Crosstalk Noise in High Density and High Speed Interconnections due to Inductive Coupling

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Abstract— Crosstalk noise in long interconnections is studied based on capacitive coupling and inductive coupling.

It is shown that pulse noise is induced due to inductive coupling in heterogeneous insulators. The pulse noise becomes predominant noise factor on the condition that lines become longer, line resistance become lower, the signal raising time becomes faster and dielectric constant of materials in the gaps on lines becomes smaller.

I. Introduction

Importance of interconnection delay, coupling, and crosstalk are in realizing high-performance circuits, as the feature size of the lines is miniaturized down to deep submicron region. The strong coupling condition between lines become in mutual inductance as well as mutual capacitance is important factor for optimizing interconnections in terms of coupling noise. as the signal rising time is short compared with the signal propagation time, it is needed to consider not only capacitive coupling but also inductive coupling.

In section II, we derive transient response for lossless LC coupling interconnections based on coupling parameters in closed formulas described in section III. In section IV, we show a new noise component due to the inductive coupling in the condition that interconnections are parallel in the heterogeneous insulators, and quantitatively estimate the inductive pulse noise in lossless LC coupling interconnections. In Section V, we also estimate the inductive pulse noise in lossy interconnections by using numerical analysis. Finally summary and conclusion are given in section VI.

II. Formulation of Transient response in Coupling interconnections

A model of two parallel interconnections is shown in Fig.1. Here, two interconnections are coupled inductively

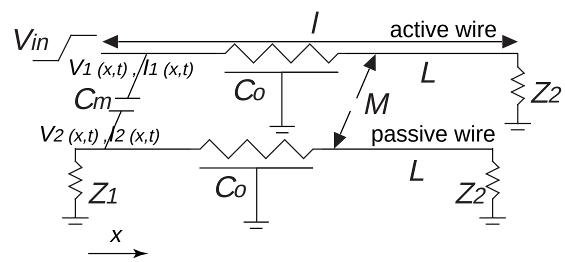

Fig. 1. Model of interconnections with inductive and iapacitiveioupling

and capacitively. Responses of far end of interconnections are given by solving the following differential equations simultaneously.

$$\begin{cases} -\dfrac{\partial I_1}{\partial x} = C\dfrac{\partial V_1}{\partial t} - C_m\dfrac{\partial V_2}{\partial t} \\ -\dfrac{\partial I_2}{\partial x} = -C_m\dfrac{\partial V_1}{\partial t} + C\dfrac{\partial V_2}{\partial t} \\ -\dfrac{\partial V_1}{\partial x} = L\dfrac{\partial I_1}{\partial t} + M\dfrac{\partial I_2}{\partial t} + RI_1 \\ -\dfrac{\partial V_2}{\partial x} = M\dfrac{\partial I_1}{\partial t} + L\dfrac{\partial I_2}{\partial t} + RI_2 \end{cases} \quad (1)$$

Here, C, C_m, L and M are self-capacitance, mutual capacitance, self-inductance and mutual inductance of interconnections per unit length, respectively. From the above equations, we obtain the telegraph equations of V_a and V_b

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as follows,

$$\begin{cases} \dfrac{\partial^2 V_a}{\partial x^2} &= ((L+M)(C-C_m)s^2 + R(C-Cm)s)V_a \\ \dfrac{\partial^2 V_b}{\partial x^2} &= ((L-M)(C+C_m)s^2 + R(C+Cm)s)V_b \end{cases}$$

here,

$$\begin{cases} V_a = V_1 + V_2 & \cdots \text{ even mode voltage,} \\ V_b = V_1 - V_2 & \cdots \text{ odd mode voltage, and} \\ s = \dfrac{\partial}{\partial t}. & \end{cases}$$

To obtain an analytical solution, we assume that interconnections are lossless wires ($R = 0$), and that the terminal impedances at far end (Z_2) is equal to the characteristic impedance ($\sqrt{\frac{L-M}{C+C_m}}$). Then we obtain the voltage response (crosstalk noise) of the wire V_2 as follows.

