Published in IET Microwaves, Antennas & Propagation
Received on 3rd September 2009
Revised on 30th March 2010
doi: 10.1049/iet-map.2009.0448

ISSN 1751-8725

High-frequency crosstalk between two parallel slotlines

V. Kotlan J. Machac

Department of Electromagnetic Field, Faculty of Electrical Engineering, Czech Technical University in Prague, Technicka 2, 166 27 Praha, Czech Republic
E-mail: vaclav.kotlan@email.cz

Abstract: A study of high-frequency crosstalk between two slotlines located in parallel on the same substrate is presented. The electric field is induced in a passive slotline because of coupling from the other closely placed slotline excited by a source. A full-wave analysis, the method of moments in the spectral domain, taking into account both the bound and the residual and leaky modes, is used to examine high-frequency crosstalk. As a result, the voltage of the total wave propagating along the coupled lines is calculated. The theory is verified by measurement and by a simulation performed in CST Microwave Studio.

1 Introduction

Open planar transmission lines have been widely investigated in the last four decades with particular attention being devoted to microstrip lines [1]. The dispersion relations, field distributions of propagating modes, together with their characteristic impedances of the various types of planar and uniplanar transmission lines, were studied using the eigenmode approach in [2]. Later, a new method of microstrip line analysis was introduced by Mesa *et al.* [3, 4] in which full-wave analysis in the spectral domain is applied. This analysis takes into account the actual amplitude of the waves excited by a defined source. This approach can incorporate not only the bound or leaky waves, but also the residual waves [3, 4], and can therefore provide clear information about the total wave exited by the source along the line. However, slotline has not usually been the main interest of researchers. Finally, full-wave analysis taking into account the source has also been applied to the slotline and has been carefully verified by measurement [5, 6].

Interesting electromagnetic field behaviour occurs when two slotlines are located very close together in parallel. These lines are coupled by crosstalk [7]. Crosstalk is an issue that is important in a wide range of microwave circuits, particularly at high frequencies, where the leakage has a great influence [8]. This crosstalk is usually undesired and microwave engineers endeavour to restrict this effect. However, transmission line theory loses accuracy with growing frequency and so full-wave analysis is needed.

This paper presents the results of an investigation of waves excited on a slotline by a real source, and also waves induced because of high-frequency crosstalk to the passive slotline situated in parallel on the same substrate.

2 Analysis

The voltage of a wave propagating along active and passive coupled lines is calculated using an accurate semi-analytical spectral domain method. The analysis takes into account all possible waves excited by the source. In order to keep the explanation of the problem as simple as possible, only the single basis function describing the electric field inside the slots is considered here. This single basis function is used in the same form for both lines 1 and 2 as shown in Fig. 1. The width of the two slots is assumed equal. However, because of the absence of structure symmetry according to the source it is necessary to expand the electric field inside the slots by using the series of basis functions (even and odd) in order to obtain proper numerical results.

The feeding current source is placed across the slot of the 'active' slotline, line 1 in Fig. 1. This source excites the wave along this infinitely long line. Owing to crosstalk some part

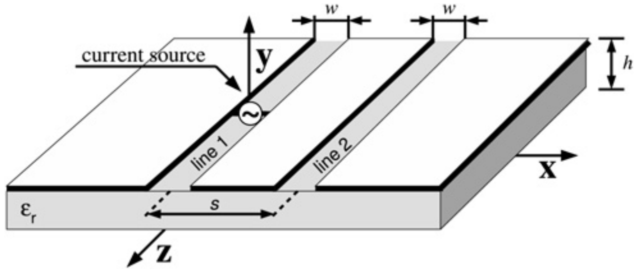

Figure 1 *Pair of slotlines of infinite length*

Line 1 is excited by a current source located at $z = 0$, $y = h$

of the excited power is coupled into the 'passive' line, line 2 in Fig. 1, which is placed at a distance s. The structure is assumed to be lossless, laterally unbounded and the metallisation is assumed to be of zero thickness.

