FDFD Full-Wave Analysis and Modeling of Dielectric and Metallic Losses of CPW Short Circuits

H. Klingbeil, K. Beilenhoff, and H. L. Hartnagel

Abstract—**A finite-difference method in the frequency domain (FDFD) is used to analyze the influence of lossy materials on the scattering behavior of CPW short ends. Not only dielectric losses but also realistic metallic losses are taken into account for the first time in an FDFD method. Both, the numerical results for the three-dimensional structure and the complex propagation constants of the homogeneous waveguide are presented. These are compared with those yielded by an analytical method and shown to be of good agreement. Finally, a simple model is presented, which describes the CPW short end with good accuracy.**

I. Introduction

In the last few years coplanar waveguides (CPW's) were more and more often used as basic transmission lines in monolithic microwave integrated circuits (MMIC's). The increase of frequency leads to significantly higher attenuation for this kind of quasi-TEM waveguide due to the skin effect. Since the skin depth is of the same order of magnitude as the conductor thickness it is not possible to apply simple approximations [1]. Consequently, it is desirable to have a powerful tool enabling one to make a full-wave analysis of arbitrary lossy structures in order to allow an accurate circuit simulation. A first step was done in [2], where comparatively small losses were analyzed by means of a FD method. Realistic metallic losses, however, could not be considered. Very few publications are known where metallic conductivities are taken into account in three-dimensional (3-D) structures. Schmidt [3], e.g., analyzed the scattering behavior of a CPW step-discontinuity with a mode matching method. This method is not suitable, however, for the analysis of arbitrarily shaped structures.

In this paper, the finite-difference method in the frequency domain (FDFD) is used to analyze the scattering behavior of a CPW short end including the effect of lossy materials. The method, however, holds for arbitrary 3-D structures. For the values of the conductivity and the loss tangent no restrictions have to be obeyed.

II. The Numerical Method

The method presented here is based on the full wave finite-difference method in the frequency domain, which was already used in [4], [5] to analyze the scattering behavior of several lossless MMIC discontinuities. It is extended to the lossy case by introducing a complex permittivity. Thus, the current distribution inside the metallizations is found by discretizing the conductors sufficiently and is not approximated by surface impedance concepts. For the analysis of the scattering behavior of a 3-D structure two different calculations have to be performed.

1) First, one or more homogeneous waveguides have to be considered that serve as ports of the 3-D structure. This analysis leads to a complex eigenvalue problem (in the lossless case only real matrices occur).
2) The scattering behavior of the discontinuity is given by a three-dimensional boundary-value problem. It leads to a sparse complex system of linear equations. This system is solved for different excitations which allows the computation of the scattering matrix of the structure.

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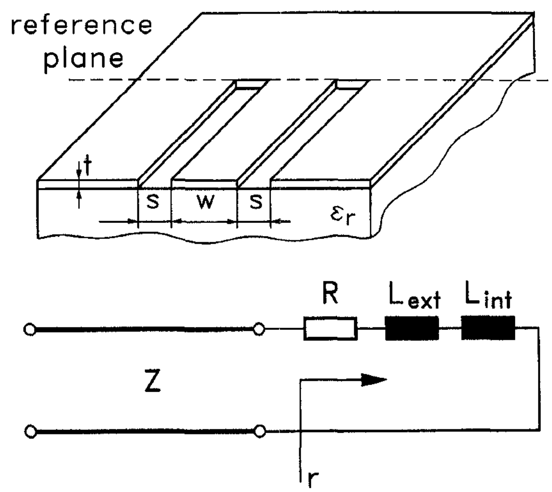

Fig. 1. CPW short end and lumped-element equivalent-circuit.

The numerical results presented in this paper clearly validate the implementation of the 2-D algorithm. The 3-D part was checked using the lossy step discontinuity presented in [3]. For $t = 10\,\mu$m and $f = 10$ GHz, e.g., the FDFD method yields $|S_{11}| = 0.194$, whereas $|S_{11}| = 0.185$ was published in [3]. After increasing the number of modes in [3] the agreement is even better.

III. Numerical Results

A simple CPW short end as shown in Fig. 1 is analyzed. The metallizations consist of Au with a conductivity of $\kappa = 3 \cdot 10^7\,\Omega^{-1}\,\mathrm{m}^{-1}$ and a thickness of $t = 3\,\mu$m. The ground-to-ground spacing $d = w + 2s$ was kept constant ($d = 50\,\mu$m), whereas the slot width s was varied in order to realize different line impedances. The ground metallizations are $w_g = 200\,\mu$m wide. GaAs is used as substrate ($\epsilon_r = 12.9$, $\tan\delta = 3 \cdot 10^{-1}$). Analyzing this structure with the FDFD method, matrix orders of 2000 for the 2-D case and 130 000 for the 3-D case are typical.

