References

1 GRAY, D. A.: 'Effect of time-delay errors on the beam pattern of a linear array', *IEEE J. Oceanic Eng.*, 1985, **OE-10**, pp. 269–277
2 BODECKER, G., RATHJEN, D., and SIEGEL, M.: 'Errors of time delay beamforming with interpolated signals', *Proc. Institute of Acoustics, Array Signal Processing*, 1985, **7**, (4), pp. 161–166
3 PRIDHAM, R. G., and MUCCI, R. A.: 'A novel approach to digital beamforming', *J. Acoust. Soc. Am.*, 1978, **63**, (2), pp. 425–434

EFFECT OF FINITE SUBSTRATE WIDTH ON HIGHER-ORDER MODE GENERATION OF ELECTRICALLY SHIELDED SYMMETRIC MICROSTRIP

J.-D. Tseng and C.-K. C. Tzuang

Indexing terms: Microstrip, Microwave integrated circuits, Microwave devices and components

The effect of finite substrate width on the higher-order mode generation of an electrically shielded microstrip is investigated. By employing the full-wave mixed potential mode-matching method the microstrip on a finite-width substrate can be accurately analysed. The dispersion characteristics of microstrip on two different substrate widths were compared in the frequency range of interest. The results indicated that the use of conventional three-dimensional simulators assuming homogeneous layered (stratified) multi-dielectric substrates may not work well above the first cutoff frequency of microstrip on a finite-width substrate.

Introduction: Rapid progress in microwave monolithic integrated circuits (MMICs) and electro-optic devices has made the microstrip on a finite-width substrate an important structure requiring rigorous analyses [1, 2]. Recently a few full-wave field-theoretic methods capable of analysing the so called proximity effect of a microstrip near the edges of substrate have been reported [3–5]. In these papers the important circuit design parameters such as the effective dielectric constant and the characteristic impedance of the dominant mode were reported. The propagation characteristics of the higher-order modes, however, are neither reported nor fully discussed. On the other hand, it is already known that the existence of higher-order modes alters the microwave circuit properties significantly using quasiplanar transmission lines on layered dielectric substrates [6, 7]. Thus this Letter investigates and compares the propagation characteristics of the higher-order modes of a microstrip on both layered and nonlayered substrates. Through the comparison of the two mode spectra, the effect of finite substrate width on the microstrip is exposed. The data obtained here enable us to estimate the upper frequency limit of the three-dimensional (M)MIC simulators [8, 9] assuming layered dielectric substrates.

Formulation: The cross-section of a shielded symmetric finite-width substrate microstrip is shown in Fig. 1. A PEC (perfect electric conductor) or a PMC (perfect magnetic conductor) can be placed at $x = 0$ to reduce the computing time for odd mode and even mode propagation, respectively. Assuming that the metal strip has infinite conductivity, and by applying the full-wave mixed potential mode-matching method [10], Fig. 1 can be divided into three regions, I, II, and III. The TE-to-x and TM-to-x Hertzian potentials are employed in region I and II, whereas the TE-to-y and TM-to-y Hertzian potentials are employed in region III. After matching the remaining boundary conditions at interfaces $y = h$ and $y = h + t$, respectively, a nonstandard eigenvalue equation can be derived for determining the complex propagation constants for the fundamental mode and higher-order modes including the complex modes.

Dispersion characteristics: Assuming that the propagation factor is $\exp(j\omega t - \gamma z)$ and $\gamma = \alpha + j\beta$, in Fig. 2 two sets of plots are superimposed for the dispersion characteristics of a symmetric microstrip on the substrate with different widths, $b = a$ (layered) and $b = 0{\cdot}5a$ (nonlayered), against frequencies ranging from 15 to 40 GHz. All the structural and material parameters for the microstrip are shown in Fig. 2. For $b = a$ the solid lines represent the propagating and evanescent modes and the dotted lines represent the complex modes. For $b = 0{\cdot}5a$, the dash–dotted lines represent the propagating and evanescent modes. For $b = a$, the substrate edges touch the sidewalls and a layered substrate is established. The dominant modes for microstrip on the two different substrate widths are almost the same and cannot be distinguished from the plots.