The current density imposed by the source across line 1 can be easily separated into the following transversal and longitudinal parts

$$J_X(x, z) = T(x)L(z) \tag{1}$$

where the transversal part is normalised as $\int_{-w/2}^{w/2} T(x)\,dx = 1$. The longitudinal part is chosen so as to obtain the fast convergence in the spectral domain, and has the following form

$$L(z) = \frac{1e^{-1/2(z/d)^2}}{d\sqrt{2\pi}} \tag{2}$$

where d is the longitudinal effective width of the source in the z direction and is set as 1 mm in all our numerical examples. The Fourier transform of (2) is then written as

$$\tilde{L}(k_z) = e^{-1/2(k_z d)^2} \tag{3}$$

The final expression for the total voltage on both lines 1 and 2 is calculated as an integral of the x-part of the electric field across the slot

$$V_1(z) = \int_{-w/2}^{w/2} E_{1x}(x, z)\,dx \tag{4}$$

$$V_2(z) = \int_{s-w/2}^{s+w/2} E_{2x}(x, z)\,dx \tag{5}$$

The electric field of each line is obtained from the corresponding spectral domain expression by the inverse Fourier transform as follows

$$E_{1x}(x, z) = \frac{E_x^{\text{bas}}(x)}{\pi}\int_{0(C_z)}^{\infty} \tilde{E}_{1x}(k_z)\cos(k_z z)\,dk_z \tag{6}$$

$$E_{2x}(x, z) = \frac{E_x^{\text{bas}}(x)}{\pi}\int_{0(C_z)}^{\infty} \tilde{E}_{2x}(k_z)\cos(k_z z)\,dk_z \tag{7}$$

where $E_x^{\text{bas}}(x)$ is the basis function of the field distribution inside the slot. $\tilde{E}_{1x,2x}(k_z)$ is the spectral domain electric field in lines 1 and 2. The field is defined to satisfy the boundary condition at the metallisation edges by a proper choice of the basis function. The basis functions are specified by the Chebyshev polynomial [9]. The integration path C_z in (6) and (7) must detour around the singularities placed on the positive real axis of the complex k_z plane [4], as sketched in Fig. 2. k_0 is the free space wave number, k_{TM} is the TM_0 surface mode wave number and $k_z^{\text{BM}_1}$, $k_z^{\text{BM}_2}$ are the even- and odd-bound mode wave numbers.

By applying Galerkin's method in the spectral domain, the Fourier transform of the electric field in the slot of both the lines is obtained as

$$\tilde{E}_{1x}(k_z) = \frac{2\pi\int_{-w/2}^{w/2} E_x^{\text{bas}}(x)J(x, k_z)\,dx}{A_{11}(k_z) - A_{12}^2(k_z)/A_{22}(k_z)} \tag{8}$$

$$\tilde{E}_{2x}(k_z) = \frac{2\pi\int_{-w/2}^{w/2} E_x^{\text{bas}}(x)J(x, k_z)\,dx}{A_{12}(k_z) - A_{22}(k_z)A_{11}(k_z)/A_{12}(k_z)} \tag{9}$$

where the numerator represents the feeding source and the denominator contains the spectral domain Green's functions. The denominators of (8) and (9) can be specified as [8]

$$A_{11}(k_z) = \int_{-\infty}^{\infty} \tilde{E}_x^{\text{bas}^2}(k_x)\tilde{G}_{xx}(k_x, k_z)\,dk_x \tag{10}$$

$$A_{22}(k_z) = A_{11}(k_z) \tag{11}$$

$$A_{12}(k_z) = \int_{-\infty}^{\infty} \tilde{E}_x^{\text{bas}^2}(k_x)\tilde{G}_{xx}(k_x, k_z)e^{-jk_x s}\,dk_x \tag{12}$$

where $\tilde{G}_{xx}(k_x, k_z)$ is the spectral domain Green's function evaluated at $y = h$, and $\tilde{E}_x^{\text{bas}}(k_x)$ is the basis function

