Influence of Adjacent Metal Cover Planes on EMC Filter Structures

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***Abstract*— This paper deals with the influence of metal cover planes close to SMD filter structures. The low-pass structures are used for EMC purpose and work in a frequency range of 100 kHz to 4 GHZ. It will be shown by measurement and simulation that a close cover plane changes the filter behavior significantly in dependency of the filter component used. For low and high impedance filter structures different influences are obtained; a parameter study was carried out in order to derive design rules.**

Key words: EMC, low-pass, conductive planes.

I. INTRODUCTION

The intensive usage of electronic systems in today's technology leads to intensive research on electromagnetic compatibility (EMC). Low-pass filters in the radio frequency (RF) area are applied to the input of electronic devices in order to attenuate the injection or emission of electromagnetic interferences (EMI).

In power electronics, EMC filters are designed to display a specific functional behavior. Unwanted parasitic behavior limits filter attenuation and therefore great effort has been spent to predict this behavior already during the development phase [1],[2]. In the automotive industry Electronic Control Units (ECUs) consist of similar filters that are mainly built with SMD components. EMC measurements are typically performed up to 1GHz or higher. The parasitic behavior of these filters depends heavily on the components used and the Printed Circuit Board (PCB). Up to now, prediction methods are mainly based on rules of thumb or quasi-stationary methods, e.g. the method of partial elements (PEEC). Although the metal housing or planes close to the filter structures are assumed to affect the filter behavior somehow, the housing is mostly neglected during the development phase.

In this paper, an ungrounded metal cover plane in parallel to the PCB is assumed to be part of the housing. The main motivation is to investigate the influence of this plane on different filter structures by measurement and simulation. Simulation will be performed by a full wave field solver [3] and will confirm on the one hand the theoretical explanation of the occurring effects. On the other hand, a parameter study in dependency of filter attenuation and components allows the derivation of design rules.

This paper is structured as follows. In Chapter II the test object and measurement setup are presented. Chapter III provides some theoretical background to explain the occurring effects. In Chapter IV a parameter study is performed.

II. TEST OBJECTS

The setup is shown in Fig. 1 and Fig. 2. It consists of a ground-plane (a) with a thickness of 60 um, a FR4 substrate (c) with a thickness of h = 1.5 mm and a filter component (f). The dimensions of the PCB are a = 160 mm and b = 100 mm. The filter behavior is characterized by means of a Vector Network Analyzer (VNA) where the measured scattering parameter S_{21} corresponds to the insertion filter function in a 50Ω system. To connect the VNA, SMA connectors (h) and a 50Ω microstrip line (e) are necessary.

In general, two different components (f) can be applied. Filter A is a capacitor or low impedance filter component connected to the ground-plane by VIAs (g). Filter B is an inductor or high impedance resistor. For each of both filter components, the influence of an ungrounded metal cover plane (b) with a thickness of 1mm is investigated.

In Fig. 3 measurements are depicted. Filter A consists of four capacitors (each 47nF) and provides an attenuation of 50dB at 1GHz [4]. Filter B consists of a 3kΩ resistor. It has to be noted that the used resistor in filter B is strictly speaking not a low pass, but shows the same parasitic behavior

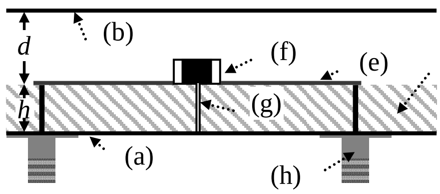

Fig. 1: Setup for the 2port measurement of a discrete filter component (f) with a metal cover plane (b).

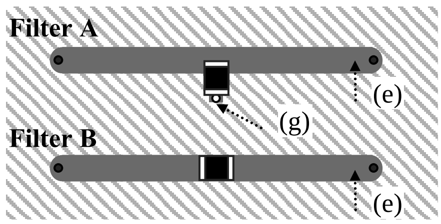

Fig. 2: Setup for the 2port measurement of a discrete filter component A (e.g. capacitor) with a VIA (g) or a filter component B (e.g. inductor).

