, transient response, victim model.

I. INTRODUCTION

Interconnect performance has experienced a major blow with the sneaking of inductive noise effects as copper becoming an adopted interconnect metal and highly increasing frequency of clock of DSM chips. Therefore, the use of interconnect model of RC has got degraded effectively with the introduction of inductive noise effects [1, 2]. RC interconnect model prove to be a failure for computing the effects of glitch and delay happening due to the phenomenon of crosstalk. In order to capture the high frequency effect such as undershoot, overshoot and ringing, the interconnect is modeled as distributed RLC network [3-5] and the accuracy in performance estimation of interconnect eventually got improved. The paper gives a mathematical equivalent RLC interconnect model for computation of glitch and delay noise effects with high precision. Simultaneous observation is required for both victim and aggressor signals on the interconnect. A highly simplified method has been put forth for the same where circuit is decoupled and the analysis of victim wire is done for all the noise effects that occur on it. Decoupling phenomenon has been discussed in the past for traditional RC interconnects by [6, 7] to approximate the results of effects of crosstalk due to capacitance. But surprisingly, decoupling of inductance is still the area untouched by [6,7].The paper thence keeps a mathematical 'equivalent decoupled interconnect model' considering all the noise effects which can be produced in the adjacent aggressor-victim lines due to mutual capacitance and mutual inductance. A decoupled equivalent model is also laid forward in the paper for the n^{th} block, computing all the noise effects for n blocks in an interconnect. Also, the victim line (victim's voltage) is represented as a function of aggressor's line (aggressor's voltage). By using this model, simulation and analysis of glitch due to crosstalk and delay noise effects on interconnect can be done successfully by feeding different combinations of input signals [8]. There are several approaches [9-13] proposed to estimate the crosstalk issue in on-chip interconnect where the interconnect is modeled as distributed RLC segment. A mathematical proof is also obtained and set forth through which the slew rates and delay effects on the decoupled victim wire can be computed easily with high accuracy and precision to say the least. The paper has successfully constructed an explicit interconnect model which gives the result accurately.

The rest of the paper is organized as follows. First the equivalent victim model is introduced along with the derivation of the victim's output voltage in Laplace domain. The time domain response for victim's output is then calculated in section III. Section IV gives the mathematical proof of computing the crosstalk glitch noise effect using this model and how this model can be used for measuring glitch and delay effects for various slew and delay combinations of aggressor signal. In the last section V, we compare the simulation results from PSPICE with our proposed equivalent victim model.

II. EQUIVALENT VICTIM MODEL

An interconnect model for single aggressor single victim is shown in the Fig.1 where following parameters are used during construction of the model such as r, l, c representing the per unit length resistance of interconnect, self-inductance and capacitance values respectively. The coupling capacitance employed in between the two interconnects is denoted as c_m & the mutual inductance by 'm', the coupling factor. Within the interconnect model, a part functions as the input driver and the other as the output driver.

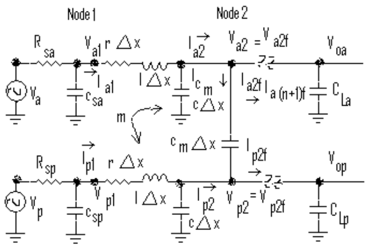

Fig.1: Interconnect Model for Single Aggressor Single Victim

When representing the voltage of victim as a function of aggressor's voltage, as a result the following cases occur. In the first case, both victim and aggressor are in the rising state, in second case aggressor is rising while the victim is falling, in the third case the opposite happens, i.e. aggressor falls while the victim rises. In the final case, both aggressor and victim experience a fall. Representing the voltages of aggressor and victim in the Laplace domain for the first case in which both aggressor and victim are rising, i.e. same state $v_p(s) = v_a(s)$ .Let us consider an exponential rise and fall time for the second and third cases in which the victim's voltage abide by the given relation-$v_a(t) + v_p(t) = v_{dd}$.The stated relation can be further simplified to obtain the value of victim's voltage as $v_p(t) = v_{dd} - v_a(t)$.The given relation for victim's voltage is represented by the following relation in the Laplace domain $v_p(s) = -v_a(s)$, where the adjustment is made in the time domain for the effect of constant.

$$v_p(s) = kv_a(s) \tag{1}$$

For opposite transitions, i.e. rising and falling & vice versa, the value of k is -1 while for the similar transitions, i.e. rising/falling for both, the value of k is +1.