$$\begin{aligned} V_2(s) &= \frac{1}{1-\alpha e^{\frac{2l}{c_e}s}} V_{in}(s) \\ &\times\{ \frac{Z_a(Z_b - Z_1)}{\beta} e^{-\frac{x}{c_e}s} - \frac{Z_b(Z_a - Z_1)}{\beta} e^{-\frac{x}{c_o}s} \\ &-\mu \frac{Z_a(Z_b - Z_1)}{\beta} e^{-\frac{2l-x}{c_e}s} \\ &-\mu \frac{Z_b(Z_a + Z_1)}{\beta} e^{-(\frac{2l}{c_e}+\frac{x}{c_o})s} \} \end{aligned} \tag{2}$$

Here, constants c_e, c_o, Z_a, Z_b, μ , respectively, α and β are,

$$\begin{aligned} c_e &= \frac{1}{\sqrt{(L+M)(C-C_m)}}, c_o = \frac{1}{\sqrt{(L-M)(C+C_m)}} \\ Z_a &= \sqrt{\frac{L+M}{C-C_m}}, Z_b = \sqrt{\frac{L-M}{C+C_m}} \\ \mu &= \frac{Z_a - Z_b}{Z_a + Z_b} \\ \alpha &= \mu \frac{Za - Z_1}{Zb - Z_1} \\ &\times \frac{2(Z_a Z_b + Z_1^2)}{(Z_a + Z_b)(Z_a - Z_b) + Z_b(Z_b - Z_1)} \\ \beta &= Za(Z_b - Z_1) + Z_b(Z_a - Z_1). \end{aligned} \tag{3}$$

Assuming that terminal impedance Z_1 is nearly equal to Z_a, considering that α is much small than 1, we obtain the principal four components of the crosstalk noise mode1, mode2, mode3 and mode4 as,

$$\begin{aligned} &V_{2,noise}(s) = \\ &\{ \frac{Z_a(Z_b - Z_1)}{\beta} e^{-\frac{x}{c_e}s} - \frac{Z_b(Z_a - Z_1)}{\beta} e^{-\frac{x}{c_o}s} \\ &-\mu \frac{Z_a(Z_b - Z_1)}{\beta} e^{-\frac{2l-x}{c_e}s} \\ &-\mu \frac{Z_b(Z_a + Z_1)}{\beta} e^{-(\frac{2l}{c_e}+\frac{x}{c_o})s} \} \times V_{in}(s) \\ &= mode1 + mode2 + mode3 + mode4. \end{aligned} \tag{4}$$

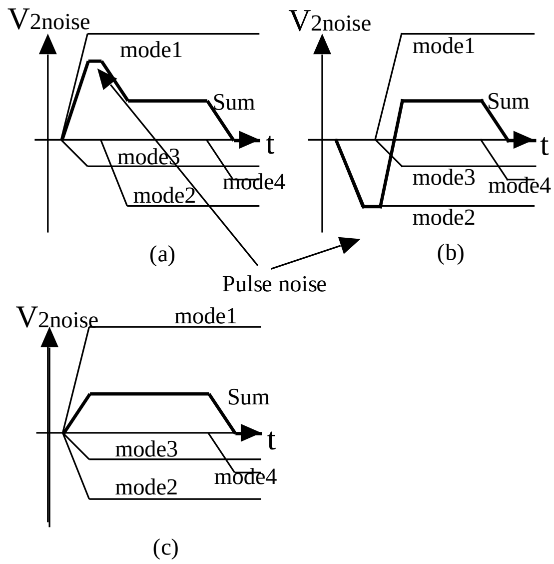


Fig. 2. Crosstalk noise appears at the end of the passive interconnection (V_2). (a):$c_e > c_o$(b):$c_e < c_o$(c):$c_e = c_o$

The four crosstalk noise components in eq.4 are waveforms similar to the input pulse V_{in} with specific magnitudes and propagation delays. Fig.2 illustrates the crosstalk noise appearing at the end of the wire ($x = l$), as described by eq.4. Delays of model1, model2, model3 and model4 are $\frac{l}{c_e}$, $\frac{l}{c_o}$, $\frac{l}{c_e}$, and $\frac{2l}{c_e} + \frac{l}{c_o}$, respectively. Fig.2 (a), (b) and (c) indicate cases of $c_e > c_o$, $c_e < c_o$ and $c_e = c_o$ respectively. Fig.2(c) shows that a revel noise appears when $c_e = c_o$. Fig.2(a), (b) shows that a pulse noise appears in addition to the level noise when $c_e \neq c_o$. We call the level and the pulse noise "inductive pulse noise," "capacitive level noise," respectively.

