THEORETICAL MOMENT METHOD FORMULATION OF AN MTL RESONATOR IN A HIGH FIELD MRI SYSTEM

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Introduction

Several works have been published on the use of the microstrip resonators in MRI systems [1]-[5]. Models based on quasi-TEM approximation and matrix capacitance and inductance calculation are applied for the analysis of microstrip transmission line (MTL) coils [3], [5]. With the increase of field strengths, rigorous computational electromagnetic techniques become an essential tool for designing and evaluating RF MTL coils. In this work, a full-wave model of a multilayer multiconductor microstrip patch antenna M3PA/MTL is developed based on the electric field integral equation (EFIE) and the spectral dyadic Green's function (SDGF) concepts. The problem of the resonance frequency, the electric current distribution on the patch conductors and the field distribution in the structure is rigorously formulated using the spectral domain integral equation method. The cross section of the cylindrical M3PA/MTL to be analyzed is shown in Fig. 1. A brief outline of the theory used to compute the results for the cylindrical M3PA/MTL is given in the following paragraph.

Theoretical formulation

The tangential field components in the spectral domain at $\rho = \rho_i$ and $\rho = \rho_{i-1}$, are related by a (4x4) transition matrix $\bar{\mathbf{T}}$ whose elements are expressed in terms of Bessel functions of the first and second kind and their derivatives. Next, one imposes the continuity conditions of the tangential electromagnetic field over the interfaces and the jump condition onto the interface conductors $\rho = \rho_M$. Thus, a relation is obtained between the tangential electric field $\tilde{\mathbf{E}}_{\nu s}$ at $\rho = \rho_M$ and the electric surface current $\tilde{\mathbf{K}}_\nu$ on the patch conductors as $\tilde{\mathbf{E}}_{\nu s} = \bar{\mathbf{G}}_\nu \cdot \tilde{\mathbf{K}}_\nu$ where $\bar{\mathbf{G}}_\nu$ is the spectral dyadic Green's function given by $\bar{\mathbf{G}}_\nu = (\bar{\boldsymbol{\Gamma}}_<^{EE} + \bar{\boldsymbol{\Gamma}}_<^{EH} \cdot \bar{\mathbf{M}}) \cdot (\bar{\boldsymbol{\Gamma}}^{EE} + \bar{\boldsymbol{\Gamma}}^{EH} \cdot \bar{\mathbf{M}})^{-1} \cdot \bar{\boldsymbol{\Gamma}}_>^{EH} \cdot \bar{\mathbf{Q}}$ in which $\bar{\boldsymbol{\Gamma}}_{<(>)} = \prod_{i=M(P)}^{2(M+1)} \bar{\mathbf{T}}_i$, $\bar{\boldsymbol{\Gamma}} = \bar{\boldsymbol{\Gamma}}_> \cdot \bar{\boldsymbol{\Gamma}}_<$ and $\bar{\mathbf{M}}$ is a (2x2) matrix linking tangential electric and magnetic field components in region 1. $\bar{\mathbf{Q}}$ is a (2x2) fixed matrix. Next, impose the two boundary conditions to obtain two vector integral equations with one from each condition; first, the tangential electric field vanishes on the patches surface and secondly, no electric surface current exists outside of the patches; $\mathbf{E}_s(\phi, z) = \frac{1}{2\pi} \sum_{\nu=-\infty}^{\infty} e^{i\nu\varphi} \int_{-\infty}^{\infty} dk_z e^{ik_z z} \bar{\mathbf{G}}_\nu(k_z) \cdot \tilde{\mathbf{K}}_\nu(k_z) = \mathbf{0}$ on the patches and $\mathbf{J}_s(\phi, z) = \frac{1}{2\pi} \sum_{\nu=-\infty}^{\infty} e^{i\nu\varphi} \int_{-\infty}^{\infty} dk_z e^{ik_z z} \tilde{\mathbf{K}}_\nu(k_z) = \mathbf{0}$ outside the patches. The Galerkin moment method is used to solve the above integral equations and we obtain a homogeneous matrix equation $\bar{\mathbf{B}} \cdot \mathbf{a} = \mathbf{0}$ for the unknown current expansion coefficients **a**. The *ij*th element of the matrix $\bar{\mathbf{B}}$ is given by $B_{ij} = \sum_{\nu=-\infty}^{\infty} e^{-i\nu(\phi_p - \phi_q)} \int_{-\infty}^{\infty} dk_z e^{-ik_z(z_p - z_q)} \tilde{J}_{\alpha n}(-\nu, -k_z) G_\nu^{\alpha\beta}(k_z) \tilde{J}_{\beta m}(\nu, k_z) = B_{ji}$ where $\tilde{J}_{\alpha n}$ ($\tilde{J}_{\beta m}$) is the FT of the *n*th (*m*th) expansion mode of the basis function along the α (β) direction. *p* and *q* are indices over the patch numbers. In order to get nontrivial solutions for **a**, we should have $\det(\bar{\mathbf{B}}) = 0$. The solutions of this determinantal equation are satisfied by complex frequencies $f = f_r + \mathbf{i} f_i$. The resonant frequency and the quality factor of the microstrip structure are f_r and $Q = f_r / 2 f_i$, respectively. For transmission line problems, the dimensions of the conductors along the *z* direction are infinite. In this case, the determinantal equation is solved for the propagation wavenumbers [6]; the integral symbol in B_{ij} is omitted. We use two sets of expansion basis functions; 1) the cavity mode sinusoidal functions without the edge condition [7] and 2) the Chebyshev polynomials with Maxwell weighting function [8]. The field components to the current distribution on the conductors, for $\rho < \rho_1$, can be expressed in the following compact forms: $\mathbf{E}_s = \frac{1}{2\pi} \sum_{\nu=-\infty}^{\infty} e^{i\nu\varphi} \int_{-\infty}^{\infty} dk_z e^{ik_z z} \tilde{\mathbf{E}}_{\nu s}$, $\mathbf{H}_s = \frac{1}{2\pi} \sum_{\nu=-\infty}^{\infty} e^{i\nu\varphi} \int_{-\infty}^{\infty} dk_z e^{ik_z z} \bar{\mathbf{M}}(\rho) \cdot \tilde{\mathbf{E}}_{\nu s}$, and $\begin{bmatrix} E_\rho & H_\rho \end{bmatrix}^T = \frac{1}{2\pi} \sum_{\nu=-\infty}^{\infty} e^{i\nu\varphi} \int_{-\infty}^{\infty} dk_z e^{ik_z z} \left[\bar{\mathbf{U}}_\nu + \bar{\mathbf{V}}_\nu \cdot \bar{\mathbf{M}}(\rho) \right] \cdot \tilde{\mathbf{E}}_{\nu s}$ where $\tilde{\mathbf{E}}_{\nu s} = \left[\left(\bar{\boldsymbol{\Gamma}}_< \cdot \bar{\mathbf{T}} \right)^{EE} (\rho, \rho_1) + \left(\bar{\boldsymbol{\Gamma}}_< \cdot \bar{\mathbf{T}} \right)^{EH} (\rho, \rho_1) \cdot \bar{\mathbf{M}}(\rho) \right]^{-1} \cdot \bar{\mathbf{G}}_\nu(k_z) \cdot \tilde{\mathbf{K}}_\nu(k_z)$. Field components can then be obtained recursively in all regions.

