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Radiated EMI from Power Converters

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Abstract: With the continuous increase of switching frequency together with the ongoing trend to higher complexity and functionality, power converters as a part of electronic systems have raised more and more electromagnetic energy pollution to the local system environment. In the same time, stringent demands are imposed on the designers of new circuits that electromagnetic interference (EMI) has to be suppressed at its source before it is allowed to propagate into other circuits and systems. In this paper, the authors present a full-wave numerical method for calculation and simulation of electromagnetic field radiated by power converter circuitry. The main objective is to analyze the layout geometry in order to obtain competitive PCB layout that will enable suitably attenuated level of the radiated electric field to safe level. By this it would be possible to ensure reliable operation of the sensitive electronic components in the proximity.

Keywords: Power Converters, EMI, Numerical modeling.

1 Introduction

With the increase of switching frequency of electronic switches, power converters have raised more and more electromagnetic energy pollution to the local environment. Electromagnetic compatibility (EMC) and electromagnetic interference (EMI) norms applied to power converters have objective to reduce conducted and radiated perturbances. Radiated EMI appears in the form of electromagnetic waves directly from the circuitry and its interface leads. The circuitry and its interface leads can liken themselves to a transmitting antenna for this radiated EMI emission. This emission is generally measured at much higher frequencies than their conducted counterparts, namely beyond 30 MHz up to several GHz [1].

The computation and analyzing the electromagnetic field distribution is the first step which will help the constructors to implement appropriate solutions in the conception process of the electric systems. Considering the high-switching

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frequency, recently the authors have developed a full-wave numerical moment method approach that is a first-stage simulation of the power converter model in order to calculate the distribution of the radiated electric field. The model is validated by comparison with commercial electromagnetic software [7].

The main objective of this paper is to present the theoretical basis of the mathematical model that is used for flexible analysis of the radiated electric field due to various configuration of the power converter circuitry. Also, some of numerous simulations of various power converter layout will be presented. The number of frequency samples required to accurately represent the spectrum across the bandwidth of interest is dynamically chosen.

2 Mathematical Model

The power converter circuitry and its interface leads behave like a transmitting antenna which radiates directly into the immediate surroundings. Considering the simplified geometry of a DC/DC converter when switch is ON the circuitry is shown on Fig. 1. Here, V_S is the source voltage, R_S is its internal resistor and R_L represents the load.

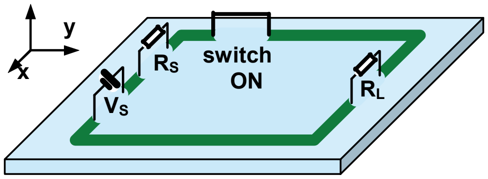


Fig. 1 - *The geometry of a simple converter (switch is ON).*

Considering the high switching frequency, from 30 MHz up to several GHz and the corresponding wavelength, λ, the mathematical model is based on rigorous formulations derived from the full set of the Maxwell's equations. In our research we use the theoretical background of microstrip antenna analysis [3] in frequency domain, involving Green's functions due to elementary HED source embedded in multilayer structure [5].

In our model, the board is represented as a dielectric layer with thickness, *d*, corresponding permittivity, $\varepsilon = \varepsilon_0 \varepsilon_r$ and permeability of air, μ_0. The currents that flow in the conducting PCB traces are assumed to have only the component in the direction of the traces (since the wavelength is much more than the trace width $\lambda ? w$). Following the thin-wire approximation of the PCB trace [2], for the final formulation of the mathematical model, the current is assumed to flow along the wire axis (for example: PCB trace width $w \sim 0.25$ mm is equivalent to radius $R \sim 0.1$ mm), as shown on Fig. 2.

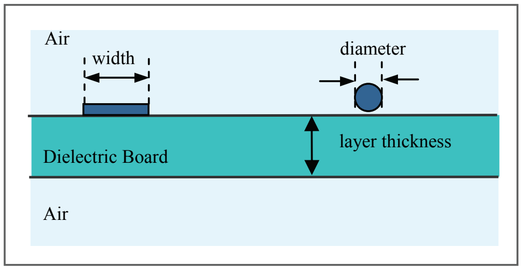


Fig. 2 – *Transformation of PCB into equivalent thin wire.*

The mathematical model is based on the mixed potential integral equation (MPIE) [3] for thin-wire structures. The physical model of the circuitry is based on the fictitious segmentation of the leads into straight tubular segments. Segment length is defined by following the thin-wire conditions in accordance with the wavelength and radius (approximately λ/10 - λ/20). The axial current along thin-wires is approximated by a linear combination of overlapping triangle dipole expansions functions positioned along two neighbor segments [4].

Following the method of moments, the currents in segments are computed by solving the following matrix equation:

$$[Z][I]=[V_S], \tag{1}$$

where $[I]$ is matrix-column of the unknown currents, $[Z]$ is the impedance matrix and $[V_S]$ is matrix-column of the voltage energization.