III. Coupling Constants

Since conventionally the exact coupling capacitances as shown in Fig.3, needs a long CPU time in calculation such as the finite element analysis, compact formulas are desirable for delay and noise analysis of interconnections. We have reported some formulas for the coupling capacitances with sufficient accuracy [4]. It is based on the following formulas for estimating wiring capacitances [2][3].

$$\begin{aligned} C_o &= \varepsilon_{ox}\{1.15(\frac{W}{H}) + 2.80(\frac{T}{H})^{0.222}\} \\ C_m &= \varepsilon_{ox}\{0.03(\frac{W}{H}) + 0.83(\frac{T}{H}) - 0.07(\frac{T}{H})^{0.222}\}(\frac{S}{H})^{-1.34} \\ C_{total} &= C_o + C_m \quad (F/\mathrm{cm}) \end{aligned} \tag{5}$$

Here, W, T, S and H is metal width, metal height, gap in two metals and thickness of insulator below metal, respectively. This formulation has an enough accuracy

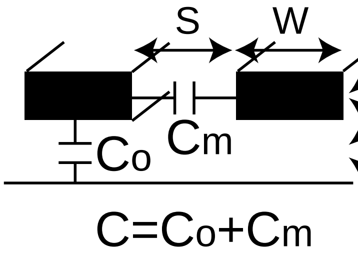


Fig. 3. Notation for coupling capacitances

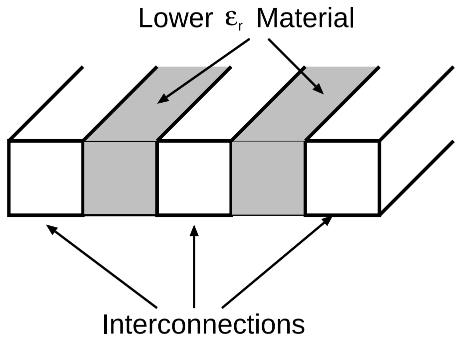


Fig. 4. Interconnection in heterogeneous insulators

for the "total" value of wiring capacitance. But when two wires are closely placed to each other, the mutual capacitance (C_m) obtained by eq.5 is considerably lower than the exact value, because a lot of the electric flux that originates from the bottom of wire are absorbed into the contiguous wire. So we proposed in [4] a new distribution function (f_{asp}) for the self-capacitance and the mutual capacitance.

$$C_o = C_o f_{asp} \tag{6}$$

$$C_m = C_o(1 - f_{asp}) + C_m \tag{7}$$

$$f_{asp} = e^{-0.614 \frac{(\frac{T}{H})^{0.018}}{(\frac{W}{H})^{0.471}(\frac{S}{H})^{0.444}}} \tag{8}$$

$$\frac{C_o}{C_{total}} = e^{-0.824 \frac{(\frac{T}{H})^{0.237}}{(\frac{W}{H})^{0.430}(\frac{S}{H})^{0.662}}}. \tag{9}$$

Errors of C_o and $C_m/(C_o + C_m)$ are less than 5% for $0.5 < (W/H), (S/H), (T/H) < 4$.

In the condition that wires are in homogeneous insulator, the self-inductance (L) is related with the wiring capacitance(C) as follows.

$$\frac{1}{\sqrt{LC}} = \frac{1}{\sqrt{\varepsilon\mu}} = \frac{c}{\sqrt{\varepsilon_r \mu_r}} \tag{10}$$

The ratio of C_m to C is equal to that of M to L.

$$\frac{M}{L} = \frac{C_m}{C} \tag{11}$$

Since μ_r of most insulators is 1, the ratio $\frac{M}{L}$ in heterogeneous insulators is considered to be equal to the ratio $\frac{C_m}{C}$ in homogeneous condition though the ratio $\frac{C_m}{C}$ in heterogeneous insulators is deviated from eq.11.

IV. Crosstalk Noise in Heterogeneous Insulators

When we can ignore the inductance, the crosstalk noise in two wires are simply described by the ratio of inter-wire capacitance to total capacitance.

$$\frac{\Delta V_{2(noise)}}{V_{1(input)}} = \frac{C_m}{C_o + C_m} \tag{12}$$

From Eq.12, and the fact that wiring capacitances are proportional to the dielectric constant of insulator, it seems that the crosstalk noise can be reduced by, heterogeneous insulators as shown in Fig.4. Here, a insulators between wired have lower dielectric constant than the other parts (upper and lower part). Many low dielectric insulators are reported [5], [6] and so on, as show in Table I.

