Analytical Crosstalk Modelling of On-Chip RLC Global Interconnects with Skin Effect for Ramp Input

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Abstract

This paper proposes a much improved, accurate, fast and simple crosstalk model for coupled interconnect line considering the skin effect. It is time for VLSI designers to examine the crosstalk noise effects in their designs, so that they are free from noise. Hence, accurate noise modelling for RLC lines is critical for timing and system integrity analysis. Skin effect alters the values of the resistance and also the inductance, which in turn affects the system integrity in particular and its response as a whole. Till now the skin effect has been neglected for modelling the on-chip interconnects. This paper addresses a novel analytical model to find the impact of skin effect on the noise variation in RLC interconnect, without considering the skin effect modelling in inductance under ramp input. In the proposed work, the resistance variation due to the skin effect is considered in a two wire transmission line model. The correlation between the skin effect and noise is also considered.

1. Introduction

In deep submicron technology, VLSI circuits usually consist of several parallel bus structures (collection of adjacent wires) which result in significant parasitic coupling effects. In recent years, increase in bandwidth requirements have led to the research into low-loss on-chip interconnects, which theoretically can achieve very high bandwidth [1]. For integrated circuits in the deep submicron (DSM) technology, interconnects play an important role in determining the chip performance and signal integrity. In deep submicron design, interconnect delay is shown to be ten to few hundred times larger than the intrinsic gate delay [2].

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In order to reduce interconnect delay, wire-sizing is found to be an effective way. On-chip interconnect analysis begins with an in-depth coverage of delay metrics, including the ubiquitous Elmore delay [3] and its many variations. The study and analysis of interconnect line has become very important issue because as integrated circuit feature sizes continue to scale well below 0.18μm, active device counts are reaching hundreds of millions [4]. The amount of interconnects among the devices tends to grow super linearly with the transistor counts, and the chip area is often limited by the physical interconnect area. Due to these interconnect area limitation, the interconnect dimensions are scaled with the devices whenever possible. Wire sizing [5] is found to be effective in reducing interconnect delays. Continuous wire sizing is a well suited approach which describes the wire by a continuous shape function. Studies on Elmore delay model have found that optimal shape function is exponential or near exponential. As frequency increases, current density within the conductor varies in such a way that it tends to exclude magnetic flux inside the conductor. This situation results in an apparent increase in resistance of the conductor because maximum current is concentrated near the surface and edges of the conductor, and it also causes the effective inductance of the conductor to decrease as frequency increases. As the technology has started working on the high frequencies, high frequency effects like skin effect and proximity effect have become significant. The reason behind the importance of considering such effect is that, these effects alters the system integrity at large scale.

2. Basic Theory

This is a well know fact that propagation delay increases with the skin effect in an on-chip interconnect. Skin effect will also affect the optimality of the system. The skin effect can be represented at the circuit level as a combination of frequency dependent resistance and inductance. However, frequency dependent circuit elements are not suitable for time-domain analysis, therefore, a circuit representation based on frequency independent elements is desirable [6]. When a transmission line model is necessary, either a SPICE compatible distributed RLC model is used or else a full transmission line model is needed. Which is chosen based on the accuracy needed and also the capability of an available circuit simulation program [7]. A wide range of models are used for interconnects depending on:

- The accuracy required nets carrying analog signals need to be modelled more,
- The amenability of the net for modelling,
- The frequency of operation.

As frequency increases, due to the magnetic field inside the conductors, the inner most shells gradually turn off and only the outer shells stay active, thus increasing the resistance and decreasing the internal inductance. This can also be achieved by parallel combination of impedance branches, where each branch will have a resistance and inductance in series as shown in the Figure (1) [8]:

The Skin Effect is the tendency of high frequency current to concentrate near the outer edge or surface of a conductor, instead of flowing uniformly over the entire cross-sectional area of the conductor. The higher the frequency, the greater the tendency for this effect to occur. There are three possible reasons we might care about skin effect. The resistance of a conductor is inversely proportional to the cross-sectional area of the conductor. If the cross-sectional area decreases, the resistance goes up. The skin effect causes the effective cross sectional area to decrease. If we are concerned about controlled impedance traces and transmission line considerations, the skin effect causes trace termination techniques to become much more complicated. If the skin effect causes the effective cross sectional area of a trace to decrease and its resistance to increase, then the trace will heat faster and to a higher temperature at higher frequencies for the same level of current [9]. There are a number of approaches available where the on-chip interconnect is modelled as distributed RLC segments for accurate performance parameters modelling [10-14]. But these models do not consider the high frequency skin effect phenomena.
3. Proposed Work

The proposed work is divided in two sub-sections. In the first sub-section, a novel analytical crosstalk model of RLC interconnect is proposed which does not included the skin effect; whereas, in the second sub-section, the skin effect is considered for interconnect modelling through simulation of the proposed RLC model. Note that the skin effect makes an adverse effect on the resistance and the inductance. However, this paper only considered the skin effect onto the resistance. The resistance increases with the skin effect; whereas, a decrease in the inductance is accounted.

3.1 Crosstalk Modelling of RLC Interconnect Analysis Without Skin Effect

In this section a new analytical model of RLC interconnects is proposed. This analysis considers the following interconnect coupling circuit:

![Analytical Model of RLC interconnect](image)

Fig. 1. Analytical Model of RLC interconnect

Applying the simple loop analysis, the following equations are obtained in terms of RLC:

For first loop:

\[ V_{in}(t) = I_1(t)R_a + L_c \left( \frac{d}{dt} I_1(t) \right) + \frac{1}{C_c} [I_1(t) - I_2(t)] \]  

(1)

For second loop:

\[ \frac{1}{C_c} [I_2(t) - I_1(t)] - \frac{1}{C_a} I_2(t) = 0 \]  

(2)

For third loop:

\[ R_c I_3(t) + R_c I_1(t) + L_c \frac{d}{dt} I_3(t) + \frac{1}{C_c} I_1(t) = 0 \]  

(3)

Taking Laplace Transform of (1) - (3), the following equations are obtained:

\[ V_{in}(s) = I_1(s)R_a + sL_a I_1(s) + \frac{[I_1(s) - I_2(s)]}{sC_c} \]  

(4)

\[ \frac{1}{sC_c} [I_2(s) - I_1(s)] - \frac{1}{sC_a} I_2(s) = 0 \]  

(5)

\[ R_c L_3(s) + R_c I_3(s) + sL_a I_3(s) + \frac{1}{sC_v} I_3(s) = 0 \]  

(6)

From (5), the following can be derived:

\[ \left[ \frac{1}{sC_v} - \frac{1}{sC_a} \right] I_2(s) = -\frac{1}{sC_c} I_2(s) \]

or,

\[ I_2(s) = I_1(s) \left( \frac{1}{sC_v} - \frac{1}{sC_a} \right) \]

so,

\[ I_2(s) = I_1(s) \left( \frac{C_a}{C_a - C_c} \right) \]

From equation (4),
\[ V_x(s) = I_1(s)R_x + sL_xI_1(s) + \frac{1}{C_v}I_1(s) - \frac{1}{C_v}I_x(s) - \frac{C_s}{C_v - C_s} \]  
(8)

or,
\[ V_x(s) = I_1(s)\left[R_x + sL_x + \frac{1}{C_v}\left(1 - \frac{C_s}{C_v - C_s}\right)\right] \]
(9)

so,
\[ I_1(s) = \frac{V_x(s)}{R_x + sL_x + \frac{1}{C_v}\left(1 - \frac{C_s}{C_v - C_s}\right)} \]
(10)

With the simple loop analysis it can be found that \[ I_3 = |I_1 - I_2| \]

Therefore, using \[ I_1 \] and \[ I_2 \], the third current can be derived which is given as,

\[ I_3(s) = V_x(s)C_v \left[R_x + sL_x + \frac{1}{C_v}\left(1 - \frac{C_s}{C_v - C_s}\right)C_v - C_x\right] \]
(11)

