COUPLING OF OPEN AND SHIELDED TRACKS ON PRINTED CIRCUIT BOARDS

Alexander P.J. van Deursen and Sergei Kapora
Department of Electrical Engineering, Eindhoven University of Technology
E-mail: a.p.j.v.deursen@tue.nl, s.kapora@tue.nl

Abstract: The DM-CM-DM coupling between tracks of a PCB and the CM circuit formed with a nearby cabinet panel is discussed based on measurements. A novel track shielding technique and its consequences for signal transport are presented; intentional partial shielding is discussed.

Key words: Printed circuit board, DM-CM coupling, resonant coupling, conformal mapping.

1. Introduction

When the signal circuits form closed loops on a printed circuit board (PCB), these may be regarded as differential mode (DM) circuits from the EMC point of view. A metal cabinet and the PCBs inside then form common mode (CM) circuits. The DM circuits can induce a CM current through a first order DM-CM coupling, and vice versa. It is well known [1], that resulting radiation due to the CM circuits often dominates the direct radiation by the DM circuits, in particular if cables are attached to the board. At the CM resonance frequencies however, a second order coupling DM-CM-DM also becomes important, which affects the DM signal transmission on the board. In order to show the importance, we present measurements for a selected model PCB, which because of its open structure clearly demonstrates the influence of track position and geometry.

One way to drastically reduce the undesired coupling is to replace the commonly used microstrip and strip lines by fully shielded track structures. We investigated such a structure, the recently developed E-Coax [2]. The shield is manufactured by creating and subsequently metalizing trench walls in three outer layers of a multilayer PCB. The middle layer contains the tracks; the top and bottom layer are short-circuited by the wall metalization. One or several tracks may be contained in a single shield. We studied the characteristic impedance $Z_0$ of a single signal track inside a shield by means of numerical Schwarz-Christoffel mapping (SC) for doubly connected regions [3]. In order to assess the importance of the fully closed shields, we also calculated the DM-DM and DM-CM coupling for situations where the metalized sidewall intentionally left a small slit open near the bottom of the trench. Here we used numerical conformal mapping for singly connected regions [4].

2. DM-CM-DM coupling measurements

Figure 1 shows the configuration and the geometry parameters of a 3-layer PCB mounted in a so-called precompliance setup, similar to the one proposed in [5]. A continuous copper ground plane (GP) has the length $l = 200$ mm, width $2w = 100$ mm and thickness $d = 30$ µm. Tracks of width $b = 1.5$ mm are placed at the distance $h = 1.5$ mm above and below the GP. The cabinet panel (CP) is a folded brass plate of $200$ mm width and $2$ mm thickness. The PCB is placed above this panel, usually at the distance $h_{CP} = 10$ mm. The vertical side B can be moved over the CP to adjust the gap width $g$, which can be varied between 1 and 20 mm. The DM circuit is formed by track 1 and continues through a via in the GP into track 2; the GP acts as return for both tracks. The CM circuit consists of the GP and the CP. This setup allows a direct measurement of the DM-CM coupling, and it bears resemblance to a PCB mounted on a back-plane.
The $s$-parameters have been measured with network/spectrum analyzer HP 4396A and $s$-parameter set HP 85046A. A pair of high quality SMA cables, selected for their low transfer impedance, connected port 1 and 2 of the PCB to the measuring equipment. The transfer from port 1 to 2 has been expressed as "available power gain" parameter $G_d$, which relates to the $s$-parameters as

$$G_d = \frac{|s_{21}|^2 (1-|\Gamma_s|^2)^2}{(1-|s_{22}|^2 + |\Gamma_s|^2 (|s_{11}|^2 - |D|^2) - 2 \text{Re}(\Gamma_s M))}$$

with $\Gamma_s = (Z_s - Z_0)/(Z_s + Z_0)$, $D = s_{11} s_{22} - s_{12} s_{21}$, and $M = s_{11} - D s_{22}$. Here $Z_0 = 50 \Omega$ is the input and output impedance of the network analyzer ports, and $Z_s$ is the characteristic impedance of the microstrip lines formed by the tracks and GP. As is apparent from above equation, $G_d$ is mainly controlled by the parameter $s_{21}$. However, it shows a smoother behavior as function of frequency; nevertheless, $G_d$ retains the DM-CM-DM resonances sought.

