Losses in GaAs Microstrip and Coplanar Waveguide

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Abstract

The finite-difference transmission line matrix (FD-TLM) method is applied to loss analysis for microstrip and coplanar waveguide (CPW). A feature of FD-TLM with a variable mesh size is adopted in this loss analysis. The analysis is validated through comparison with a frequency domain method. The numerical results provide a clear picture of frequency dependence of losses up to $100GHz$ for microstrip and coplanar waveguide.

1. Introduction

Microstrip (Fig.1) and coplanar waveguide (Fig.2) have been used extensively in microwave circuits as transmission lines for wide ranging applications. The dependence of line losses on frequency has only been reported by few papers. There are several approaches that can be adopted for analyzing the losses of microstrip and CPW. The primary one is the quasi-static approach [1], with which it is difficult to find the accurate current distribution on the conductors. Another popular approach is based on the frequency domain full-wave method [2]. Using this method, people have found that a standard $0.1-mm$ GaAs semiconductor CPW has significantly less loss than a microstrip at higher values of characteristic impedance. In this paper the finite-difference transmission line matrix (FD-TLM) method [3] is used to study the losses of microstrip and CPW. As well, a comparison between the losses of microstrip and CPW with a characteristic impedance of around 50Ω is made to extract design data. Since this method allows the mesh size to be varied, the current distribution can be derived at locations that are very close to conductor surfaces. This unique feature guarantees the accuracy of the numerical results. The results show a clear picture of losses for microstrip and CPW over a wide range of frequency and provide a practical guide for the system design.

2. Formulations

A. FD-TLM Method

The finite-difference transmission line matrix (FD-TLM) method [3] is a hybrid technique which combines the transmission line matrix (TLM) method [4, 5] and finite-difference time-domain (FD-TD) method [7] for solving time-domain electromagnatic problems [8, 9]. The FD-TLM method offers both TLM's physical mechanism for wave propagation and FD-TD's computational efficiency. Fig.3 shows the three dimensional variable-mesh FD-TLM cell. Cells represented by six nodes can have sides of different length and are stacked together to fill the entire space being simulated. In this method, E nodes and H nodes are assumed as $E_x = V_x/u, E_y = V_y/v, E_z = V_z/w$ and $H_x = I_x/\overline{u}, H_y = I_y/\overline{v}, H_z = I_z/\overline{w}$. Here V is the voltage at an E node and I is the current at an H node. The quantities u, v, w are the mesh sizes in the x, y, z directions, respectively. According to the above assumption, the variable-mesh TLM method can be expressed in a finite difference form.

The corresponding finite-difference equation for E_z node is

$$\begin{aligned} V_z^{n+1} &= (1-\frac{2\,Y_{Lzn}}{K_z})V_z^n \\ &+ \frac{2\,Z_0}{K_z}(I_{x,j-1} - I_x + I_y - I_{y,i-1})^{n+\frac{1}{2}} \end{aligned} \quad (1)$$

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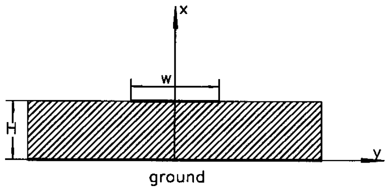

Figure 1: Microstrip.

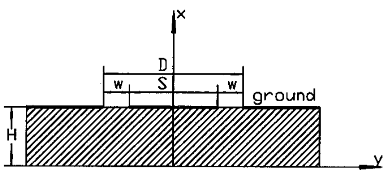

Figure 2: Coplanar waveguide.

where

$$K_z = Y_{yzn} + Y_{yzn,i-1} + Y_{xzn} + Y_{xzn,j-1} + Y_{Szn} + Y_{Lzn} \quad (2)$$

is the sum of the normalized admittances at the E_z node (Fig.4 (a)), Z_0 is the impedance of free space.

The corresponding finite-difference equation for H_x node is

$$I_x^{n+\frac{1}{2}} = I_x^{n-\frac{1}{2}} + \frac{2}{Z_0 M_x}(V_z - V_{z,j+1} + V_{y,k+1} - V_y)^n. \quad (3)$$

where

$$M_x = Z_{xzn,j+1} + Z_{xzn} + Z_{xyn} + Z_{xyn,k+1} + Z_{Pxn} \quad (4)$$

is the sum of the normalized impedances at the H_x node (Fig.4 (b)).