Elements of the matrix $[Z]$ are mutual impedances z_{mn} calculated between pair of segments: source segment *n* with current I_n, and test segment *m*:

$$z_{mn}=\frac{v_{mn}}{I_n}=\frac{1}{I_n}\int_{\Delta\ell_m}\boldsymbol{W}_m\cdot(-\boldsymbol{E}_{n\text{tang}})d\ell=\frac{1}{I_n}\int_{\Delta\ell_m}\boldsymbol{W}_m\cdot(j\omega\boldsymbol{A}_n+\nabla\phi_n)_{\text{tang}}d\ell\,. \tag{2}$$

The boundary conditions are tested by using the triangular weighting functions [4]. In the above equation, the mixed potential integral equation of the electric field vector due to elementary source horizontal electric dipole (HED) is used by following the basic idea originally developed for analysis of multi-layer microstrip antennas:

$$\boldsymbol{E}=-j\omega\mathbf{A}-\nabla\phi\ , \tag{3}$$

$$\boldsymbol{A}=\int_{wire}\underline{\boldsymbol{G}}_A\boldsymbol{I}dx'\ \text{ and} \tag{4}$$

$$\phi=\int_{wire}G_\phi q dx'\ \text{ with }\ j\omega q=-\nabla'\boldsymbol{I}\,. \tag{5}$$

Here, $\underline{\mathbf{G}}_A$ is the dyadic Green's function for the magnetic vector potential, and G_ϕ is the scalar potential Green's function due to elementary horizontal electric dipole (HED) source embedded in stratified medium. The traditional form for the vector potential dyadic Green's function is used [5]:

$$\underline{\boldsymbol{G}}_A = (\boldsymbol{xx} + \boldsymbol{yy})G_{xx}^A + \boldsymbol{zx}G_{zx}^A + \boldsymbol{zy}G_{zy}^A \, . \tag{6}$$

The Green's functions are obtained by numerical integration of the following Sommerfeld-type integrals:

$$G_{A,\phi}(\boldsymbol{r},\boldsymbol{r}') = \frac{1}{2\pi}\int_0^\infty \tilde{G}_{A,\phi}(k_\rho; z, z') J_0(k_\rho \rho) k_\rho dk_\rho \, . \tag{7}$$

Here, $\tilde{G}_A$ and $\tilde{G}_\phi$ are corresponding spectral domain Green's functions for the vector and scalar potentials due to HED over dielectric substrate [6]:

$$\tilde{G}_{xx}^A = \frac{\mu_0}{2jk_{z0}}[e^{-jk_{z0}(z-z')} + \tilde{R}_{01TE} e^{-jk_{z0}(z+z')}] \, , \tag{8}$$

$$\tilde{G}_{zx}^A = \frac{-\mu_0}{2k_\rho^2}[k_x(\tilde{R}_{01TE} + \tilde{R}_{01TM})e^{-jk_{z0}(z+z')}] \, , \tag{9}$$

$$\tilde{G}_\varphi = \frac{1}{2\varepsilon_0}[\frac{e^{-jk_{z0}(z-z')}}{jk_{z0}} + \frac{k_{z0}^2 \tilde{R}_{01TE} + k_0^2 \tilde{R}_{01TM}}{jk_{z0}k_\rho^2} e^{-jk_{z0}(z+z')}] \, , \tag{10}$$

$$\tilde{R}_{01TE} = \frac{R_{01TE} + R_{10TE} e^{-jk_{z1}2d}}{1 + R_{01TE} R_{10TE} e^{-jk_{z1}2d}} \, , \quad \tilde{R}_{01TM} = \frac{R_{01TM} + R_{10TM} e^{-jk_{z1}2d}}{1 + R_{01TM} R_{10TM} e^{-jk_{z1}2d}} \, , \tag{11}$$

$$R_{01TE} = -R_{10TE} = \frac{k_{z0} - k_{z1}}{k_{z0} + k_{z1}} \, , \quad R_{01TM} = -R_{10TM} = \frac{k_1^2 k_{z0} - k_0^2 k_{z1}}{k_1^2 k_{z0} + k_0^2 k_{z1}} \, , \tag{12}$$

with $k_{zi}^2 = k_i^2 - k_\rho^2$ for $i = 1,0$. Here, $k_0 = \omega\sqrt{\varepsilon_0 \mu_0}$ and $k_1 = \sqrt{\varepsilon_r} k_0$ are corresponding propagation constants of air and dielectric substrate respectively.