As shown in the Fig.1, the infinitesimal part Δx is assumed for each interconnects. The victim's and aggressor's voltage can be written in the Laplace domain as-:

$$v_{a1} = v_{a2} + (r+sl)\Delta x I_{a1} + sml\Delta x I_{p1} \tag{2}$$

$$v_{p1} = v_{p2} + (r+sl)\Delta x I_{p1} + sml\Delta x I_{a1} \;. \tag{3}$$

Using the value of equation (1) in equation (3), we get

$$kv_{a1} = kv_{a2} + (r+sl)\Delta x I_{p1} + sml\Delta x I_{a1} \tag{4}$$

Multiplying equation (2) by k, we obtain

$$kv_{a1} = kv_{a2} + (r+sl)k\Delta x I_{a1} + smlk\Delta x I_{p1}$$

Now subtracting it by equation (4), we get

$$(r+sl)\Delta x k I_{a1} - sml\Delta x I_{a1} = (r+sl)\Delta x I_{p1} - sml\Delta x I_{p1}$$

Or, $I_{a1}\left((r+sl)k\Delta x - sml\Delta x\right) = I_{p1}\left((r+sl)\Delta x - smlk\Delta x\right)$

$$I_{a1} = \left[\frac{(r+sl) - skml}{k(r+sl) - sml}\right] I_{p1} \tag{5}$$

By putting equation (5) in (3), we simplify a relation for victim's voltage as

$$V_{p1} = V_{p2} + k\Delta x \left\{ \frac{(r+sl)^2 - (sml)^2}{k(r+sl) - sml} \right\} I_{p1} \tag{6}$$

Equation (6) can be simplified as k can only have 2 values

$$V_{p1} = V_{p2} + \Delta x \left(r + sl\left(1 + km\right)\right) I_{p1} \;. \tag{7}$$

The current at Node 1 of victim can be given as-:

$$I_{p1} = sc\Delta x V_{p2} + I_{p2} \;. \tag{8}$$

Using equation (8) in (7), we have

$$V_{p1} = \left[1 + (sc)\left(r + sl(1+km)\right)(\Delta x)^2\right] V_{p2} + \left(r + sl(1+km)\right)\Delta x I_{p2} \tag{9}$$

Now, representing the current at Node 2 on aggressor and victim sides as-:

$$I_{a1} = sc_m\Delta x\left(V_{a2} - V_{p2}\right) + I_{a2f} \tag{10}$$

$$I_{p2} = -sc_m\Delta x\left(V_{a2} - V_{p2}\right) + I_{p2f} \;. \tag{11}$$

Since k can only possess 2 values, the current at Node 2 as given by the equation (11) can be simplified as-:

$$I_{p2} = sc_m(1-k)\Delta x V_{p2} + I_{p2f} \;. \tag{12}$$

Now using the equation (12) in (9); we obtain a simplified equation which looks like:

$$V_{p1} = \left[1 + \left((sc) + sc_m(1-k)\right)\left(r + sl(1+km)\right)(\Delta x)^2\right] V_{p2} + \left(r + sl(1+km)\right)\Delta x I_{p2f} \;. \tag{13}$$

To obtain the current at Node 1, we substitute equation (12) in (8) and get-:

$$I_{p1} = \left((sc) + sc_m(1-k)\right)\Delta x V_{p2} + I_{p2f} \;. \tag{14}$$

As shown in the Fig.1 also, $V_{p2} = V_{p2f}$; the voltage and current at Node1 can be represented in terms of Node2 [14] as-:

$$\begin{bmatrix} V_{p1} \\ I_{p1} \end{bmatrix} = \begin{bmatrix} 1 + \left((sc) + sc_m(1-k)\right)\left(r + sl(1+km)\right)(\Delta x)^2 & \left(r + sl(1+km)\right)\Delta x \\ \left((sc) + sc_m(1-k)\right)\Delta x & 1 \end{bmatrix} \begin{bmatrix} V_{p2f} \\ I_{p2f} \end{bmatrix}$$

