A Full-Wave Approach to the Modeling of Discontinuities of Real Conductors in Planar Lossy Lines for MMIC Applications

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ABSTRACT

We present a full-wave approach to the analysis of discontinuities of real conductors in planar lines, where conductor losses, as well as their finite thickness, are rigorously taken into account. The computational load is quite independent of the number of the dielectric layers composing the substrate, making the model particularly suitable to the analysis of discontinuities in MMIC circuits.

INTRODUCTION

Conductor losses of the interconnecting lines are a significant design parameter in MMIC's, due to the miniaturized dimensions: metallization thickness ranges in the order of the skin-depth and the usual perturbation approach becomes questionable. Moreover MMIC devices are usually grown in highly stratified substrates, excluding the possibility of quasi-TEM approaches, that are limited to low frequencies anyway, while the dielectric stratification also works against the use of classical mode matching approaches. The answer to the need for modeling in such a context is usually provided by means of highly numerical methods, either in the frequency or in the time-domain; however the computational effort required may be large when the analysis involves objects of very different cross-sections, e.g. when conductors are extremely thin with respect to the substrate thickness.

In this context, we propose a Green function-based three-dimensional approach, that is able to take into account finite losses and thickness. As a validation of the proposed approach, simulated data are compared with measurements for a directly coupled resonator in thick, lossy microstrip.

ANALYSIS

The method works as a 3D generalization of the approach that we have proposed in [1] for uniform planar waveguides: the strategy is to build the 3D-dyadic Green's function - $\tilde{\mathbf{Z}}(\mathbf{r},\mathbf{r}')$ for the substrate dielectric layers, *without* the conducting strips. The details of this process are given in [1]: all the information concerning the dielectric stack are build into a scalar transverse impedance, allowing to easily account for an arbitrary dielectric stratification. In order to consider the shielding effects, the stack is assumed to be enclosed in a conducting box: this choice also reduces the computational effort.

An integral equation is then recovered by imposing Ohm's law to hold in the conducting regions, that is

$$\iiint_V d\mathbf{r}' \left[\tilde{\mathbf{Z}}(\mathbf{r},\mathbf{r}') + \rho(\mathbf{r})\tilde{\mathbf{I}}(\mathbf{r},\mathbf{r}')\right]\mathbf{J}(\mathbf{r}') = \mathbf{0} \qquad (1)$$

where $\rho(\mathbf{r})$ is the conductor resistivity; the integration domain V is restricted to regions where sources are non-vanishing, that is, in conducting regions. At this point the procedure introduced by Jackson and Pozar [2],[3] in 2D for lossless lines, is extended to 3D by replacing surface currents with volume currents; condition (1) replaces the classical constraint requiring the vanishing of the tangential electric fields at the conductor surfaces. The integration domain is divided in three subdomains: the domain V_1, at the left-hand side of the discontinuity, where an electric current of unit amplitude is assumed to travel from (minus) infinity up to the discontinuity:

$$\mathbf{J}^{(1)}(\mathbf{r}) = \left[\mathbf{J}_t^{(1)}(x,y) + J_z^{(1)}(x,y)\hat{\mathbf{z}}\right] e^{-j\beta_1(z-z_{01})} \quad if \quad \mathbf{r}\in V_1 \qquad (2)$$

In definition (2) $\mathbf{J}_t$ and J_z are assumed to be the current distributions of the fundamental mode obtained by means of a previous 2D dispersion analysis, as applied to the left-side feed line. This preliminary analysis also provides the propagation constant β_1. Two points have to be remarked:

- just the real part of β_1 is used at this step, that is, the feeding line of the device under study is considered lossless
- practically, the feeding lines are truncated after some wavelengths, constraining the minimum longitudinal dimensions of the box.

The first consideration turns out to be reasonable, as the feeding and loading lines of a good measurement tool have usually to be as lossless as possible. The second consideration relies on the fact that integral equation (1) is discretized and solved just near the discontinuity region. Note that the discontinuity region may include more successive discontinuities, linked by means of an arbitrary number of accessible modes.

It is also assumed that in region V_1 there is a reflected wave of unknown amplitude Γ traveling backward to minus infinity

$$\mathbf{J}^{(R)}(\mathbf{r}) = \Gamma\left[\mathbf{J}_t^{(1)}(x,y) - J_z^{(1)}(x,y)\hat{\mathbf{z}}\right] e^{j\beta_1(z-z_{01})} \quad if \quad \mathbf{r}\in V_1 \qquad (3)$$

z_{01} in (2) and (3) defines a reference plane for Γ.

A second region V_2 is defined at the right-side of the discontinuity, being a loading line of infinite length where a current of unknown amplitude T flows:

$$\mathbf{J}^{(T)}(\mathbf{r}) = T\left[\mathbf{J}_t^{(2)}(x,y) + J_z^{(2)}(x,y)\hat{\mathbf{z}}\right] e^{-j\beta_2(z-z_{02})} \quad if \quad \mathbf{r}\in V_2 \qquad (4)$$

where z_{02} defines the reference plane at which transmission in the fundamental mode is given by T. The discontinuity region is denoted as V_D: in this region the components of the current are expanded by means of suitable scalar expanding $f_i(\mathbf{r})$ with unknown amplitudes X_i , that is

$$\mathbf{J}^{(D)}(\mathbf{r}) = \sum_{i=1}^{N}[X_i^{(X)} f_i^{(X)}(\mathbf{r})\hat{\mathbf{x}} + X_i^{(y)} f_i^{(y)}(\mathbf{r})\hat{\mathbf{y}} + X_i^{(z)} f_i^{(z)}(\mathbf{r})\hat{\mathbf{z}}] \quad if \quad \mathbf{r}\in V_D \qquad (5)$$

Usually the domain V_D partially overlaps V_1 and V_2. The expressions (2-5) are then substituted in (1) and the resulting integral equation in the unknown (Γ,T,$\mathbf{X}$) is solved by means of Galerkin's method.

