FDTD Simulation of Signal Degradation in Lossy and Dispersive Coplanar Waveguides for High-Speed Digital Circuits

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ABSTRACT

The signal degradation stemming from the conductor loss and the substrate dispersion of coplanar waveguides for the high-speed digital circuits is analyzed by a. combination of the original and frequency-dependent versions of the FDTD method. The metallic strips with the finite thickness are described by the equivalent Z-parameters. The backed conductor, if any, is modelled by the surface impedance. The dispersive property of the GaAs substrate is formulated by using the Kronig-Kramers relationship. The simulation results demonstrate the signal degradation due to modal dispersion, conductor loss, substrate dispersion and leakage effects in a conductor backed CPW structure.

Introduction

There has been increasing interest in the use of coplanar waveguides (CPW) in monolithic microwave and millimeter wave integrated circuits and high speed digital circuits. For operating speeds higher than gigabit per second, the modal dispersion and losses, including the conductor loss, the substrate dispersion and the radiation loss, will become major factors that lead to the amplitude and the waveform distortion. It is, therefore, of tremendous practical benefit to understand these properties.

This paper analyzes the signal degradation stemming from the conductor loss and the substrate dispersion of coplanar waveguides for the high-speed digital circuits. The analysis is based on the use of a combination of the original and frequency-dependent versions of the FDTD method. The significant contributions of this paper include: The metallic strips with the finite thickness are described by the equivalent Z-parameters, instead of the thin surface impedance sheet; The backed conductor, if any, is modelled by the surface impedance; The dispersive property of the GaAs substrate is formulated by using the Kronig-Kramers relationship, inclusive of the lattice vibration and the free-carrier absorption; and the algorithm for the FDTD scheme developed allows the fields to be updated recursively. Some interesting results are presented and discussed for a typical conductor backed CPW structure.

Formulation

Let us consider the conductor backed coplanar waveguide shown in Fig.1. The formulation will be briefly summarized as follows, although the extension of the present formulation to the more complicated structures including the discontinuities is straight-forward.

In essence, our analysis is based on the use of a combination of the original and frequency-dependent versions of the FDTD method. In the air region, the conventional Yee's FDTD scheme is used. But elsewhere, significant modifications to the original version must be made to account for their frequency-dependent properties.

(a) For the metallic strips

Analogous to the theory of transmission lines, the electromagnetic behavior of the metallic strips with the finite thickness can be described by the equivalent Z parameters as

$$\begin{bmatrix} \overline{E_{td}} \\ \overline{E_{tu}} \end{bmatrix} = \begin{bmatrix} Z_{11} & Z_{12} \\ Z_{21} & Z_{22} \end{bmatrix} \cdot \begin{bmatrix} \overline{H_{td}} \\ \overline{H_{tu}} \end{bmatrix} \tag{1}$$

where the subscriptions td and tu represent the lower and upper tangential field components. The elements of matrix [Z] are

$$Z_{11} = Z_{22} = Z_{s1}\coth(\gamma_1 t_c)$$

$$Z_{12} = Z_{21} = Z_{s1}\operatorname{csch}(\gamma_1 t_c)$$

$$Z_{s1} = \sqrt{\frac{j\omega\mu}{\sigma_1}}\ , \quad \gamma_1 = \sqrt{j\omega\mu\sigma_1}$$

Eqn.(1) can be rewritten as

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$$\begin{bmatrix} \overline{E_{td}} \\ \overline{E_{tu}} \end{bmatrix} = \begin{bmatrix} Z'_{11} & Z'_{12} \\ Z'_{21} & Z'_{22} \end{bmatrix} \cdot \begin{bmatrix} j\omega \overline{H_{td}} \\ j\omega \overline{H_{tu}} \end{bmatrix} \qquad (2)$$

where

$$Z'_{11} = Z'_{22} = \frac{1}{j\omega} Z_{s1} \coth(\gamma_1 t_c)$$

$$Z'_{12} = Z'_{21} = \frac{1}{j\omega} Z_{s1} \operatorname{csch}(\gamma_1 t_c)$$

