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Reduction of Far-End Crosstalk on Coupled Microstrip PCB Interconnect

Bertalan Eged, Assistant Professor (C-EGED@...) (*)
Ferenc Mernyei, MIEEE (**)
István Novák, Associate Professor, SIEEE (T_NOVAK@...) (*)
Péter Bajor, Postgraduate Student (PETER@...) (*)

(*) Department of Microwave Telecommunications, Technical University of Budapest
1111 Budapest, Goldmann tér 3, HUNGARY
Phone/FAX: +36-1-181-2968, E-mail: ...NOV.MHT.BME.HU
(**) TERATEC, Corp. R&D Dept. T, Sect 2.
2-11-13 Nakacho, Mushasio-Shi, TOKYO, 180 JAPAN
Phone: +81-422-37-2483, FAX: +81-422-37-2483, E-mail: MERNYEI@teratec.co.jp

Abstract - Crosstalk among interconnects and PCB traces is a major limiting of signal quality in high-speed digital equipment. The paper evaluates a possible way of far-end crosstalk reduction with a special geometry of traces. This paper shows that increasing capacitive coupling along coupled surface microstrip traces may be used to reduce far-end crosstalk.

I. INTRODUCTION

With lumped representation, n coupled transmission lines (n+1 conductors) are described by L, C, R and G matrices of size of *n* x *n*. Two lossless, symmetrical, reciprocal transmission lines are described by

$$L = \begin{vmatrix} L_{11} & L_{12} \\ L_{21} & L_{22} \end{vmatrix} = \begin{vmatrix} L & L_M \\ L_M & L \end{vmatrix} \qquad C = \begin{vmatrix} C_{11} & C_{12} \\ C_{21} & C_{22} \end{vmatrix} = \begin{vmatrix} C & -C_M \\ -C_M & C \end{vmatrix} \tag{1}$$

where L_M and C_M are the mutual inductance, and capacitance, respectively.

This structure may be described by its eigenvalues, even and odd mode characteristics. The wave can be decomposed to an even-mode and odd-mode wave. Each wave has its own characteristic impedance, and propagation constant. The even-mode and odd-mode characteristic impedances and phase constants are:

$$Z_{oe} = \sqrt{\frac{L + L_M}{C - C_M}} \qquad t_{pde} = \sqrt{(L + L_M)(C - C_M)} \tag{2}$$

$$Z_{oo} = \sqrt{\frac{L - L_M}{C + C_M}} \qquad t_{pdo} = \sqrt{(L - L_M)(C + C_M)} \tag{3}$$

We can define the relative capacitive (c_c) and inductive (c_m) coupling as

$$c_c = \frac{C_M}{C} \qquad c_m = \frac{L_M}{L} \tag{4}$$

Rearranging t_{pde} and t_{pdo}, we can obtain the following expressions:

$$t_{pde} = \sqrt{(L + L_M)(C - C_M)} = \sqrt{LC}\sqrt{(1 + c_m)(1 - c_c)} \tag{5}$$

$$t_{pdo} = \sqrt{(L - L_M)(C + C_M)} = \sqrt{LC}\sqrt{(1 - c_m)(1 + c_c)} \tag{6}$$

In homogeneous medium the capacitive and inductive couplings are equal, $c_c = c_m = c$, therefore the two modes propagate with the same speed: $t_{pde} = t_{pdo}$. If the two modes propagate with the same speed, no far-end crosstalk will appear with matched terminations. This is the case of coupled pair of planar striplines. In surface microstrip, however, due to the inhomogenity of the medium, t_{pde} and t_{pdo} are different. In surface microstrips, since the dielectric fills only the volume below the strips, the capacitive coupling is weaker than inductive coupling: $c_c<c_m$. In the time domain, this results in far-end crosstalk waveform with an opposite slope with respect to the main edge [1], [2].

To reduce far-end crosstalk on microstrip structures, we should increase the capacitive coupling. We can do it in different ways. In some cases, like backplanes, there are discrete loading and mutual capacitances on the lines. We can increase the coupling with these additional discrete capacitances [3]. Another way is to build more layers with different relative permittivities with the highest permittivity material being closest to the lines [4], [5]. This structure increases the coupling while the main capacitances remain nearly unchanged. Another method is to use overlay layer above the strips to increase coupling [6], [8]. All of the above methods have a disadvantage that they require additional elements and/or technological steps. The method described in this paper will show a technique, which requires only the modification of the layout of the lines.

