Paper

Reduction of Q-factor of resonance in power/ground planes of multilayer PCBs by using resistive metal films

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The problem of suppressing resonances of the power/ground planes in a printed circuit board (PCB) by coating high resistive metal films onto the inner sides of copper layers of the two planes has theoretically been studied. The Q-factor that determines the strength of the resonance, is derived from the surface resistance of the composite conductor. To achieve the smallest Q-factor of a given resonance mode, there is an optimum thickness for a fixed conductivity or reversibly an optimum conductivity for a fixed thickness, for the coating metal films. The numerical results for the reduction rate of the quality factor and the input impedance of a test board have demonstrated the effectiveness of the high resistive metal coating to suppress the resonances of the power/ground planes in PCBs.

Keywords: EMI, PCB, resonance of power/ground planes, Q-factor, high resistive metal

1. Introduction

With continuous increase of the clock rate of digital signal system on a printed circuit board (PCB), it seems reasonable to consider the power distribution system on a PCB as a dynamic electromagnetic system in which the propagation effects are important. In fact, the power/ground planes of a multilayer PCB must be considered as a parallel plate waveguiding system [1]. Most of the EM energy associated with the transient process remains captured within the two planes, which actually form an edges-open resonator, or in other words, a patch antenna. It has been known that the resonances of the power/ground planes not only cause radiated emission as EM interference, but also affect ground bounce due to switching noise in a digital system.

As well known, the performances of a resonator or an antenna, such as the input and transfer impedances as well as the radiation efficiency, are greatly affected by its quality factors Q for the operating modes. In our present problem, the Q-factors of the modes are expected as smaller as possible in order to suppress these resonances, that in turn results in reduction of the radiated emission (EMI) at these resonant frequencies from the PCB. We here propose a method to reduce the Q-factors of resonant modes of power/ground planes in PCBs by introducing a larger conductor loss, which can be realized by coating high resistive metal films onto the inner sides of copper layers commonly used for the power/ground planes. In fact, the effectiveness of using high resistive metal to suppress the unintentional resonances in PCBs has recently been confirmed by the measurement results [2], where the high resistive conductor patterns are placed around the "edges" of solid power/ground planes.

2. Basic Theory

To investigate the power/ground plane resonances, a full cavity-mode resonator model [3][4] has been developed to characterize the physical structure as a planar multi-port microwave circuit, and in fact, such a model has already been applied in designing microwave planar circuits [5][6]. The resonator model is an analytical description of impedance matrix (Z-parameters) of an unloaded power/ground plane structure. Since most PCBs are electrically thin, by selecting a Green's function [6] of the 2-D Helmholtz equation with the boundary condition of the second kind (the perfect magnetic conductor sidewalls), it is possible to express the impedance matrix in terms of the eigenfunctions and eigenvalues of the Helmholtz problem. Each"mode"in the Z-parameter expression corresponds to a pole in the impedance. The full-mode representation of the Z-parameters of the power/ground plane structure is an infinite summation of modes, and results in an infinite numbers of poles.

To predict exactly the resonance properties of the power/ground plane structure, it is necessary to take into account the effect of losses due to the power/ground conductors, and dielectric and radiation. The radiation loss is usually small enough to be ignored, compared to the dielectric and conductor losses in normal power/ground plane geometries. The dielectric loss naturally appears in the imaginary part of the dielectric constant, while the conductor loss should be incorporated by considering a change in the two-dimensional (2-D) transverse wavenumber due to finite conductivity. Furthermore, a summation formula is applied to the mode-expansion series to achieve a rapid calculation

for the elements of Z-matrix. In this way, an improved expression for the impedance matrix has recently been developed by us [7]. Each element of the matrix, Z_{ij}, represents the transfer impedance from port i to port j, and for $i = j$, Z_{ii}, represents the input impedance of the cavity at the port i. For a rectangular PCB structure with length a and width b, the final expression for the Z-matrix elements is given as follows:

$$Z_{ij} = \sum_{n=0}^{\infty} \frac{\omega\mu_d ha}{\mathrm{j}\,2b} C_n \cos(k_{yn}y_i)\cos(k_{yn}y_j) \times \mathrm{sinc}^2(k_{yn}\mathrm{w}) \frac{[\cos(\alpha_n x_-) + \cos(\alpha_n x_+)]}{\alpha_n \sin\alpha_n} \quad (1)$$