The equation stated above is in the form

$$\begin{bmatrix} V_{p1} \\ I_{p1} \end{bmatrix} = \begin{bmatrix} 1 + ab & a \\ a & 1 \end{bmatrix} \tag{15}$$

where

$$a = \left(r + sl(1+km)\right)\Delta x \tag{16}$$

$$b = s\left(c + c_m(1-k)\right)\Delta x \tag{17}$$

When comparing equation (15) with ABCD model; we can obtain the following relations which can be represented as

$$L = l(1 + km)\Delta x \quad (18)$$

$$C = (c + c_m(1 - k))\Delta x. \quad (19)$$

Equation (15) can be represented as given below for n number of divisions-:

$$\begin{bmatrix} V_{p1} \\ I_{p1} \end{bmatrix} = \begin{bmatrix} 1+ab & a \\ b & 1 \end{bmatrix}^n \begin{bmatrix} V_{p(n+1)} \\ I_{p(n+1)} \end{bmatrix}. \quad (20)$$

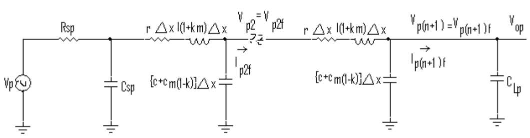

Fig.2: Equivalent victim decoupled model with the n[th] block

On simplifying the above matrix values, we obtain the following result-:

$$\begin{bmatrix} V_{p1} \\ I_{p1} \end{bmatrix} = \begin{bmatrix} M_1 & M_2 \\ M_3 & M_4 \end{bmatrix} \begin{bmatrix} V_{p(n+1)} \\ I_{p(n+1)} \end{bmatrix} \quad (21)$$

where

$$M_1 = \cosh(\theta, L) \quad (22)$$

$$M_2 = Z_1 \sinh(\theta, L) \quad (23)$$

$$M_3 = \frac{1}{Z_1}\sinh(\theta, L) \quad (24)$$

$$M_4 = \cosh(\theta, L) \quad (25)$$

with $\theta_1 = \frac{\sqrt{ab}}{x}$ $Z_1 = \frac{\sqrt{a}}{\sqrt{b}}$.

As $V_{p(n+1)} = V_{op}$ and $I_{p(n+1)} = I_{op} = sc_{Lp}V_{op}$

Hence, we can represent in terms of input parameters and output parameters by using (21), we obtain

$$V_{p1} = (M_1 + sc_{Lp}M_2)V_{op} \quad (26)$$

$$I_{p1} = (M_3 + sc_{Lp}M_4)V_{op} \quad (27)$$

where c_{Lp} is known as the capacitive load of the receiver.

We can employ the same action on the input side voltage which can be represented as-:

$$V_{ip} = (1 + sR_{sp}C_{sp})V_{p1} + R_{sp}I_{p1} \quad (28)$$

where the input driving circuit of the victim net is modeled as a voltage source which is driving a resistance R_{sp} and an output capacitance C_{sp}. After substituting the equation (26) and (27) in equation (28) and then solving it for output voltage on that side, i.e.V_{op}, we get-:

$$V_{ip} = (1 + sR_{sp}C_{sp})(M_1 + sC_{Lp}M_2)V_{op} + R_{sp}(M_3 + sC_{Lp}M_4)V_{op}$$

Or, $$V_{op} = \frac{V_{ip}}{(1 + sR_{sp}C_{sp})(M_1 + sC_{Lp}M_2) + R_{sp}(M_3 + sC_{Lp}M_4)} \quad (28.a)$$

$$V_{op} = \frac{V_{ip}}{(M_1 + R_{sp}M_3) + s(C_{Lp}M_2 + R_{sp}C_{sp}M_1 + R_{sp}C_{Lp}M_4) + s^2(R_{sp}C_{sp}C_{Lp}M_2)} \quad (28.b)$$

To obtain the solution in the time domain (since the calculation is done in Laplace domain),we need to take the fourth order approximation of M_1-M_4 since it gives a sixth order polynomial in the denominator which can be solved further resulting in a solution as stated in the following section.

III. TRANSIENT RESPONSE OF EQUIVALENT VICTIM MODEL

The transient response for the equivalent decoupled victim interconnect model can be obtained by approximating the value of k and having one of the either mentioned conditions- a rising or falling tranisition (change of state, from 0 to 1 or 1 to 0).The value of k is taken as 1 at both victim and aggressor when rising state is confronted and as a result input voltage is replaced by the relation as- $V_{ip} = V_{dd}\frac{\beta}{s(s+\beta)}$ in equation (28), where β is given as the inverse of the time constant, τ .Now the poles can be computed by using numerical ways, similarly residues can be calculated by employing the partial fraction method. Now to obtain the output voltage of victim at the receiver side, Inverse Laplace transformation is applied and hence represented in the time domain as-:

$$V_{op}(t) = \frac{V_{dd}\beta}{q}(d_1 + d_2e^{-\beta t} + d_3e^{p_1 t} + \ldots + d_8e^{p_6 t}) \quad (29)$$

where

q is coefficient of highest degree element; 6 here as sixth order polynomial is obtained, d_i's-:residues p_i's-:poles of sixth order polynomial obtained.

