FINITE ELEMENT ANALYSIS OF SKIN EFFECT IN COPPER INTERCONNECTS AT 77K AND 300K

Uttam Ghoshal and L.N. Smith

Microelectronics and Computer Technology Corporation
Austin, TX 78727

ABSTRACT

We present the methodology for calculating normal skin effect in complex geometries using finite elements. We have used these results to analyze the performance of copper interconnects at 77K and 300K for both digital and microwave applications. This includes attenuation per unit length, phase velocity and characteristic impedance as a function of frequency from d.c. to 10 GHz. For digital signal propagation, skin effects are important for predicting rise time degradation for times less than 1.2 times the time of flight delay, while for larger times the d.c. resistance corresponding to the cross section of the signal line is adequate for explaining the lossy characteristics.

INTRODUCTION

Liquid nitrogen temperature (77K) operation of CMOS VLSI circuits has recently been introduced for high performance mainframe computers [1]. Among the major advantages of low temperature operation are smaller system volume, better computing speed and higher reliability. Wafer scale packaging technologies, such as the copper-polyimide "medium film" technology, are now becoming mature enough to provide dense interconnects and efficiently pack chips at the first level of system interconnections [2]. The recent discoveries of high temperature superconductivity in oxide materials with critical temperatures in the 90K regime and the possibility of superconducting devices at 77K suggest that replacement of copper conductors by superconductors in wafer scale interconnects might remove important constraints on signal delay, attenuation and dispersion in off-chip communication, especially as interconnects are scaled to finer geometries to provide more communication paths. In order to make a good comparison between the performance of copper interconnects and its high Tc superconductor replacement, it is necessary to analyze skin effects.

Our work is different from the earlier, seminal work by Kautz [3] and recent literature [4] in three important regards. First, we consider a realistic narrow microstrip geometry of practical importance for digital applications, where impedance is dominated by fringing fields. This is done by finite element methods. Secondly, we calculate a linear equivalent circuit for the lines using multiple inductors and resistors, by fitting to the frequency-dependant complex impedance obtained from our finite element simulations. Third, we simulate the line response for a variety of different circuit configurations and inputs using SPICE, including step functions with finite rise time and matched source impedance, and scaled CMOS drivers. This is more relevant than the Gaussian pulse propagation which has been previously analyzed using an analytic approach. We present simple physical interpretations of these simulations, discuss the circumstances under which the skin effect can be neglected, and comment on possible windows of opportunity for superconducting interconnects.

We have not incorporated the anomalous skin effect into our our finite element model . This effect is important when the classical skin depth δ is much smaller than the electronic mean free path l_e , leading to a non-local relation between the current density **J** and the electric field **E**. For copper at 77K with conductivity $\sigma = 450$ (S / μm), the mean free path can be calculated from the material constant (σ / l_e) $= 1.55 \times 10^{15}\ \Omega^{-1}\text{m}^{-1}$ [5]. Thus $\delta \geq l_e$ (= 0.29 μm) for frequencies $f \leq 6.6$ GHz. This neglect of anomalous skin effect was justified by calculating the anomalous surface impedance of a wide line. We found that it makes less than 1 % difference at 1 GHz, 6 % difference at 10 GHz and about 10 % difference at 20 GHz.

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SIMULATION APPROACH

We have developed in-house techniques to analyze the frequency-dependant impedances of general 2D and 3D lossy interconnect, using finite element techniques. There are many sophisticated finite element software packages such as ANSYS [6], for analysis of thermo-mechanical problems; our work has been to cast the electromagnetic problem into a form which can make use these finite element solvers. In this section we briefly describe our method (see also [2]).