2.1 Calculation of the electric field vector

Once the currents $I_1, I_2, \ldots, I_N$ are computed by solving the system of linear equations (1), the electric field vector $\boldsymbol{E}$ in arbitrary point above the loop may be calculated by using the superposition principle:

$$\boldsymbol{E} = \sum_{n=1}^{N} \boldsymbol{E}_n = \sum_{n=1}^{N} (-j\omega \boldsymbol{A}_n - \nabla \phi_n) \, , \tag{13}$$

where

$$\boldsymbol{E}_n = -j\omega \int_{\Delta\ell_n} \bar{\boldsymbol{G}}_A \cdot \boldsymbol{I}_{\ell n} d\ell' + \nabla \frac{1}{j\omega} \int_{\Delta\ell_n} \frac{dI_{\ell n}}{d\ell'} G_{\phi} d\ell' \, . \tag{14}$$

When the solution of above equation due to all source current segments is evaluated, then a very accurate 3-D profile of the radiated electric field can be obtained.

3 Numerical Results

In this paper the authors present some numerical results that treat the problem of how various layout of the power converter circuitry affect the distribution of the electric field in frequency domain.

It is assumed that the circuit perimeter is fixed at $L = 40\,\text{cm}$, and the equivalent radius of the PCB is $R = 0.5\,\text{mm}$. Three various dimensions of the circuits are tested: (a) square circuitry (10 cm x 10 cm), (b) rectangle circuitry (7.5 cm x 12.5 cm), and (c) rectangle circuitry (5 cm x 15 cm), as shown on Fig. 2. The dielectric layer is assumed with relative permittivity $\varepsilon_{\text{r}} = 4.7$ and thickness $d = 2.5\,\text{mm}$ The amplitude of the harmonic source generator is $\text{V}_{\text{S}} = 1\,\text{V}$ with internal resistor $\text{R}_{\text{S}} = 50\,\Omega$ placed at the center of: (a) one side of the square circuit, and (b-c) centre of the shortest side of the rectangle circuit. The load resistor is $\text{R}_{\text{L}} = 100\,\Omega$ placed in the centre of the opposite side.

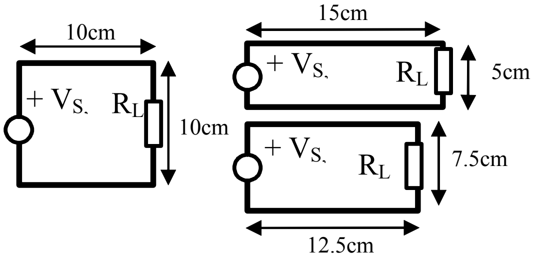
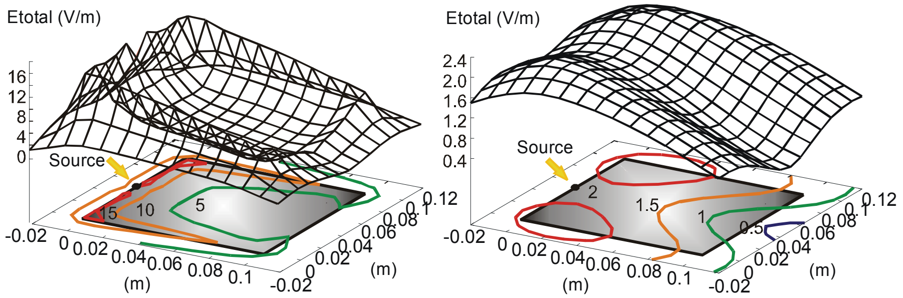


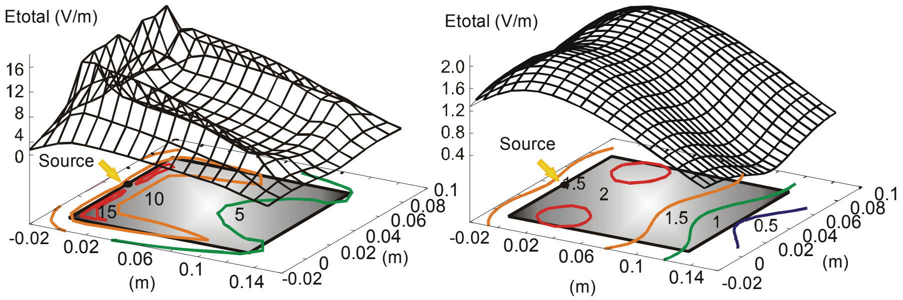
Fig. 3 - *The analyzed grounding configurations in two-layer soil.*

Fig. 3: (a), (b) and (c) show the distribution of the total electric field at points 1 cm and 5 cm above the configurations under consideration at 500 MHz (the corresponding wavelength in the air is ~0.6 m). We are focused on the calculation of the electric field at observation points 1 cm and 5 cm above the circuits.

