

# CLOSED FORM MODELING OF CROSSTALK FOR DISTRIBUTED RLCG ON-CHIP INTERCONNECTS USING DIFFERENCE MODEL APPROACH

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## Abstract

*On chip interconnect plays a dominant role on the circuit performance in both analog and digital domains. Interconnects can no longer be treated as mere delays or lumped RC networks. Crosstalk, ringing and reflections are just some of the issues that need to be addressed for the efficient design of high speed VLSI circuits. In order to accurately model these high frequency effects, inductance had been taken into consideration. Within this frequency range, the most accurate simulation model for on-chip VLSI interconnects was the distributed RLC model. Unfortunately, this model has many limitations at much higher of operating frequency used in today's VLSI design. This can lead to inaccurate simulations if not modeled properly. At even higher frequency the conductance metrics has become a dominant factor and has to be taken into consideration for accurate modeling of the different on-chip performance parameters. The traditional analysis of crosstalk in a transmission line begins with a lossless LC representation, yielding a wave equation governing the system response. With the increase in frequency and interconnection length due to the increase in the number of on-chip devices, the lossy components are prevailing than the lossless components. With the reduction of pitch between the adjacent wires in deep sub-micron technologies, coupling capacitances are becoming significant. This increase in capacitances results the introduction of noise which is capable of propagating a logical fault. An inaccurate estimation of the crosstalk could be the origin of the malfunction of the circuit. Cross talk can be analyzed by computing the signal linkage between aggressor and victim nets. The aggressor net carries a signal that couples to the victim net through the parasitic capacitances [13]. To determine the effects that this cross talk will have on circuit operation, the resulting delays and logic levels for the victim nets must be computed. This paper proposes a difference model approach to derive crosstalk in the transform domain. A closed form solution for crosstalk is obtained by incorporating initial conditions using difference model approach for distributed RLCG interconnects. We have proposed an explicit expression for the estimation of cross-talk noise. Our model considers both lossless components (i.e. L, C) and lossy components (i.e. R, G). The SPICE simulation justifies the accuracy of our proposed approach.*

## Keywords:

Crosstalk Calculation, Distributed RLCG Model, Interconnect, Noise, VLSI

## 1. INTRODUCTION

The design techniques in sub-micron technologies increase effects of coupling in interconnections [1]. Indeed, in deep sub-micron technology, the order of capacitive coupling between lines reach to some severe values so that we can't be indifferent to the ampleness of the noise due to this coupling [2]. As integrated circuit feature sizes continue to scale well below 0.18 microns, active device counts are reaching hundreds of millions [3]. 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 [14]. Several

factors bound to the technology contribute to the increase of crosstalk problems: the increase of the number of metal layers [4], the increase of the line thickness, the density of integration and the reduction of the spacing between lines. This set of new challenges is referred as signal integrity in general. Among all these problems, capacitive coupling induced cross talk is the issue that has been seen by an increasing number of backend vendors [14]. Cross talk typically happens between two adjacent wires when their cross coupling capacitance is sufficiently large to influence each other's electrical characteristic. Especially for an on-chip bus, crosstalk noise is a serious problem for VLSI design. In bus structure, crosstalk immunity is more important because long interconnect wires often run together and in parallel. Interconnect lines may be coupled and subjected to the effects of mutual inductive and capacitive coupling, such as crosstalk. It is possible to use both a distributed and a lumped model for these macro models.

In this paper, we have proposed a closed form expression for the coupling noise by analyzing the interconnect using RLCG model. The major drawback of the proposal made in [8] is that it does not consider the shunt lossy component for estimation of the coupling noise. Our model is a generic one in the sense that we can easily derive the model proposed in [8] by just neglecting the shunt lossy component term (i.e. G).

The rest of the paper is organized as follows: Section 2 discusses the basic theory, transmission line model, crosstalk, glitch and modes. Section 3 describes the difference model and our proposed method for noise calculation. Section 4 shows the experimental and simulation results. Finally section 5 concludes the paper.

## 2. BASIC THEORY

### 2.1 TRANSMISSION LINE MODEL

Defining the point at which an interconnect may be treated as a transmission line and hence reflection analysis applied, has no consensus of opinion. A rule of thumb is that when the delay from one end to the other is greater than risetime/2, the line is considered electrically long [15]. If the delay is less than risetime/2, the line is electrically short. A transmission line [5] can be described at the circuit level using series inductance and resistance combined with shunt capacitance and conductance. An infinitesimal unit length of the transmission line looks like the circuit as shown in Figure 1. The parameters are defined as follows.