When we consider both the inductive and capacitive coupling, the inductive pulse noise appears due to the difference in the propagation speed of even mode (c_e) and the odd mode (c_o). If the insulators between wires are lower dielectric material than the other parts, c_e become lower than c_o because electric field of odd mode propagates in the lower dielectric insulators while the magnetic field of are not influenced by the lower dielectric insulators, as shown in Fig.5.

As is shown in Fig.6, magnitude of the inductive pulse noise becomes larger as the dielectric constant becomes far off and the signal rising time becomes shorter. The magnitude of inductive pulse noise is plotted with the ratio of dielectric constants of inter-wire part to the other part as a parameters in Fig.7. There is a trade off point for each signal rising speed between the magnitude of the inductive pulse noise and that of capacitive level noise.

V. Lossy Interconnections

Actual interconnections have parasitic resistance. When a signal propagate the interconnections lose a signal energy in high frequency component. Therefore, it is indispensable to estimate the inductive pulse noise with consider the wiring resistance. Since it is hard to solve the differential equations (eq.1-1) in a lossy condition, we have developed a simulator for the lossy interconnections to numerically solve these differential equations. Simulation results on the condition of Fig.8 are shown in Fig.9 and 10. Fig.9 and 10 show responses of lossless and lossy interconnections, respectively. The inductive pulse noise grows large as the distance from the input increases. The pulse noise on the lossy interconnections is smaller than

TABLE I
Example of low dielectric insulators

	MATERIAL	Dielectric Constant
1	$-[0-\mathrm{SiMe_2}-]-$	2.3
2	$-[-\mathrm{SiPh_2CH_2CH_2CH_2}-]-$	3.2
3	SPI 100siloxane polyimide (Genefal Electric)	3.7
5	$\mathrm{CH}=\mathrm{CH_2}$ $[-\mathrm{OSiMe_2}-\mathrm{Ph}-\mathrm{SiMe_2}-$ $-\mathrm{O}-\mathrm{SiMe}-\mathrm{O}-]-$	3.5
6	$-[-\mathrm{SiMe_2CH_2}-]-$	2.8
7	Poly(trimethylsilylnorborane)	2.3
8	$-[-\mathrm{SiMe_2}-\mathrm{CH_2}-]-$	2.8
9	PMDA-ODA/Polydimethyl-siloxane coplymer	2.8

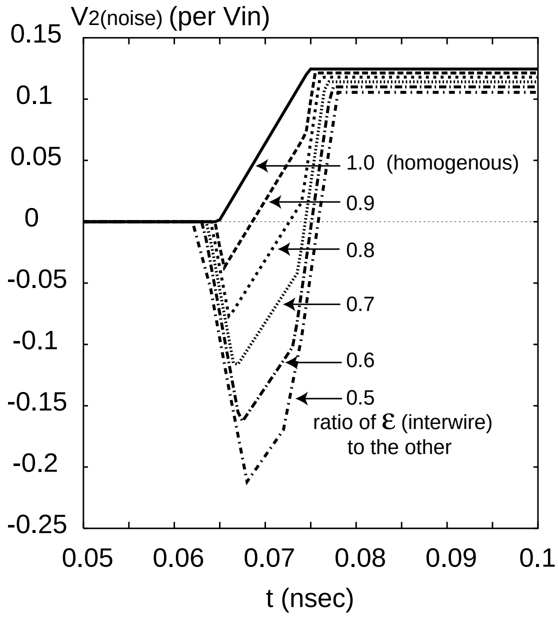


Fig. 6. Crosstalk noise at the end of the passive wire. (eq.4) W=4.0μm,S=4.0μm,T=1.0μm,H=1.0μm and line length= 1cm