II. COUPLED MICROSTRIPS

A frequently used trace structure on high-speed SMT PCBs and multichip modules is the microstrip configuration. As mentioned earlier, this coupled structure can be modeled by its L and C matrices. Our aim is to increase the capacitive coupling, so let's look at the capacitances of a coupled pair of microstrip lines; [9], [10].

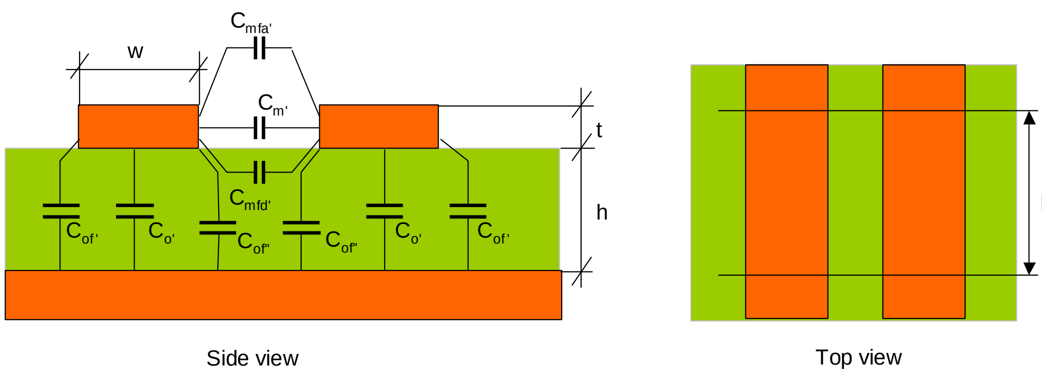

Fig. 1.: *Coupled pair of microstrip lines.*

The self and mutual capacitance of the structure:

$$\frac{C}{l} = \varepsilon_0\,\varepsilon_r\,K_{c1}\left(\frac{w}{h}\right) \tag{7}$$

$$\frac{C_m}{l} = \varepsilon_0\,\varepsilon_r\,K_{c2}\left(\frac{t}{s}\right) \tag{8}$$

where K_{c1} and K_{c2} are modeling the fringing effect on the strips ($C_{of'}$, $C_{mfd'}$, $C_{mfa'}$). For crosstalk calculations, we are interested in the mutual capacitance.

The mutual capacitance between lines can be calculated as a function of metal thickness, gap and the correction for fringing capacitances.

The main capacitance of the lines can be calculated as a function of substrate height, line width and a correction for fringing capacitances.

III. COMPENSATION METHOD

In the compensated structure we will increase the length of the gap without increasing the length of coupled section.

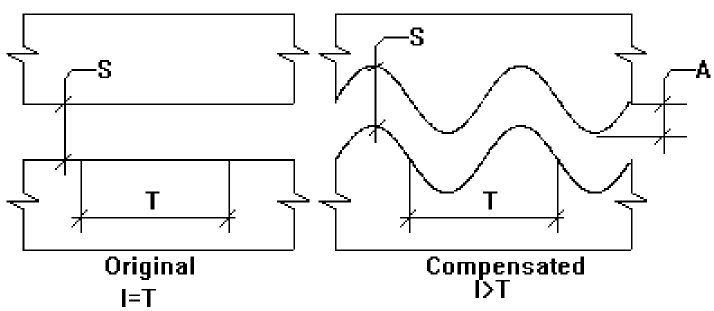

Fig. 2.: *Increasing mutual capacitance with gap profile function.*

The new mutual capacitance is:

$$\frac{C_m^{'}}{l^{'}} = \varepsilon_0 \, \varepsilon_r \, K_{c2}\left(\frac{t}{s}\right) \tag{9}$$

The *Increasing Factor* (IF) of the mutual capacitance is equal to:

$$IF = \frac{C_m^{'}}{C_m} = \frac{l^{'}}{l} \tag{10}$$

The main capacitances of the lines remain nearly constant because the surface areas of the lines were not changed. The fringing capacitances, on the other hand, will increase because the contours of the lines become longer.