where $\mathrm{sinc}(x) = \sin(x)/x$, $k_{yn} = n\pi/b$, $\alpha_n = a\sqrt{\kappa^2 - k_{yn}^2}$, $x_\pm = 1-(x_i \pm x_j)/a$, x_i, x_j, y_i and y_j are the coordinates of the center of the ith and jth ports in the x- and y-directions, respectively, w is much less than the wavelengths of interest and represents the port half width (we have assumed for simplicity that the port sizes in the x- and y-directions for the ith and jth ports are the same), h is the dielectric thickness between the power/ground planes, ω is radian frequency, μ_d is the permeability of the dielectric, and $\mathrm{j} = \sqrt{-1}$. The constant $C_n = 1$ if $n = 0$, and $C_n = 2$ if $n \neq 0$. For lossy conductor, the complex transverse wavenumber κ is approximately obtained as [7] (see also the appendix)

$$\kappa^2 = \omega^2\mu_d\varepsilon_d - \mathrm{j}2\omega\varepsilon_d Z_s/h \quad (2)$$

where the permeability and permittivity of the dielectric are denoted by μ_d and $\varepsilon_d = \varepsilon_0\varepsilon_r$, respectively, and ε_r stands for the relative dielectric constant. Z_s represents the surface impedance of the power/ground conductors. For a metal layer whose thickness is much larger than its skin-depth at the frequencies of interest, the internal surface impedance for a unit length and unit width is given as [8],

$$Z_s = (1+\mathrm{j})\,R_s,\ R_s = \frac{1}{\delta_s\,\sigma_c},\ \delta_s = \sqrt{\frac{2}{\omega\,\mu_c\,\sigma_c}} \quad (3)$$

where R_s represents the surface resistivity of the metal layer (conductor), σ_c is the conductivity, and δ_s the skin depth of field penetration into the metal.

At cavity resonance frequencies, the magnitudes of the impedance are related to the quality factors (Q-factors) of the resonances. Higher losses in the cavity result in lower quality factors, which in turn lead to lower power bus impedances at the resonance frequencies. As mentioned above, real power/ground plane structures exhibit loss due to the finite resistance of the conductor walls, dielectric loss, radiation loss, and losses due to surface waves induced on the outer surface of the conductors. Radiation loss and surface wave losses are usually small enough to be ignored. Formulas for conductive loss and dielectric loss are well documented [9][10]. For very thin dielectric layer between the power and ground planes, the quality factors due to conductive loss in the top and bottom planes and the dielectric loss are given by [9],

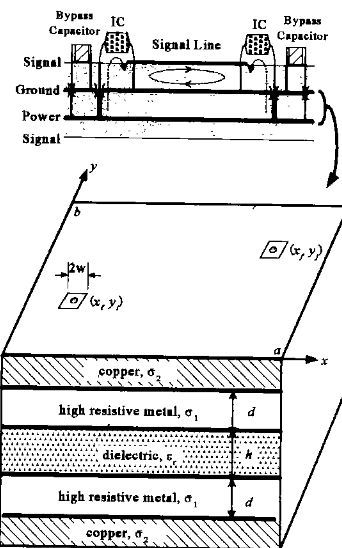

Fig. 1. Geometry of the problem.

$$Q_c = \frac{h}{\delta_s} = \frac{\omega\mu_c h}{2R_s},\quad Q_d = \frac{1}{\tan\delta} \quad (4)$$

where $\tan\delta$ is the loss tangent of the dielectric. The overall quality factor can then be approximated by,

$$\frac{1}{Q} = \frac{1}{Q_c} + \frac{1}{Q_d} \quad (5)$$

Eq.(4) shows that Q_d can be reduced by increasing the loss tangent of the dielectric and Q_c by reducing the separation distance between the two conductor planes (a thinner dielectric layer) and/or increasing the surface resistivity of the metal planes. In a recent study called NCMS Embedded Capacitance Project, Hubing *et.al.* [10] have confirmed that the board resonances can be completely damped by the conductive loss in the planes when the dielectric spacing between the two planes is on the order of a skin depth in the conductor (for example, a few microns for copper, in this case $Q_c \sim 1$). However, it is not all the cases that a few microns spacing between the planes could be adapted. Here we propose an alternative to reduce Q_c by increasing the surface resistivity of the conductors for the power/ground planes, which is realized by coating thin films of high resistive metal (conductivity σ_1) onto the inner sides of copper (conductivity σ_2, $\sigma_2 > \sigma_1$) layers of the power/ground planes, as shown in Fig.1.

