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CROSSTALK REDUCTION ON STRIPLINE PRINTED CIRCUIT BOARDS WITH ADDITIONAL CENTER TRACES

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***ABSTRACT* - Reduction of crosstalk on microstrip transmission lines with additional center traces has been widely covered in the literature. This paper extends simulation and measurement results to stripline configurations, and shows that in spite of its homegenious medium, stripline configurations with more than two coupled traces will exhibit inherent far-end crosstalk. It is also shown that similarly to crosstalk reduction in microstrips, the grounded traces behave like resonators and may produce unwanted ringings on the waveforms. Both time-domain and frequency-domain responses are addressed.**

I. INTRODUCTION

Unwanted electromagnetic coupling among interconnects can seriously affect signal quality, and reduce noise margin in today's high-speed digital systems. Distortion of high-speed pulses, and crosstalk of multiconductor microstrip and stripline structures are covered in the literature from the early years of computers [1], [2]. A recent summary of state-of-the-art knowledge can be found in [1]. Reduction of coupling in fine-pitch dense PCBs and Multichip Modules, and consequently, reduction of crosstalk pulse magnitude may be most readily accomplished by increased physical separation of coupled conductors. In multiconductor flat cables the typical solution has been to introduce grounded conductors between signal conductors. The same solution on microstrip printed-circuit-boards is analyzed in [3], [4], [5], and [6]. In [7] it is shown both in the time and frequency domain that the grounded center traces on microstrip structures will act like undamped resonators, hence there is an increasing chance of ringing if the signal bandwidth is not adequately limited. So far, reduction of crosstalk in multiconductor striplines by additional grounded center traces is not covered in the literature.

The paper gives simulated and measured results of crosstalk reduction by additional center traces in stripline.

II. MODEL

The mechanical dimensions of the analysed structure is shown in Fig. 1. It consists of three coupled stripline traces, with equal spacing and line width on epoxy-glass (FR4) dielectric material. Copper thickness and dielectric heights were 17 microns, and 1.5 mm, respectively. In order to enable easy acces to the internal joint points, upper and lower halves were not glued together, thus leaving a narrow airgap between them.

The outer lines may be viewed as two active lines of a bus. To allow the effect of a centre line to be investigated, the structure is broken down into ten segments of identical length and dimensions, plus two identical short pieces to allow the mechanical connections of SMA connectors. The center strip can be connected to ground at the joints of segments. The total length of the structure was selected so that the first peaks both in the near-end and far-end frequency responses can be expected around or below 1GHz, which allows us to neglect the dispersion of the dielectric material. With the selected geometrical data, losses of DUT can not be neglected. During simulation, the DC and skin losses of traces were taken into account, the parallel loss of epoxy-fiber material was neglected.

Table I. and Table II. show the capacitance and inductance matrices used to simulate the DUT. Data for LC matrices, as well as for losses was obtained and simulation was done by the RLGC planar structure simulator and Signal Integrity software package of Contec Microelectronics U.S.A. Inc. The simulator package uses the method of

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characteristics for the coupled transmission lines (see [8] and [9]), thus inherently eliminates the ringing problems of simulations which may often arise when using lumped LC transmission-line equivalents.

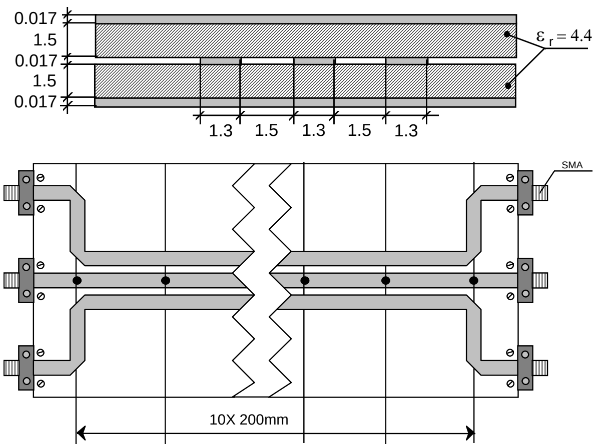

Figure 1.: *Mechanical dimensions of the Device Under Test. Top sketch: cross-sectional view. Lower sketch: top view with cover removed. Dimensions are in mm. Nominal trace impedance: 50 Ω.*

```
.model STRREFmod TRA nlines =  2
+CMATRIX =
+  1.382e-013     -5.062e-016
+                              1.382e-013
+LMATRIX =
+  3.519e-010     1.345e-012
+                              3.519e-010
```