At first we approximate the *Increasing Factor* (IF) by the line integral of the coupling profile. For the sake of simple computation, a sinusoidal profile was chosen so that the line integral and increasing factor are:

$$l^{'} = \int_T \sqrt{1 + \left[f^{'}(x)\right]^2}\, dx \tag{11}$$

$$IF = \frac{\int_T \sqrt{1 + \left[f^{'}(x)\right]^2}\, dx}{T} \tag{12}$$

It means that the new mutual capacitance is equal the capacitance of a planar capacitor with length *l'*, height *t* and gap *s*. However, the actual capacitance will be higher, because the electric field in the gap will take the shortest path. We can model this effect with a correction factor (K_p), which depends on the gap size, gap profile function and its amplitude.

$$IF^{'} = K_p \, IF \tag{13}$$

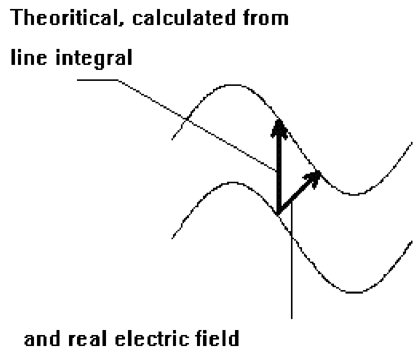

Fig. 3. *Electric field in the gap.*

The predicted *Increasing Factor* as a function of profile amplitude is shown in Figure 4.

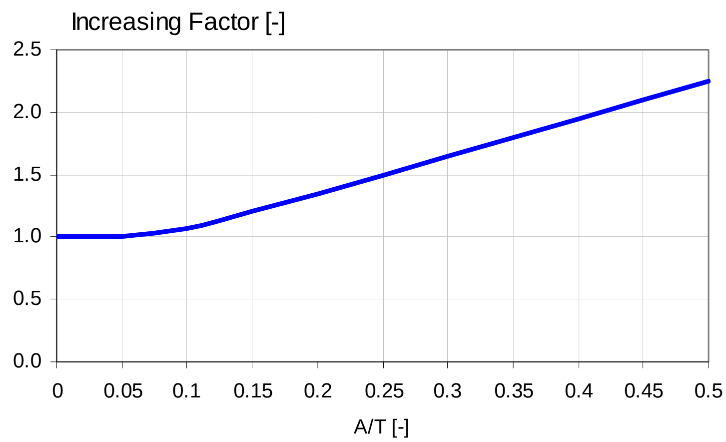

Fig. 4. *Increasing Factor (IF) computed from line integral of the gap profile function.*

IV. DEVICE UNDER TEST

For measurement purposes ten DUTs were built on epoxi fibreglass (FR4) material with copper-etching technology with h = 1.6mm, t = 0.018 mm and $\varepsilon_r = 4.3$. The total length was L = 200 mm, the width before patterning was w = 2.5 mm and the gap was s = l.5 mm. The sinusoidal function was chosen as a gap profile with a period of T = 2.5mm. On the ten different samples the relative amplitude of the profile function (A/T) varied in increments of 0.05. (A = 0..1.125 mm, A/T = 0..0.45), see Figure 5.

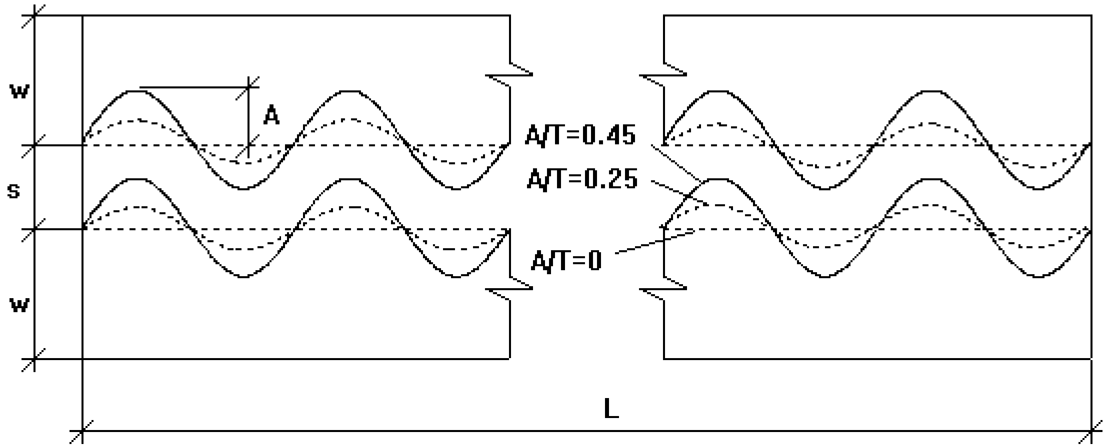

Fig. 5*. Top view of the device under test (DUT) used to measure the C and L matrices and time- and frequency domain far-end crosstalk.*

V. MEASURED AND SIMULATED RESULTS

First the low frequency capacitance and inductance matrices were measured with an HP4285A precision RLC meter. The values are shown in Figure 6.