In the third case as described before, i.e. victim falling and aggressor rising, we add V_{dd} to output voltage at receiver side, $V_{op}(t)$ to adjust the consequences in the time domain.

IV. CROSSTALK GLITCH NOISE COMPUTATION

Since mentioned before that the interconnect model gives way to only those cases in which victim's voltage is represented as the function of aggressor's voltage. But in addition to this, we can also compute the glitch effect due to the aggressor signal in either rising or falling state on a majorly static victim signal as there exists no such clear relation between the two stated above by simple calculations on the equivalent interconnect model constructed. For single aggressor single victim, a slightly changed output voltage at victim is obtained and its mathematical proof is also generated in the following relation given below-:

$$V_{op}(s) = \frac{2T_1(s)}{T_1(s)T_4(s) - T(s)T_3(s)}V_{ia}(s) + \frac{2T_2(s)}{T_1(s)T_4(s) - T_2(s)T_3(s)}V_{ip}(s) \quad (30)$$

where following assumptions are made in order to obtain the glitch effect such as-:

$$T_1(s) = -R_{sp}\left[\left(A_3 - B_3\right) + sC_{La}\left(A_4 - B_4\right)\right] - \left[\left(A_1 - B_1\right) + sC_{La}\left(A_2 - B_2\right)\right]\left(1 + sC_{SP}R_{SP}\right)$$
$$T_2(s) = R_{sa}\left[\left(A_3 + B_3\right) + sC_{La}\left(A_4 + B_4\right)\right] + \left[\left(A_1 + B_1\right) + sC_{La}\left(A_2 + B_2\right)\right]\left(1 + sC_{sa}R_{sa}\right)$$
$$T_3(s) = -R_{sp}\left[\left(A_3 + B_3\right) + sC_{Lp}\left(A_4 + B_4\right)\right] + \left[\left(A_1 + B_1\right) + sC_{Lp}\left(A_2 + B_2\right)\right]\left(1 + sR_{sp}C_{sp}\right)$$
$$T_4(s) = R_{sa}\left[\left(A_3 - B_3\right) + sC_{Lp}\left(A_4 - B_4\right)\right] + \left[\left(A_1 - B_1\right) + sC_{Lp}\left(A_2 - B_2\right)\right]\left(1 + sC_{sa}R_{sa}\right)$$

Now the A's and B's can be represented in the matrix form by the given relation-:

$$\begin{bmatrix} A_1 & A_2 \\ A_3 & A_4 \end{bmatrix} = \begin{bmatrix} \cosh\left(\gamma_c L\right) & Z_c \sinh\left(\gamma_c L\right) \\ \frac{1}{Z_c}\sinh\left(\gamma_c L\right) & \cosh\left(\gamma_c L\right) \end{bmatrix}$$
$$\begin{bmatrix} B_1 & B_2 \\ B_3 & B_4 \end{bmatrix} = \begin{bmatrix} \cosh\left(\gamma_f L\right) & Z_f \sinh\left(\gamma_f L\right) \\ \frac{1}{Z_f}\sinh\left(\gamma_f L\right) & \cosh\left(\gamma_f L\right) \end{bmatrix}. \qquad (31)$$

By using the above parameters we obtain a glitch effect positive in nature on output of victim (input signal at victim is assumed to be zero since it is static) which is given as under-:

$$V_{opG}(s) = \frac{2V_{ia}\left[R_{sp}\left(\left(A_3 - B_3\right) + sC_{La}\left(A_4 - B_4\right)\right) + \left(\left(A_1 - B_1\right) + sC_{La}\left(A_2 - B_2\right)\right)\left(1 + sC_{sp}R_{sp}\right)\right]}{\left[R_{sp}\left(\left(A_3 - B_3\right) + sC_{La}\left(A_4 - B_4\right)\right) + \left(\left(A_1 - B_1\right) + sC_{La}\left(A_2 - B_2\right)\right)\left(1 + sC_{sp}R_{sp}\right)\right]\left[R_{sa}\left(\left(A_3 - B_3\right) + sC_{Lp}\left(A_4 - B_4\right)\right) + \left(\left(A_1 - B_1\right) + sC_{Lp}\left(A_2 - B_2\right)\right)\left(1 + sC_{sa}R_{sa}\right)\right]}$$