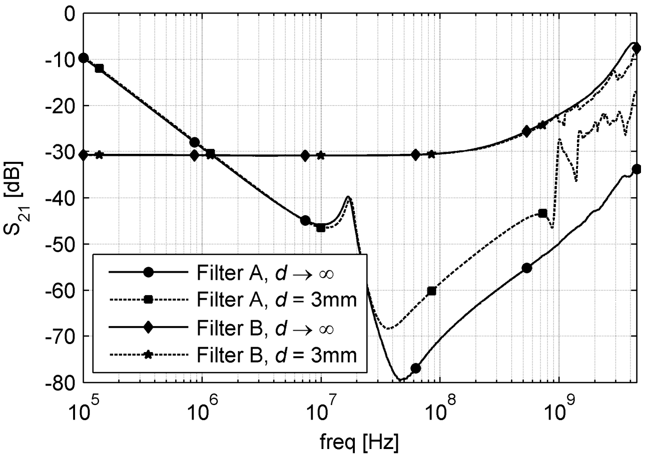

Fig. 3: Measurement of S_{21} of a filter A (4 capacitors) and filter B (3kΩ resistor) without cover plane and with cover plane.

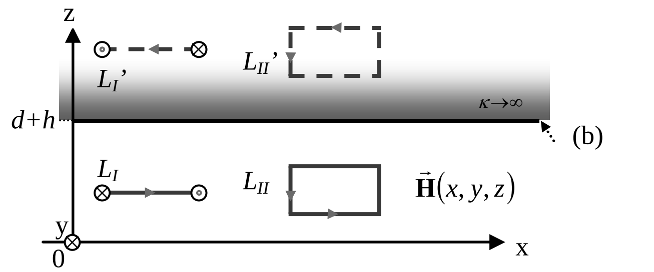

Fig. 4: With the help of the mirror current theory the magnetic field of two different oriented loops L_I and L_{II} in the area z < d+h can be calculated.

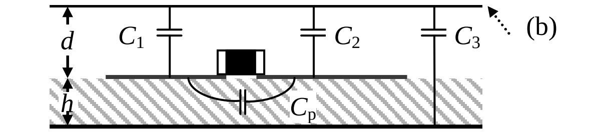

Fig.5: To take a metal cover plane for filter component B into account, stray capacitances to the plane have to be considered.

compared to an inductor. It can be seen for both filter components that a cover plane at a distance d = 3 mm shows influence on the parasitic behavior of the filters. For filter A, the filter attenuation is decreased by 10dB between 50 MHz and 0.9 GHz. Above this area, additional 10dB are observed due to resonances and thus the filter attenuation is reduced significantly. Filter B seems to be more robust and shows only slight deviation due to the resonances starting at 0.9 GHZ. Compared to filter A, the attenuation seems to be increased.

III. UNDERSTANDING

As for 0.9 GHz the used metal cover plane is not longer electrically small (dimensions ≈ λ), one has to distinguish between quasi-stationary and wave propagation effects.

A. Quasi-stationary Effects

At first, the attenuation decrease in filter A is analyzed in detail. As known from [4], the Equivalent Series Inductance (ESL) can be calculated out of the mutual inductance between input and output loop. These loops are created by the traces and the capacitors. In Fig. 4 two loops with different orientations are depicted. The influence of the metal cover plane on the mutual inductance is explained as follows:

- Magnetic field $\vec{\mathbf{H}}$ within z < d+h increases, mutual inductance of similar oriented loops increase.
- Magnetic field $\vec{\mathbf{H}}$ within z < d+h decreases, mutual inductance of similar oriented loops decrease.

Loop L_I is oriented in parallel to the metal cover plane (b). It is well known from literature that the induced eddy current in the metal plane must exhibit the opposite direction. With the aid of the mirror current theory, this loop is mirrored at z = d+h to L_I' and the magnetic field decreases.

The same law is now applied to a loop L_{II} with an orientation perpendicular to the metal plane. The direction of the mirror loop L_{II}' is not changed and the resulting magnetic field is increased. Due to mirroring, for both loops at z = d+h the boundary condition $H_n = H_z = 0$ is fulfilled.

According to [4], the used partial loops to model the filter are perpendicular to the ground-plane and also to the metal cover plane. Thus, the mutual inductance (ESL) of the capacitor filter is increased due to a metal cover plane. For practical simulation it has to noted, that the ground-plane must also be included in the mirroring process. This leads to an infinite number of mirror loops in z-direction and increases the computational effort significantly.