A/T	C [F/mm]	C_M [F/mm]	L [H/mm]	L_M [H/mm]
0.00	1.200E-13	-4.390E-15	2.973E-10	3.123E-11
0.05	1.199E-13	-4.854E-15	3.038E-10	3.303E-11
0.10	1.197E-13	-4.807E-15	3.062E-10	3.123E-11
0.15	1.200E-13	-5.910E-15	3.128E-10	3.165E-11
0.20	1.215E-13	-7.355E-15	3.188E-10	3.315E-11
0.25	1.238E-13	-1.008E-14	3.237E-10	3.238E-11
0.30	1.248E-13	-1.217E-14	3.310E-10	3.235E-11
0.35	1.283E-13	-1.476E-14	3.347E-10	3.078E-11
0.40	1.330E-13	-1.844E-14	3.351E-10	2.458E-11
0.45	1.373E-13	-2.478E-14	3.413E-10	2.375E-11

Fig. 6*. Measured low frequency static capacitances and inductances of the DUTs with different A/T values.*

From these values we can calculate the capacitive and inductive couplings and the real value of *Increasing Factor*. The capacitances and inductances were measured on a 200 mm long DUT and the values were normalized to one mm length. We can calculate the capacitive (c_c) and inductive (c_m) coupling factors with Equation (4). If we compare the two coupling factors, we will find the particular gap profile amplitude, which gives equal capacitive and magnetic couplings, hence this will be the optimum compensation.

Figure 7 shows the computed and measured *Increasing Factor* values for the A/T range covered by the ten DUTs. For the same A/T values Figure 8 shows the capacitive coupling and inductive coupling values.

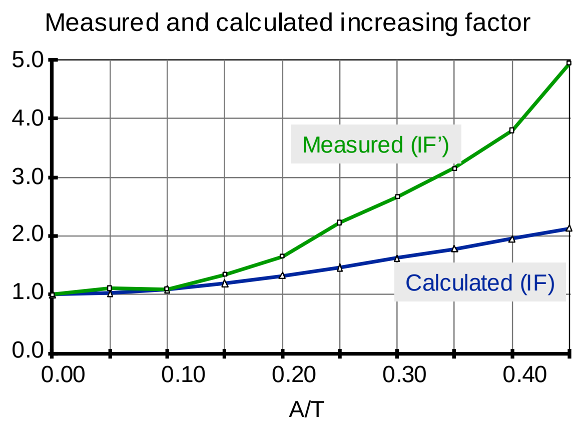

Fig. 7. *Coupling increasing factor claculated from line integral of gap profile function (IF), and calculated from measured capacitances (IF'). The real coupling increasing is higher than a calculated from line integral.*

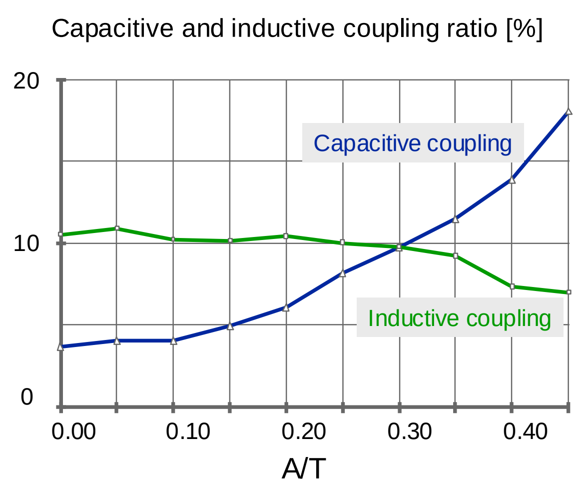

Fig. 8. *Calculated capacitive and inductive couplings. The two couplings are equal at A/T = 0.3 we can expect the maximally far-end crosstalk cancellation with this profile function.*

With the measured capacitance and inductance matrices we can simulate the far-end crosstalk on the lines. The calculation shows that the A/T = 0.3 relative gap profile amplitude is the best case for far-end crosstalk compensation. The simulation was done with MicroSim Pspice. The simulation results show that the A/T = 0.3 value causes a little overcompensation, while the A/T = 0.25 value is the best case. Figure 9 shows the simulated far-end crosstalk for three different gap profiles, assuming 0.1 ns input rise time and 1V source magnitude.