A. The Homogenous CPW

First of all, the homogeneous coplanar waveguide (cross section of Fig. 1) is analyzed. In Fig. 2 the finite-difference results of both the relative effective permittivity $\epsilon_{r,eff}$ and the attenuation α are compared with those obtained by an analytical method [6]. The results of the two methods show good agreement. The rising values of $\epsilon_{r,eff}$ for high frequencies occur due to a slight dispersion which, however, cannot be considered in the analytical model.

As Fig. 3 shows not only the propagation constants $\beta - j\alpha$, but also the complex values of the characteristic impedance yield good results. Furthermore, the field distribution inside the conductors exhibits the same behavior as computed with the mode-matching technique [1].

B. The Short End

The scattering behavior of a CPW short end is usually described by a length-extension $l_{ext,\beta}$ [7]. The length-extension $l_{ext,\beta} =$

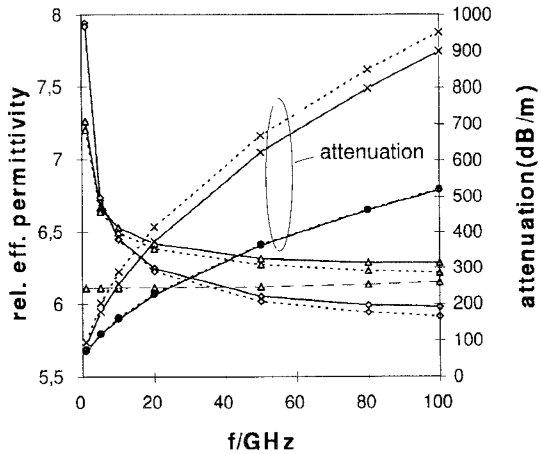

Fig. 2. Relative effective permittivity $\epsilon_{r,eff} = (\beta/\beta_0)^2$ (◇: $s = 5\,\mu$m; △: $s = 15\,\mu$m) and attenuation α (×: $s = 5\,\mu$m; •: $s = 15\,\mu$m) of a homogeneous coplanar waveguide (solid lines: lossy structure; dotted lines: [6]; dashed line: lossless structure).

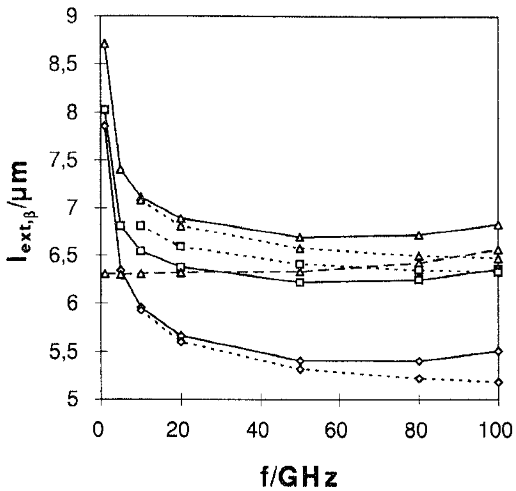

Fig. 4. Effective length extension $l_{ext,\beta}$ (◇: $s = 5\,\mu$m; △: $s = 15\,\mu$m; □: $s = 20\,\mu$m) of a CPW short end (solid lines: lossy structure, dotted lines: model [see (1)]; dashed line: lossless structure).

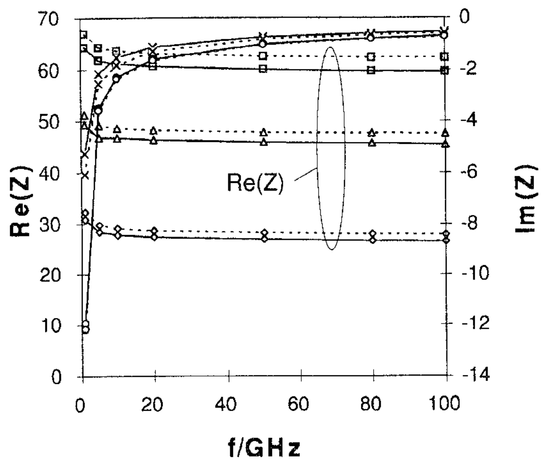

Fig. 3. Real part (◇: $s = 5\,\mu$m; △: $s = 15\,\mu$m, □: $s = 20\,\mu$m) and imaginary part (×: $s = 5\,\mu$m; o: $s = 20\,\mu$m) of the characteristic impedance Z_W of a homogeneous coplanar waveguide (solid lines: lossy structure, dotted lines: [6]).