In the time domain, eqn.(2) can be expressed in terms of the convolution form as

$$\overline{E_{td}}(t) = \int_0^t \phi_{11}(t-\tau) \frac{\partial \overline{H_{td}}(\tau)}{\partial \tau} d\tau + \int_0^t \phi_{12}(t-\tau) \frac{\partial \overline{H_{tu}}(\tau)}{\partial \tau} d\tau \qquad (3a)$$

$$\overline{E_{tu}}(t) = \int_0^t \phi_{21}(t-\tau) \frac{\partial \overline{H_{td}}(\tau)}{\partial \tau} d\tau + \int_0^t \phi_{22}(t-\tau) \frac{\partial \overline{H_{tu}}(\tau)}{\partial \tau} d\tau \qquad (3b)$$

where $\phi_{ij}(t)$ (i,j=1,2) have the explicit expressions as [1]

$$\phi_{11}(t) = \phi_{22}(t) = \frac{1}{\sigma t_c} \theta_3(0 \,|\, t\mu\sigma t_c^2) \qquad (4a)$$

$$\phi_{12}(t) = \phi_{21}(t) = \frac{1}{\sigma t_c} \theta_4(0 \,|\, t\mu\sigma t_c^2) \qquad (4b)$$

where θ_3 and θ_4 are theta functions. They can be found in terms of a series of exponential functions [2]. Substituting eqn.(4) into (3), the tangential electric fields can be solved in a recursive manner.

(b) For the backed conductor

Due to the finite conductivity, a fraction of fields will penetrate into the backed conductor. This means the interface between the substrate and the conductor is no longer an electric wall. Such an interface should be modelled by a surface impedance as

$$\overline{E_{tc}} = Z_{s2} \overline{H_{tc}} \qquad (5)$$

Similarly, eqn.(5) can be written in the time domain as

$$\overline{E_{tc}}(t) = \int_0^t \phi_c(t-\tau) \frac{\partial \overline{H_{tc}}(\tau)}{\partial \tau} d\tau \qquad (6)$$

where

$$\phi_c(t) = \sqrt{\frac{\mu}{\sigma}} \cdot \frac{1}{\sqrt{\pi t}} \qquad (7)$$

When $\phi_c(t)$ is approximated in terms of a decaying series of exponential functions, eqn.(6) is evaluated recursively.

It should be mentioned that eqns.(3) and (5) are found based on the assumption that the electric and magnetic fields are collocated in space. In accordance with the FDTD scheme, the tangential magnetic fields on the interfaces are approximated by those that offset half spacial cell to the interfaces.

(c) For the GaAs substrate

According the theory of solid state physics [3], the dispersive property of the GaAs substrate is attributed mainly to the lattice vibration and the free-carrier absorption. The complex permittivity is formulated as

$$\epsilon = \epsilon_0 \epsilon_\infty \left(1 + \frac{\omega_L^2 - \omega_T^2}{\omega_T^2 - \omega^2 + j\omega\Gamma} + \frac{\omega_p^2}{\omega(j\nu_c - \omega)} \right) \qquad (8)$$

where ϵ_∞ is the lattice relative dielectric constant as $\omega \to \infty$, ω_L and ω_T are the transverse- and the longitudinal- optic- phonon frequencies. Γ is the phonon damping constant, ω_p defines the plasma frequency, and ν_c is the free-carrier damping constant. In the time domain, we have

$$\overline{D}(t) = \epsilon_0 \epsilon_\infty \overline{E}(t) + \epsilon_0 \int_0^t \chi(t-\tau) \overline{E}(\tau) d\tau \qquad (9)$$

$\chi(t)$ can be found as

$$\chi(t) = \chi_1(t) + \chi_2(t) \qquad (10)$$

where

$$\chi_1(t) = \frac{\epsilon_\infty (\omega_L^2 - \omega_T^2)}{\beta} e^{(-\frac{\Gamma}{2}t)} \sin(\beta t)$$

$$\chi_2(t) = \frac{\epsilon_\infty \omega_p^2}{\nu_c} (1 - e^{(-\nu_c t)})$$

$$\beta = \sqrt{\omega_T^2 - (\frac{\Gamma}{2})^2}$$

Following the procedure suggested in [4], the E-field components can be recursively evaluated from eqn.(9).