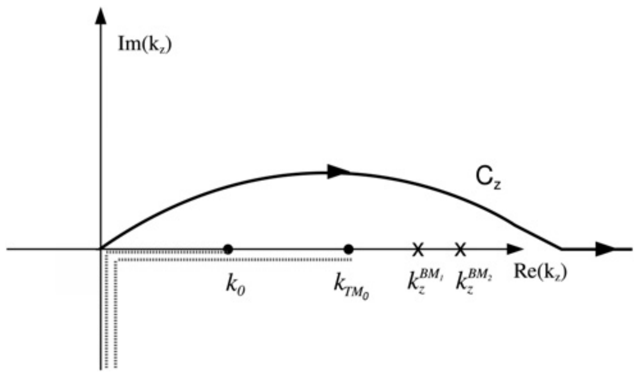

Figure 2 *Path of integration in (6) and (7) detours above the singularities that appear on the positive real axis of the complex k_z plane*

Branch points at k_0 and $k_{\text{TM}0}$ are shown, together with bound mode – even and -odd poles k_z^{BM1}, k_z^{BM2}. Branch cuts are sketched by the dashed lines describing the field distribution inside the slot in the x-direction expressed in the spectral domain. The path of integration in integrals (10)–(12) is the real axis of the complex k_x plane [5].

3 Numerical results

3.1 Slotline on a thin high-permittivity substrate

A computer code has been developed on the basis of the results of our theoretical analysis explained above. The series of basis functions (even and odd) are used for numerical calculations. Three basis functions in the x-direction and two basis functions in the z-direction are used.

The numerical results are shown for the line with substrate permittivity of $\varepsilon_r = 9.9$, thickness of the substrate $h = 1$ mm, line width $w = 0.25$ mm and the distance between centres of the lines $s = 0.75$ mm. First, we studied the voltage of the wave excited by the source on a single slotline. This single slotline has been analysed in detail in [5] and our results are presented there. We have described three important frequency ranges for this configuration of the line. At 10 GHz, only the bound mode propagates, so the voltage is constant. The second leaky wave starts to propagate together with the bound mode at about 30 GHz, so the voltage decreases rapidly close to the source because of strong leakage, but stays constant at further distances because of the propagating bound mode. Finally, above 50 GHz only the first and the second leaky waves propagate along the line [5] and the field is completely attenuated because of leakage.

Adding the second coupled line close to the active line, the voltage distributions along these lines are modified by the crosstalk. Fig. 3 shows the voltages along the two lines at 10 GHz. The nearly constant amplitude of the bound wave voltage predicted on the single slotline is varied because of the coupling of the lines. The interesting point is that the maxima and minima of the voltages are more or less the same for these two lines, the line with the source (line 1) as well as the line without any source (line 2). Owing to the absence of leaky modes at this low frequency, we suppose that the oscillations are due to the bound mode coupling only. These results are in full accordance with the results presented in [8, 9] for the current of coupled microstrip lines.

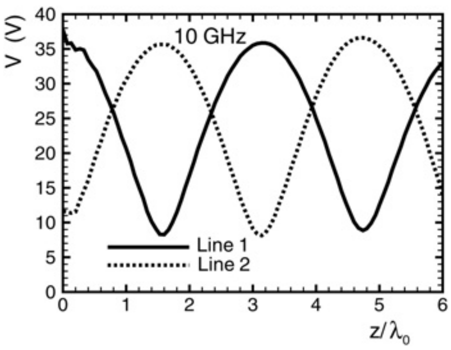

Figure 3 *Voltages along the parallel slotlines with $w = 0.25$ mm, $h = 1$ mm, and $\varepsilon_r = 9.9$*

Distance between the lines is $s = 0.75$ mm. The current source working at 10 GHz is placed at $z = 0$ on line 1

Fig. 4 shows the voltage distributions on both lines calculated at 30 GHz. The energy starts to leak on the slotline in the form of a leaky wave at about this frequency, but further propagates in the form of the bound mode. It should be pointed out that, unlike in the case shown in Fig. 3, the energy propagating along the two lines does not have the same level in this working regime. However, a high percentage of the energy still propagates along the passive line because of the coupling. It should be mentioned that the shape of the voltage distribution along the active line is similar to that for the single slotline excitation, as shown in Fig. 3.