Table I. *Unit-length capacitances and inductances of the reference structure, as used in SPICE simulation. Unit length: one mm.*

```
.model STRSHLDmod TRA nlines =  3
+CMATRIX =
+  1.393e-013     -1.052e-014         -5.142e-017
+                 1.404e-013          -1.052e-014
+                              1.393e-013
+LMATRIX =
+  3.510e-010     2.694e-011          2.218e-012
           +        3.502e-010        2.694e-011
           +                   3.510e-010
```

Table II. *Unit-length capacitances and inductances of the DUT as used in SPICE simulation. Unit length: one mm.*

III. CROSSTALK IN STRIPLINE BUSSES

The simplest approach to crosstalk analysis covers two symmetrical lossless coupled lines with matched terminations. Because of the symmetry and reciprocal nature of the coupled lines, the two-by-two inductance and capacitance matrices can be described by two independent values for each: L, L_M, C, and C_M. Assuming weak coupling, the crosstalk can be described in a simplified way by the crosstalk coefficients:

$$K_n = (L_M / L - C_M / C) / 4$$
$$K_f = (C_M Z_0 - L_M / Z_0) / 2$$

where Z_o is the characteristic impedance of line(s):

$$Z_0 = \sqrt{L / C}$$

For piece-wise-linear excitation and weak coupling, the approximate near-end (V_n) and far-end (V_f) crosstalk voltage magnitudes are given by:

$$V_n = K_n V_{in} \qquad (t_{tr} < 2t_{pd})$$
$$V_f = K_f \, l \frac{dv_{in}}{dt}$$

where t_{tr}, V_{in}, and dv_{in}/dt are the transition (rise or fall) time, the magnitude and slew rate of input step waveform, respectively, and t_{pd} is the one-way propagation delay of the lines. If the dielectric is nonhomogenious, the normalized capacitive and inductive couplings have different magnitudes, hence the far-end crosstalk coefficient is nonzero. This is the case in microstrips. In stripline structures, the dielectric is homogenious, the normalized capacitive and inductive couplings have the same magnitude, hence the far-end crosstalk coefficient is zero. Under matched-terminated conditions, zero far-end crosstalk coefficient will result in zero magnitude of the far-end crosstalk waveform.

It is important to note that the above simplified theory does not necessarily imply that far-end crosstalk is always zero in stripline configurations. While it is true that homogenious medium (stripline configuration) results in no dispersion of mode velocities, far-end crosstalk, as it is shown below, with more than two signal traces and with single-ended terminations to ground is generally not zero.

Figure 2 shows the equivalent schematic and node numbering for simulation runs and measurements. The upper schematic refers to the reference arrangement with only the two outer traces of the DUT in place (Table I), the lower schematic shows the actual DUT (Table II).

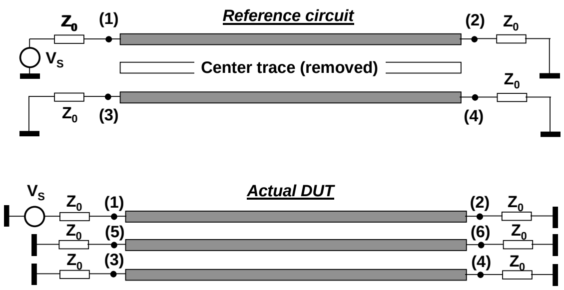

Figure 2.: *Equivalent schematic and node numbering.*

The calculated normalized capacitive and inductive couplings for the reference structure: L_M/L = 0.003822, C_M/C = 0.003663. The slight difference is due to the 0.017mm air gap between the two dielectric layers. Simulated waveforms of Fig. 3. show the near-end and far-end crosstalk waveforms. Note that as it is expected, in spite of the long coupled length, far-end crosstalk is very small.

In case of three coupled stripline traces, L and C values from Table II are taken. For two adjacent traces the normalized inductive coupling and capacitive coupling is L_M/L = 0.07675, and C_M/C = 0.07552, respectively. Note that these values are close, and the difference can be again due to the narrow air gap. Compared to the previous values, the stronger coupling is evident because of the reduced separation between the traces.