Applying the current divider rule, the current in the victim line capacitor may be found. This finally yields to \[ I_x = I_1R_2/(R_x + R_2) \]

where \[ R_1 = (R_x + R_y + sL_x) \] or, \[ R_2 = \frac{1}{sC_v} \]

\[ I_x = \frac{V_x(s)C_v}{R_x + sL_x + \frac{1}{C_v}\left(1 - \frac{C_s}{C_v - C_s}\right)C_v - C_x} \]
(13)

The output voltage is given as,
\[ V_{xv}(s) = I_x \times \frac{1}{sC_v} \]
(14)

From (13) and (14) we have,
\[ V_{xv}(s) = \frac{V_x(s)}{sC_v}\left[C_v(R_x + sL_x) + \frac{1}{C_v}\left(1 - \frac{C_s}{C_v - C_s}\right)C_v - C_x\right] \]
(15)

Some important assumptions that have been made in this paper are as follows:

\[ R_y + R_v = A; \quad R_x + \frac{1}{C_v}\left[1 - \frac{C_s}{C_v - C_s}\right] = B \]
(16)

For ramp input, \[ V_{in} = V_0/s^2 \]
(17)

Using the above assumptions and applying partial fraction theory yields,
\[ V_{xv}(s) = \frac{V_x}{s^2C_v}\left[\frac{C_v(A + sL_x)}{B + sL_v + 1 + A + sL_x}\right] \]
(18)

\[ S(s) = \frac{F_1}{s} + \frac{F_2}{s^2} + \frac{F_3}{s^3} + \frac{F_4}{s^5} + \frac{F_5}{s^7} \]
(19)

\[ F_5 = (s(s)^*)|_{s=\omega} = \left[\frac{C_v(A + sL_x)}{(B + sL_v)(1 + A + sL_x)}\right]|_{s=\omega} = \frac{C_vA}{B(A + 1)} \]
(20)

\[ F_4 = -F_2L_x + F_3 L_x \]
(21)

\[ F_4 = B + AB + F_1(B + L_x + L_v + AL_x) = C_vL_x \]
(22)

\[ F_3 = \left(C_vL_x - F_1(B + L_x + L_v + AL_x)\right) = \frac{C_vL_x}{B(1 + A)} - \frac{C_vA(B + L_x + L_v + AL_x)}{[B + AL_x]^2} \]
(23)
Taking the Laplace inverse transforms of (18) and with the help of (19)-(24), one can derive the explicit expression for $V_{co}(t)$ and it is given in (25).

$$V_{co}(t) = \frac{V_o}{(C_C - C_s)}(\alpha_t + \beta_t + \chi_t + \zeta_t)$$

where

$$\alpha_t = \left[ C \left( \frac{A - B}{L_s} L_t \right) \left( \frac{C}{A} \right) - \frac{C_s A}{C C_s} \right] \mu(t)$$

$$\beta_t = \left[ \frac{C_s A}{C C_s} \right] \left( 1 + A + \frac{A_L}{L_s} \right)$$

$$\chi_t = \frac{C_s A}{C C_s} \left( \frac{A + 1}{B + 1} \right) \frac{A_L}{L_s}$$

$$\zeta_t = \frac{C_s}{(A + 1)} \frac{A_L}{L_s} \left( \frac{1}{L_s} \right) e^{-\frac{1}{L_s}}$$

The equation (25) describes the coupling noise voltage without the presence of skin effect phenomena.

3.2 Crosstalk Modelling of RLC Interconnect With Skin Effect

The skin effect is the tendency of high frequency current density to be highest at the surface of a conductor and then to decay exponentially towards the centre [15]. The resistance of a conductor is inversely proportional to the cross sectional area of the conductor. If the cross sectional area decreases, the resistance goes up. The skin effect causes the effective cross sectional area to decrease. Therefore, the skin effect causes the effective resistance of the conductor to increase [16]. It is simple to analyze the skin effect in a homogeneous conducting half space with the current parallel to the interface. Let the current is in z-direction and y axis is normal to the interface as shown in the Figure 2.

![Fig. 2. Homogeneous conducting half space](image)

If the angular frequency of the current is $\omega$, and the medium has a conductivity $\sigma$, and permeability $\mu$, the complex current density is found to be $J_y(y) = J_{y0} e^{-\kappa y}$, where, $k = \sqrt{\frac{\omega \sigma \mu}{2}}$.