Finally, the voltage of the waves propagating at 50 GHz is plotted in Fig. 5. The voltage along line 1 decreases rapidly, because there is no bound mode and all energy is radiated from the slot because of leakage [5, 6]. However, the meaningful magnitude of the voltage is also induced on the second line in the vicinity close to the source of line 1. Similarly as in the case of line 1, the voltage falls rapidly with distance, because the bound mode cannot be transmitted at this frequency.

As we have pointed out, very interesting wave behaviour has been found at 10 GHz. Only the bound wave propagates and the coupling of the slotlines is the strongest at this frequency of all studied regimes. Fig. 3 shows this coupling for very closely located slotlines. This coupling could be detrimental for the proper functioning of microwave circuits. Hence we have studied this effect in more detail. Fig. 6*a* shows the voltage excited along line 1, considering the different distances from line 2. It is shown that the wave excited along line 1 is very strongly influenced by the closeness of the second line, up to a distance of about 10 mm, which means 10 h. The wave

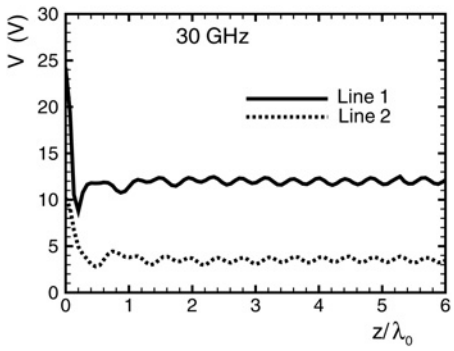

Figure 4 *Voltage distribution along parallel slotlines defined in Fig. 3*

Current source working at 30 GHz is placed at $z = 0$ on line 1

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doi: 10.1049/iet-map.2009.0448

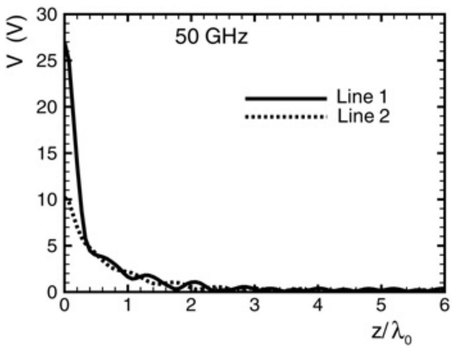

Figure 5 *Voltage along the lines of parallel slotlines defined in Fig. 3*

Current source of 50 GHz is placed at z = 0 on line 1

along line 1 has almost the same characteristics for distances from line 2 higher than 10 mm as the wave on the single slotline plotted in Fig. 3.

Fig. 6*b* presents the voltage distribution along coupled line 2 for various distances from line 1. The plots in Fig. 6*b*, together with the plots in Figs. 6*a* and 3, show that the two voltage distributions along the two coupled lines are complementary. The voltage on line 1 has maxima at points where the voltage on line 2 has minima and vice versa. Naturally, the voltage induced on line 2 decreases with increasing line distance and finally it has a negligible value for distances exceeding 10 mm and the lines are not coupled any more.