For filter B, the parasitic equivalent circuit is given in Fig 5. At first glance, an increase of the parasitic capacitance C_p is expected due to the parallel stray capacitances C_1 and C_2. However, with small filter structures compared to the cover plane, also the stray capacitance

$$C_3 = \varepsilon_0 \varepsilon_{r,eff} \frac{A}{d+h}, \quad \varepsilon_{r,eff} = \frac{(d+h)\varepsilon_{r,subs}}{d\,\varepsilon_{r,subs}+h} , \qquad (1)$$

has to be taken into account. The variable A is the area of the metal cover plane (b) and $\varepsilon_{r,eff}$ is the effective relative dielectric constant calculated by [5].

The capacitance C_3 leads for higher frequencies to the same electrostatic potential of the cover plane and the ground-plane. Due to the equivalent circuit of the parasitic filter behavior, the filter attenuation is increased. In Fig. 3 no effects at all are observed, which states for very small stray capacitances.

B. Wave Propagation Effects

In order to investigate the resonances in Fig. 3 for higher frequencies (f > 0.9 GHz), measurements with a new test object are performed. This device (Fig. 6) consists of two SMA connectors only. The measured S_{21} is the coupling

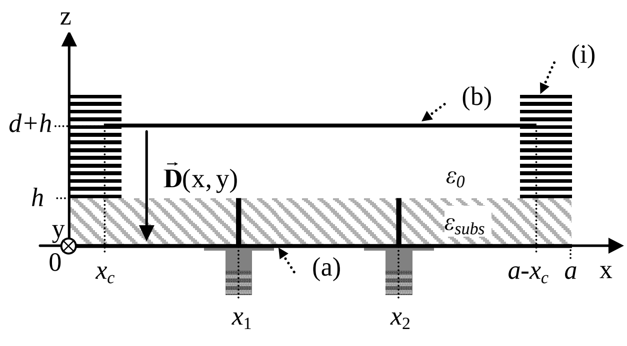

Fig. 6: Exact dimensions of the PCB and the metal cover in the x-z directions. The whole configuration has a length of b in y-direction.

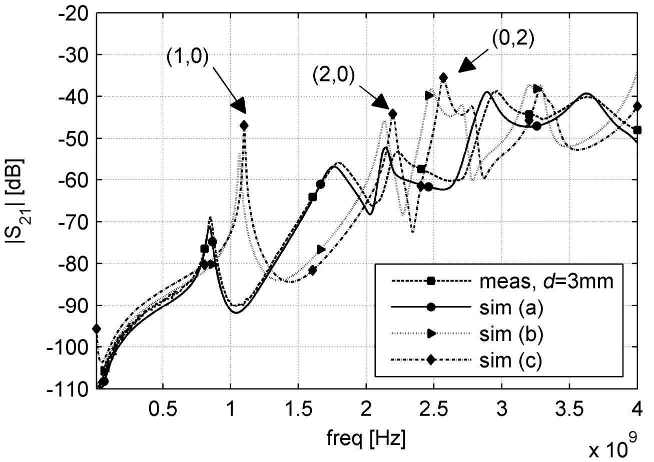

Fig. 7: Measurement and simulation of S_{21} between two SMA-connectors with d = 3 mm. Three different simulation models (a/b/c) are used.

between the two connector stubs as the microstrip line and the filter components are removed. It has to be noted that the cover plane (b) is smaller than the ground-plane (a) and a plastic test fixture (i) is necessary to fix the cover plane. In Fig. 7 the measurement is shown and a full wave simulation by [3] of the whole test setup shows good agreement. The details for the model (a) are given Table I.

In Fig. 8 the electric field in z-direction at the first resonance frequency $f = 0.8$ GHz is analyzed at z = 1 mm. Clearly, it can be seen, that standing wave effects occur and the depicted mode is the first mode (λ/2) in x-direction. Applying the parallel plane system from [6], the resonance frequencies are calculated with help of Eq. (1) to

$$f_{mn} = \frac{1}{2\pi\sqrt{\varepsilon_{r,eff}\ \varepsilon_0\ \mu_0}}\sqrt{\left(\frac{m\pi}{a-2x_c}\right)^2+\left(\frac{n\pi}{b}\right)^2}\ , \qquad (2)$$

where m is related to the modes in x-direction and n is related to the modes in y-direction. The electric field and the steady electric flux density $\vec{\mathbf{D}}$ between the plates are oriented in z-direction. It has to be mentioned that the appearance and intensity of the modes also depends on the position of the excitation. In this paper, the first modes in y-direction are not excited, as both SMA connectors at y = 0.5b are located in the minimum of the first mode n.