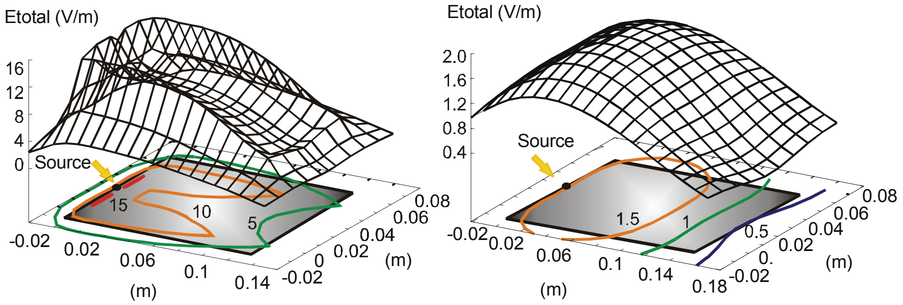
Fig. 4: (a), (b) and (c) show the distribution of the total electric field at points 1 cm and 5 cm above the circuitry at 1 GHz (the corresponding wavelength in the air is ~0.3 m).



(a) 10 cm x 10 cm circuit.

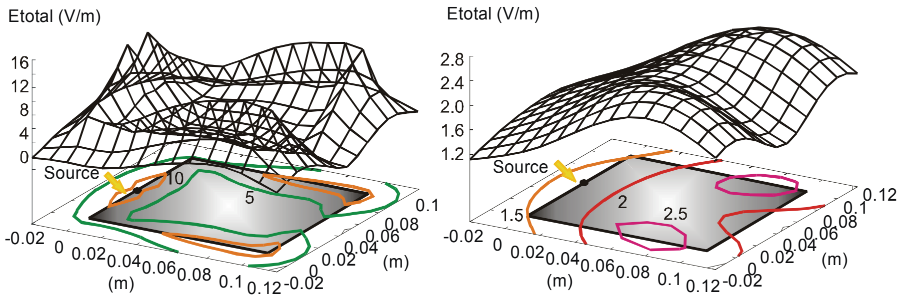


(b) 7.5 cm x 12.5 cm circuit.

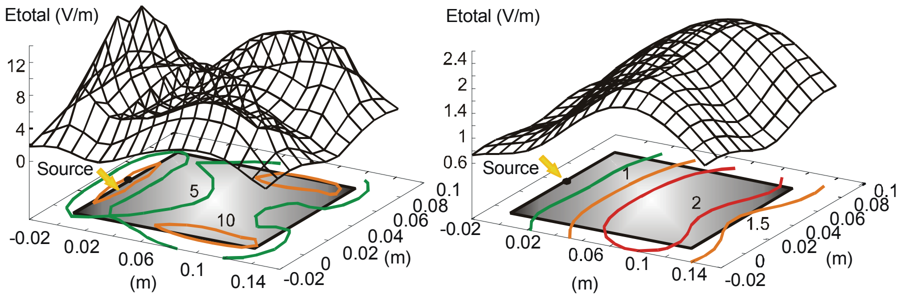


(c) 5 cm x 10 cm circuit.

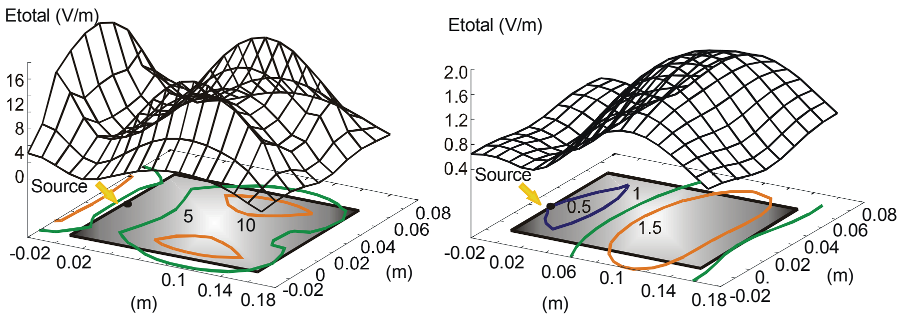
Fig. 3 – *Electric field distribution at 500 MHz calculated at observation points at 1 cm (left) and 5 cm (right) above the circuits under consideration.*



(a) 10 cm x 10 cm circuit.



(b) 7.5 cm x 12.5 cm circuit.



(c) 5 cm x 15 cm circuit.

Fig. 4 – *Electric field distribution at 1 GHz calculated at observation points at 1 cm (left) and 5 cm (right) above the circuits under consideration.*

5 Conclusion

In this paper, we applied frequency domain full-wave approach to calculate the radiated electric field by of power converters. The main objective is to analyze the layout geometry in order to obtain competitive PCB layout that will enable suitably attenuated level of the radiated electric field to safe level. Also, by this it would be possible to investigate the location of “safe islands” where the design will ensure reliable operation of the sensitive electronic components. In the paper, some of numerous investigations of the influence of the circuit geometry on the distribution of the electric field in proximity are presented.

6 References

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