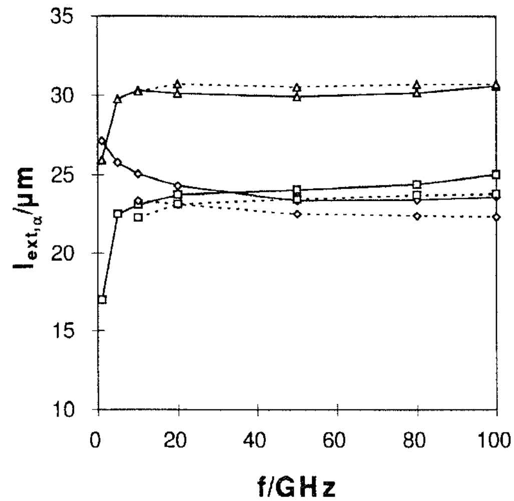

Fig. 5. Effective length extension $l_{ext,\alpha}$ (◇: $s = 5\,\mu$m; △: $s = 15\,\mu$m, □: $s = 20\,\mu$m) of a CPW short end (solid lines: lossy structure; dotted lines: model [see (1)].

$-(\varphi/2\beta)$ is defined as the length difference between a real short end and an ideal one where the two reflection coefficients have the same phase. φ is the phase of the reflection coefficient r. The modulus of r is equal to one if losses are neglected. However, if realistic metallic and dielectric losses are introduced, the reflection coefficient decreases. To describe this behavior, a second length-extension $l_{ext,\alpha} = -(\ln|r|/2\alpha)$ can be defined. These parameters were chosen because both length-extensions are almost constant over a wide frequency range.

The values of $l_{ext,\alpha}$ and $l_{ext,\beta}$, computed with the FDFD method, are presented in Figs. 4 and 5.

As it is shown in Fig. 4, the length extension $l_{ext,\beta}$ increases for decreasing frequencies. This result is interesting since it differs from the lossless case [7]. For frequencies higher than 10 GHz, this behavior occurs due to internal inductances in the lossy conductors which do not exist in ideal conductors. A reasonable equivalent circuit for a lossy CPW short end is shown in Fig. 1. For frequencies f higher than 10 GHz, the external inductance L_{ext} remains almost constant, whereas the internal inductance L_{int} decreases with $\sqrt{f}^{-1}$ [8]. The resistance R equals the imaginary part of the impedance of the internal inductance: $R = 2\pi f L_{int}$. For the mentioned frequency range, $d = 50\,\mu$m, $t = 3\,\mu$m and the considered materials the following modeling formulas hold:

$$\frac{R}{\sqrt{\frac{f}{\text{GHz}}}} = \left(6.3 + 0.25\,\frac{s}{\mu\text{m}}\right)\text{m}\Omega$$

$$L_{ext} = \left(0.35 + 0.13\,\frac{s}{\mu\text{m}}\right)\text{pH}. \quad (1)$$

The dotted curves in Figs. 4 and 5 show the length-extensions computed by this model. For frequencies lower than 10 GHz, the external inductance L_{ext} becomes frequency dependent because the skin depth δ is of the same order of magnitude as the metallization thickness, which leads to a change of the magnetic field distribution. For an application of the element below 10 GHz this effect has to be considered.

IV. Conclusion

The FDFD method, which had been shown to yield reliable results for nearly arbitrarily shaped structures, was expanded in order to allow the analysis of structures with lossy materials. A simple coplanar short circuit was investigated with this method. In order to verify the improved FDFD method the transmission line parameters (β, α, and Z_W) of a homogeneous waveguide were compared with results obtained by a highly sophisticated and well-proven model. Good agreement was observed. The effects already known from homogeneous coplanar waveguides were also observed for the short circuit, e.g., the external inductance changes significantly for low frequencies. This is due to conductor losses whereas the influence of the dielectric losses is comparatively small. Finally, a simple model was presented that describes the parasitics of a lossy short end with good accuracy.

