COUPLING EFFECTS AND RADIATION OF A 3-LAYER PCB

ON TOP OF A METAL CABINET PLANE

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ABSTRACT

Metal cabinets form common mode (CM) circuits with the printed circuit boards (PCB) inside. Those CM circuits can affect the differential mode (DM) signal transmission and lead to the higher radiated emissions from the PCB. In this contribution we study the effect of a single nearby metal cabinet plane (CP) on the DM transmission characteristics and radiation of a 3-layer PCB with a ground plane (GP). The CM structure induces several resonances in the DM transfer over 0.1 – 1.8 GHz frequency range, depending on track geometry. The corresponding increase of radiation at resonant frequencies is also observed. Simulations confirm the experimental results.

INTRODUCTION

Metal cabinets are often used to protect electronic equipment against electromagnetic interference. However, the cabinet and the printed circuit boards (PCB) inside form common mode (CM) circuits, which may influence the differential mode (DM) signal transmission on the boards and cause higher radiated emissions from the PCB. We study this coupling effect both experimentally and numerically.

SETUP DESCRIPTION

The 3-layer PCB is mounted in a so-called precompliance setup, similar to the one used in [1]. The configuration and the parameters of geometry are shown in Fig.1.

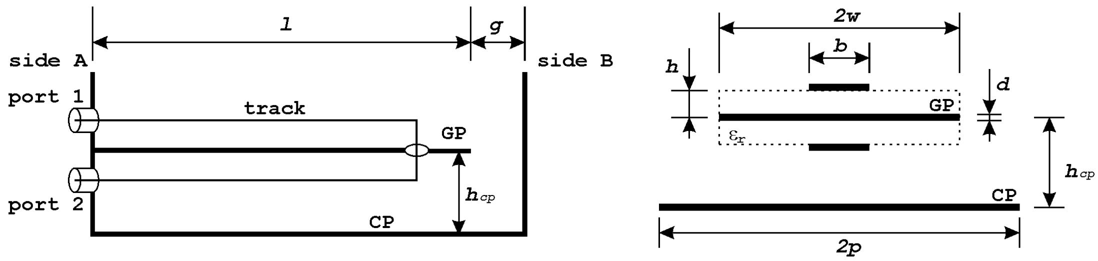

Fig. 1. Side-view and cross-section of a PCB in a precompliance setup

A continuous copper ground plane (GP) of PCB has length l = 20 cm, width $2w$ = 10 cm and thickness d = 0.03 mm. Tracks of width b = 1.5 mm are placed at height h = 1.5 mm above and below the GP. Track 1 above meanders over the surface, while track 2 below runs parallel to the length of the PCB. The tracks can be interconnected at the far end (side B) through a hole in the GP, in order to determine the DM signal transfer. Alternatively, both tracks can be connected to the ground plane at the far end. The dielectric constant of the board material (FR-4) ε_r = 4.7. The cabinet plane (CP) is folded from a brass plate of 200 mm width, 300 mm length and 2 mm thickness. The PCB is placed above the CP at distance h_{CP} = 10 mm. The gap width g between the edge of the PCB and the vertical part of cabinet at the side B can be varied between 1 and 20 mm. In this configuration the DM circuits consist of track 1 and the GP, and track 2 and the GP, while the GP and the CP form the CM circuit.

The S-parameters of the system were measured with the network/spectrum analyzer HP 4396A and the S-parameters test set HP 85046A. The PCB was connected to the measuring equipment by a pair of high quality SMA cables. Each measurement session was preceded by a full 2-port calibration.

Simulations were performed with the transient solver of CST Microwave Studio 3.0 [2], which uses the finite integration method. A Gaussian-shaped pulse was used as an excitation signal. The frequency domain S-parameters were obtained by applying the Discrete Fourier Transform (DFT) to the time domain results. Simplified SMA connectors and waveguide ports were used to excite and absorb power. The tracks, the ground and cabinet planes were modeled as surface impedances with $\sigma = 5.6 \cdot 10^7$ S/m and $\mu_r = 1$. The open boundary conditions were applied at the distance of several cell sizes. The "lines per wavelength" parameter, that defines the typical cell size, was chosen to be 10. Other settings were: cell ratio limit of 40, finer discretization in critical regions of the system (i.e. connectors and cross-sections of microstrip and board dielectric). The accuracy settings required for time-domain signals to sufficiently decrease to zero led to about 125 000 mesh cells and 24 hours of calculation (Pentium III 1GHz) for the typical configuration (g = 10 mm, frequency range 0 – 2 GHz).