$$R_{sa}\left(\left(A_3 + B_3\right) + sC_{La}\left(A_4 + B_4\right)\right) + \left(\left(A_1 + B_1\right) + sC_{La}\left(A_2 + B_2\right)\right)\left(1 + sC_{sa}R_{sa}\right)\Big] \Big[R_{sp}\left(\left(A_3 + B_3\right) + sC_{Lp}\left(A_4 + B_4\right)\right) + \left(\left(A_1 + B_1\right)\right.$$

Now let us consider some equivalents on both aggressor and victim net such as $R_{sa}=R_{sp}$, $C_{sa}=C_{sp}$, & $C_{La}=C_{Lp}$.

Hence we can simplify the above relation as given below-:

$$V_{opG}(s) = \frac{V_{ia}(s)}{2}\left[\frac{1}{R_{sp}\left(A_3 + sC_{Lp}A_4\right) + \left(A_1 + sC_{Lp}A_2\right)\left(1 + sC_{sp}R_{sp}\right)} - \frac{1}{R_{sp}\left(B_3 + sC_{Lp}B_4\right) + \left(B_1 + sC_{Lp}B_2\right)\left(1 + sC_{sp}R_{sp}\right)}\right] \qquad (32)$$

Using the relations in equations (22) to (25) and considering negative value of k and hence obtaining the relation in the matrix form as-:

$$\begin{bmatrix} M_1 & M_2 \\ M_3 & M_4 \end{bmatrix} = \begin{bmatrix} B_1 & B_2 \\ B_3 & B_4 \end{bmatrix}. \qquad (33)$$

Similarly employing k=1, i.e. for both victim and aggressor in the same state, we can obtain the following relationship in matrix form-:

$$\begin{bmatrix} M_1 & M_2 \\ M_3 & M_4 \end{bmatrix} = \begin{bmatrix} A_1 & A_2 \\ A_3 & A_4 \end{bmatrix}. \qquad (34)$$

Now we consider the rising effect as to obtain a positive glitch effect noise and feed a rising exponential input signal at the input of the victim's net and by using the set of matrices obtained in equation (33), (34), we can re-write equation (32) as follows-:

$$V_{opG}(s) = \left[\left[\frac{V_{ipR}(s)}{\left(M_1 + sC_{Lp}M_2\right)\left(1 + sC_{sp}R_{sp}\right) + R_{sp}\left(M_3 + sC_{Lp}M_4\right)}\right]_{k=1} - \left[\frac{V_{ipR}(s)}{\left(M_1 + sC_{Lp}M_2\right)\left(1 + sC_{sp}R_{sp}\right) + R_{sp}\left(M_3 + sC_{Lp}M_4\right)}\right]_{k=-1}\right] \qquad (35)$$

From above equation and equation (28.a), we can conclude that output voltage glitch on the victim line is nothing but the voltage output on victim line in case of rising state, i.e. k=1 minus the voltage output on victim line when the opposite transition occurs, i.e. k=-1 and the complete equation divided by a factor of 2.

Considering a slew rate 's' and delay as 'd', the glitch noise on victim line by an aggressor signal can be represented as G_{sd}. Let us assume similar transition of aggressor and victim signals to be Si_{nsd} and opposite transition of aggressor and victim signals to be Op_{nsd} where n denotes the victim signals' state which can either be rising 'r' or falling 'f', equation (35) is simplified as

$$G_{sd} = \frac{Si_{rsd} - Op_{rsd}}{2} \qquad (36)$$

$$Si_{rsd} = 2G_{sd} + Op_{rsd} \qquad (37)$$

$$Op_{rsd} = Si_{rsd} - 2G_{sd} \qquad (38)$$

where equations (37) & (38) are the relations for computation of similar and opposite transitions on both aggressor and victim line respectively.

The above equations hence forth, can be used for calculating the glitch effect on victim signal when either of the two parameters varies, i.e. slew rate or delay.