R = Series resistance per unit length

L = Series inductance per unit length

G = Shunt conductance per unit length

C = Shunt capacitance per unit length.

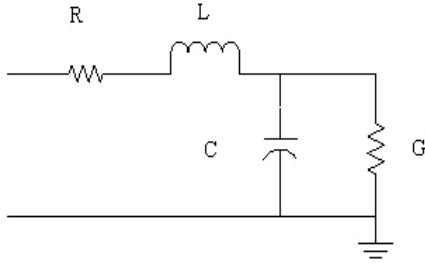


Fig.1. RLCG parameters for a segment of a transmission line

It is critical to model the transmission path when designing a high-performance, high-speed serial interconnect system. The transmission path may include long transmission lines, connectors, vias and crosstalk from adjacent interconnect. Values for R, L, C, and G are extracted from a given layout, designed in 0.18 $\mu$ m technology.

## 2.2 CROSS TALK

Crosstalk is undesired energy imparted to a transmission line due to signals in adjacent lines. The magnitude of the crosstalk induced is a function of rise time, signal line geometry and net configuration (type of terminations, etc.) [15]. To overcome the problems faced at high frequency of operation, shielding techniques have been employed [11]. A common method of shielding is placing ground or power lines at the sides of a victim signal line to reduce noise and delay uncertainty [16]. The crosstalk between two coupled interconnects is often neglected when a shield is inserted, significantly underestimating the coupling noise. The crosstalk noise between two shielded interconnects can produce a peak noise of 15% of VDD in a 0.18  $\mu$ m CMOS technology [12]. An accurate estimate of the peak noise for shielded interconnects is therefore crucial for high performance VLSI design. In the complicated multilayered interconnect system, signal coupling and delay strongly affect circuit performances. Thus, accurate interconnect characterization and modeling are essential for today's VLSI circuit design. Two major impacts of cross talk are:

(I) Crosstalk induces delays, which change the signal propagation time, and thus may lead to setup or hold time failures.

(II) Crosstalk induces glitches, which may cause voltage spikes on wire, resulting in false logic behavior. Crosstalk affects mutual inductance as well as inter-wire capacitance.

When the connectors in high speed digital designs are considered, the mutual inductance plays a predominant role compared to the inter-wire capacitance [8]. The effect of mutual inductance is significant in deep submicron technology (DSM) technology since the spacing between two adjacent bus lines is very small. The mutual inductance induces a current from an aggressor line onto a victim line which causes crosstalk between connector lines.

In multi-conductor systems, crosstalk can cause two detrimental effects: first, crosstalk will change the performance of the transmission lines in a bus by modifying the effective characteristic impedance and propagation velocity. Secondly, crosstalk will induce noise onto other lines, which may further degrade the signal integrity and reduce noise margins [8].

## 2.3 GLITCH

Crosstalk Glitch (CTG) is a glitch signal provoked by coupling effects among interconnects lines which have unbalanced drivers and loads [6]. The magnitude of the glitch depends on the ratio of coupling capacitance to line to ground capacitance. When a transition signal is applied at a line which has a strong line-driver while stable signals are applied at other lines which have weaker drivers, the stable signals may experience a coupling noise due to the transition of the stronger signal [8]. A glitch may be induced in connector 'j' in which the signal is static, due to neighboring connector lines in which the signal is varying [7]. This is given by the equation (1),

$$V_{glitch}^j = \sum_j \pm L_{jk} \frac{dj_k}{dt} \forall j \neq k \quad (1)$$

where,  $L_{jk}$  represents mutual inductance between  $j^{\text{th}}$  and  $k^{\text{th}}$  connector. The sign of the coupled voltage is positive or negative depending upon whether the  $k^{\text{th}}$  neighboring connector undergoes a rising or a falling transition.

## 2.4 ODD MODE

When two coupled transmission lines are driven with voltages of equal magnitude and 180 degree out of phase with each other, odd mode propagation occurs. The effective capacitance of the transmission line will increase by twice the mutual capacitance, and the equivalent inductance will decrease by the mutual inductance [15]. In Fig.2, a typical transmission line model is considered where the mutual inductance between aggressor and victim connector is represented as  $M_{12}$ .  $L_1$  and  $L_2$  represent the self inductances of aggressor and victim nodes while  $C_c$ ,  $C$ , denote the coupling capacitance between aggressor and victim, self capacitance respectively.