DM TRANSFER

The DM signal transmission characteristics of the 3-layer PCB were studied in a number of configurations. With both tracks interconnected at the far end through a hole in GP, the signal transfer from port 1 to port 2 depends on the presence of a nearby cabinet plane and the orientation of the PCB (or geometry of the track that is located between GP and CP). This is clearly seen if the transfer is expressed in the available power gain parameter G_A, defined as the ratio of the power available from the network to the power available from the source. G_A is related to the well-known S-parameters as:

$$G_A = \frac{|s_{21}|^2\left(1-|\Gamma_S|^2\right)}{\left(1-|s_{22}|^2\right)+|\Gamma_S|^2\left(|s_{11}|^2-|D|^2\right)-2\,\mathrm{Re}\left(\Gamma_S M\right)} \tag{1}$$

with $\Gamma_S = (Z_S - Z_0)/(Z_S + Z_0)$, $D = s_{11}s_{22} - s_{21}s_{12}$, and $M = s_{11} - Ds_{22}^*$. Here Z_S =50 Ohm is the input impedance of the network analyzer, and Z_0 =68 Ohm – the characteristic impedance of the microstrip lines formed by the tracks and GP.

The numerical and experimental results for the PCB with a straight track 2 between GP and CP are shown in Fig.2. The gap width is g = 10 mm and the distance between GP and CP is h_{CP} = 10 mm.

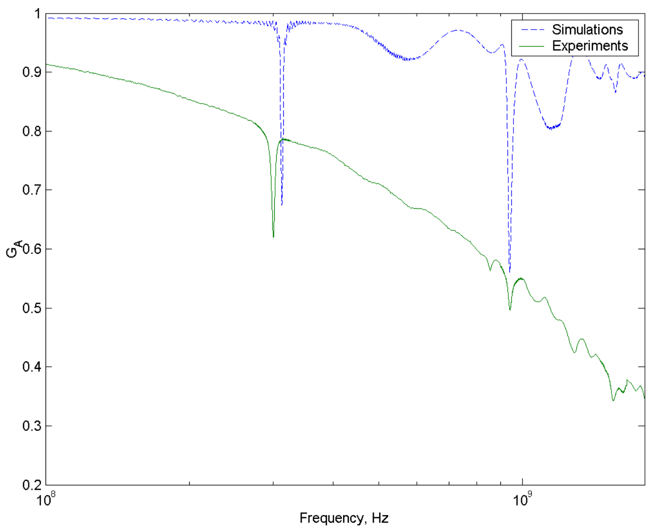

Fig. 2. Available power gain. Straight track 2 between GP and CP

The structure shows its resonant behavior at about 300Mhz, 900Mhz and 1.5 GHz. These frequencies correspond to odd multiples of the $\lambda/4$ resonance in the CM circuit. The calculations also show the corresponding increase of the local fields in expected areas of the setup (i.e. electric field inside the gap and of the magnetic field between GP and CP). The exact values of the resonance frequencies slightly depend on the gap width g, as illustrated in Fig.3 for the first resonance near 300 MHz. One can see that increasing the gap, and as a result decreasing the parasitic capacitance C_g between the far edge of the PCB and vertical part B of the cabinet [1], leads to the increase of the resonance frequency.

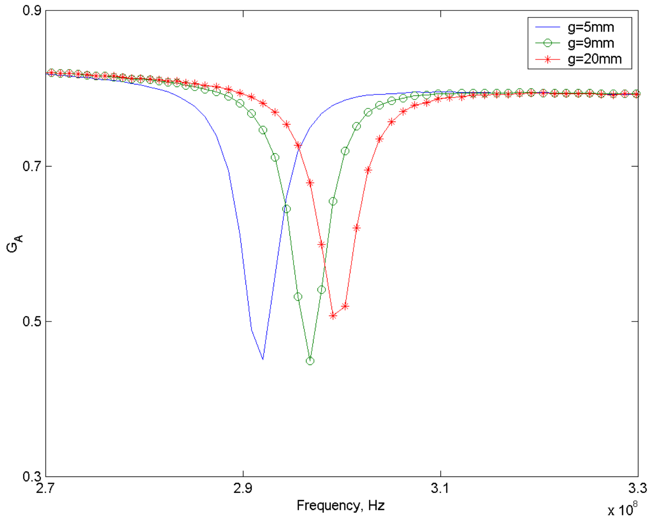

Fig. 3. Effect of the gap width g on the frequency of first resonance

With the board turned upside down (meandering track 1 is located between GP and CP), the intensity of first resonance becomes less, while the intensity of the third one becomes much higher (see Fig.4). This behavior is observed in both numerical and experimental results.