The other four equations for E_x, E_y and H_y, H_z can be obtained similarly.

B. Loss Analysis

It is well known that the losses per unit length can be represented in terms of an attenuation factor α in the expression for transmitted power $P(z)$

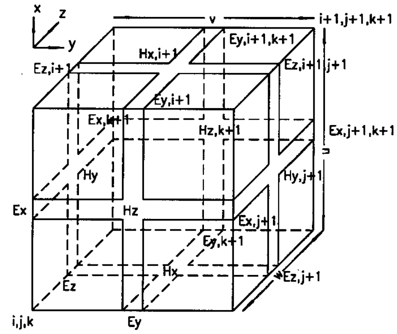

Figure 3: Variable-mesh FD-TLM cell.

$$P(z) = P_0 e^{-2\alpha z}, \quad (5)$$

where z denotes a point along the direction of propagation parallel to the strip conductor and P_0 the transmitted power at an earlier point $z = 0$. Letting $\alpha = \alpha_d + \alpha_c$, the sum of a dielectric attenuation factor α_d and a conductor attenuation factor α_c, we get

$$\alpha = \frac{P_c + P_d}{2P(z)} \quad (6)$$

or

$$\alpha_c = \frac{P_c}{2P(z)} \quad (7)$$

$$\alpha_d = \frac{P_d}{2P(z)} \quad (8)$$

here P_c and P_d denote the average conductor power loss and the average dielectric power loss per unit length.

The magnitude of the conductor surface current is calculated by

$$J^{u,l}(f) = \sqrt{|J_y^{u,l}|^2 + |J_z^{u,l}|^2} \quad (9)$$

where u and l superscripts refer to the current on the upper and lower sides of the conductor,

$$J_y^{u,l}(f) = \int_0^\infty H_z^{u,l}(t) e^{-j2\pi ft} dt \quad (10)$$

$$J_z^{u,l}(f) = \int_0^\infty H_y^{u,l}(t) e^{-j2\pi ft} dt. \quad (11)$$

where $H_y^{u,l}$ and $H_z^{u,l}$ are the tangential magnetic field components on upper and lower sides of the conductor. Then, based on equation (7), the conductor loss factor can be estimated by the equation:

$$\alpha_c(f) = \frac{R_s(f)}{2Z_0(f)I^2(f)}\left[\int_{-a}^{a} dy + 2\int_{b}^{\infty} dy\right] \cdot [|J_y^u|^2 + |J_z^u|^2 + |J_y^l|^2 + |J_z^l|^2] \quad (12)$$

where for microstrip $a = w/2, b = 0$, and for CPW $a = s/2, b = s/2 + w$. R_s is the surface skin resistivity calculated by $R_s(f) = \sqrt{\pi f \mu \rho_c}$ and $I(f)$ is obtained by the Fourier transform of the total current on the strip definded as the loop integral of the magnetic field around the metal strip. The characteristic impedance $Z_0(f)$ is obtained from the ratio of $V(f)/I(f)$. Here $V(f)$ is the Fourier transform of the line integral of vertical electric field under the center of the strip.

Dielectric loss is given by [6]

$$\alpha_d(f) = \frac{q\epsilon_r tan\delta}{\epsilon_{r_{eff}}(f)\lambda_g(f)} \quad (13)$$

where q is the filling factor for the dielectric constant, ϵ_r is the substrate relative dielectric constant, $tan\delta$ is the substrate loss factor. The effective dielectric constant $\epsilon_{r_{eff}}$ and the guide wavelength λ_g are obtained from the transfer function for a length, L, of microstrip or CPW which is

$$e^{-\gamma(f)L} = \frac{E_x(f, z = L)}{E_x(f, z = 0)} \quad (14)$$

where $\gamma(f) = \alpha(f) + j\beta(f)$. The permittivity $\epsilon_{r_{eff}}(f)$ is defined through $\beta(f)$ as $\beta(f) = 2\pi f\sqrt{\mu_0\epsilon_0\epsilon_{r_{eff}}(f)}$, or

$$\epsilon_{r_{eff}}(f) = \frac{\beta^2(f)}{\omega^2\epsilon_0\mu_0}. \quad (15)$$

The guide wavelength λ_g is defined as $\lambda_g = 2\pi/\beta(f)$.

