ANALYSIS OF COPLANAR ON-CHIP INTERCONNECTS ON LOSSY SEMICONDUCTING SUBSTRATE

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Abstract

In this paper, a method for analysis and modeling of coplanar transmission interconnect lines that are placed on top of silicon-silicon oxide substrates is presented. The potential function is expressed by series expansions in terms of solutions of the Laplace equation for each homogeneous region of layered structure. The expansion coefficients of different series are related to each other and to potentials applied to the conductors via boundary conditions. In plane of conductors, boundary conditions are satisfied at N_d discrete points with N_d being equal to the number of terms in the series expansions. The resulting system of inhomogeneous linear equations is solved by matrix inversion. No iterations are required. A discussion of the calculated line admittance parameters as functions of width of conductors, thickness of the layers, and frequency is given. The interconnect capacitance and conductance per unit length results are given and compared with those obtained using full wave solutions, and good agreement have been obtained in all the cases treated.

Keywords

Lossy transmission lines, silicon, semiconductor substrate, frequency dependent capacitance, conductance

1. Introduction

Conducting substrates cause different effects, e.g. coupling ones. For low substrate conductivity (up to 10 S/m), there are strong capacitive coupling effects between adjacent lines due to very small line-to-ground capacitances in comparison to line-to-line capacitances. In this case, the capacitive coupling dominates the overall coupling behaviour. Transmission interconnect lines on MIS structures have been investigated for many years. There are many techniques for computing interconnect shunt admittance parameters. In [2], [3] Hasegawa et. al. presented an analysis of microstrip line on a Si-SiO2 system using parallel-plate waveguide model. In [1], the new model is developed to represent fin line and wide line interconnect behaviour over a 20 GHz frequency range and includes the substrate conductance effects. In [4], propagation properties of multilayer coplanar lines on different types of silicon substrates are investigated. In [5], [6], quasi-analytical analysis of broadband properties of multiconductor transmission lines on semiconducting substrates is done, and the calculated results for line parameters as function of frequency are discussed. Numerous electromagnetic approaches have been published which contains results of numerical full-wave or quasi-TEM analyses [7] – [12]. We can mention, the method based upon the classical mode-matching procedure [7], the spectral-domain analysis method [8], [9], and the finite element method [10] have been investigated for this structure. Recently quasi-TEM analysis on coplanar structure has made the incorporation of metallic conductor losses in the analysis possible and has provided a physical basis for the construction of equivalent circuits [11]. In [12], the CAD-oriented equivalent-circuit modeling procedure based on a quasi-stationary spectral domain approach that takes into account the skin effect in the silicon semiconducting substrate is presented.

The purpose of this paper is a slight modification of a recently proposed series expansion method [13] – [15], developed for the electrical modeling of lossy-coupled multilayer interconnection lines, that does not involve iterations and yields solutions of sufficient accuracy for most practical interconnections as used in common VLSI chips. We use here a Fourier series restricted to cosine functions. The solution for the layered medium is found by matching the potential expressions in the different homogeneous layers with the help of boundary conditions. In the plane of conductors, boundary conditions are satisfied only at a finite, discrete set of points (point matching procedure) [16].

2. Method of Analysis

The new modeling procedure is described for typical on-chip interconnects on a lossy silicon substrate with permittivity ε_s and conductivity σ, as shown in Fig. 1. It consists of signal lines and two ground lines in the same plane.