The microstrip geometry is shown in Figure 1. We assumed a copper conductivity of 450 (S / µm); this is a value we have measured for electroplated copper and is close to bulk values. The relative dielectric constant of the polyimide was assumed to be 3.5. The simulation we perform corresponds to driving a short differential segment of the line with a voltage step, while the end of the line is mathematically shorted. The skin parameters are calculated by solving the partial differential equation for the time evolution of the magnetic vector potential **A** :

$$-\nabla^2 \mathbf{A} + \sigma\mu \frac{\partial \mathbf{A}}{\partial t} + \mu\varepsilon \frac{\partial^2 \mathbf{A}}{\partial t^2} = \mathbf{J_o}$$

Proper current continuity and boundary conditions were applied to maintain current continuity and ensure that the fields decay to zero at large distances. The impedance as a function of frequency is obtained from the current-voltage definitions based on power and energy considerations and then fitted to a linear circuit consisting of a series array of parallel resistor-inductor segments. This equivalent circuit can then be used in SPICE or equivalent circuit simulation programs to simulate the transmission line response in complex circuits. The frequency we use extends from low frequencies up to about 10 GHz. This upper limit is somewhat artificial: the method can be extended to much higher frequencies, but would require finer spatial resolution and increased computer time and expense. 10 GHz was chosen as the upper frequency limit for the reasons discussed above.

SIMULATION RESULTS

Impedances were obtained for the line geometry shown in Figure 1. We started with the "nominal" geometry of 10 µm wide and 5 µm thick lines with 10 µm thick dielectric and also considered the same geometry scaled down by a factor of 2 and 5. Results from the finite element simulations for the resistances of such interconnects as a function of frequency and the ones obtained by the wide line approximation (WLA) used by Kautz [3] are shown in Figure 2. WLA grossly overestimates the resistances by a factor of 2 at low frequencies and by a factor of 4 at higher frequencies. This is because at low frequencies, the WLA includes an equal resistance contribution from the ground plane which we find to be relatively small while at high frequencies, there are additional conducting paths on the sides of the interconnects and the ground plane. At high frequencies, the resistance can be expressed as $R_{sn} = \frac{G}{\sigma\delta}$ where G is a geometric factor independent of material parameters. This factor G is closely related to the shape factor discussed in [7]. To obtain an interpretation of G, we assume that fields decay exponentially from the surface into the conductor at high frequencies in which case the general formula of resistance based on (I^2R / I^2)-type power relation can be simplified as follows:

$$R_{sn} = \frac{\int_{C+G} \sigma |E|^2 \, dS}{\left|\int_{C/G} \sigma E \, dS\right|^2} \rightarrow \frac{1}{\sigma\delta} \frac{\int_{C+G} |E|^2 dl}{\left|\int_{C/G} E \, dl\right|^2} = \frac{G}{\sigma\delta} \text{ as } \omega \rightarrow \infty$$

The integrals in the numerator are evaluated for the conductor as well as the ground plane whereas the ones in the denominator are evaluated either for the conductor or the ground plane, because the current in the conductor is same as the current in the ground plane. The surface S is the cross sectional surface and the contour l is the path bounding S. It can be directly inferred that for wide lines G = (2 / width) and that G scales up by the scaling factor as the geometries are scaled down. From simulations it was found that the ground plane contributed about one-third the total resistance at high frequencies and G ≈ (0.5 / width) for the nominal geometry. Simulation results indicate that the internal impedance of the copper interconnect per unit length closely follows the approximation :

$$Z_{sn} = G\sqrt{\frac{j\omega\mu}{\sigma}} \coth(\sqrt{j\omega\mu\sigma}\, G\, W\, t)$$

Figure 3 and 4 show the variation of attenuation per unit length $\alpha = 20 \log_{10} e \, \mathrm{Re}(\gamma)$ and phase velocity $v = 2\pi f / \mathrm{Im}(\gamma)$ as a function of frequency , γ being the propagation constant for the frequency f. Figure 5 shows the real and imaginary part of the surface impedance (both normal and anomalous) for wide copper microstrips, which were calculated based on the theory of Reuter and Sondheimer [5].