Assuming that  $L_1 = L_2 = L_0$ , the currents will be of equal magnitude but flow in opposite direction [7]. Thus, the effective inductance due to odd- mode of propagation is given by equation (2).

$$L_{odd} = L_1 - L_2 \quad (2)$$

The magnetic field pattern of the two conductors in odd-mode is shown in fig 3 [15].

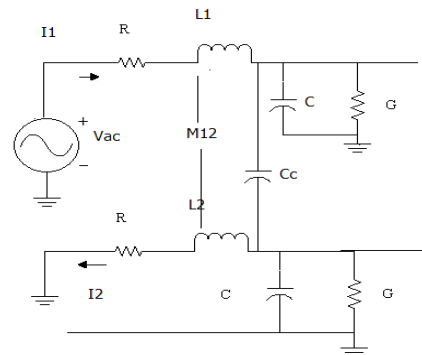


Fig.2. An Example for two line Transmission line model

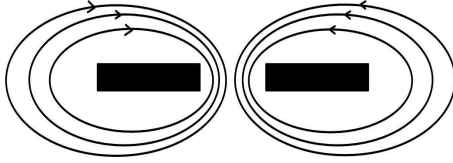


Fig.3. Magnetic Field in Odd Mode

## 2.5 EVEN MODE

When two coupled transmission lines are driven with voltages of equal magnitude and in phase with each other, even mode of propagation occurs. In this case, the effective capacitance of the transmission line will decrease by the mutual capacitance and the equivalent inductance will increase by the mutual inductance. Thus, in even-mode propagation, the currents will be of equal magnitude and flow in the same direction [7]. The effective inductance, due to even mode of propagation is then given by equation (3).

$$L_{\text{even}} = L_1 - L_2 \quad (3)$$

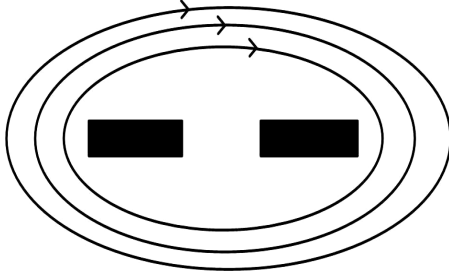


Fig.4. Magnetic Field in Even Mode

## 3. MODELING OF CROSS TALK IN RLCG INTERCONNECT

### 3.1 DIFFERENCE MODEL

The frequency-domain difference approximation [10] procedure is more general, because it can directly handle lines with arbitrary frequency-dependent parameters or lines characterized by data measured in frequency-domain. The time-domain difference approximation procedure should be employed only if transient characteristics are available. For a single RLCG line, the analytical expressions are obtained for the transient characteristics and limiting values for all the modules of the system and device models. The difference approximation procedure is applied to both the characteristic admittances and propagation functions and the resulting time-domain device models have the same form as the frequency-domain models. The difference approximation procedure involves an approximation of the dynamic part of the system transfer function, given by equation (12), with the complex rational series or distorted part of the transient characteristic with the real exponential series. This criterion results in simple and efficient approximation algorithms, and requires a minimal number of the original-function samples to be available, which is important if the line is characterized with delay and crosstalk [10].

### 3.2 ANALYSIS OF CROSSTALK USING DIFFERENCE MODEL

Let us consider the interconnect system which consists of a single uniform line and ground line as shown in Figure 5, and assume the length of the line is  $d$ .

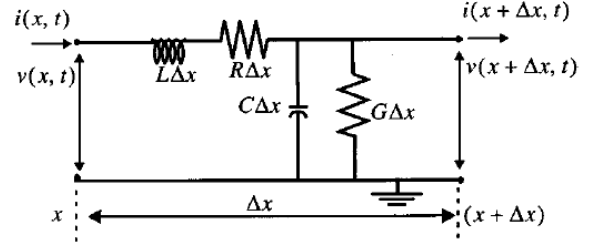


Fig.5. Equivalent circuit of each uniform section

The electrical parameters for each sub section are  $R\Delta x$ ,  $L\Delta x$ ,  $C\Delta x$  and  $G\Delta x$ , respectively, where  $R$ ,  $L$ ,  $C$  and  $G$  are per-unit length resistance, inductance, capacitance and conductance of the line, respectively.