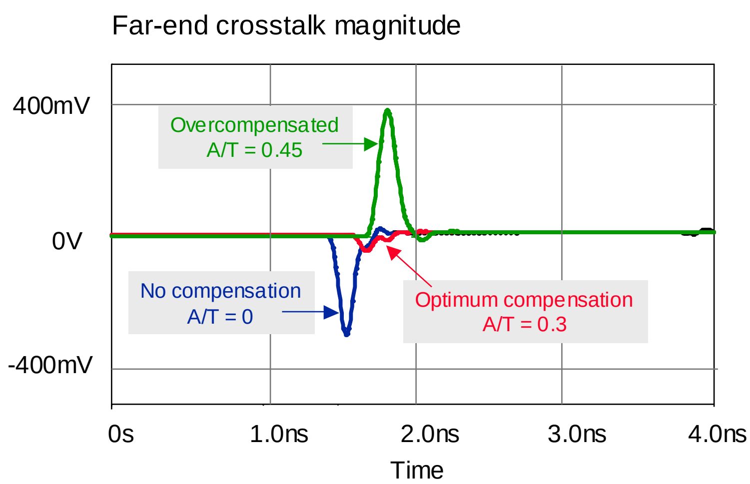

Fig. 9. *Simulated time-domain far-end crosstalk with three different gap profile amplitude.*

The DUTs were measured with an HP 8750 vector network analyzer with time-domain option. Figure 10 shows the measured time domain far-end crosstalk.

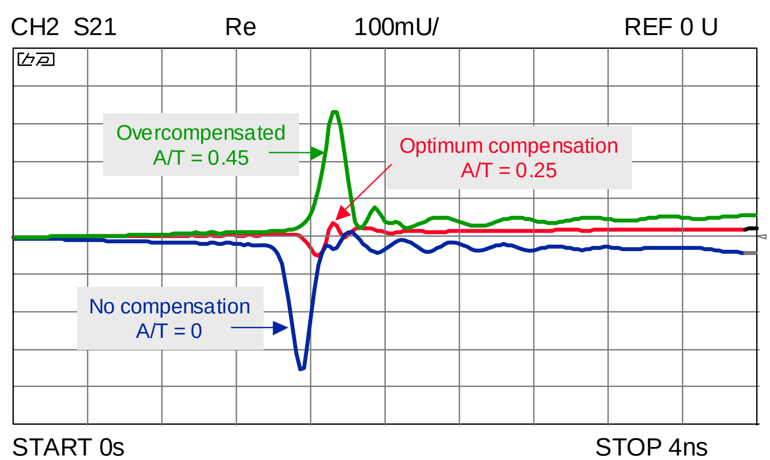

Fig. 10. *Measured time-domain far-end crosstalk.*

With the same network analyzer the far-end crosstalk was also measured in the frequency domain. Figure 11 shows the far-end crosstalk in the uncompensated case and on the optimally compensated lines. The magnitude response of an overcompensated DUT was nearly equal to magnitude response of the uncompensated DUT.

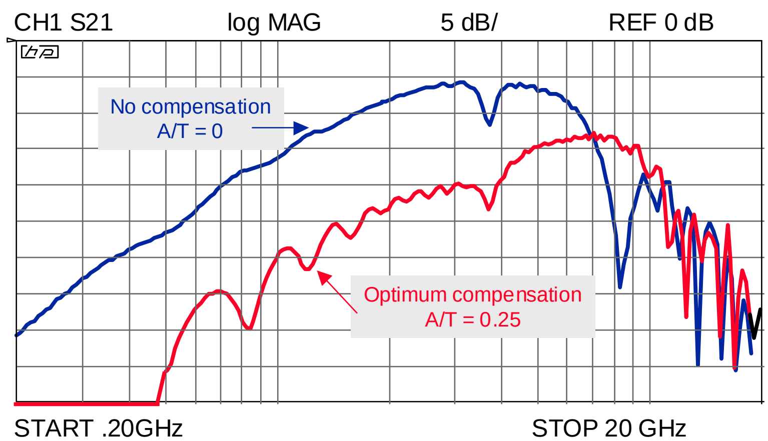

***Fig. 11**. Measured frequency-domain far-end crosstalk.*

VI. CONCLUSIONS

On coupled microstrip structures the capacitive coupling is weaker then magnetic coupling. This effect generates large far-end crosstalk. This crosstalk can be compensated with a modified trace geometry. The crosstalk reduction depends on the gap profile function parameters. With the actual dimensions used for the series of DUTs, it was possible to span a full range from undercompensation, optimum compensation, and also overcompensation.

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