With x_c = 22 mm and ε_r = 4.5 (FR4) an effective dielectric constant $\varepsilon_{r,eff}$ of 1.34 is calculated. The first resonance f_{10} is calculated with Eq. (2) to 1.13 GHz, which deviates significantly from the measured resonance f_{10} = 0.8 GHz. This is probably caused by the dielectric influence of fixture (i) and different plane dimensions expressed by x_c.

TABLE I
DIFFERENT SIMULATION MODELS

Model	x_c / mm	a / mm	b / mm	Item (i)	T / min	Boundaries
(a)	22	160	100	yes	23	Open (add space)
(b)	0	116	100	no	13	Open (add space)
(c)	0	116	100	no	2	x,y: H_{tan}=0, z: Open

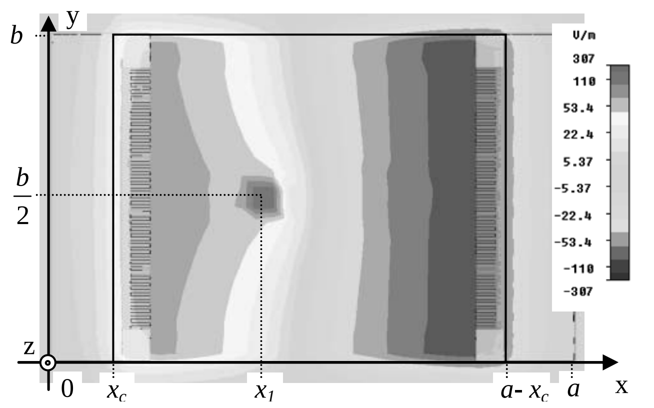

Fig. 8: E-Field plot in z-direction for model (a) at z = 1 mm and a frequency of f =0.8 GHz (first resonance).

To achieve better agreement, the fixture (i) is removed in model (b) and x_c is set to zero. This leads to good agreement of model (b) and Eq. (2). For model (c) the boundary conditions in x and y-direction are changed to $H_{\tan} = 0$, according to [6]. Although fringing field effects are neglected, only small deviations compared to model (b) in Fig. 7 are obtained. In addition, the computing time is reduced considerably. As long as the distance between the planes $d+h$ is small compared to the plane dimensions, these boundary conditions are valid and model (c) is well suited for the parameter study.

IV. PARAMETER STUDY

In this study the influence of a metal cover plane on filter A and B is investigated depending on filter attenuation and distance d of the cover plane. According to Fig 2., full wave models of Filter A and Filter B have been created in [3] with different parasitic behavior.

Fig. 9 shows the attenuation of Filter A. The solid lines (o) show attenuations of 50dB and 30dB at 1 GHz without any cover plane. The solid curve (o) in Fig. 10 illustrates the behavior of filter B with attenuations of 30dB and 25dB at 1 GHz without any cover plane.

The influence of finite cover planes at d = 3 mm and d = 7 mm is demonstrated in both plots with the help of model (c) in Table I. Furthermore, the influence of an infinite cover plane at $d = 2h = 3$ mm is shown (open boundaries) which accords to the quasi-stationary explanations.

For filter A, the filter attenuation is decreased in general. The influence of a cover plane with $d \geq 4h$ can be neglected within 3dB when the attenuation is <30dB at 1GHz. For an attenuation >30dB up to 12dB attenuation decrease (ESL) and up to 20dB attenuation decrease due to the resonance peaks can be observed. A doubled distance can halve the influence on the attenuation of the filter.

For filter B, the filter attenuation is increased in general. For infinite planes, the ground-plane and the cover plane show the same electrostatic potential and thus the filter attenuation is increased due to the stray capacitances. The influence of a metal cover plane with $d \geq 2h$ can be neglected for higher frequencies when the attenuation is <25dB at 1GHz. For an attenuation >25 dB up to 5dB decrease can be observed.