Using Kirchoff's Voltage Law (KVL), we can write,

$$v(x,t) = i(x,t)R_{\Delta x} + L_{\Delta x} \frac{di(x,t)}{dt} + v(x+\Delta x,t) \quad (4)$$

Using Kirchoff's Current Law (KCL), we can write,

$$i(x,t) = G_{\Delta x}v(x+\Delta x,t) + c_{\Delta x} \frac{dv(x+\Delta x,t)}{dt} + i(x+\Delta x,t) \quad (5)$$

Simplifying the equations (4) and (5) and after applying Laplace transform, we get,

$$-\frac{\partial V(x)}{\partial x} = (R + sL)I(x) \quad (6)$$

$$-\frac{\partial I(x)}{\partial x} = (G + sC)V(x) \quad (7)$$

Differentiating equations (6) and (7) with respect to the  $x$ , and after simplifying we get,

$$\frac{\partial^2 V(x)}{\partial x^2} = P^2 V(x) \quad (8)$$

And

$$\frac{\partial^2 I(x)}{\partial x^2} = P^2 I(x) \quad (9)$$

where the  $P$  is the propagation constant and is defined as,

$$P = \sqrt{(R + sL)(G + sC)} \quad (10)$$

The general solution of equation (8) is given by

$$V(x) = A_1 e^{-Px} + A_2 e^{Px} \quad (11)$$

Where  $A_1$  and  $A_2$  are the constants determined by the boundary conditions. From equations (8) and (11) we get,

$$-\frac{\partial}{\partial x} [A_1 e^{-Px} + A_2 e^{Px}] = (R + sL)I(x) \quad (12)$$

After simplifying we get,

$$I(x) = \frac{1}{Z_0} [A_1 e^{-Px} - A_2 e^{Px}] \quad (13)$$

where  $Z_0$  is the characteristic impedance. Assuming at  $x=d$ , the termination voltage and current are  $V(d) = V_2$  and

$I(d) = I_2$  then we get,

$$V_2 = A_1 e^{-Pd} + A_2 e^{Pd} \quad (14)$$

$$I_2 = \frac{1}{Z_0} [A_1 e^{-Pd} - A_2 e^{Pd}] \quad (15)$$

After solving equation (14) and (15) for  $A_1$  and  $A_2$  we get,

$$A_1 = \frac{1}{2} [V_2 + I_2 Z_0] e^{Pd} \quad (16)$$

$$A_2 = \frac{1}{2} [V_2 - I_2 Z_0] e^{-Pd} \quad (17)$$

Substituting these values of  $A_1$  and  $A_2$  in equation (11)

$$V(x) = \left[ \frac{[V_2 + I_2 Z_0] e^{P(d-x)}}{2} + \frac{[V_2 - I_2 Z_0] e^{P(x-d)}}{2} \right] \quad (18)$$

Now substituting the values of  $A_1$  and  $A_2$  from (16) and (17) in equation (13) we get,

$$I(x) = \frac{1}{Z_0} \left[ \frac{[V_2 + I_2 Z_0] e^{P(d-x)}}{2} - \frac{[V_2 - I_2 Z_0] e^{P(x-d)}}{2} \right] \quad (19)$$

Let at  $x=0$ ,  $V(x) = V_1$  and  $I(x) = I_1$  then from equation (18) and (14), we can write:

$$V_1 = \cosh(Pd) V_2 + Z_0 \sinh(Pd) I_2 \quad (20)$$

$$I_1 = \frac{1}{Z_0} \sinh(Pd) V_2 + \cosh(Pd) I_2 \quad (21)$$

Since ABCD parameters are defined as

$$\begin{bmatrix} V_1 \\ I_1 \end{bmatrix} = \begin{bmatrix} A & B \\ C & D \end{bmatrix} \begin{bmatrix} V_2 \\ -I_2 \end{bmatrix} \quad (22)$$

So we can write ABCD matrix from equation (20) and (21)

$$\begin{bmatrix} V_1 \\ I_1 \end{bmatrix} = \begin{bmatrix} \cosh(Pd) & -Z_0 \sinh(Pd) \\ \frac{1}{Z_0} \sinh(Pd) & -\cosh(Pd) \end{bmatrix} \begin{bmatrix} V_2 \\ -I_2 \end{bmatrix} \quad (23)$$