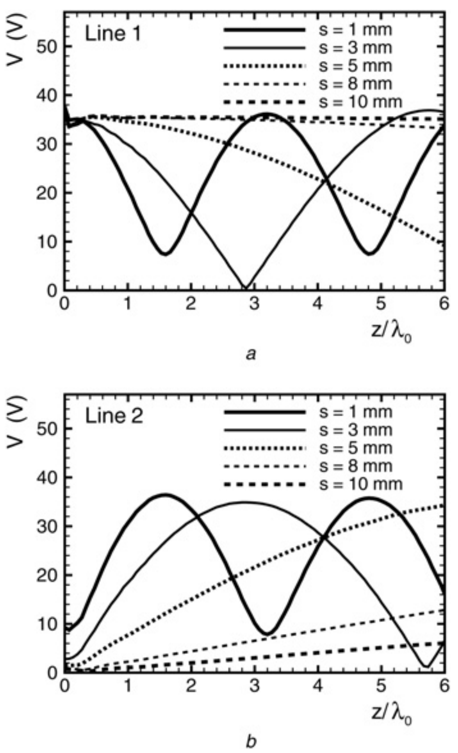

Figure 6 *Voltage defined in Fig. 3 for various distances s*

a Line 1
b Line 2
Frequency is 10 GHz

The two coupled lines in fact represent one fictive line transmitting the total wave excited by the source. The total voltage of this wave can be defined as the sum of the two particular voltages with respect to their phases. This total voltage calculated for all cases of waves represented in Figs. 6*a* and *b* is nearly equal, and is plotted in Fig. 7 for the case of 10 GHz, taken here for $s = 5$ mm. The same procedure was performed for 30 and 50 GHz, as shown in Fig. 7. These voltage distributions are very similar to the distributions plotted for a single slotline as also shown in Fig. 7. This is due to the fact that the same power must be transported along the single slotline as well as along the two coupled slotlines.

In order to validate our results we have employed the Matrix pencil method (MPM) [10]. Using this method we have decomposed the total voltage wave along the slot into a sum of complex exponential functions whose exponents can be correlated with the propagation constants of the relevant waves appearing in the structure. Those exponents have been compared with the propagation constants obtained from the dispersion plot of our structure (a coplanar waveguide, CPW). At 10 GHz we have obtained from the MPM two normalised propagation constants corresponding to even and odd bound modes: $k_{z1}/k_0 = 1.99$, $k_{z2}/k_0 = 2.31$, and independently we have obtained: $k_{z1}/k_0 = 2.00$, $k_{z2}/k_0 = 2.31$ from dispersion characteristics of CPW. At 30 GHz we have calculated one bound mode with $k_z/k_0 = 2.41$ using the MPM. An identical value of bound mode propagation constant $k_z/k_0 = 2.41$ has been obtained from the dispersion characteristics of CPW. The algebraically decreasing residual wave [4, 6] with its influence increasing with raised frequency naturally represents the difference between the total field and the above-mentioned modes. Next, we have obtained the leaky mode propagation constant $k_z/k_0 = 2.55 - \mathrm{j}0.05$ from dispersion characteristics of CPW at 50 GHz. An identical $k_z/k_0 = 2.55 - \mathrm{j}0.05$ has been

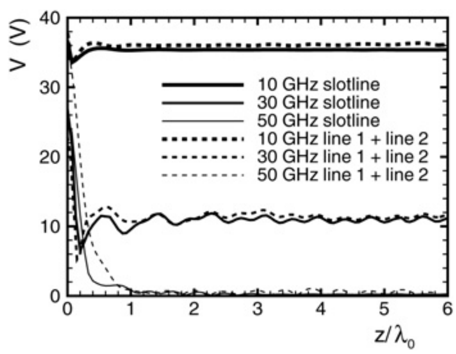

Figure 7 *Voltage along the single slotline defined with $w = 0.25$ mm, $h = 1$ mm, and $\varepsilon_r = 9.9$ plotted together with the sum of both waves along lines 1 and 2 for coupled lines with distance $s = 5$ mm*

calculated using the MPM. Additionally, we have also obtained a significant magnitude for the rest wave that certainly corresponds to an algebraically decreasing residual wave [4, 6]. The very good agreement between the propagation constants obtained by the decomposition of the total wave and also by independent method from dispersion characteristics validates our results for all the frequencies considered here.

3.2 Slotline on thick and low-permittivity substrate

The structure analysed next is defined as follows: substrate permittivity $\varepsilon_r = 2.6$, substrate thickness $h = 14.6$ mm, and line width $w = 5.6$ mm. Results are figured out for two frequencies and for various distances between the lines. The voltage at 3 GHz represents the bound mode only and the voltage at 5 GHz represents the bound mode together with the leaky mode [5, 6]. The measured and calculated curves fit together well, despite the fact that the measured data are degraded by the standing wave, because of improper line termination.

The following figures show the behaviour of the voltage excited along the two coupled lines for various distances *s*. Fig. 8*a* shows the progress of the voltage on line 1 at 3 GHz. The influence of the second line is not as strong as in the case of the line with high permittivity, as seen in Fig. 6*a*. The voltage along the line has more or less the same shape for each distance *s*. The voltage along line 1 decreases, and thus the voltage along line 2 increases, as shown in Fig. 8*b*, for small distances *s*. The lines are nearly not coupled for distances greater than about 80 mm.