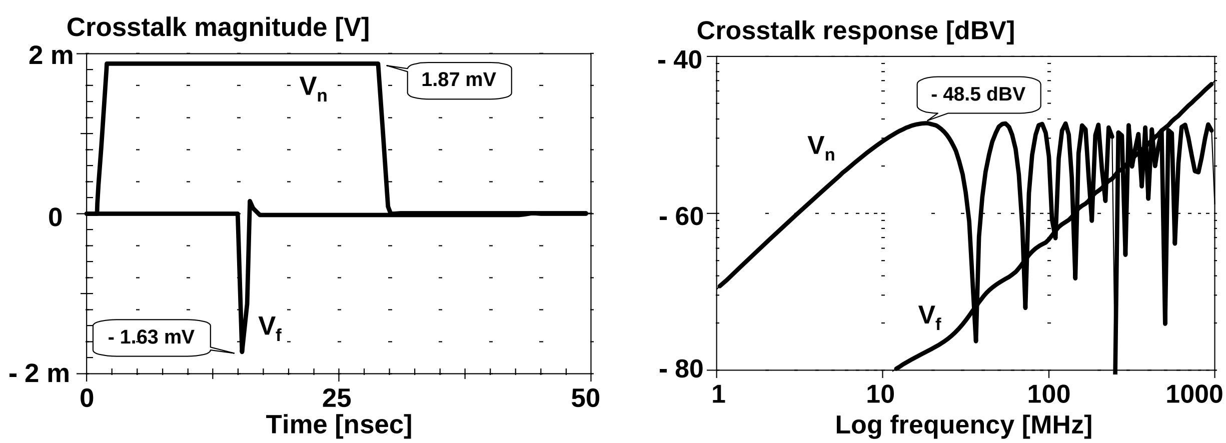

Figure 3.: *Simulated crosstalk response of the reference structure with two outer traces and matched terminations. Source voltage for time-domain response: 0-to-2 V piece-wise linear ramp. Source voltage for frequency-domain response: 2 Vrms sinewave.*

However, if we consider the two outer traces with the third trace in place, normalized inductive and capacitive coupling becomes L_M/L = 0.006319, and C_M/C = 0.000369, respectively. This indicates that the capacitive coupling between the outer traces is much smaller than the inductive coupling. Also, the fact that the center trace is in its place, increased the inductive coupling almost by a factor of two.

With single-ended matched terminations on all six nodes, the resulting crosstalk responses are shown in Figure 4 for the two adjacent traces and for the two outer traces. Note that the near-end and far-end frequency responses between the two outer traces are almost identical, which may be in contradiction with the general expectations. This is because the coupling between these traces is essentially just magnetic coupling. As the magnetically coupled current flows through the input (node 3) and output (node 4) terminations, the resulting time-domain waveforms have similar shapes and magnitudes. Responses between the adjacent traces are according the usual expectations.

To reduce the crosstalk, the center line can be grounded. Figure 5 shows the simulated response with different combinations of extreme termination at the far ends of center trace (short, open, short-open). Note that on the left-hand side graph in Figure 5, the plateau on waveform A_1 (center trace shorted) and A_2 (center trace open) is 0.2 mV and 3.2 mV, respectively, which compares with 1.65 mV for matched termination (see Fig. 4).

