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A WH/GSMT BASED FULL-WAVE ANALYSIS OF THE POWER LEAKAGE FROM CONDUCTOR-BACKED COPLANAR WAVEGUIDES†

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ABSTRACT

A full-wave analysis using the Wiener-Hopf procedure and the generalized scattering matrix technique (WH/GSMT) is applied to study the EM field propagation of conductor-backed coplanar waveguides. The power leakage from the dominant mode and the effect of finite extent lateral ground planes will be emphasized.

INTRODUCTION

The power leakage of propagating modes from some planar microwave waveguides has recently attracted more attention in the microwave community [1-3]. The usually undesired leakage can influence the performance of high-speed, high-density microwave circuits, by producing cross talk between neighboring lines. Some of these planar transmission lines, like microstrip lines, have a dominant mode whose propagation constant is real which means the field is guided and bounded around the line and no power leakage in the transverse direction occurs for this mode. The leakage phenomenon for these type of planar transmission lines only exists for the higher order modes below or near cut-off frequencies. However, there are other planar transmission lines, such as slotlines and coplanar waveguides, which always have power leakage even for the dominant mode. The propagation constant thus contains both real and imaginary parts, implying propagation and leakage, respectively. Any analytical method neglecting the imaginary part of the propagation constant for these waveguides is incomplete and may be inaccurate in certain situations. However, the complex propagation constant sometimes make the analysis more difficult. For example, in the spectral domain approach the branch point is no longer on the real axis of the integration plane and the integration path has to be detoured [3]. Also, the search of the propagation constant becomes two-dimensional instead of one dimensional.

Here, a full-wave analysis scheme, referred to as the Wiener-Hopf/Generalized scattering matrix technique (WH/GSMT) is used to study the propagation of planar transmission lines embedded in multi-layer dielectric slabs. A similar idea has been employed in [5] where a simplified model is used. Although this method can be applied to many different types of planar transmission lines, the leakage and the effect of finite extent lateral ground planes of a conductor-backed coplanar waveguide on a single dielectric layer will be emphasized here. One reason for choosing this geometry is that despite the importance and special features of coplanar waveguides in microwave integrated circuits, with a few exceptions, most of the results published in the literature (to the knowledge of the authors) are based on certain assumptions, like quasi-TEM approximation or neglecting the leakage. Therefore, numerical results from a rigorous full-wave analysis may provide some new information. Another reason is that the WH/GSMT developed here is especially suitable for studying planar transmission lines with conductors of finite lateral extent.

ANALYSIS

The method used here consists of two steps [5]. First, a canonical problem where the scattering of obliquely incident plane waves by a perfect electric conducting (PEC) half-plane embedded in multilayer dielectric slabs is formulated and solved by means of the Wiener-Hopf technique [4]. Note that in the geometry of the canonical problem a PEC top cover and a PEC backing ground plane are placed in order to simplify the factorization process in the Wiener-Hopf analysis. The details of solving the canonical problems will not be addressed here but it is worth noting that the edge condition is incorporated in this canonical solution. One of the important properties of the Wiener-Hopf analysis relevant to the present problem is that the edge condition has to be enforced to obtain a unique solution. Once the Wiener-Hopf procedure is completed, a transverse resonance relation is then constructed via the generalized scattering matrix technique in a self-consistent manner, taking into account all the interactions between the edges. Compared with other commonly used methods, such as the spectral domain and mode matching techniques, the WH/GSMT is more analytically oriented. That is, the analysis required in the first step makes the numerical computation in the second step very efficient. Besides, this method gives a good physical picture of the plane waves which bounce back and forth and build the resonance.

†This work was supported in part by the Joint Service Electronics Program (contract N00014-89-J-1007) and by the Ohio State University Research Foundation.

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RESULTS AND DISCUSSIONS

For all the curves presented here, β denotes the phase constant and α the attenuation constant. Note that sometimes the logarithm or a scaling factor is used to illustrate the curve better.

First, to verify the WH/GSMT presented here, the normalized phase and attenuation constants vs. slot width for a conductor-backed slotline is depicted in Figure 1 along with the results from [1] and [3] where mode matching and the spectral domain approach are used, respectively. As one can see, the results agree quite well.