The output crosstalk voltage is given by

$$V_2(s) = \frac{V_1(s)}{\cosh(Pd)} \quad (24)$$

For the step input voltage we get,

$$V_2(s) = \frac{V_0}{s \cosh(Pd)} \quad (25)$$

$$\text{Or, } V_2(s) = \frac{V_0}{s \cosh(\sqrt{(R+sL)(G+sC)})} \quad (26)$$

After simplification, we get from equation (26)

$$V_2(s) = \frac{2V_0}{s \left( s + \frac{R}{L} \right) \left( s + \frac{G}{C} \right)} \quad (27)$$

After taking inverse Laplace transform of equation (27), we get,

$$v_2(t) = v_0 \left[ \frac{2LC}{RG} - \frac{2LC}{GL - RC} \left( \frac{L}{R} e^{-\frac{Rt}{L}} + \frac{C}{G} e^{-\frac{Gt}{C}} \right) \right] \quad (28)$$

This is our proposed model for noise voltage induced by the aggressor line onto the victim line.

Now we will consider two typical cases of frequency of operation.

#### CASE -1(For Very Low Frequency)

For very low frequency, where  $R \gg \omega L$ , equation (26) reduces to

$$V_2(s) = \frac{V_0}{s \cosh \sqrt{RG}} \quad (29)$$

After taking inverse Laplace transform of equation (29), we get,

$$v_2(t) = \frac{V_0}{\cosh \sqrt{RG}} u(t) \quad (30)$$

#### CASE -2 (For very High Frequency)

For high frequency, where  $R \ll \omega L$ , equation (19) reduces to

$$V_2(s) = \frac{V_0}{s \cosh(s\sqrt{LC})} \quad (31)$$

After taking inverse Laplace transform of equation (24), we get

$$v_2(t) = v_0 \left[ 1 - \cos \left( \sqrt{\frac{2}{LC}} t \right) \right] \quad (32)$$

## 4. EXPERIMENTAL RESULTS

Most of the earlier research and reduction techniques consider only capacitive coupling [2, 9]. But in the case of very high frequencies as in GHz scale, inductive crosstalk comes into the important role and it should be included for complete coupling noise analysis. The configuration of circuit for simulation is shown in Figure 2. The high-speed interconnect system consist of two coupled interconnect lines and ground and the length of the lines is  $d = 10$  mm. The sample dimensions of the cross sections of a minimum sized wire in a  $0.18\mu\text{m}$  technology are given in figure 6.

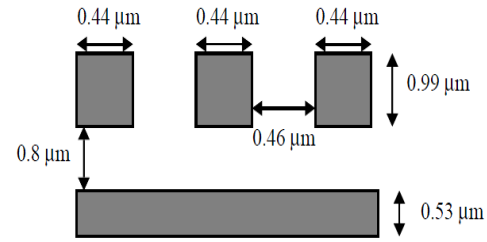


Fig.6. Sample Dimensions of Cross-sections of minimum sized wire in a  $0.18\mu\text{m}$  technology

The extracted values for the parameters  $R$ ,  $L$ ,  $C$ , and  $G$  are given in Table 1.

Table1: RLCG parameters for a minimum- sized wires in a  $0.18\mu\text{m}$  technology. Where the conductance is a function of frequency,  $f$

Parameter(s)	Value/m
Resistance(R)	120 kΩ/m
Inductance(L)	270 nH/m
Conductance(G)	15f pS/m
Capacitance(C)	240 pF/m

The left end of the first line of figure 2 is excited by 1-V trapezoidal form voltage with rise/fall times 0.5 ns and a pulse width of 1 ns. Other parameters of lumped elements are  $R_1=R_2=50$  ohms and  $C_1=C_2=1\text{pF}$ . Figure 7 and Figure 8 correspond to the frequency responses of the magnitude of voltage at the near end of victim line and the far end of aggressive line, respectively. The operating frequency is taken as 2GHz.

From figures 7 and 8, we find that, as the frequency increases, crosstalk noise becomes very severe, and hence the signal of interest is attenuated. At very high frequency, of the order of GHz, oscillation occurs sharply, which is a bit different from the transfer function of lumped parameter systems. Figure 9 and Figure 10 correspond to the waveforms of voltage responses at the far end of the aggressive line and both ends of victim line.