Now, let us consider an instance where (s_1, d_1) are the slew rate and delay of the aggressor signal while (s_2, d_2) are the slew rate and delay of the victim signal. Supporting the above argument, the effect of rising aggressor on rising victim can be computed by using the following relations where slew rates and delays happen to be different, using equation (37), we get-:

$$Si_{rs_2d_2} / _{s_1d_1} = G_{s_1d_1} + G_{s_2d_2} + Op_{rs_2d_2}\,. \qquad (39)$$

Equation (39) shows that response of rising victim signal due to the rising aggressor is equal to the sum of glitch effect with slew rate s_1 and delay d_1, glitch effect with slew rate s_2 and delay d_2 and response of rising victim signal when the value of k is taken to be -1.

Similarly a relation is obtained using equation (38) where the effect of falling aggressor is seen on rising victim signal with different slew rates and delays as

$$Op_{rs_2d_2} / _{s_1d_1} = Si_{rs_2d_2} - G_{s_1d_1} - G_{s_2d_2}\,. \qquad (40)$$

Hence glitch effect on victim signal having delay d_1 and slew rate s_1 can be calculated as

$$G_{s_1d_1} = \frac{Si_{rs_1d_1} - Op_{rs_1d_1}}{2}\,. \qquad (41)$$

V. SIMULATION RESULTS AND DISCUSSION

Consider two parallel RLC interconnect lines shown in Fig.1, in 90 nm technology. All wires are 1500 µm long. Such bus structures are typical in high performance CPU designs. The extracted values [15] for the parameters R, L, C, M and C_c for the 90 nm process technologies are given in Table I.

Table 1: Extracted Values of Interconnect Parameters for 90 nm Process Technology

Parameters	Value/mm
Resistance (R)	13Ω/mm
Inductance (L)	0.16nH/mm
Capacitance (C)	0.45pF/mm
Mutual Inductance (M)	0.028nH/mm
Coupling Capacitance (C_c)	642.49fF/mm

Table 2 altercates about the comparison of Equivalent victim's proposed model and SPICE models' simulation of victim's output waveform when both aggressor and victim are rising, aggressor and victim signals have identical parameters. It is assumed that the input for all experiments is a pulse waveform with 1.0 V V_{PP} and 2ns pulse width. As can be seen from the Fig. 3, the equivalent victim model is consistent with the PSPICE waveform.

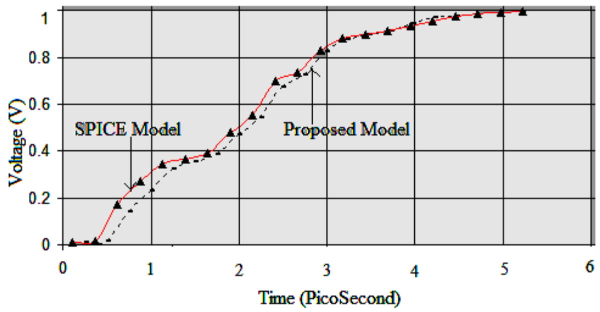

Fig.3: Comparison of Equivalent Victim's Proposed Model and SPICE Models' Simulation of Victim's Output Waveform

Table 2: Comparison of Equivalent Victim's Proposed Model and SPICE Models' Simulation of Victim's Output Waveform

Time (10^{-12} sec)	Proposed Victim Model (V)	SPICE Model (V)
0.0	0.008	0.009
0.1	0.014	0.017
0.3	0.14	0.17
0.6	0.23	0.27
0.8	0.323	0.345
1.0	0.354	0.365
1.3	0.386	0.389
1.5	0.467	0.478
1.75	0.543	0.553
2.1	0.674	0.698
2.7	0.727	0.734
2.9	0.826	0.828
3.3	0.874	0.882
3.7	0.895	0.898
3.9	0.912	0.914
4.2	0.943	0.932
4.5	0.969	0.957
4.7	0.978	0.975
4.9	0.986	0.988
5.0	0.994	0.993
5.23	0.995	0.998

The glitch phenomena on a victim wire due to rising aggressor signal is analyzed and shown in Fig.4. The simulations were also carried out by varying the mutual inductance and coupling capacitance parameters, but in all the cases the output waveform of the equivalent victim model was quite comparable with the PSPICE model. Table 3 discusses about the comparison of effect of glitch on the Equivalent victim's proposed model and SPICE models' simulation of victim's output waveform when aggressor is rising. From the Fig.4 we can analyze that the equivalent victim model is consistent with the PSPICE waveform.