The intensity of the current density vector decrease exponentially with increasing $y$. At a distance $\delta'$,

$$\delta' = \frac{1}{k} = \frac{2}{\omega \sigma \mu}$$

The amplitude of the current density vector decreases 1/e of its value $J_{y0}$ at the boundary surface. This distance is known as the “skin depth”. Skin depth for Copper is $\sigma = 57 \times 10^6 S/m, \mu = \mu_0$ for iron $\sigma = 10^7 S/m, \mu = 1000$, for sea water $\sigma = 4S/m, \mu = \mu_0$, for wet soil $\sigma = 0.01S/m, \mu = \mu_0$.

It is mentioned earlier that we have only considered the skin effect on resistance. So for calculating the total effect on the impedance we have to calculate the output impedance.

From the Figure 1,
The total output resistance in RLC interconnects is given as bellow

\[
Z_o = \left[ \frac{\left( R + R_c + sL_c \right) \frac{1}{sC_c}}{R + R_c + sL_c + \frac{1}{sC_c}} \right] \left[ \frac{\left( R + R_s + sL_s \right) \frac{1}{sC_s}}{R + R_s + sL_s + \frac{1}{sC_s}} \right] \left( R + R_c + R_s + sL_c + sL_s + \frac{1}{sC_c} + \frac{1}{sC_s} \right)
\]

(27)

Skin effect on resistance can be calculated by using the following equations:

\[
R_{\text{total}} = R_{\text{DC}} + \sqrt{f} R_{\text{AC}},
\]

where,

\[
R_{\text{DC}} = \frac{\rho L}{W d}, \quad R_{\text{AC}} = \frac{\rho L}{A_{\text{cross, interconnect, wire}}},
\]

\[
= \frac{L \sqrt{\rho}}{2 \omega \sqrt{d}} = \frac{L \sqrt{\rho}}{2 \omega \sqrt{d}} \sqrt{f}
\]

(30)

Therefore,

\[
R = R_{\text{total}} = \frac{L \sqrt{\rho}}{W d}, \quad \frac{L \sqrt{\rho}}{2 \omega \sqrt{d}} \sqrt{f}
\]

(31)

It is evident from the above equation that resistance increases as a function of the square root of the frequency due to the skin effect.

Thus equation (25) can be used to find the closed form expression of the cross talk noise with skin effect.

\[
V_{\text{crosstalk}}(t) \approx \frac{V_0}{(C_o - C_i)} \left( \alpha_2 + \beta_2 + \gamma_2 + \xi_2 \right)
\]

(32)

where,

\[
\alpha_2 = \left[ \frac{2 \omega_0}{\omega^2} \right] \left[ \frac{C_o}{L_o} \right]; \quad \beta_2 = \left[ \frac{C_o}{L_o} \right] \left[ \frac{L_o}{C_o} \right]; \quad \gamma_2 = \left[ \frac{C_o}{L_o} \right] \left[ \frac{L_o}{C_o} \right]; \quad \xi_2 = \frac{C_o}{L_o} \left( 1 + \frac{L_o}{C_o} \right)
\]

4. Simulation Results And Discussions

The motive of this paper is to make a reflection on the skin effect’s impact on the on-chip interconnect and its performance under ramp input. It can be analyzed by the simulation result that the crosstalk noise voltage has been increased. This is due to the fact that skin effect has an adverse effect on the resistance, so with the existence of the skin effect the resistance is increased. This increase in the resistance in turn increases the delay which is undesirable in terms of speed of the chip. We now develop a simple model for distributed RLC interconnect line. The high-speed interconnect system consist of two coupled interconnect lines and ground and the length of the lines is d = 10 mm. These lines are excited by the voltage source of 1.8 V with driver resistance of Rs. The extracted values for the parameters R, L, and C are given in Table 1 [17]. Figure 3 and 4 demonstrate the behaviour of output node crosstalk noise voltage curve with and without considering the skin effect, respectively.
Table 1. RLC parameters for a minimum-sized wires in a 0.18 μm technology.