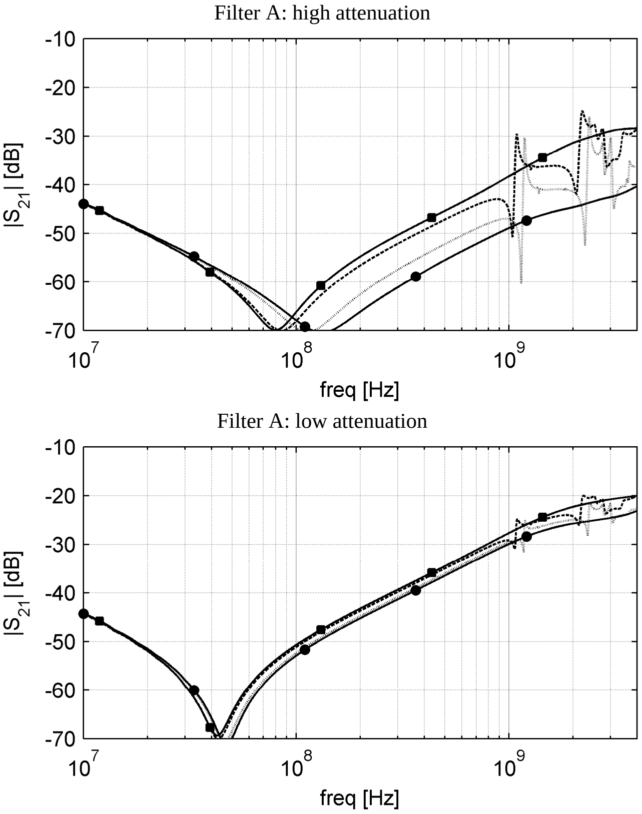

Fig. 9: Two different filter attenuations for Filter A.

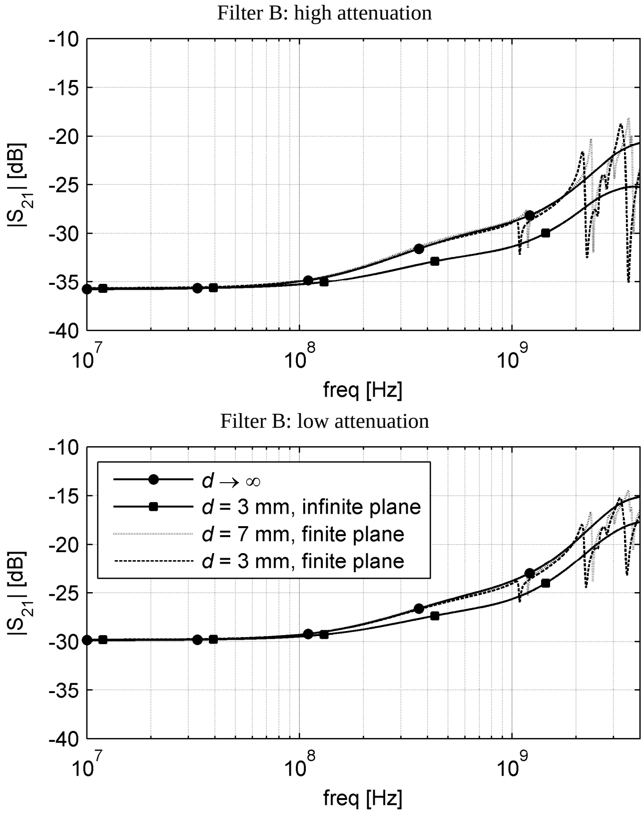

Fig. 10: Two different filter attenuations for Filter B.

Furthermore it is demonstrated, that the infinite plane simulation allows a good behavioral approximation of Filter A. This seems to be also valid for higher frequencies, what recommends the usage of the PEEC method and the mirror theory for simulation purpose.

V. Conclusion

In this paper, the influence of a metal cover plane on the parasitic behavior of SMD low-pass filter structures used in the automotive sector has been investigated.

It has been shown that, in general, filter attenuation under the influence of metal covers is increased for inductive filter structures and is decreased for capacitive filter structures. Several reasons for this behavior have been found. For electrically small dimensions of the cover plane and the PCB, two main effects are responsible for the behavior. As capacitive filter structures can be modeled with loops, the orientation of the loops to the metal plane plays an important role. Due to the given orientation, the magnetic coupling (ESL) between the loops is increased. For inductive filter structures, the stray capacitances of the metal plane to the components and to the ground-plane affect their behavior. At higher frequencies, the metal cover plane and the PCB ground-plane act as parallel plane system with cavity resonances. The filter excites standing wave modes and the attenuation is decreased in general.

A parameter study has demonstrated that the influence of metal covers on simple filter structure with less than 30 dB attenuation at 1 GHz can be neglected when the plane is located not closer than two times of the substrate height to the PCB surface. Future work could target the investigation of the influence of perpendicular plane structures, the dependency of the filter position and the influence of a metal cover plane on combinations of capacitive and inductive filter structures.

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