![Fig. 2. Measured $G_d$ for the straight track between the GP and CP (lower curve and left scale) and for the straight track on top of the PCB (upper curve and right scale).](image)

The DM transmission characteristic of the combined track 1 and 2 on the PCB has been studied for different configurations. The lower curve in Fig. 2 presents the experimental $G_d$ when the straight track 2 faces the cabinet panel. The gap width is $g = 10$ mm. The transfer shows its resonant behavior at about 0.30, 0.95 and 1.6 GHz. These frequencies correspond to odd multiples of the first quarter wavelength resonance over the length of the CM circuit. The resonance frequencies depend only slightly on the gap width $g$. A larger gap leads to a higher resonance frequency, because of the smaller gap capacitance $C_g$ terminating the CM circuit at side B. When the board is turned upside down and meandering track 1 faces the CP, the intensity of the first resonance (dip in $G_d$ at 300 MHz) becomes less pronounced, while the intensity at the third one becomes larger; see the upper curve in Fig. 2. In order to determine the relative coupling strength of both tracks, we replaced the straight track 2 between the GP and CP by a well shielded semi-rigid coaxial cable. The resonances in $G_d$ were absent over the entire 0.3 – 1.8 GHz frequency range. A similar effect is observed when the cabinet panel is placed at a large distance $h_{CP} > 30$ mm, or when it is absent. The addition of 150 $\Omega$ CM load over the gap, as in the original precompliance setup [5], damps the CM resonances sufficiently to make them no longer visible in the DM transfer for both PCB orientations.

The dielectric loss of the FR4 material is responsible for the overall non-resonant decrease of $G_d$ with frequency. A fit of the non-resonant part to a simple transmission line (TL) model for the 0.8 m long track resulted in a frequency-independent loss tan($\delta$) of 0.019, which is a common value for FR4.

Summarizing, at lower frequencies the DM circuit with the straight track between CP and GP couples stronger with the CM circuit in comparison to the meandering track between CP and GP. This difference can be attributed to two causes. First, the $z$-directed wave velocity of the current through the straight track has a closer match to the one in the CM circuit; the meander has intermediate $x$-oriented delay lines. Secondly, the ratio of the DM-CM coupling for the straight track between or on top of the GP is about a factor $\pi w/2h_{CP}$, or 8 in the present case [5, Eq. 4 and 8]; the coupling for the meandering track is then neglected because of its mismatch in electrical length. The factor $\pi w/2h_{CP}$ also explains the reduction in resonant coupling when $h_{CP}$ is made larger.

![Fig. 3. Saw-cut cross section of shielded track structure.](image)

3. Shielded tracks characteristic impedance

As mentioned in the introduction, fully shielded PCB tracks can be manufactured nowadays. Figure 3 shows a saw-cut cross section. We investigated the dependence of the characteristic impedance $Z_0$ on manufacturing tolerances for the position of a single signal track by numerical Schwarz-Christoffel mapping (SC) for doubly connected regions [3]. The parameters are shown in Fig. 4: base width $b$, height $h$.
$h$, width of the track $wt$, thickness $t$, slope angle between base and sidewall $\alpha$ and also coordinates of the track’s center. The track and shield were regarded as perfect electrical conductors.

Fig. 4: Parameters of the cross-section of the shielded E-Coax with single signal conductor.

The cross-section of the E-Coax is smaller than 1 mm; the implicitly assumed TEM approximation is then valid up to many GHz. In the example the height $h = 450 \mu m$, the base width $b = 700 \mu m$, the angle $\alpha = 80$ degrees, the track width $wt = 200 \mu m$, track thickness $t$ is $18 \mu m$ and $\varepsilon_r = 3.5$. This results in a largest $Z_0$ of 49.91 $\Omega$ when the track is located at the center, i.e. $x_{\text{center}} = 0, y_{\text{center}} = h/2$. Figure 5 shows the contours of deviations of 1, 2, 5 and 10 percent from the maximum $Z_0$-value when the position of the center of the signal track is varied in both $x$ and $y$ directions, with the other parameters constant.

For the four points (• × + o) indicated in Fig. 5, we compared the SC results with $Z_0$ obtained by method of moments (MoM) with adaptive discretization to improve accuracy [6]. The results were identical up to at least the fourth significant digit, or to within 0.025 percent.