(d) Source consideration and boundary treatment

The successive excitation similar to [5] is used. First, we calculate the static field distribution using the finite difference method in conjunction with the asymptotic boundary condition. The static solution is then used as the initial excitation of 2D FDTD simulation. For an estimated propagation constant, record the simulation results for E-fields at a certain time. The dynamic fields are further used as the spatial distribution of 3D FDTD excitation. Load plane, the open air region and the side walls are terminated by the first Mur's ABC.

Results and discussions

A typical conductor backed coplanar waveguide structure has been simulated. Where the central strip width w= 2.2mm, the slot width s= 1mm, the substrate thickness h= 2mm, the conductivity σ_1 of the metallic strip and the conductivity σ_2 of the backed conductor: $\sigma_1 = \sigma_2 = 3.33\times10^7$ s/m, the high frequency relative permittivity of the GaAs substrate $\epsilon_\infty = 12.85$, and the main optical parameters of the GaAs substrate are found from [3].

Fig.2 show comparison of temporary time responses of a launched Gaussian pulse with and without the conductor loss or the substrate dispersion. Where the FDTD cell Δx= 0.2mm, Δy= 0.2 mm, Δz= 0.25mm. It is obvious that the waveform distortion is mainly due to the modal dispersion, whereas the conductor loss and the substrate dispersion reduce the pulse amplitude.

Fig.3 shows the effect of the strip thickness on the pulse propagation. For operating speeds lower than or equal to gigabit per second, the strip thickness expects to make the greater contribution to the signal degradation, because the skin depth is comparable to the strip thickness.

Fig.4 shows the frequency domain transmission coefficient for a single CPW resonator structure including the coupling gaps. Where the resonator length l=5mm, the gaps of the resonator coupling with the source line and load line $G_1=G_2$= 0.4mm, the strip thickness t_c= 5mm. The FDTD cell Δx= 0.2mm, Δy= 0.2 mm, Δz= 0.2mm.

The dynamic distributions of the E_y-field component directly under the metallic strips along the propagation direction are shown in Fig.5. It is found that a fraction of fields leak along the transverse direction.

Conclusions

The signal degradation stemming from the conductor loss and the substrate dispersion of coplanar waveguides for the high-speed digital circuits has been analyzed by a combination of the original and frequency-dependent versions of the FDTD method. The present formulation has involved in several novel features. The results presented for the conductor backed CPW show that (1) The signal degradation is attributed mainly to the modal dispersion, the conductor loss and the substrate dispersion; (2) The strip thickness affect the signal amplitude, depending on the operating speed; and (3) There is the power leakage along the transverse direction.

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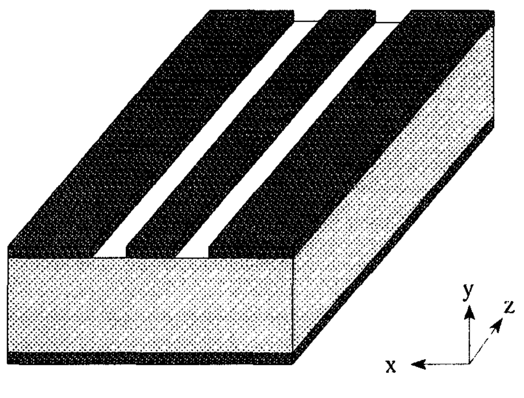
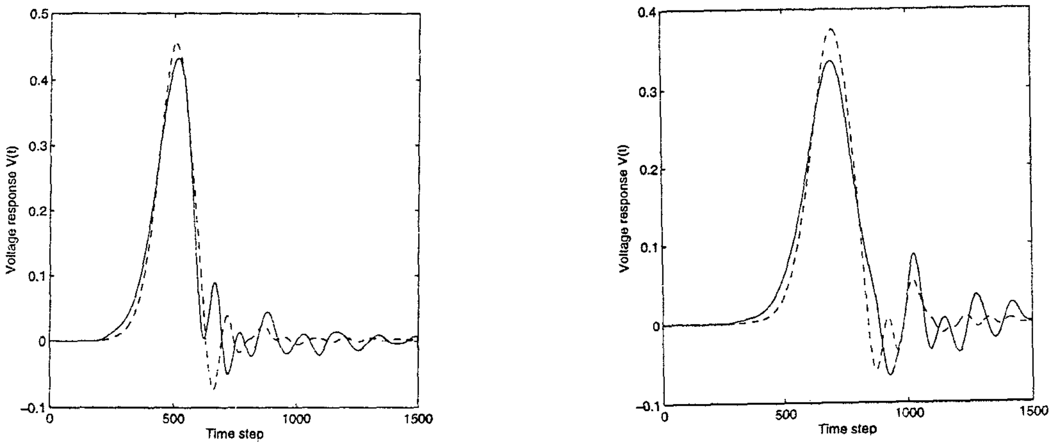