Fig. 9*a* and *b* show the voltage distributions along both lines at their particular distances calculated at 5 GHz. The bound mode together with the leaky mode propagates at this frequency [5, 6]. The field distributions plotted in Figs. 8*a* and *b*, 9*a* and *b* have the same characteristics as those for the line on the high-permittivity substrate, but at a different scale of coordinates. This is because of the longer wavelength of the wave guided along the line on the low permittivity substrate.

The voltage along the single slotline and the sum of the two voltages representing the fictive wave propagating along the coupled slotlines are shown in Fig. 10. The plots are comparable. The only difference is in the amplitude of the voltages. This is because the characteristic impedance is different for a single slotline and for the transmission line composed of two separate slots.

4 Experimental verification

The theoretical results of our spectral domain code were compared with the results of simulations performed using

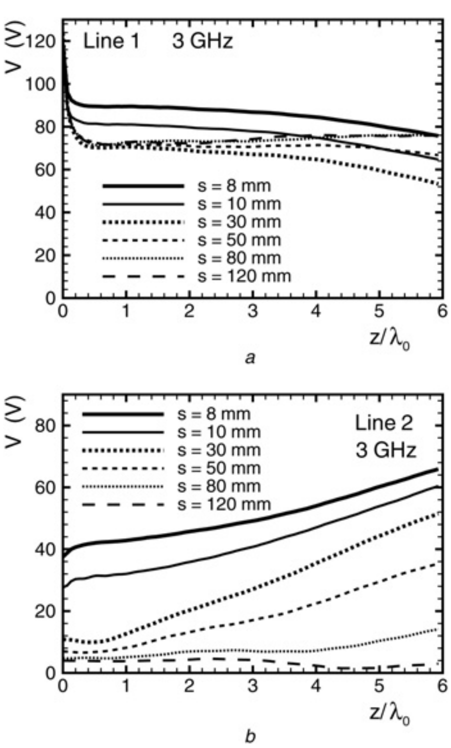

Figure 8 *Voltage distribution defined by $w = 5.6$, $h = 14.6$, and $\varepsilon_r = 2.6$ for various distances s*

a Line 1
b Line 2
Frequency is 3 GHz

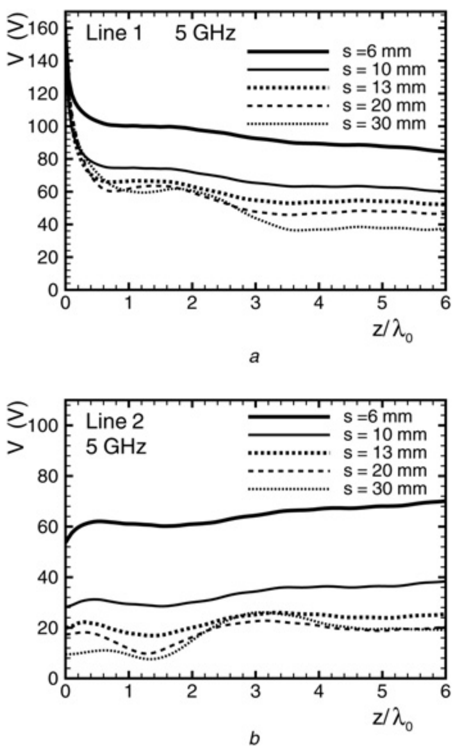

Figure 9 *Voltage along defined in Fig. 8 for various distances s*

a Line 1
b Line 2
Frequency is 5 GHz

doi: 10.1049/iet-map.2009.0448

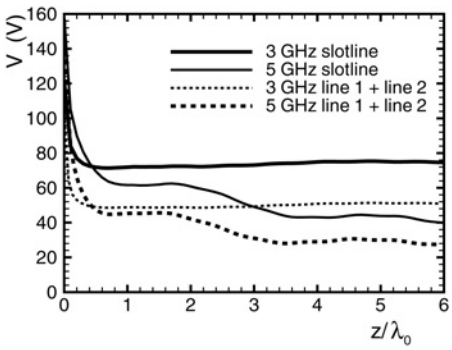

Figure 10 *Voltage along the single slotline defined in Fig. 8 and the sum of both waves along lines 1 and 2 for coupled lines with distance s = 10 mm*