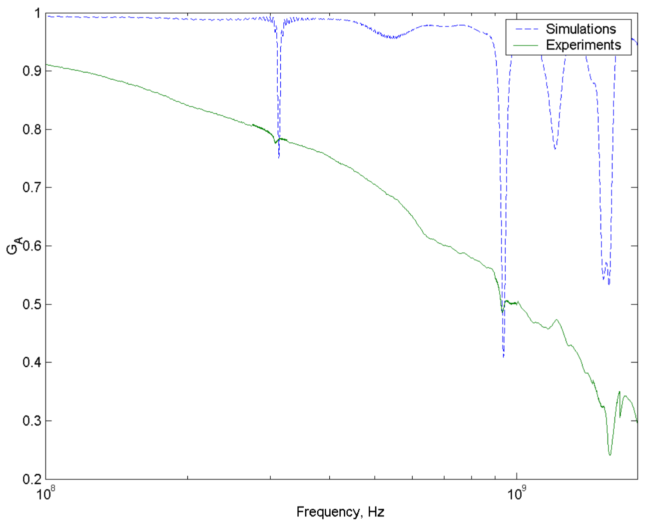

Fig. 4. Available power gain. Meandering track 1 between GP and CP

In another configuration the straight track 2 is shielded (i.e. replaced by a semi-rigid coaxial cable). The resonances in G_A are absent over the full 0.1 – 1.8 GHz frequency range. The similar effect is observed when the cabinet plane is absent or located at a large distance $h_{CP} > 30$ mm. These experimental results are shown in Fig. 5. Addition of 150 Ohm CM load over the gap, as in the original precompliance setup, damps the CM resonances, which are then no longer visible in the DM circuit for all configurations.

The overall non-resonant reduction of G_A with frequency in experimental data is most likely caused by the dielectric losses of the PCB material. This type of losses was not included in the model.

Summarizing, at lower frequencies DM circuit with the straight track between CP and GP couples stronger with the CM circuit in comparison to the meandering track. The same behavior at frequencies up to 650 MHz was observed in [1] by using a simple transmission line (TL) model for a different setup configuration. On the other hand, the meandering track between the CP and GP couples much stronger at the higher frequencies of the range investigated. This cannot be explained by a simple TL approach

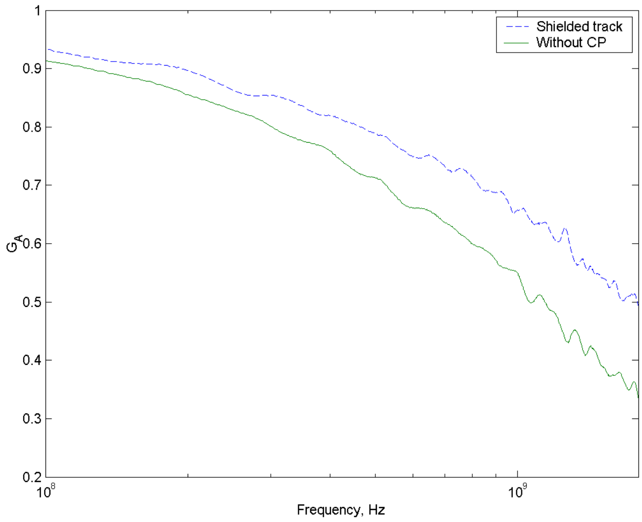

Fig. 5. Available power gain. Shielded track 2 and board without CP. Experimental results

RADIATION

The aforementioned resonant coupling also leads to the increase of radiation from the setup. The results of first experiments show that the nearby CP causes a substantial increase of the radiation at the resonance frequencies. When the straight track 1 is located between GP and CP, the signal measured at 3 m distance is about 40 dB larger compared to the configuration without CP at the frequency of first resonance (around 300 MHz). The less pronounced influence of the proper track termination is also observed. Initial simulations confirm this effect at higher resonance frequencies as well.

CONCLUSIONS

The experimental and numerical results of DM-CM coupling of a 3-layer PCB in so-called precompliance setup have been presented. Good agreement is observed, however calculations required much computational time. The DM circuit with the straight track between CP and GP couples strongly with the CM circuit at lower frequencies, whereas the meandering track between CP and GP couples more at the higher frequencies. This resonant coupling not only affects the signal transmission on the PCB, but also leads to the increase of radiation from the PCB at the resonance frequencies. The introduction of additional load in the CM circuit or shielding around track reduces the coupling and can be used as a preventive measure against possible interference problems.

REFERENCES

[1] F.B.M. van Horck, A.P.J. van Deursen, P.C.T. van der Laan, "Common-Mode Currents Generated by Circuits on a PCB – Measurements and Transmission Line Calculations", *IEEE Trans. Electromagn. Compat.*, vol. 43, no. 4, November 2001, pp. 608 – 617

[2] http://www.cst.de (on-line)