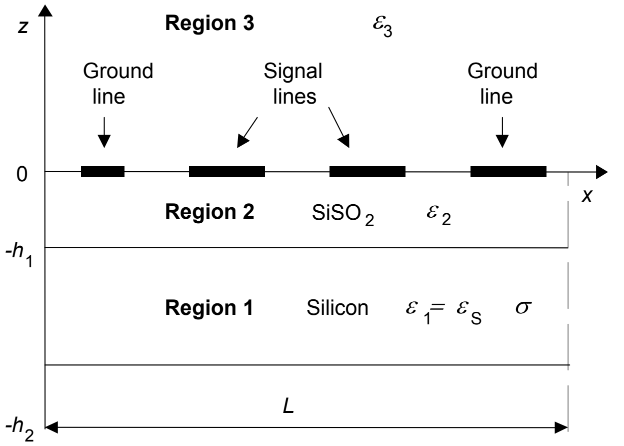


Fig. 1 Cross-sectioned view of the coplanar interconnect lines on lossy silicon substrate

The whole structure in x direction is bounded by open surfaces (Neumman type boundary conditions) $x = 0$ and $x = L$, respectively.

If the conductivity is small enough or the frequency is high enough but still well below the quasi-stationary frequency limit, in each layer of the structure, the electric potential must be a solution of the Laplace equation

$$\frac{\partial^2 \varphi}{\partial z^2} + \frac{\partial^2 \varphi}{\partial x^2} = 0. \tag{1}$$

In the most general case, the field variation in the x direction could be described as a Fourier integral. However, for numerical computations, discrete series are easier to handle than integrals. For this reason, we use an expansion of the potential function in terms of cosine functions [13] – [15]. The functions $\cos[m(\pi/L)x]$ form a complete orthogonal set [17] over the domain $0 < x < L$ for the integer values of m. Thus, we may express the potential function φ in regions 3 and 2 as follows:

$$\varphi_3(x,z) = a_0 + \sum_{m=1}^{\infty} a_m \exp\left[-(m\pi/L)z\right]\cos(m\pi x/L) \tag{2}$$

for $0 \le z < \infty$, and

$$\varphi_2(x,z) = b_0 + c_0 z + \sum_{m=1}^{\infty} \left\{ b_m \exp\left[-(m\pi/L)z\right] + \right. \\ \left. + c_m \exp\left[(m\pi/L)z\right] \right\} \cos(m\pi x/L); \tag{3}$$

for $-h_1 \le z \le 0$, respectively.

Results obtained from the full-wave analysis [6] have shown that influence of the finite substrate thickness h_2-h_1 can be neglected for practical dimensions, i.e., $(h_2$-$h_1) >> h_1, s, w$, for the case without ground plane.

In region of silicon substrate, which is assumed to be infinitely thick ($h_2 \to \infty$), we may write the potential as

$$\varphi_1(x,z) = d_0 + \sum_{m=1}^{\infty} d_m \exp\left[\frac{m\pi}{L}(z+h_1)\right] \cos\left(\frac{m\pi x}{L}\right) \tag{4}$$

for $z \le -h_1$. The different functional forms of the series expansions for potential distribution in the structure given by eqns. (2) to (4) are dictated by the boundary conditions.

3. Boundary Conditions and Shunt Admittance Parameters Calculation

In order to determine the potential distribution in the structure the boundary conditions must be satisfied. At interface between dielectric layers, two boundary conditions need to be satisfied, i.e., the potential and the normal component of electric induction vector must be continuous. This leads to the following expression for the structure shown in Fig. 1:

$$b_m = \frac{1}{2}\left(1 - \frac{\varepsilon_1}{\varepsilon_2}\right) \exp\left[-m\frac{\pi}{L}h_1\right] d_m \quad \text{for} \quad m = 1,2,... \tag{5}$$

$$c_m = \frac{1}{2}\left(1 + \frac{\varepsilon_1}{\varepsilon_2}\right) \exp\left[m\frac{\pi}{L}h_1\right] d_m \quad \text{for} \quad m = 1,2,... \tag{6}$$

$$b_0 = d_0 \quad \text{and} \quad c_0 = 0 \tag{7}$$

The boundary conditions in the plane of the infinitesimally thin coplanar conductors at $z = 0$ are slightly more complicated. Here, we require again that potential function φ_i assume the same value on either side of the interface. This requirement leads to the following conditions