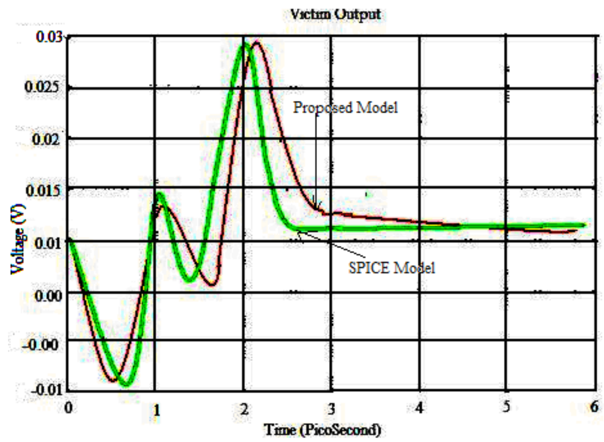

Fig.4: Comparison of Glitch Phenomena on Equivalent Victim's Proposed Model and SPICE Models' Simulation of Victim's Output Waveform

Table 3: Comparison of Glitch Phenomena on Equivalent Victim's Proposed Model and SPICE Models' Simulation of Victim's Output Waveform

Time (10^{-12} sec)	Proposed Victim Model (V)	SPICE Model (V)
0.0	0.000	-0.0002
0.1	-0.001	-0.001
0.4	-0.002	-0.004
0.6	-0.005	-0.0006
0.8	-0.0008	-0.0006
1.0	0.007	0.008
1.4	0.008	0.007
1.6	0.011	0.010
1.75	0.017	0.018
2.0	0.027	0.026
2.4	0.021	0.018
2.9	0.013	0.011
3.3	0.013	0.011
3.7	0.013	0.011
3.9	0.013	0.011
4.2	0.0125	0.012
4.5	0.0125	0.012
4.7	0.012	0.012
4.9	0.011	0.012
5.0	0.011	0.012
5.2	0.012	0.012
5.4	0.011	0.012
5.7	0.011	0.013

Fig.5 shows the effect on the rising victim due to rising aggressor having higher slew rate, the slew rate of aggressor and victim. Several other simulations were carried out for crosstalk noise by varying the slew rates and delays of aggressor signal. Table 4 discusses about the comparison of Equivalent victim's proposed model and SPICE models' simulation of victim's output waveform when aggressor is rising, aggressor and victim signals have identical parameters. As can be seen from the Fig. 5, the equivalent victim model is consistent with the PSPICE waveform.

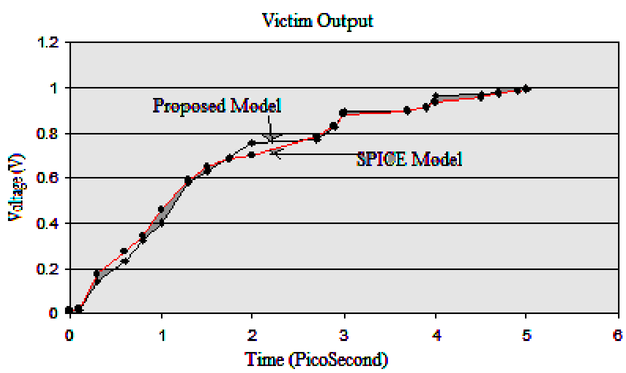

Fig.5: Comparison of Equivalent Victim's Proposed Model and SPICE Models' Simulation of Victim's Output Waveform

Table 4: Comparison of Equivalent Victim's Proposed Model and SPICE Models' Simulation of Victim's Output Waveform

Time (10^{-12} sec)	Proposed Victim Model (V)	SPICE Model (V)
0.0	0.008	0.009
0.1	0.014	0.017
0.3	0.14	0.17
0.6	0.23	0.27
0.8	0.323	0.345
1.0	0.4	0.46
1.3	0.58	0.59
1.5	0.63	0.65
1.75	0.693	0.683
2.0	0.756	0.698
2.7	0.77	0.78
2.9	0.826	0.828
3.0	0.894	0.882
3.7	0.895	0.898
3.9	0.912	0.914
4.0	0.963	0.932
4.5	0.969	0.957
4.7	0.978	0.975
4.9	0.986	0.988
5.0	0.994	0.993

VI. CONCLUSION

This paper analyzes and discusses the problem of crosstalk, glitches and delay in a decoupled RLC transient equivalent model for victim net. A mathematical expression based model is developed for a decoupled victim interconnects with all its coupling effects intact. The proposed model is also capable of computing the glitch and delay noise effects for varying slew rates and delays much faster. The SPICE simulations verify the accuracy of the equivalent victim model.