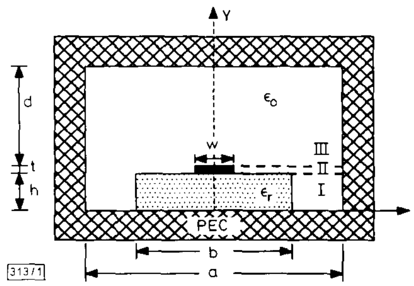

Fig. 1 *Cross-sectional view of symmetric shielded microstrip line on finite-width substrate*

In contrast the mode spectra of the higher-order modes change drastically. First, the total number of the propagating modes at $f = 40$ GHz reduces from 8 to 6. The number of the evanescent modes also decreases at $f = 40$ GHz for $b = 0{\cdot}5a$. Secondly, all the complex modes which occur when $b = a$ disappear when $b = 0{\cdot}5a$ in the mode spectrum region of interest. The distribution of the complex modes in the mode spectrum for the layered case ($b = a$) is quite complicated. The reduction in substrate width apparently has strong effect on the complex modes. Third, the cutoff frequencies of all the higher-order modes shift toward higher frequencies as the value of the substrate width decreases from $b = a$ to $b = 0{\cdot}5a$. The higher the order of mode, the greater the deviation in the cutoff frequency. The last important observation is that, in the lower frequency range, or approximately below the first cutoff frequencies, 22·1 GHz for $b = a$ and 23·1 GHz for $b = 0{\cdot}5a$, the normalised propagation constants of the first six evanescent modes are very close to one other. The seventh higher-order (evanescent) mode of the nonlayered ($b = 0{\cdot}5a$) case is below the ninth higher-order mode of the layered ($b = a$) case.

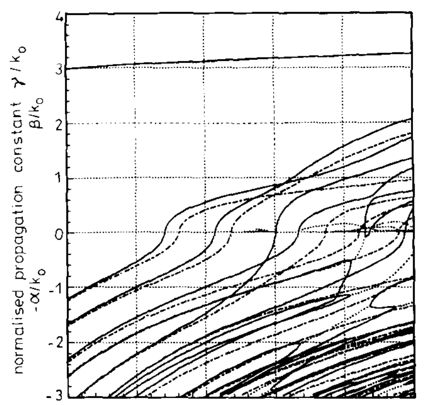

Fig. 2 *Superimposed dispersion characteristics including propagating modes, evanescent modes, and complex modes, of two different substrate widths, $b = a$ and $b = 0{\cdot}5a$*

Structural and material parameters are: $\varepsilon_r = 12{\cdot}9$, $h = 0{\cdot}635$ mm, $t = 0{\cdot}03556$ mm, $d = 4{\cdot}32944$ mm, $w = 0{\cdot}6$ mm, $a = 6{\cdot}0$ mm
——— propagating and evanescent modes for $b = a$
—·—· propagating and evanescent modes for $b = 0{\cdot}5a$
······ complex modes for $b = a$

Conclusions: The effect of finite substrate width on the higher-order mode generation of an electrically shielded microstrip has been presented. By varying the value of the substrate width b, from a to $0{\cdot}5a$, it is found that, for the particular case of a symmetric microstrip in the frequency range of interest, the dominant modes of propagation are indistinguishable from one other whereas the mode spectra of the higher-order modes change drastically above the first cutoff frequencies. These drastic changes in higher-order mode spectra suggest that the conventional three-dimensional simulators, assuming layered substrate ($b = a$) will result in very questionable results beyond the first cutoff frequency in our particular case study. On the other hand, when the operating frequency is below the first cutoff frequency, the variation of the substrate width has little influence on the dominant and the first six higher-order (evanescent) modes. This suggests that the conventional three-dimensional simulators may work well for the particular case study. Furthermore, this Letter reports for the first time that the complex modes are subject to the strong influence by varying the substrate width of the microstrip while the substrate dielectric constant is kept the same.

Acknowledgment: This work was supported by the National Science Council, Republic of China, under Grant NSC 81-0404-0-

14th November 1991

J.-D. Tseng and C.-K. C. Tzuang (*Institute of Communication Engineering, Center for Telecommunication Research, National Chiao Tung University, No. 1001 Ta Hsueh Road, Hsinchu, Taiwan, Republic of China*)

References

1 PUCEL, R. A.: 'MMICs, modelling, and CAD—where do we go from here?'. Proc. 16th European Microwave Conf. (Dublin), 1986, pp. 61–70

2 SUETA, T., and IZUTSU, M.: 'Integrated optic devices for microwave applications', *IEEE Trans.*, 1990, **MTT-38**, (5), pp. 477–482

3 THORBURN, M., AGOSTON, A., and TRIPATHI, V. K.: 'Computation of frequency-dependent propagation characteristics of microstriplike propagation structures with discontinuous layers', *IEEE Trans.*, 1990, **MTT-38**, (2), pp. 148–153

4 TSENG, J.-D., and TZUANG, C.-K. C.: 'Full-wave analysis of proximity effects on microstrip lines near the edge of a packaged MMIC'. Proc. 20th European Microwave Conf. (Hungary), 1990, pp. 1322–1327

5 ENGEL, A. G., JUN., and KATEHI, L. P. B.: 'Analysis of microstrip structures on and near dielectric ridges using an integral equation-mode matching technique'. 1991 IEEE MTT-S Int. Symp. Dig., Session D-4, pp. 135–138