<table>
<thead>
<tr>
<th>Parameter(s)</th>
<th>Value/m</th>
</tr>
</thead>
<tbody>
<tr>
<td>Resistance (R)</td>
<td>120 kΩ/m</td>
</tr>
<tr>
<td>Inductance (L)</td>
<td>270 nH/m</td>
</tr>
<tr>
<td>Capacitance (C)</td>
<td>240 pF/m</td>
</tr>
<tr>
<td>Coupling Capacitance (C_c)</td>
<td>682.49 fF/m</td>
</tr>
</tbody>
</table>

Table 2. Experimental result under Step input without skin effect

<table>
<thead>
<tr>
<th>Ex</th>
<th>R_s (KΩ)</th>
<th>C_L (fF)</th>
<th>SPICE (nV)</th>
<th>Proposed Model (nV)</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>1</td>
<td>10</td>
<td>6.02</td>
<td>6.12</td>
</tr>
<tr>
<td>2</td>
<td>5</td>
<td>50</td>
<td>18.26</td>
<td>18.21</td>
</tr>
<tr>
<td>3</td>
<td>10</td>
<td>750</td>
<td>43.87</td>
<td>44.03</td>
</tr>
<tr>
<td>4</td>
<td>50</td>
<td>1000</td>
<td>57.34</td>
<td>58.98</td>
</tr>
<tr>
<td>5</td>
<td>100</td>
<td>1500</td>
<td>64.72</td>
<td>63.26</td>
</tr>
</tbody>
</table>

Table 3. Experimental result under Step input with skin effect

<table>
<thead>
<tr>
<th>Ex</th>
<th>R_s (KΩ)</th>
<th>C_L (fF)</th>
<th>SPICE (nV)</th>
<th>Proposed Model (nV)</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>1</td>
<td>10</td>
<td>43.21</td>
<td>45.98</td>
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<tr>
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<tr>
<td>3</td>
<td>10</td>
<td>750</td>
<td>167.54</td>
<td>165.25</td>
</tr>
<tr>
<td>4</td>
<td>50</td>
<td>1000</td>
<td>197.45</td>
<td>186.83</td>
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<tr>
<td>5</td>
<td>100</td>
<td>1500</td>
<td>207.87</td>
<td>208.11</td>
</tr>
</tbody>
</table>

![Fig. 3. Output node noise with considering skin effect for R=1560 KΩ](image1)

![Fig. 4. Output node noise without considering skin effect for R=120 KΩ](image2)

We compare the crosstalk noise voltages obtained from SPICE with those proposed model in both the cases i.e., with and without considering the skin effect on the interconnect lines. Table 2 and Table 3 compare the noise voltages obtained from SPICE with those found using the proposed model without and with considering skin effect. Note that the difference between the proposed model and the SPICE noise voltage is about 2% in both the cases. The simulation shows that failure to account for the skin effect leads to three different errors. First, since the internal inductance is included in the line parameters, the crosstalk voltage is overestimated. This error could be removed by including only the external inductance in the RLC model, but this would underestimate the inductive effect at low frequencies. Second, the attenuation suffered by the signal due to line resistance is underestimated. Finally, the higher inductance leads to a higher characteristic impedance and incorrectly predicts loss of matching at the input.
5. Conclusion

We proposed a new methodology for skin effect equivalent circuits, based on the fitting of the transfer function to numerical simulations. It is shown that the methodology allows arbitrary accuracy in the modelling of the skin effect, and can be adapted to different situations and modelling requirements. We propose a method for the calculation for the skin effect of RLC interconnects. The proposed method of computing skin effect relies on the poles and residues of the transfer function and can be used in any kind of model order reduction technique. In this paper, we have also proposed a novel analytical crosstalk noise voltage model to find the impact of the skin effect on the noise in RLC interconnect without considering the skin effect on inductance because the value of the resistance increases dramatically with comparison to the value of inductance. This paper also reflects that the skin effect and the crosstalk noise increases rapidly in comparison to the ramp input. Simulation results demonstrate the validity and correctness of our proposed model.

References