$$a_m = b_m + c_m \quad \text{for} \quad m = 1,2,... \tag{8}$$

$$a_0 = b_0. \tag{9}$$

But the continuity of the normal component of the electric induction vector now holds only in the gaps between the coplanar conductors and not on the perfectly conducting interconnect conductors themselves. On the surface of the interconnect conductors the potential function need to be equal to the applied voltages $V(x)$. We write the conductor potential as a function of x to indicate that its value is different on different conductors in the structure even though $V(x)$ is constant on each conductor. Thus, we obtain using the relations (8) and (9), the following set of equations:

$$b_0 + \sum_{m=1}^{N-1} (b_m + c_m) \cos\left[\frac{m\pi}{L}x_j\right] = V(x_j) \tag{10}$$

on the conductors

$$\sum_{m=1}^{N-1} m\left[\left(\frac{\varepsilon_3}{\varepsilon_2} - 1\right) b_m + \left(\frac{\varepsilon_3}{\varepsilon_1} + 1\right) c_m\right] \cos\left[\frac{m\pi}{L}x_j\right] + \\ + \frac{L}{\pi}c_0 = 0 \quad \text{in the gaps} \tag{11}$$

The subindex j to the x coordinate means that we satisfy the boundary conditions in the conductor plane $z = 0$ only at a finite set of discrete points $x = x_j$, which may be chosen arbitrarily. In our case, for convenience, discrete points are spaced equidistantly. There are N points in the conductor plane $z = 0$ if there are N terms in the series expansion in order to provide as many equations (one for each j) as we have unknown coefficients d_m. It should be clear that the only remaining undetermined coefficients are the d_m, since the a_m, b_m and c_m all depend unambiguously on d_m via eqns. (5) to (9). It is also important to note that the sets of eqns. (10) and (11) cannot be counted as furnishing $2N$ equations since (10) is used only on the conductor surfaces, while (11) is used only in the inter-conductor space (gaps) at $z = 0$.

Point matching procedures simplifies the computational complexity considerably. There is no need for evaluating integrals over products of the potential $V(x)$ with the orthogonal set of functions $\cos[m(\pi/L)x]$. Other methods of analysis [13], [15] require that the value of the voltage in the inter-conductor space be obtained by iteration, starting from a suitable initial choice. None of these complications occurs in the point matching procedure. We need only to compute elements of the matrix **A** in the equation system

$$\sum_{m=0}^{N-1} A_{jm}\, d_m = \begin{cases} V(x_j) & \text{on conductors} \\ 0 & \text{in interconductor gaps} \end{cases} \tag{12}$$

defined by (5) to (11). Inversion of the matrix leads to the computation of the unknown coefficients via

$$d_m = \sum_{j=0}^{N-1} \left(A^{-1}\right)_{mj} \begin{cases} V(x_j) & \text{on conductors} \\ 0 & \text{in interconductor gaps} \end{cases} \tag{13}$$

Once the coefficients d_m are determined, the other expansion coefficients can be obtained using (5) to (9) and the potential function distribution follows from (2) to (4), respectively.

Once the potential distribution is available, it is easy to calculate the capacitance C and conductance G per unit length of the examined transmission line structure. The lossy semiconducting substrate is taken into account by the complex permittivity

$$\varepsilon_{cs} = \varepsilon_s - j\sigma/\omega \tag{14}$$

where ε_s is the permittivity and σ conductivity of the semiconducting substrate (silicon).

Due to the quasi-TEM character of the electromagnetic fields in examined structure the frequency dependent distributed admittance per unit length Y is calculated as

$$Y = G + j\omega C = j\omega Q/\Delta V \tag{15}$$

where Q is the total charge per unit length and ΔV denotes the voltage difference between the conductors. Since we can calculate the capacitance and conductance per unit length of the examined structure very easily with the new procedure, all quasi-stationary propagation parameters of multiconductor transmission lines may be obtained.

4. Numerical Results

In order to demonstrate the suitability of the new formulation for the potential distribution computation, we present some examples. The numerical results and all graphs are calculated and constructed by computing the potential distribution at those x values that coincide with the points used for point matching; 150 terms are used in the series expansions.