0 Based on electromagnetic theory [8], the surface impedance Z_s for the composite conductor (the high resistive metallic film plus the copper layer) is given as

$$\frac{Z_s}{R_{s1}} = (1+\mathrm{j})\left[\frac{\sinh x + (R_{s2}/R_{s1})\cosh x}{\cosh x + (R_{s2}/R_{s1})\sinh x}\right] \quad (6)$$

where $x = (1+\mathrm{j})(d/\delta_{s1})$, d is the thickness and δ_{s1} the skin depth of the coating material. R_{s1} and R_{s2} ($R_{s1} > R_{s2}$) are the surface resistivities of the coating (high resistive metal) and coated (copper) materials, respectively. From Eq.(4), the reduction rate of the quality factors due to the conductive loss with and without

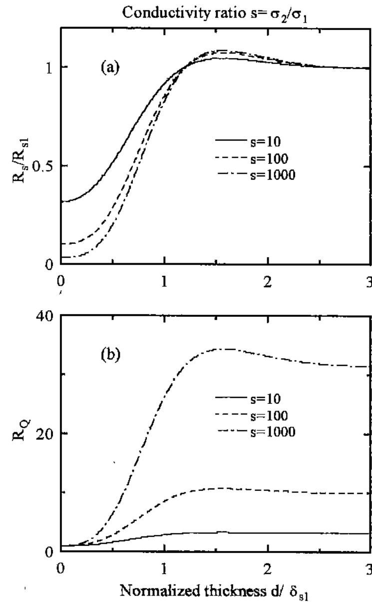

Fig. 2. (a) The normalized resistivity $R_s/R_{s1} = \Re e\{Z_s/R_{s1}\}$, and (b) the Q-factor reduction rate R_Q, as a function of the normalized thickness d/δ_{s1}, with the conductivity ratio $s = \sigma_2/\sigma_1$ as a parameter.

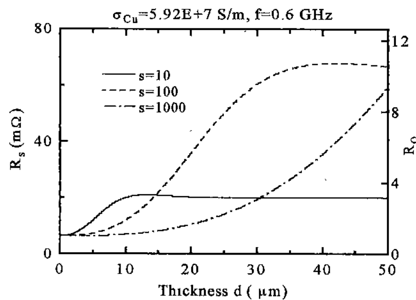

Fig. 3. R_s and R_Q as a function of the thickness d, with the conductivity ratio s as a parameter.

the high resistive metallic films is easily obtained as

$$R_Q = \frac{Q_{c0}}{Q_c} = \Re e\left\{\frac{Z_s}{R_{s2}}\right\} \cdots\cdots\cdots\cdots\cdots\cdots (7)$$

According to Eq.(5), it is obvious that the total Q is mainly governed by the smaller one of Q_c and Q_d. Thus, for the reduction of Q_c due to large loss of the high resistive metal to be really effective, it is necessary that Q_{c0} due to the loss of only the copper should be comparable or less than Q_d due to the dielectric loss.

3. Results and Discussion

To investigate how the high resistive metallic films work for reduction of Q-factors of the power/ground resonances, numerical calculations have been performed. Curves of resistivity ratio of the composite conductor to the coating high resistive material and the corresponding reduction rate for Q_c are plotted in Fig.2 as a function of the coating thickness d normalized to the skin-depth δ_{s1} of the coating material, with the conductivity ratio of the coated material to the coating material $s = \sigma_2/\sigma_1$ as a parameter. There is an optimum value for the normalized thickness d/δ_{s1} (~ 1.55) to achieve a maximum resistivity $R_s = \Re e\{Z_s\}$ for the composite conductor, and the larger the ratio s is, the larger the reduction rate $R_Q = Q_{c0}/Q_c$ can be achieved when the coating thickness is greater than its skin-depth. However, a larger ratio s means a large skin-depth of the coating material, and usually the permissible thickness for the power/ground planes is around 30μm. Fig.3 shows the composite resistivity R_s and the reduction rate R_Q as a function of directly the thickness d for a given frequency, then we can find that there is an optimum value of s to obtain a maximum reduction rate $(R_Q)_{max}$. This is more clearly observed from Fig.4, where R_s and R_Q are plotted as a function of s with the thickness d or the frequency as a parameter. The larger the thickness d is for a given frequency f or the larger the frequency f is for a given thickness d, the larger the conductivity ratio s can be used for the materials, which in turn leads to a greater reduction rate R_Q achieved.