The next set of results are meant to evaluate the performance of copper interconnects in digital wafer scale circuits at 77 K. The linearized circuit representation of a 1.0 cm long transmission line segment is shown in Figure 6. Two circuit configurations were investigated using SPICE; the first configuration consisted of the lossy transmission line with linear matched source and open far end and the second configuration had a non-linear mismatched CMOS driver ($\beta_p = 350$, $\beta_n = 175$) instead of the linear source. Figure 7 compares the response on copper lines of different lengths (10 cm, 30 cm and 50 cm) in the first configuration, calculated using the linearized circuit model which included skin-effect to the one neglecting skin effect elements but retaining the d.c. resistance contribution due to the finite cross section of the signal conductor (dashed curve). When the skin effect terms are neglected, the leading edge of the signal at the receiver-end is attenuated by $\exp(-\frac{R_t}{2 Z_{0\infty}})$, Rt being the total line resistance. However when skin effect terms are included, the signal starts from zero level after the time of flight delay and quickly follows the approximate characteristics obtained by neglecting the skin effect. In all our circuit simulations, skin effects were determined to be important for predicting the response for times less than 1.2 times the time of flight delay, while for larger instants the approximation of neglecting skin effects yield characteristics which were in close agreement with the exact results. This is somewhat fortuitous, as the skin resistance is partially compensated by the increased inductance at low frequencies. The far end of the transmission lines were kept open because most CMOS or BiCMOS circuits (which generally have driver output impedances greater than the transmission line characteristic impedance) currently designed adopt this approach. The open far end configuration doubles the receiver signal amplitude and dissipates no power under static conditions. The SPICE parameters for the CMOS inverter corresponded to an advanced 0.5 μm technology [8] with a large signal transconductance (i.e. transconductance at Vds=Vgs=1.2 V) of 86 mS / mm for n-channel and 43 mS / mm for p-channel. Both the transistors had a threshold voltage of 0.3 V. For the CMOS driver, the rise time of the input pulse at the last stage of the driver cascade was taken to be 0.2 ns. Figure 8 shows the effects of the non-linear CMOS driver on the response of 10 cm long copper interconnects at 77 K.

CONCLUSIONS

We have analyzed in detail the signal propagation through copper lines of different cross sections and lengths at 77K and 300K by including normal skin effect and effects of fringing fields. Copper wafer scale interconnects at 77K provide performance comparable to superconductors even when the present-day interconnect densities are scaled by a factor of two. However beyond this limit, as the need for higher densities for the first level of system interconnects grows and higher performance devices are available, superconductors will play an increasingly important role in alleviating the constraints posed by signal attenuation and dispersion. Calculation of crosstalk without neglecting skin effect, power distribution systems, cost etc. also need to be analyzed as part of an evaluation of the use of superconductors in digital systems.

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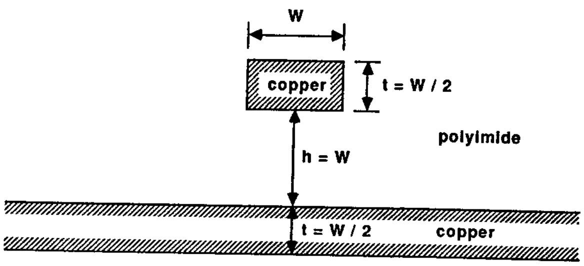


Figure 1. Microstrip geometry

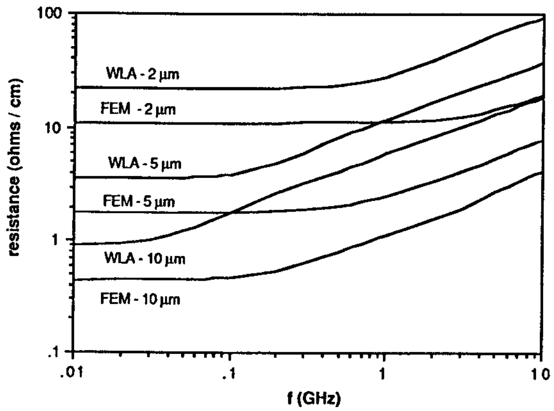


Figure 2. Resistance versus frequency (77 K)

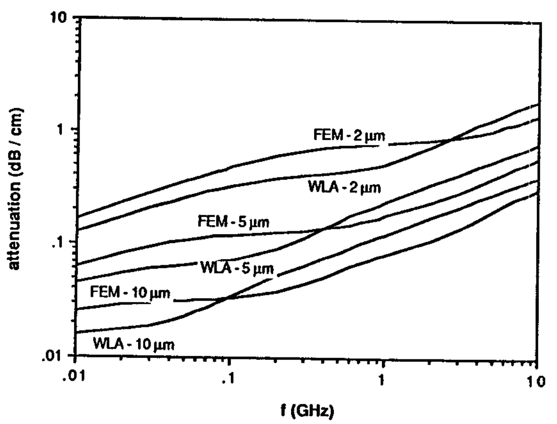


Figure 3. Attenuation versus frequency (77 K)