Fig.1 Conductor backed coplanar waveguide



(a) $z = 75\ \Delta z$ (b) $z = 100\ \Delta z$

Fig. 2 Comparison of the temporary time responses of a launched Gaussian pulse with (——) and without (----) the conductor loss or the substrate dispersion. the strip thickness t_c=10μm, the free-carrier density $n_e = 1\times10^9\ cm^{-3}$

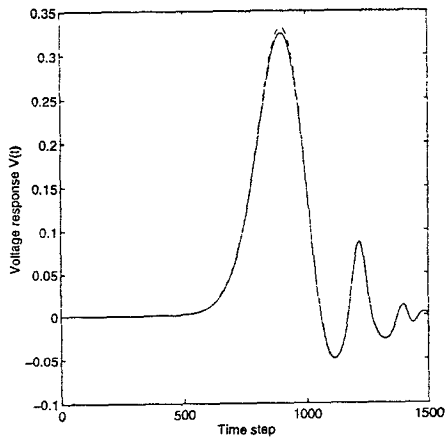


Fig.3 Effect of the strip thickness on the pulse propagation. —— t_c = 0.5 μm, ---- t_c = 5 μm

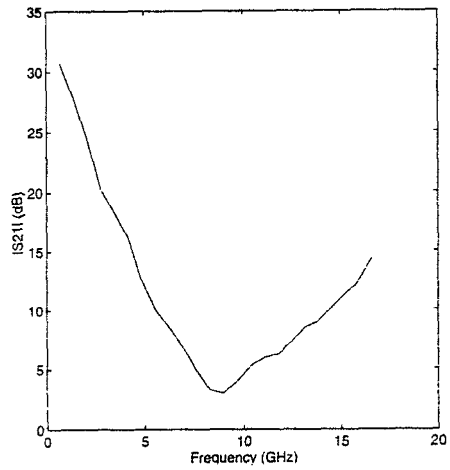
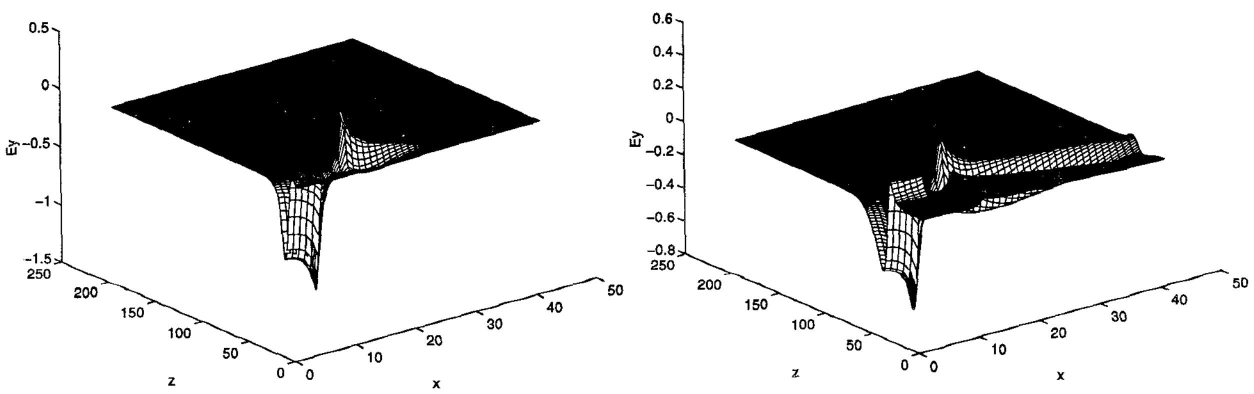


Fig.4 Frequency domain transmission coefficient of a single resonator including the coupling gaps



(a) $t = 250\ \Delta t$ (b) $t = 500\ \Delta t$

Fig.5 Dynamic distributions of the E_y-field component directly under the strips along the propagation direction