In Figure 3, results are shown for β/k_o and α/k_o as a function of frequency for the conductor backed coplanar waveguide illustrated in Figure 2 from 1.0 GHz to 50.0 GHz. Several curves are shown where the coplanar waveguide has different center conductor and gate widths, but the ratio between W and G is kept the same. Since the TEM wave in the parallel plate waveguide region does not have a cut-off frequency, there will always be some power leakage in the lateral direction for all frequencies. In Figure 3, curves (1) and (2) which correspond to smaller values of W and G show very slight dispersion for the phase constant through all the frequencies under consideration. The attenuation constants of curves (1) and (2) are almost linear functions of frequency in the range considered here. However, curve (3), which corresponds to a coplanar waveguide with larger values of W and G, behaves quite differently. It is dispersive for the phase constant and the attenuation constant of curve (3) increases linearly at low frequencies but begins to decrease after it reaches a peak around 8 Ghz. This behavior can be explained because for the coplanar waveguide corresponding to curve (3), the center conductor and the two slots become electrically large at higher frequencies such that the fields concentrate below the center conductor and get weaker in the slot region. Thus weaker fields are excited in the parallel plate waveguides on both sides of the transmission line away from the two slots. This also explains why the phase constant of curve (3) approaches the limit of $\sqrt{\epsilon_r} = 3.16$ as the frequency goes higher. In other words, the conductor-backed coplanar waveguide for curve (3) behaves more like a microstrip line rather than a coplanar waveguide.

In [2], an expression for the attenuation constant α is derived from the reciprocity theorem. Here, a calculation of the propagation constant of a conductor-backed coplanar waveguide as a function of ground-to-ground spacing (W+2G in Figure 2) is performed and the result is given in Figure 4 where the substrate thickness and the ratio between center conductor width W and slot width G are fixed. When W+2G is small, the attenuation constant has a linear to square frequency dependence which generally agrees with the prediction in [2]. But as W+2G gets larger, the coplanar waveguide tends to behave more like a microstrip line. Note that in the cases considered in Figures 3 and 4, the normalized attenuation constant (α/k_o) tends to have a linear frequency dependence when the normalized phase constant (β/k_o) is nondispersive, whereas, it has a more complicated frequency dependence when the phase constant becomes more dispersive.

Finally, the effect of the finite lateral ground planes of a conductor-backed coplanar waveguide is studied. Although quantitative results are not included in this manuscript because of limited space, they will be presented at the conference. The coplanar waveguide with finite lateral ground planes can have purely bounded modes without lateral power leakage as long as the mode propagates at much slower speed than the surface wave. Compared with a coplanar waveguide that has infinite lateral ground planes, there are many more modes for a coplanar waveguide with finite lateral ground planes.

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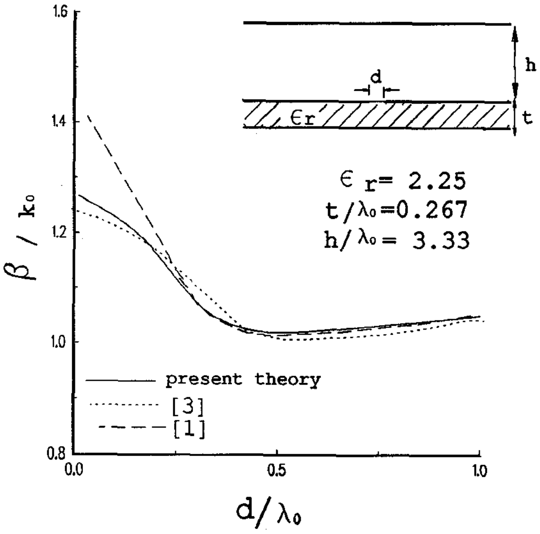

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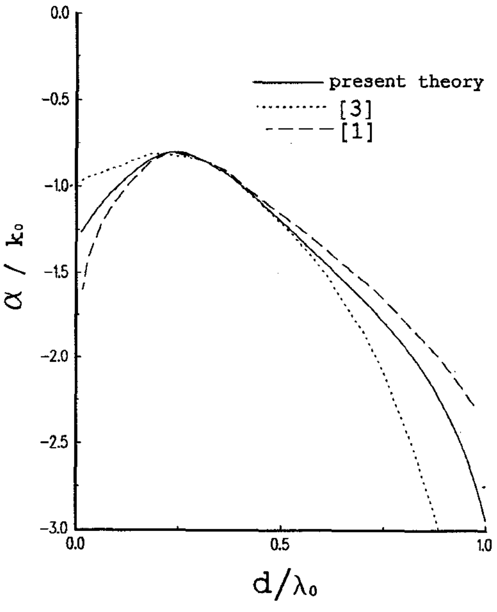


(b)

Figure 1: Normalized phase constant β/k_0 (a) and normalized attenuation constant α/k_0 (b) of a conductor-backed slotline vs. slot width compared with the results of [1,3]. k_0 is the free space wavenumber.