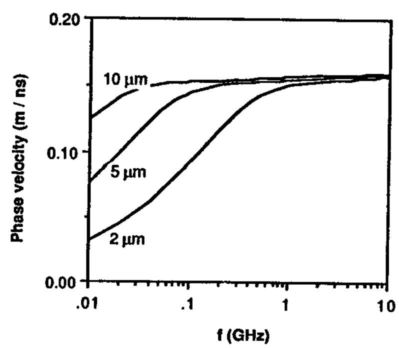


Figure 4. Phase velocity versus frequency (77K)

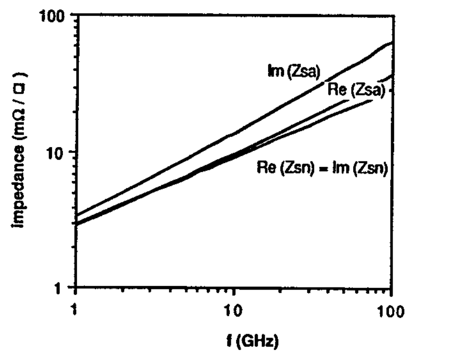
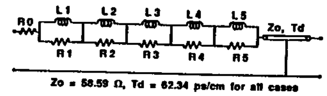


Figure 5. Anomalous surface impedance



Case	77 K 10 µm x 5 µm	77 K 5 µm x 2.5 µm	77 K 2 µm x 1 µm	300 K 10 µm x 5 µm	300 K 2 µm x 1 µm
R0 (Ω/cm)	0.45	1.77	11.1	3.57	86
R1 (Ω/cm)	0.46	1.11	9.11	5.42	0
L1 (nH/cm)	0.14	0.18	0.19	0.17	0
R2 (Ω/cm)	0.32	3.57	9.45	13.91	0
L2 (nH/cm)	0.08	0.12	0.09	0.12	0
R3 (Ω/cm)	1.60	6.44	0	0	0
L3 (nH/cm)	0.06	0.07	0	0	0
R4 (Ω/cm)	1.93	0	0	0	0
L4 (nH/cm)	0.04	0	0	0	0
R5 (Ω/cm)	2.41	0	0	0	0
L5 (nH/cm)	0.02	0	0	0	0

Figure 6. Linear circuit model

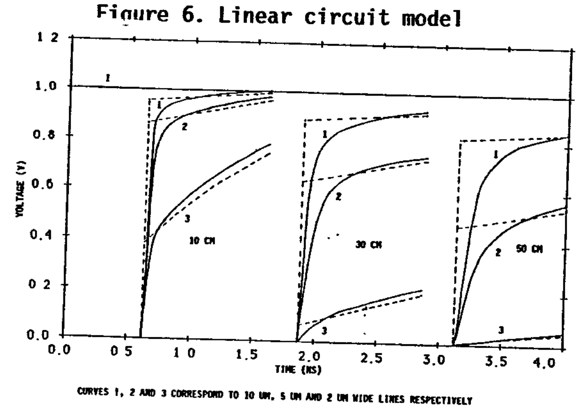


Figure 7. Line response for different lengths

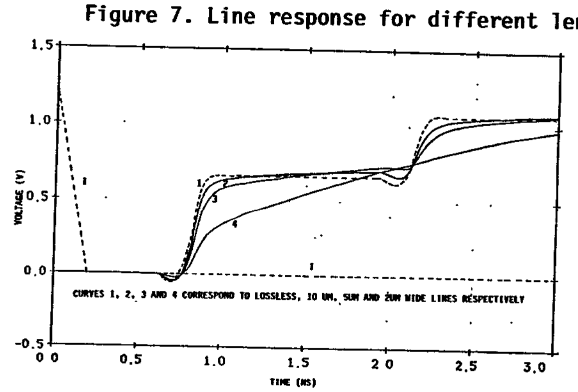


Figure 8. Response with a CMOS driver (77K)
Curves in 7 and 8 were smoothened- # of segments =20