References

[1] W. C. Elmore, "The transient response of damped linear networks with particular regard to wide-band amplifiers", Journal of Applied Physics, **19**, 55–63 (1948).

[2] Kahng A B, Muddu S. "An analytical delay for RLC interconnects", IEEE Trans. Computer-Aided Design of Integrated Circuits and Systems,**16,**.1507-1514 (1997).

[3] K. Banerjee, A. Mahrotra, "Analysis of on-chip inductance effects for distributed RLC interconnects", IEEE Transactions on computer aided design of integrated circuits and systems, **21**, 904-915 (2002).

[4] Vikas Maheshwari, Shruti Gupta, Kapil Khare,Vimal Yadav,Rajib Kar, Durbadal Mandal,Anup Kr. Bhattacharjee, "Efficient Coupled Noise Estimation for RLC On-Chip Interconnect", IEEE Symposium on Humanities, Science and Engineering Research (SHUSER-2012), Kuala Lumpur, Malaysia, June 24-27, pp.1125-1129.

[5] P.V.Hunagund,A.B.Kalpana, "Crosstalk noise modeling for RC and RLC interconnects in deep submicron VLSI circuits", Journal of computing, **2**, 60-65 (2010).

[6] P. Chen, D. Kirkpatrick, K. Keutzer, "Miller Factor for Gate Level Coupling Delay Calculation", EEE/ACM International Conference on Computer-Aided Design (ICCAD), San Jose, California, USA , November 5-9 (2000), pp. 68-74.

[7] A. Kahng, S. Muddu, E. Sarto, "On Switch Factor Based Analysis of Coupled RC Interconnects", Design Automation Conference, Bergen, Norway, August 2000, pp.79-84.

[8] Shehzad Hasan, Ajoy K. Palit and Walter Anheier, "Equivalent Victim Model of the Coupled Interconnects for Simulating Crosstalk Induced Glitches and Delays", IEEE Workshop on Signal propagation on Interconnect (SPI), Strasbourg, France, 12-15 May 2009, pp. 1-4.

[9] Sourajeet Roy, Anestis Dounavis, "Efficient Delay and Crosstalk Modeling of RLC Interconnects Using Delay Algebraic Equations", IEEE Trans. VLSI System, **19**, 342-345 (2011).

[10] V. Maheshwari, K. Khare, R. Kar, D. Mandal, A. K. Bhattacharjee, "Crosstalk Noise Estimation for Generic RLC Trees

with Capacitive Coupling", Proc. IEEE Prime Asia 2012, Hyderabad, India, December 2012, pp.150-154.

[11] Vikas Maheshwari, Naomi Joshi, Er Anushree, Rajib Kar,Durbadal Mandal, Anup Kr. Bhattacharjee, “4-π Crosstalk Noise Model for Deep Submicron VLSI Global RC Interconnects”, IEEE Symposium on Humanities, Science and Engineering Research (SHUSER-2012), Kuala Lumpur, Malaysia, June 2012, pp. 355-360.

[12] Rajib Kar, V. Maheshwari, V. Agarwal, A. K. Mal, A. K. Bhattacharjee, “Modelling of RLC Interconnect Delay for Ramp Input Using Diffusion Model Approach,” 2010 IEEE Symposium on Industrial Electronics and Applications (ISIEA 2010), 3-6 October 2010, Penang, Malaysia, pp. 436-440.

[13] Susmita sahoo,Madhumanti Dutta,Rajib kar, “Accurate Crosstalk Analysis for RLC on-chip VLSI interconnect”, International Journal of Electrical and Electronics Engineering, WASET, **5**, 302-310 (2011).

[14] A. K. Palit, V. Meyer, W. Anheier, J. Schloeffel, “ABCD Modeling of Crosstalk Coupling Noise to Analyze the Signal Integrity Losses on the Victim Interconnect in DSM Chips”, 18th Int. Conf. on VLSI design, Kolkata, India, 3-7 Jan. 2005, pp. 354-359.

[15] Semiconductor Industry Association, National Technology Roadmap for semiconductors, 2012.