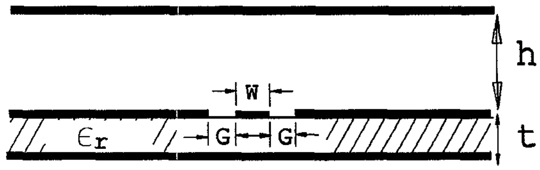
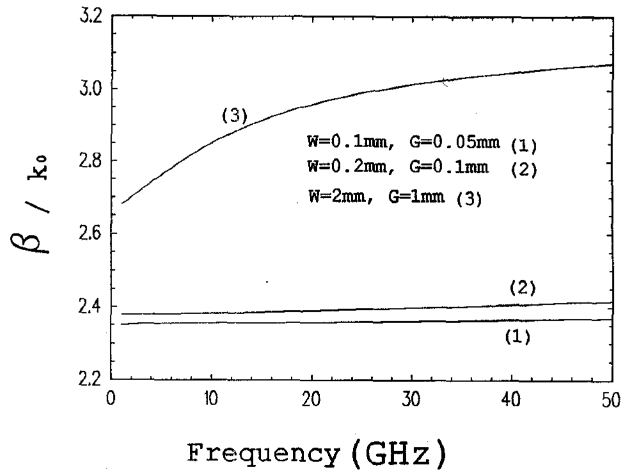


Figure 2: A conductor-backed coplanar waveguide with infinite lateral ground planes.



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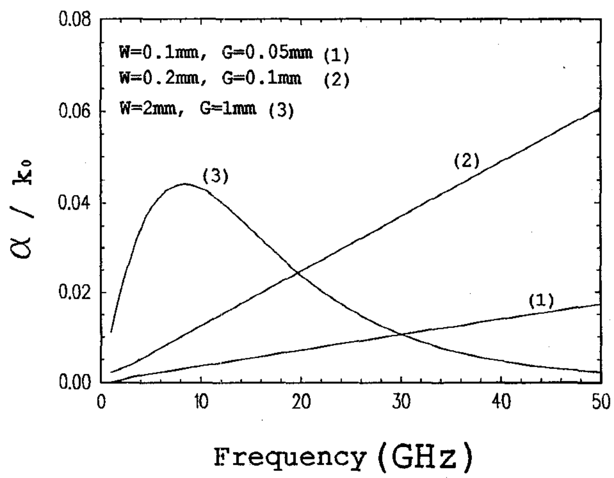
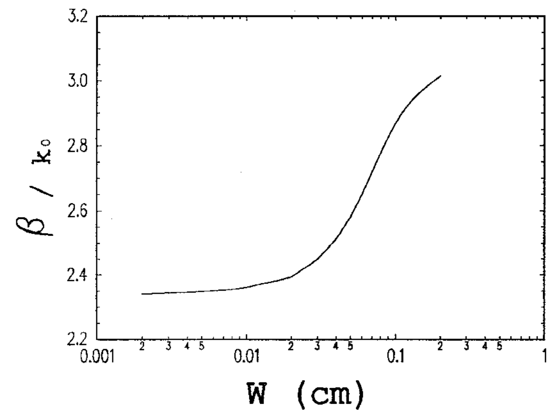


(b)

Figure 3: Normalized phase constant β/k_0 (a) and normalized attenuation constant α/k_0 (b) of a conductor-backed coplanar waveguide with infinite lateral ground planes vs. frequency. Center conductor width and slot width are (1) W=0.1mm, G=0.05mm; (2) W=0.2mm, G=0.1mm; (3) W=2mm, G=1mm. Substrate dielectric constant ϵ_r=10.0, substrate thickness t=0.04cm, top cover height h=0.11cm, k_0 is the free space wavenumber.



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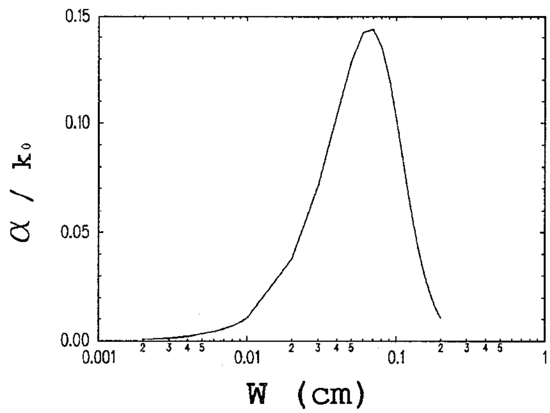

(b)

Figure 4: Normalized phase constant β/k_0 (a) and normalized attenuation constant α/k_0 (b) of a conductor-backed coplanar waveguide with infinite lateral ground planes vs. center conductor and slot width. Substrate dielectric constant ϵ_r=10.0, substrate thickness t=0.04cm, top cover height h=0.11cm, frequency=30.0GHz, W/G=2.0, k_0 is the free space wave number.