CST Microwave Studio. The line investigated in Section 3.2 was fabricated and examined experimentally by measuring the electrical field distribution along the two lines. Fig. 11 shows the voltage of the waves propagating along the lines defined in Fig. 9 at frequency 5 GHz. The slotlines were modeled in CST Microwave Studio using 'open' boundary conditions at the substrate edges and therefore without reflections from the substrate edges. However, the fabricated line has finite dimensions and because of non-perfect termination by an absorbing material at the substrate edges, the results are degraded by a standing wave. The voltage distributions calculated by our code are also plotted in Fig. 11, for the sake of comparison. The agreement between all three plots in Fig. 11 is fairly good.

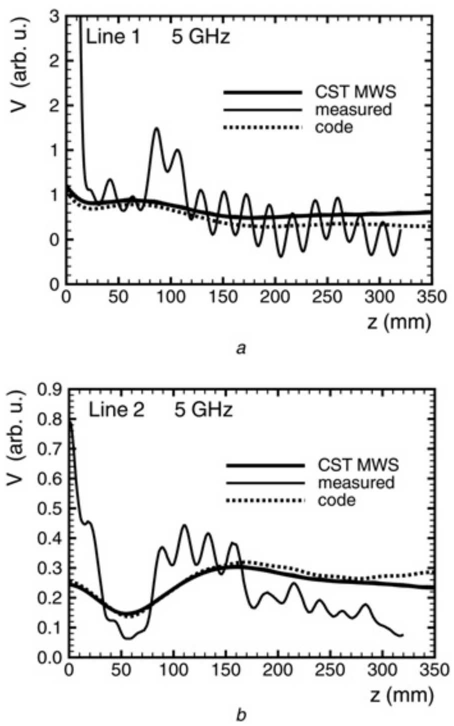

Figure 11 *Measured, simulated using CST Microwave Studio and calculated voltage defined in Fig. 9 for distance s = 20 mm*

a Line 1
b Line 2
Frequency is 5 GHz

The data are conveniently normalised for the purpose of comparison, and thus it is represented in arbitrary units.

5 Conclusions

The crosstalk between two slotlines has been studied in this paper. An analytical description of the voltage excited along the line by the source was derived, together with a description of the induced voltage along the passive coupled line. The three basic working regimes of the high permittivity substrate of the slotline were analysed. The first working regime studied here covers the frequency band of propagation of only the bound mode. The second regime is at those frequencies where both the bound and the second leaky modes propagate. Finally, the last regime is at high frequencies, where the bound mode does not propagate on either of the lines. Simulations performed using CST Microwave Studio and also measurements verified the theory sufficiently. The same investigation was also made, with adequate results, in the case of slotlines designed on a low-permittivity thick substrate.

The voltage distributions along the coupled lines are substantially modified by crosstalk. Strong maxima and minima appeared in the voltage distributions, and these two voltage distributions are complementary. The two coupled lines in fact represent one fictive line transmitting the total wave excited by the source. The total voltage of this wave is defined as the sum of the two particular voltages with respect to their phases. This total voltage is very similar to the distribution along a single slotline, as the same power must be transported both through this single slotline and along the two coupled slotlines. The crosstalk naturally loses its influence as the distance between the lines increases.

The presence of the crosstalk between the two closely located slotlines has to be taken into account in microwave design, because it can have a destructive influence in some configurations, in both analogue and digital transmissions guided through these lines.

6 Acknowledgments

The experiment was supported by the Czech Grant Agency under grant no. 102/09/0314 'Investigation of metamaterials and microwave structures with the help of noise spectroscopy and magnetic resonance', and the evaluation of the results and numerical simulations was supported by the Czech Grant Agency under grant no. 102/08/H018 'Modeling and simulation of fields', and by the Czech Technical University in Prague under grant No. SGS10/271/OHK3/3 T/13 'Artificial electromagnetic structures for high frequency technology.'

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