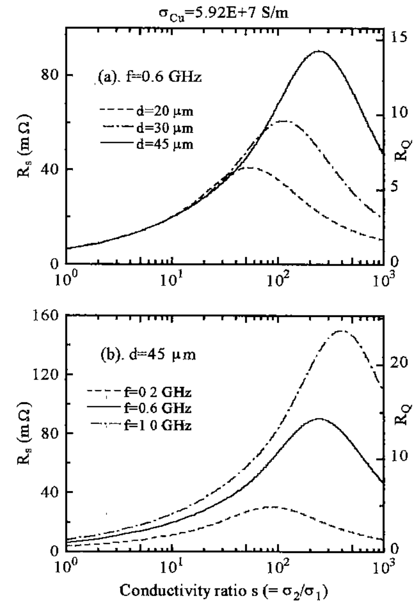

Fig. 4. R_s and R_Q as a function of the conductivity ratio s, with (a) the thickness d and (b) the frequency f as a parameter.

The above results have shown how the coating high resistive metal films affect the quality factor Q_c due to conductive loss of the power/ground planes. Now we will show the effect of the coating films on the input impedance Z_{in} at a feed point on the power/ground planes of a PCB. Calculations are made by Eq.(1) for the impedance element Z_{ij} with $i = j$. The sizes of the board are taken to be a = 80mm and b = 56mm, and the feed point is located at (10, 10).

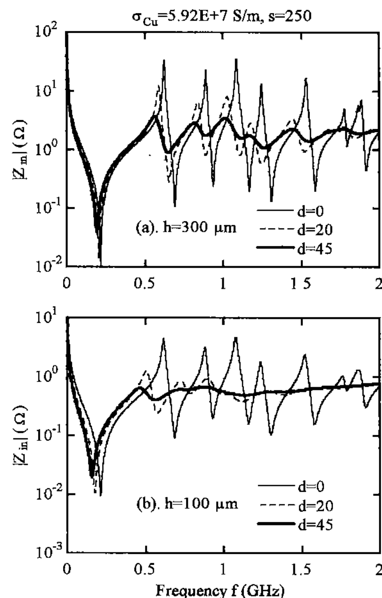

Fig. 5. The input impedance Z_{in} as a function of the frequency f with the thickness d as a parameter, for the power/ground plane spacing (a) $h = 300$ and (b) $h = 100$ μm.

The relative dielectric constant of the dielectric material between the power/ground planes is assumed to be $\varepsilon_r = 9.0\,(1 + \mathrm{j}\,0.002)$, so the quality factor due to the dielectric loss is $Q_d = 500$. The coated good conductor is assumed to be copper whose conductivity is $\sigma_2 = \sigma_{Cu} = 5.92 \times 10^7$ S/m. At the first resonance frequency $f = 0.625$ GHz of the board, the quality factor due to copper loss is $Q_{c0} \simeq 120$ and 40 for the spacing between the two planes $h = 300$ and 100 μm, respectively. The input impedance $|Z_{in}|$ for these two values of the spacing is plotted in Fig.5(a) and 5(b) as a function of frequency with the coating thickness $d = 0$, 20 and 45 μm, respectively. The conductivity ratio s of copper to the high resistive metal has been taken as 250 for all the cases, which corresponds to its optimum value at the first resonance frequency of the board as the coating thickness being $d = 45$ μm. It is observed that, compared to the case without the high resistive metal coating ($d = 0$), a reduction of about 20 dB at the peak of the impedance around the first resonance frequency can be achieved for both the two spacing cases when the coating thickness $d = 45$ μm, corresponding to a reduction rate of $R_Q \simeq \sqrt{s} \simeq 15$. The reduction at the impedance peak for a higher resonance frequency should be more remarkable, since R_Q becomes larger for a higher frequency as shown in Fig.4(b). Finally, the impedance resonances have almost been damped for the case of $h = 100$ μm and $d = 45$ μm because Q_c has been reduced to a small value of about 3 even though at the first resonance frequency.