References

[1] W. Heinrich, "Conductor loss on transmission lines in monolithic microwave and millimeter-wave integrated circuits," *Int. J. Microwave and Millimeter-Wave Computer-Aided Engineering,* vol. 2, no. 3, pp. 155–167, 1992.

[2] S. Haffa, D. Hollmann, and W. Wiesbeck, "The finite difference method for s-parameter calculation of arbitrary three-dimensional structures," *IEEE Trans. Microwave Theory Tech.,* vol. 40, no. 8, pp. 1602–1610, Aug. 1992.

[3] R. Schmidt and P. Russer, "Full wave analysis of coplanar waveguide discontinuities by partial wave synthesis," in *ACES,* 1994.

[4] A. Christ and H. L. Hartnagel, "Three-dimensional finite-difference method for the analysis of microwave-device embedding," *IEEE Trans. Microwave Theory Tech.,* vol. 35, no. 8, pp. 688–696, Aug. 1987.

[5] K. Beilenhoff, W. Heinrich, and H. L. Hartnagel, "Improved finite-difference formulation in frequency domain for three-dimensional scattering problems," *IEEE Trans. Microwave Theory Tech.,* vol. 40, no. 3, pp. 540–546, Mar. 1992.

[6] W. Heinrich, "Quasi-TEM description of MMIC coplanar lines including conductor-loss effects," *IEEE Trans. Microwave Theory Tech.,* vol. 41, pp. 45–52, Jan. 1993.

[7] K. Beilenhoff, H. Klingbeil, W. Heinrich, and H. L. Hartnagel, "Open and short circuits in coplanar MMIC's," *IEEE Trans. Microwave Theory Tech.,* vol. 41, no. 9, pp. 1534–1537, Sept. 1993.

[8] H. Klingbeil and W. Heinrich, "Calculation of CPW A.C. resistance and inductance using a quasistatic mode-matching approach," *IEEE Trans. Microwave Theory Tech.,* vol. 42, no. 6, pp. 1004–1007, June 1994.

A New Extraction Method for Noise Sources and Correlation Coefficient in MESFET

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***Abstract*—A new extraction method for noise sources and correlation coefficient in the noise equivalent circuit of GaAs metal semiconductor field effect transistor (MESFET) is proposed. It is based on the linear regression, which allows us to extract physically meaningful parameters from the measurement in a systematic and straightforward way. The confidence level of the measured data can also be easily examined from the linearity, *y*-intercept of the linear regression, and the scattering from the regression line. Furthermore, it is found that the time constant of correlation coefficient whose value is almost the same as that of the transconductance should be considered to model noise parameters accurately. The calculated values of minimum noise figure, optimum impedance, and noise resistance using above approach, show excellent agreement with measurement for a typical MESFET device studied in this paper.**

I. Introduction

The noise characteristics of a linear two-port system can be completely described by 2-port parameters such as Y, Z, S-parameters and the additional four noise parameters. There are several equivalent sets of these four noise parameters, depending on how to represent the two-port circuit [1], [2]. In addition to the high frequency equivalent circuit parameters, the most important noise parameters from the viewpoint of circuit designers are the minimum noise figure, the optimum impedance (real and imaginary parts), and the noise resistance. They are essential in low noise circuit design. However, it is difficult to incorporate them into the small signal equivalent circuit and the physical relationship between them is not obvious. Therefore, purely empirical representations such as Y- or H-representations are popularly used in the noise equivalent circuit.

Recently, it is reported by Pospieszalski [3] that in the case of H-representation of metal semiconductor field effect transistor (MESFET), the correlation between the noise voltage source at the gate and the noise current source at the drain can safely be ignored in expressing the measured noise characteristics of MESFET with a decent accuracy. Granted it is true in the intrinsic device, the parasitic gate-source capacitance, C_{gsp}, invokes the correlation between gate noise voltage and drain noise current sources [4]. Since it is very difficult to discriminate C_{gsp} from the intrinsic capacitance C_{gsi} by S-parameter measurement, and according to Anholt's estimation of C_{gsp} [5], C_{gsp} is comparable to C_{gsi} especially in the case of low current where the better noise characteristics is obtained, the correlation coefficient has to be considered in the noise equivalent circuit.

We found that the increase of the imaginary part of correlation coefficient with frequency affects the noise characteristics considerably at high frequencies as will be shown in Section IV. The time constant of correlation coefficient is introduced to explain this. But this introduction does not add the number of parameters because the time delay of transconductance is found to be almost the same as time constant of the correlation coefficient. The physical reason for this justification will be explained in Section II.

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