4.1 Example 1

To illustrate and validate the new proposed formulation, a coplanar strip interconnects structure shown in Fig. 2 is considered. The technological and geometrical parameters of this structure are:

w = 9.6 μm, w_g = 20 μm, s_g = 100 μm, t_{ox} = 0.58 μm, t_{si} = 500 μm, $\varepsilon_{ox} = 3.9\ \varepsilon_0$, $\varepsilon_{si} = 11.8\ \varepsilon_0$, σ = 15.5 S/m.

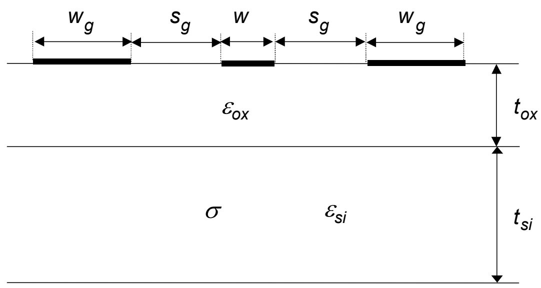
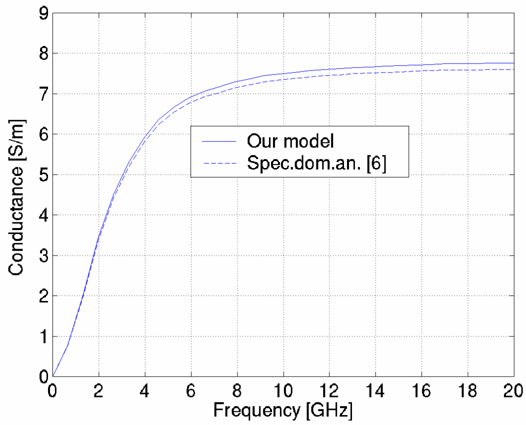
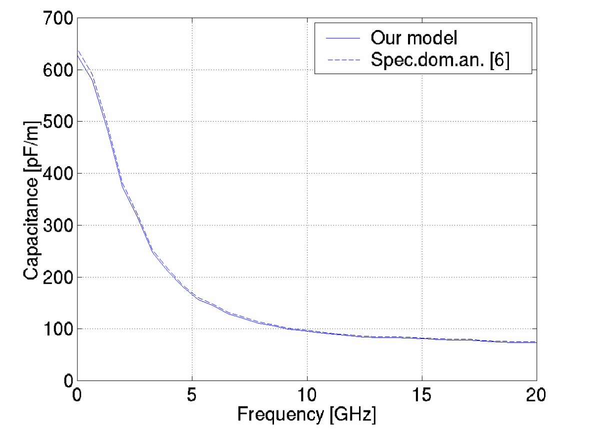


Fig. 2 Symmetric coplanar strip interconnects configuration

The frequency-dependent per unit length capacitance and conductance parameters for coplanar strip interconnects structure are shown in Fig. 3.



(a)



(b)

Fig. 3 The frequency response of (a) conductance per unit length and (b) capacitance per unit length. The solid lines in the figures are obtained with our model, and the dashed lines are the result from the spectral domain approach.

It can be seen that the frequency response calculated by using new formulation (point matching method with cosine Fourier series) is in very good agreement with that computed from the quasi-analytical analysis [6]. As expected, the lossy silicon substrate has a significant impact on the frequency-dependence of the line parameters of coplanar strip interconnect. It can be seen that the conductance per unit length G rapidly increases (see Fig. 3a) in the lower frequency range while the capacitance per unit length C decreases (see Fig. 3b).

4.2 Example 2

As the second application, the distributed capacitance and conductance per unit length for two coplanar coupled interconnect lines shown in Fig. 4 is considered.