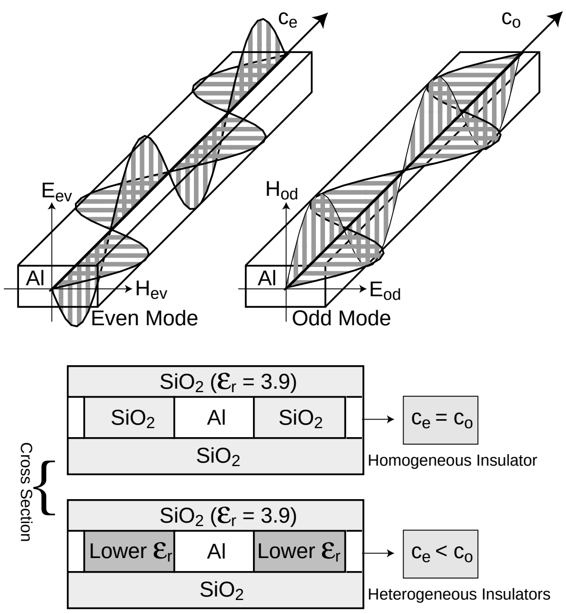


Fig. 5. Propagation of electromagnetic wave through interconnections

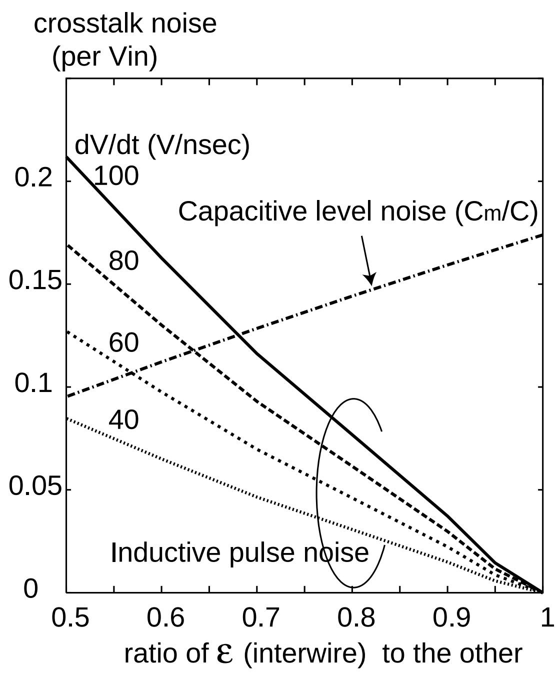


Fig. 7. Magnitude of inductive pluse noise(at lossless interconnections ($R=0$))

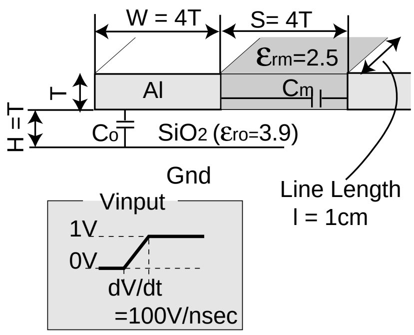
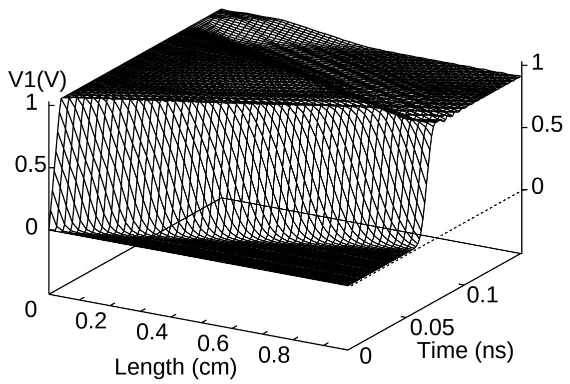
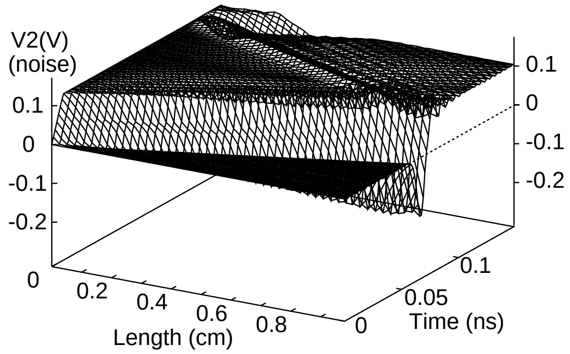