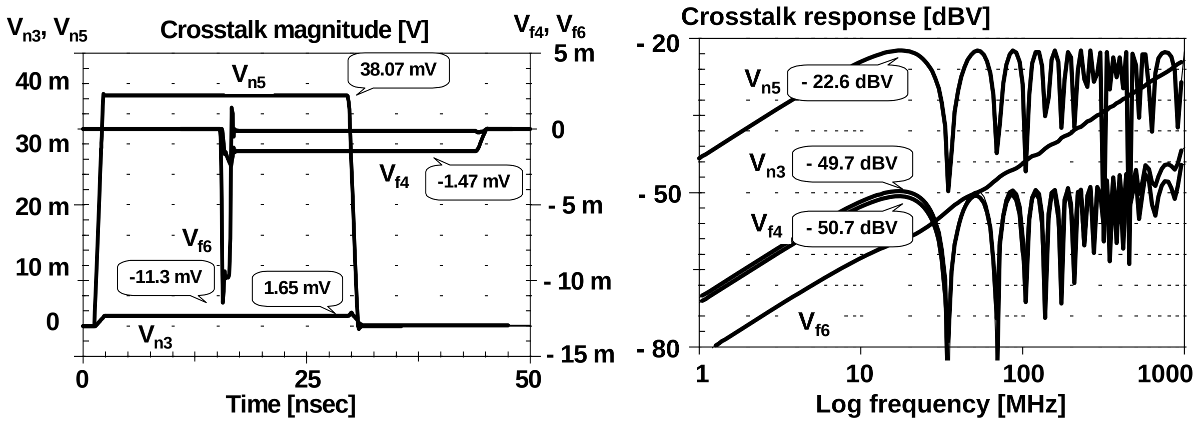

Figure 4.: *Time-domain (on the left) and frequency-domain (on the right) crosstalk responses of the DUT with single-ended matched terminations (50-ohms) for the two adjacent traces (V_{n5}, V_{f6}), and for the two outer traces (V_{n3}, V_{f4}). Note that the frequency responses between the two outer traces are very similar at the near and far ends, which is in contradiction with the general expectations when the simplified crosstalk model is used. Responses between the adjacent traces follow the outcome of the simple model.*

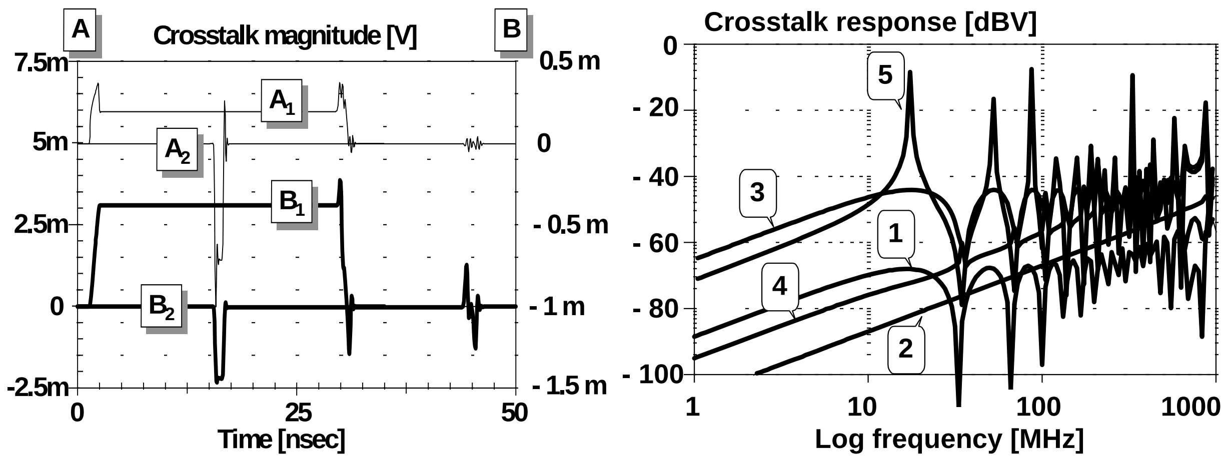

Figure 5.: *Simulated responses with extreme termination on the center trace, time-domain responses on the left, frequency-domain responses on the right. Time-domain responses are shown for open terminations at nodes 5 and 6 on scale A (trace A1: near-end, trace A2: far-end), and for short terminations at nodes 5 and 6 on scale B (trace B1: near-end, trace B2: far-end). Traces in the frequency-domain graph: traces 1 and 2: near-end and far-end responses with short terminations at nodes 5 and 6, traces 3 and 4: near-end and far-end responses with open terminations at nodes 5 and 6, trace 5: near-end and far-end responses (same trace) with short termination at node 5 and open termination at node 6.*