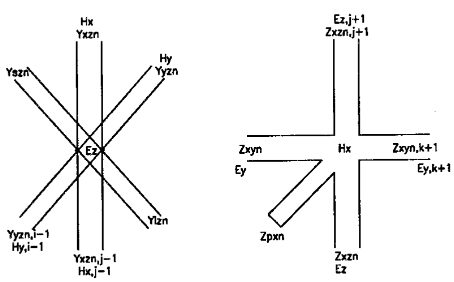

Figure 4: (a) E_z shunt node. (b) H_x series node.

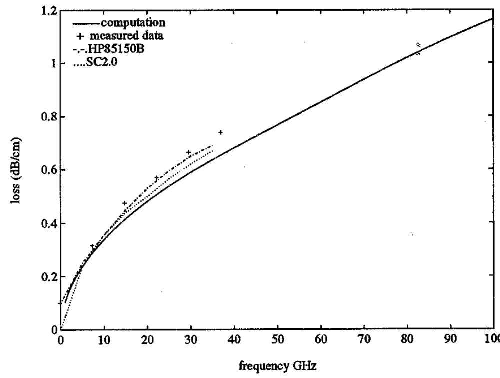

Figure 5: Microstrip losses ($W = 0.07mm, Zo = 50\Omega$).

3. Numerical Results

For a given substrate thickness and permittivity, the microstrip impedance is varied by only changing the strip width. In contrast to this, the CPW impedance depends roughly on the ratio of the inner conductor width to the total cross section. Thus, CPW with different sizes can have the same impedance. Since the purpose of this paper is to provide data for choosing optimum guiding structures for GaAs monolithic circuits, all results have been obtained for the typical values of $\epsilon_r = 12.8$, $tan\delta = 0.002$, gold resistivity $\rho = 2.4 \times 10^{-6}\Omega \cdot cm$ and substrate thickness $h = 0.1mm$. Conductor thickness is not considered in the computation.

Typical microstrip line was investigated in this paper. The calculated results for microstrip losses (Fig.5) show good agreement with the simulation results that were obtained by using SUPERCOMPACT 2.0, HP85150B and with the measurement results [10]. The CPW is also analyzed for four different sizes (see Table 1 for the size list), which have different combinations of S and D, but have the same characteristic impedance of around 50Ω. At a frequency of $60GHz$, the results show good agreement with that calculated by [2]. From the comparison in Fig.6 of losses for microstrip and CPW, it is interesting to note that for a characteristic impedance of 50Ω, some CPW combinations of S and D may have more loss, while, some combinations may have less loss than the microstrip. It can be observed that smaller cross sections have higher losses, and thus a trade-off exists between the size and loss.

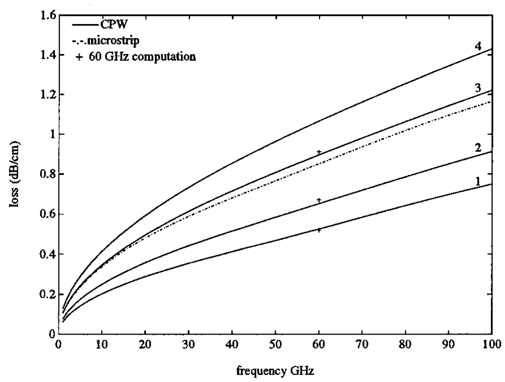

Figure 6: Comparison of the losses between microstrip and CPW with the same characteristic impedance of 50Ω (size shown in Table 1).

	D(mm)	S(mm)	W(mm)	K(S/D)
case 1	0.4	0.232	0.084	0.58
case 2	0.3	0.18	0.06	0.60
case 3	0.2	0.104	0.048	0.52
case 4	0.125	0.069	0.028	0.55

Table 1: CPW geometrical sizes for the computation

4. Conclusions

Two typical microwave transmission lines, microstrip and coplanar waveguide, are analyzed using the FD-TLM method to determine the losses over a wide frequency range. The FD-TLM method permits one to derive the current distribution on the conductor surfaces. It is observed that different combinations of S and D for 50Ω CPW have different losses. They may be greater or less than the losses for a microstrip up to $100GHz$. The line loss is just one parameter that must be considered when choosing a transmission structure. Other factors, such as size and structure complexity, should also be taken into account in practice.

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