4. Partly shielded track
To assess the relative importance of the fully closed shields, we calculated several coupling situations where the metalized sidewall of the shield intentionally left a slit open at the bottom of the trench. The slit in one sidewall was varied between fully open ($g = 1$) and nearly closed ($g = 0$); see Fig. 6. This factor $g$ is to be distinguished from the one in Sect. 2. In the calculations a single wire source replaced the single track; the wire was positioned at the track center, which is at half height in the E-Coax, now of $h = 468 \mu m$. The current through the wire is assumed to return through the shield and the connected ground plane (GP). The static mutual inductance $M$ for two wire tracks at 0.8 mm distance has been taken as measure for the coupling.

Table 1 shows the $M$-values for 4 basic configurations, a microstrip and a stripline, versus an E-Coax with “open” and “half-open” ($g = 0.5$) sidewall. Even for half-closed sidewalls of E-Coax, $M$ is much smaller than for the usual transmission lines on a PCB (upper two in Table 1).

<table>
<thead>
<tr>
<th>Configuration</th>
<th>$M$, nH/m</th>
</tr>
</thead>
<tbody>
<tr>
<td>• • • • • • •</td>
<td>29.433</td>
</tr>
<tr>
<td>• • • • • • •</td>
<td>1.861</td>
</tr>
<tr>
<td>• • • • • • •</td>
<td>2.181</td>
</tr>
<tr>
<td>• • • • • • •</td>
<td>0.1746</td>
</tr>
</tbody>
</table>

Table 1: Mutual inductances for 4 basic configurations. The wire distance is 800 $\mu m$, the height above the plane is 234 $\mu m$.

Fig. 6: Configuration for cross-talk modeling. The parameter $g$ indicates the open fraction of the sidewall.

More detailed study considered three different situations. The first situation was the track-to-track cross talk with one track inside an E-Coax and the other track close to it above a ground plane indicated by the thick solid line in Figure 6. The ground plane (GP) extended to infinity in both $x$-directions. The cross-section of the GP and the E-Coax wall in the complex plane $z = x+iy$ were mapped onto the inside of the unit circle in the complex $z$-plane, with the source wire at the center. The numbers in Fig. 6 indicate the consecutive vertices for this situation. The mutual inductance $M_1$ with respect to the sec-
ond wire then follows from \( M_1 = -\mu_0/2\pi \ln(|\zeta_2|) \) with \( \zeta_2 \) the \( \zeta \)-position of the second wire. The upper line in Fig. 7 presents \( M_1 \) as function of the open wall fraction \( g \). For smaller values \( g < 0.5 \), \( M_1 \) is about proportional to \( g^{-2} \).

The second situation concerned the track-to-track cross talk when both sidewalls between the tracks had the same slit \( g \) (additional dashed shield in Fig. 6). The highly accurate and stable cross-ratio SC procedure [4] directly gave accurate results even for narrow slits or small \( g \). The GP had to be modified into a large box around the E – Coax (dashed line in Fig. 6) to provide the closed structure required by the existing Toolbox software. The size of this box was chosen at least 5 times the size of the E – Coax structure; this ensured that the M-results did not depend on the box size to within an accuracy of \( 10^{-3} \) desired here. The resulting \( M_2 \) is again shown in Fig. 7 by the lower line. For small \( g \), \( M_2 \) varied proportional to \( g^{-7} \).

The third situation deals with the coupling towards the external world, or with the mutual inductance \( M_{cm} \) between a track circuit and the circuit formed by the ground plane, now of finite width \( 2w \), and the environment. The \( M_{cm} \) describes the flux between the ground plane and infinity due to a current through the track, which returns through the GP as above. The GP with the E – Coax is again mapped onto the unit circle and its exterior region. A line dipole is formed by the transform \( \zeta_{w} \) of the current carrying wire and its mirror image with respect to the circle at \( 1/\zeta_{w} \). The value of \( M_{cm} \) is then obtained by \( -\mu_0/2\pi \ln(|\zeta_{w}|) \) [7, Eq. A5]. In the numerical example, the width \( 2w \) of the GP is taken equal to 4 mm, with the E – Coax in the middle. In [7, Eq. A1] it is also shown that \( M_{cm} \) varies with \( 1/w \). As could be expected, \( M_{cm} \) varies for small \( g \) in a similar way as \( M_1 \).

The accuracy of the calculated \( M \)-values has been estimated from the accuracy \( a \) of the transforma-

5. Conclusions

Resonant DM-CM-DM coupling has been demonstrated in a precompliance setup for a three layer PCB. A shielded straight track showed a strongly reduced coupling. A novel PCB shielding technique has been analyzed by Schwarz-Christoffel transformation for doubly connected polygonal regions', ACM Trans. Math. Soft., 24(3): 317-33, 1998

References