Fig. 8. Simulating condition of Fig.9 and 10

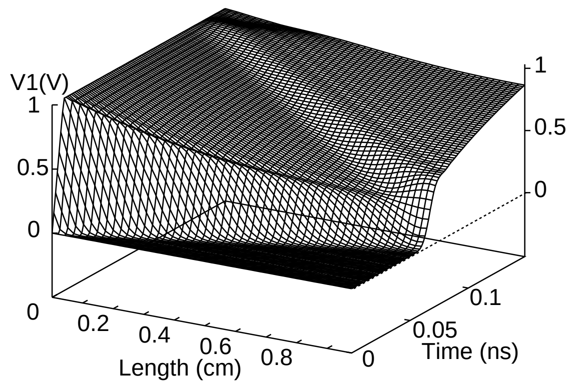


(a) Response of the activate wire

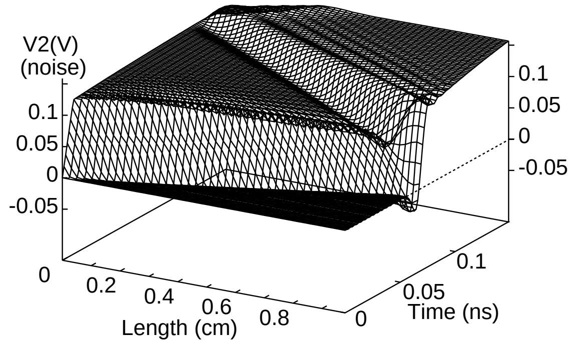


(b) Response of the passive wire (crosstalk noise)

Fig. 9. Response of lossless interconnections (numerical analysys) @dV/dt = 100V/nsec



(a) Response of the activate wire



(b) Response of the passive wire (crosstalk noise)

Fig. 10. Response of lossy ($R = 100\Omega$/cm) interconnections (numerical analysys) @dV/dt = 100V/nsec

the lossless interconnections because the energy of pulse noise in high frequency is lost in propagation.

Simulation results are summarizes in Fig.11, with inter-wire dielectric materials, rising speeds of input signal and the wiring resistance as parameters.

The pulse noise become a predominant noise factor as the wiring resistances become low, the signal rising time become fast and the dielectric constant of materials between wires becomes small. as the

VI. Summary and Conclusion

In this paper, we have described,

1. the crosstalk noise model due to both the inductive and capacitive couplings,

2. the inductive and capacitive coupling constants described as closed formulas for closely placed interconnections,

3. the inductive pulse noise in lossless interconnections, and

4. the interconnection simulator to evaluate the inductive pulse noise in lossy interconnections.

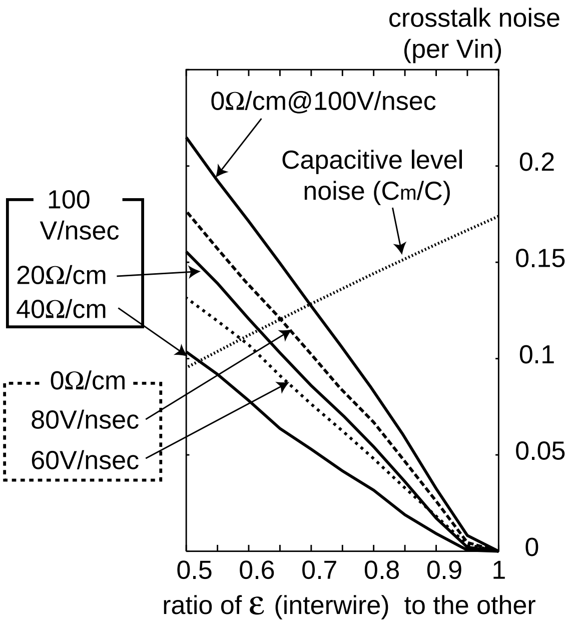


Fig. 11. Magnitude of inductive pulse noise (numerical analysis) @same cross sectional aspect ratio as Fig8

It is necessary to treat long interconnections as lines with both the parasitic inductance and parasitic capacitance, when signal delay time is larger than signal raising time. We evaluated the crosstalk noise for the 1 cm interconnections as an example. We found out a new component of pulse noise in heterogeneous insulators using low dielectric materials in gaps of interconnections. The pulse noise becomes a predominant noise factor as lines become longer, line resistances become lower, the signal raising time becomes faster and dielectric constant of materials in the gaps on lines becomes smaller.

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