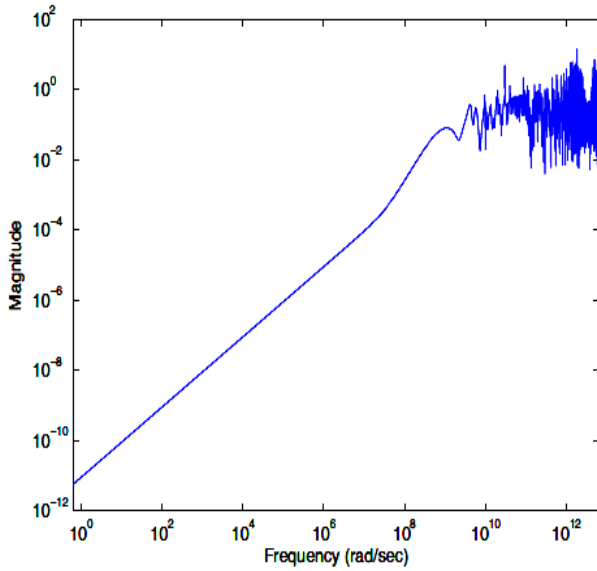


Fig.7. Frequency Response at near end of victim line

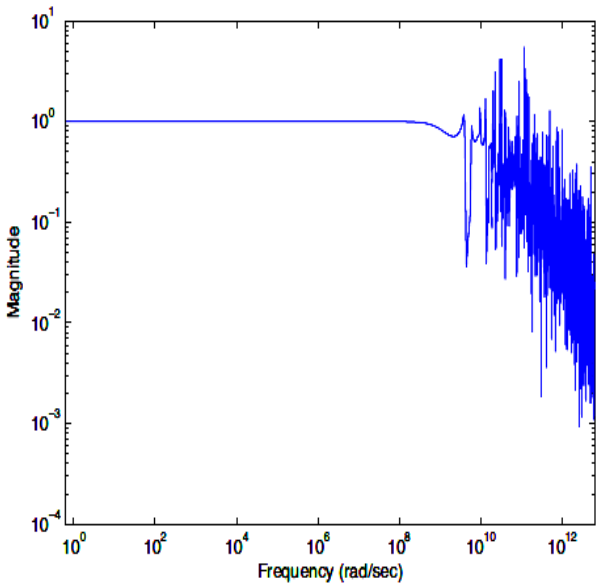


Fig.8. Frequency Response at far end of aggressive line

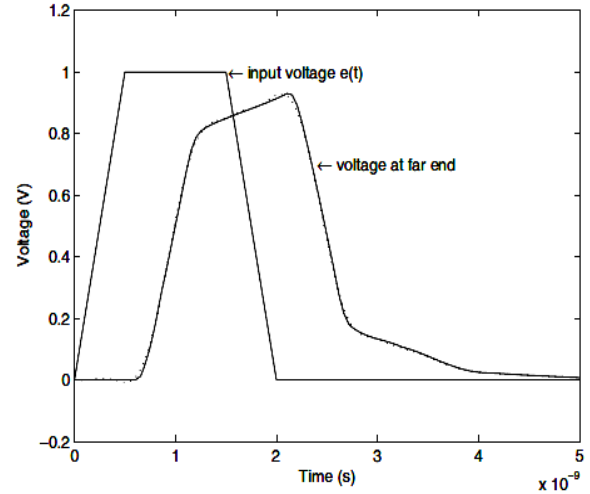


Fig.9. Waveform of voltage far end of aggressive line

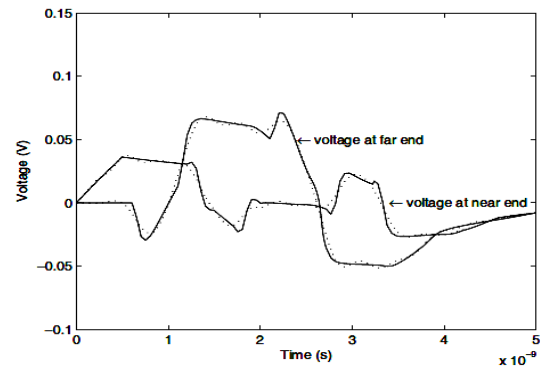


Fig.10. Waveform of voltage at both end of victim line

## 5. CONCLUSIONS

In this paper, we discussed the problem of crosstalk noises in high-speed coupled interconnect systems. We proposed a distributed RLCG transmission line model of interconnects using difference model approach. Result shows that, at low frequencies, the model exhibits a RC behavior but at high frequencies has a substantially different behavior due to the effects of inductance. On the basis of Laplace transformation of distributed parameter model deduced in time domain, transfer functions of crosstalk noises are built, and crosstalk noise response is analyzed theoretically. Simulation results demonstrate the validity and correctness of our method.

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