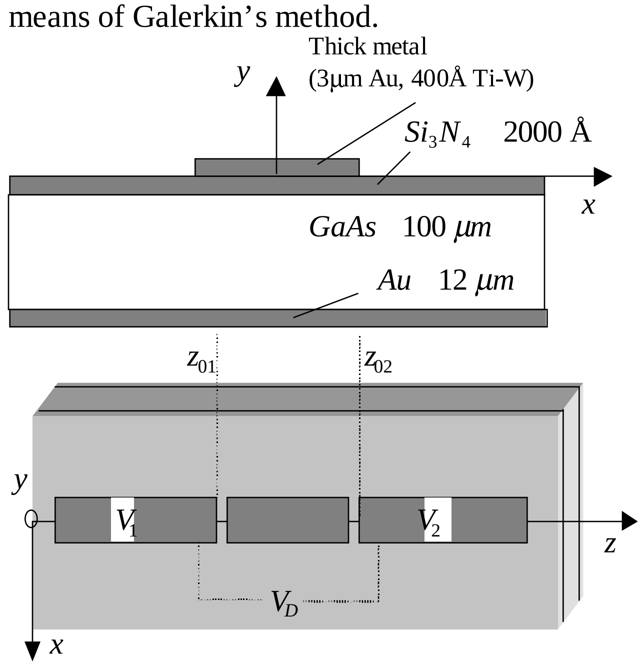

Figure1: Microstrip cross-section and resonator top view; the microstrip is 70 µm wide; the resonator is 7mm long, with 5 µm spacing gaps.

RESULTS

In order to validate it, the proposed approach was applied to the directly coupled lossy resonator reported in [4] and depicted in figure 1. The domains V_1 and V_2 are the feeding and loading lines respectively, whereas the discontinuity domain V_D is assumed to be the resonator itself and part of the feeding and loading lines, extending half a wavelength from their ends. As previously said, the feeding and loading lines are truncated after several wavelengths -usually five to eight, but the results are unaffected-. The functions $f_i(\mathbf{r})$ are chosen to be piecewise constant along the x and y directions, and piecewise sinusoidal (PWS) along the z-direction for all the current components. The Galerkin's weight functions are once more the $f_i(\mathbf{r})$ and the weighting procedure is just applied in the discontinuity domain. The metallic box is made large, the x-dimension being 1 mm while the z-dimension depends on the frequency for which the computation has to be performed, as the reported measurements [4] refer to open resonators. A comparison between measured and simulated data for the attenuation constant and the transmission coefficient is reported in figure 2 and 3 respectively, showing an encouraging agreement in spite of the assumed enclosure, also considering the margin of error in this kind of measurements.

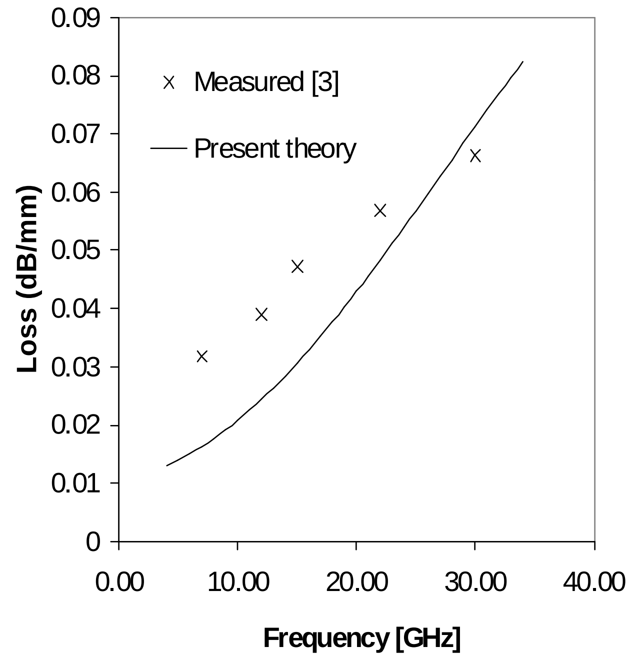

Figure 2: Comparison between nominal measured data and computed data for microstrip attenuation constant.

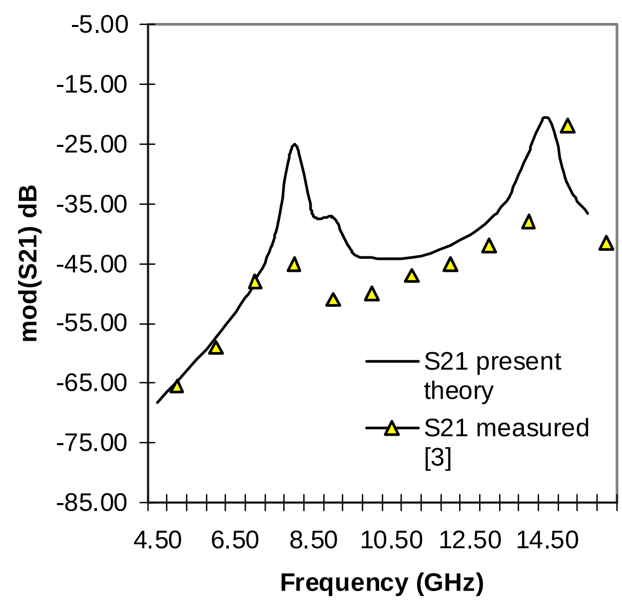

Figure 3: Comparison between measured and computed data for the directly coupled lossy resonator depicted in fig. 1.

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