6 WU, S.-C., YANG, H.-Y., and ALEXOPOULOS, N. G.: 'A rigorous dispersive characterization of microstrip cross and tee junctions'. 1990 IEEE MTT-S Int. Symp. Dig., Session II-3, pp. 1151–1154

7 OMAR, A. S., and SCHÜNEMANN, K. F.: 'The effect of complex modes at finline discontinuities', *IEEE Trans.*, 1986, **MTT-34**, (12), pp. 1508–1514

8 RAUTIO, J. C., and HARRINGTON, R. F.: 'An electromagnetic time-harmonic analysis of shielded microstrip circuits', *IEEE Trans.*, 1987, **MTT-35**, (8), pp. 726–730

9 HAROKOPUS, W. P., JUN., and KATEHI, P. B.: 'Characterization of microstrip discontinuities on multilayer dielectric substrates including radiation losses', *IEEE Trans.*, 1989, **MTT-37**, (12), pp. 2058–2066

10 TZUANG, C.-K. C., and TSENG, J.-D.: 'A full-wave mixed potential mode-matching method for the analysis of planar or quasi-planar transmission lines', *IEEE Trans.*, 1991, **MTT-39**, (10), pp. 1701–1711

SLOT-COUPLED EXCITATION OF MICROSTRIP DIPOLE ANTENNAS

A. Ittipiboon, D. Roscoe, M. Cuhaci and Y. M. M. Antar

Indexing terms: Microstrip, Dipole antennas

The slot-coupled method is proposed as an alternative to the popular electromagnetically coupled technique (EMC) for feeding microstrip dipole antennas. This Letter describes the basic design procedure and some key advantages of this method over EMC. Experimental results confirm the effective and practical performance of the proposed structure.

Introduction: Microstrip dipoles are one attractive type of radiating element with a potential application in future active phased array antennas. Their desirable features include simplicity, small size and linear polarisation. The electromagnetically coupled (EMC) technique, first proposed by Oltman [1], is a feed mechanism of microstrip dipoles which is compatible with the integration of active devices in a monolithic microwave integrated circuit (MMIC) environment. The popularity of the EMC dipoles has increased and they have become the subject of many investigations in recent years [2].

The EMC dipole is usually a multilayer structure with the microstrip feedline printed on the bottom substrate and the dipole printed on the upper. The dipole excitation is achieved through electromagnetic coupling to the feedline with no direct contact. As a result of the feedline and the dipole being located on the same side of the ground plane, the feedline radiation contributes significantly to ripples in the radiation patterns. This parasitic radiation increases with the mismatch at the dipole [3]. This Letter proposes the slot-coupled technique as an alternative dipole excitation method which eliminates the problem of feedline radiation by using the ground plane to isolate the feedline from the antenna. The coupling from the feedline to the antenna is then accomplished through a slot in the common ground plane. This technique has been used to feed microstrip patch antennas [4–6]. In this Letter, it is demonstrated that the slot-coupled technique can be successfully adapted for use with dipoles.

Geometry and configuration: The physical configuration of the structure is shown in Fig. 1. The microstrip dipole simply consists of a thin-strip conductor printed on a substrate of low permittivity to maintain high radiation efficiency. Increasing the thickness of the substrate decreases efficiency, but increases the operating bandwidth. In practice, these parameters can be selected to suit the design requirements. In this study, however, printed dipoles were fabricated on a substrate of $\varepsilon_r = 2{\cdot}5$ with a thickness of 0·794 mm. The length of all dipoles is 26 mm, and the widths vary from 1 to 3·6 mm. The printed dipole is isolated from the feedline, which is printed on the substrate on the other side of the common ground plane. The feedline is Duroid 1060 with $\varepsilon_r = 10{\cdot}2$ and a thickness of 0·635 mm. The coupling from the feedline to the printed dipole is accomplished through a narrow slot which is cut in the common ground plane perpendicular to the feedline. For each dipole length and width, the slot length can be optimised to obtain good coupling. The open circuit tuning stub L_s in the feedline can be used to remove the reactive part of the input impedance of the dipole to improve the match.

Results: Measurements were performed for a printed dipole with a fixed width and length using different coupling slot lengths. It was found that as the slot length increased the dipole can be changed from undercoupled to matched and finally to overcoupled. In practice, however, it is desirable to

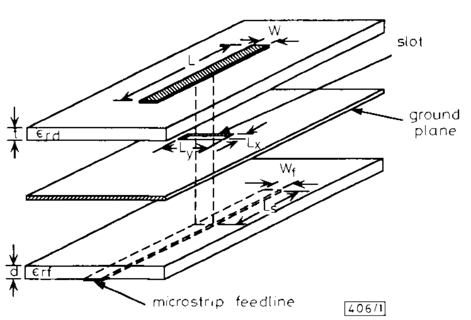

Fig. 1 *Slot-coupled microstrip dipole configuration*