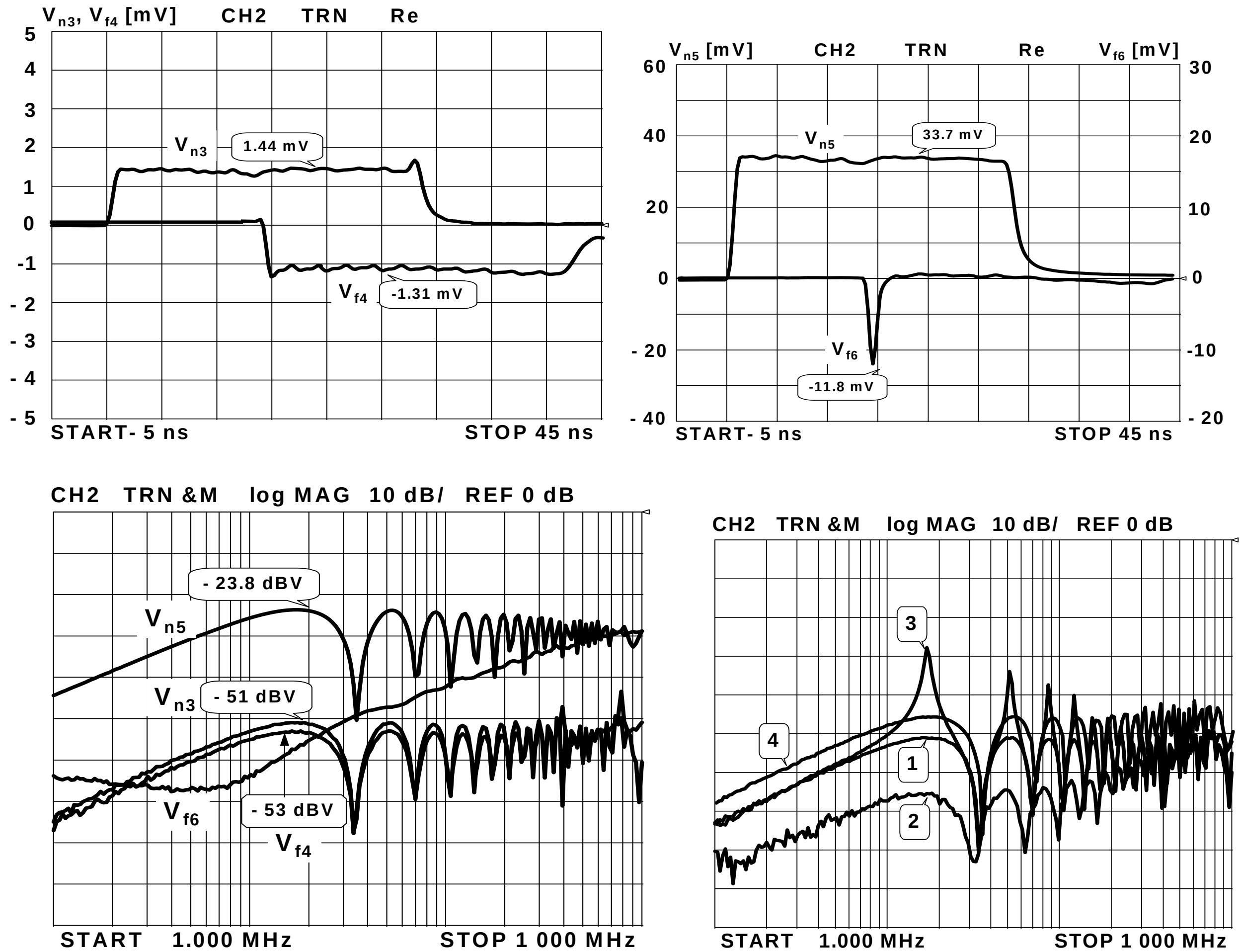

Figure 6.: *Measured responses of the DUT. Upper left graph: near-end and far-end time-domain responses on the two outer traces with matched terminations on all nodes. Upper right graph: near-end and far-end time-domain responses on the center trace, with matched terminations on all nodes. Lower left graph: near-end and far-end frequency-domain responses on the center and outer traces with matched terminations on all nodes. Lower right graph: near-end frequency-domain responses at node 3 with the center trace matched terminated (trace 1), with the center trace shorted to ground at the far ends (trace 2), with the center trace left open at the far ends (trace 3), and with the center trace grounded at node 5 and shorted at node 6 (trace 4).*

IV. MEASUREMENT RESULTS

Figure 6 shows measured responses of the DUT in different configurations. Frequency-domain and time-domain responses were measured by an HP8752A Vector Network Analyzer with TDR option. The first three graphs of Fig. 6 compares with the values of Figure 4, while the fourth graph compares with Figure 5. Note the good correspondence between simulated and measured data. On the lower left-hand side graph of Figure 6, the far-end response between the two adjacent traces (V_{f6}) has a plateau at low frequencies. This is due to the finite DC resistance of the ground planes. Also, the near-end and far-end responses on the two outer traces are slightly different (-51 dB and -53 dB) which is due to the losses.

V. CONCLUSIONS

For multiple-trace stripline busses, capacitive coupling between non-adjacent traces will be significantly smaller than magnetive coupling, thus the far-end crosstalk is not zero. It is shown that both the near-end and far-end crosstalk can be reduced by grounded center traces. Increase of crosstalk occurs only when the shield trace is grounded at one end and its other end left open.

Acknowledgment

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