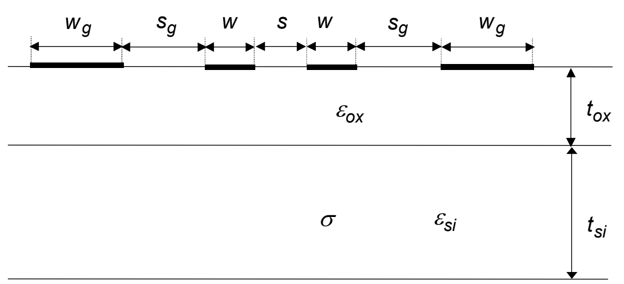
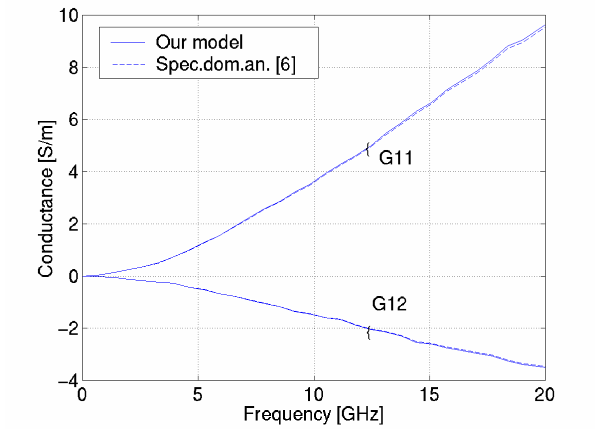


Fig. 4 Symmetric coplanar coupled interconnect structure with two narrow signal lines and two wide ground metallization

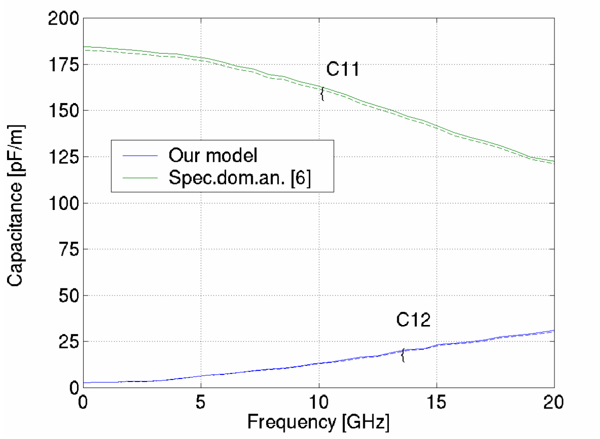
This symmetric interconnect geometry has the following electrical and geometrical parameters:

w = 2.0 µm, w_g = 20 µm, s = 2.0 µm, s_g = 100 µm, t_{ox} = 0.50 µm, t_{si} = 500 µm, ε_{ox} = 3.9 ε_0, ε_{si} = 11.8 ε_0, σ = 100 S/m.

For comparison, the same symmetric coupled strip coplanar interconnect problem is also rigorously solved by using spectral domain approach [6] with Chebyshev polynomial basis functions weighted by appropriate edge factors. The conductance and capacitance per unit length of the coupled interconnects are calculated by using our point matching method as a function of frequency (f = 0 – 20 GHz), and compared with those of the spectral domain approach (full-wave solvers). We can see that the calculated results by our method are in very good agreement with the rigorous full-wave method solutions for whole frequency range. A comparison of the frequency response of the point matching-cosine Fourier series approach with that computed by the spectral domain technique [6] (Fig. 5a and b) shows that our approach yields very good results with little computation efforts.



(a)



(b)

Fig. 5 Self and mutual shunt admittance components: (a) conductance per unit length and (b) capacitance per unit length.

5. Conclusion

We have presented a point matching method and cosine Fourier series approach of single and coupled coplanar interconnect lines on lossy silicon substrate based on the quasi-stationary field analysis. Frequency-dependent values for the capacitance and conductance per unit length matrices have been calculated by a very simple formulation (no iterations needed) of the problem that is well suited for computer simulations with little programming effort.

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