4. Conclusions

In conclusion, based on a closed-form expression for the impedance Z-matrix elements of the power/ground planes in multilayer PCBs recently developed by us, the problem of suppressing resonances of the power/ground planes in a PCB by coating high resistive metal films onto the inner sides of copper layers of the two planes has theoretically been studied. The Q-factor that determines the strength of the resonance, is derived from the surface resistance of the composite conductor. It is seen that the composite conductor becomes about as good, or as bad, as though the coating were of infinite depth when the coating thickness is greater than the skin-depth of the coating material. This fact means that for the coating films there is an optimum thickness for a fixed conductivity or reversibly an optimum conductivity for a fixed thickness to achieve the smallest Q-factor of a given resonance mode, due to the conductive loss. In the practical application, the thickness of the films has to be designed as large as possible, and then the optimum conductivity should be determined at the first resonance frequency of the power/ground planes, because the problems of EMI and ground bounce occurred at the first resonance frequency are usually more serious than those at higher resonance frequencies. Moreover, the suppressing effect on the resonance due to the conductive loss becomes stronger as the frequency goes higher, since the surface resistivity of the composite conductor increases due to the smaller skin-depth. The numerical results for the reduction rate of the quality factor and the input impedance of a test board have demonstrated the effectiveness of the high resistive metal coating to suppress the resonances of the power/ground planes in PCBs.

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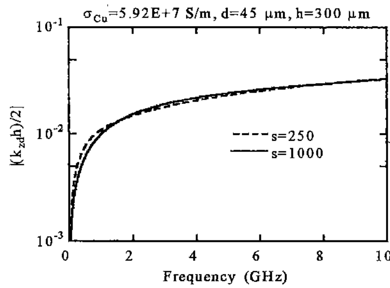

app. Fig. 1. $|(k_{zd}h)/2|$ as a function of the frequency f with the conductivity ratio s as a parameter, for the power/ground plane spacing $h = 300$ and the coating thickness $d = 45$ μm.

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Appendix

1. Justification of Eq.(2)

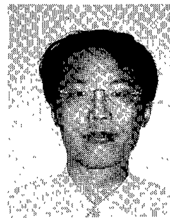
As shown in [7], Eq.(2) is obtained from the dispersion equation

$$\tan(\frac{k_{zd}h}{2}) = \frac{j\omega\varepsilon_d Z_s}{k_{zd}} \quad \cdots\cdots\cdots\cdots\cdots\cdots\cdots \text{(A1)}$$

by making use of the approximation:

$$\tan(\frac{k_{zd}h}{2}) \simeq \frac{k_{zd}h}{2} \quad \cdots\cdots\cdots\cdots\cdots\cdots\cdots\cdots \text{(A2)}$$

where $k_{zd}^2 = k_d^2 - \kappa^2$ with $k_d = \omega\sqrt{\mu_d\varepsilon_d}$ being the wavenumber in the dielectric. Mathematically, the approximation $\tan(z) \simeq z$ is good enough if the condition $|z| < 0.1$ is satisfied. Thus, to validate that the approximation Eq.(A2) is reasonable, we have to calculate the value of $|k_{zd}h|$ with the use of Eq.(2). The calculated result of $|(k_{zd}h)/2|$ is plotted in app.Fig.1 as a function of the frequency with the conductivity ratio s being a parameter, for the case of that the power/ground plane spacing is $h = 300$ μm and the coating thickness of the high resistive metal is $d = 45$ μm , which is the worst case that the approximation may be broken. However, it can be found from app.Fig.1 that the approximation Eq.(A2) and then Eq.(2) are justifiable to be used, even for the frequency up to 10 GHz.



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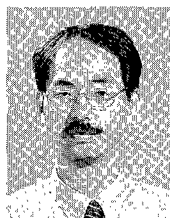

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