Sample numerical results

To verify the validity of the present formulation, the effective dielectric constant of a cylindrical strip mounted inside a ground cylindrical surface is computed using the determinantal equation. The results are presented in Fig. 2 along with the available results [6]. There is a very good agreement between the two results. It is evident from Fig.2 that the proposed formalism is rigorous and accurate. The B_ϕ field component under the curved strip of an MTL versus the radial distance ρ ($0<\rho<\rho_2$) at $(\phi,z)=(0,0)$ and f=200 MHz, is shown in Fig. 3. The parameters of the microstripline are [3] ρ_2= 7.25, ρ_3= 7.53, ρ_4= 10.22, ρ_5= 10.5 all in *cm*, $\varepsilon_{2r}=\varepsilon_{4r}=1$, $\varepsilon_{3r}=\varepsilon_{5r}=2.9-0.0348i$, *w*=6.4 *mm*. The electrical characteristics of the phantom are varied from low dielectric/low conductivity (ε_{1r}=2.3) to high dielectric/high conductivity (ε_{1r}=57-54i). From our numerical simulations, we summarize the main results as follows: the *B* field variations with dimension and electrical characteristics of the sample are not very significant except when the load presents high permittivity/high conductivity and a dimension very close to the strip interface $\rho = \rho_2$. At an arbitrary observation point within the structure, the field components along the *z* direction are not negligible and depend on the *z* coordinate variations. In this case, the quasi-TEM analysis can lead to inaccurate results. For a better analysis and design of multilayer MTLs, the full-wave methods are more rigorous and since they make few approximations.

Conclusion

An efficient full-wave moment method for the analysis of cylindrical multilayer multiconductor patch coil/MTL is presented. The model is validated for the single strip MTL structure by comparison with numerical data available from other studies. The numerical results concerning the multistrip MTL are under processing and will be reported later. The proposed method is intended to be an alternative to the quasi-TEM approach for the analysis/design of phased array microstrip coils.

References

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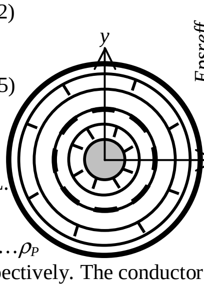

Fig. 1. Cross section of the generalized M3PA/MTL. The dielectric layers (from the origin to the outer grounded cylinder) have radii $\rho_1, \ldots, \rho_{M-1}, \rho_M, \rho_{M+1}, \ldots \rho_P$ and permittivities $\varepsilon_{1r}, \ldots, \varepsilon_{M-1r}, \varepsilon_{Mr}, \varepsilon_{M+1r}, \ldots, \varepsilon_{Pr}$ respectively. The conductors are located at the interface $\rho = \rho_M$.

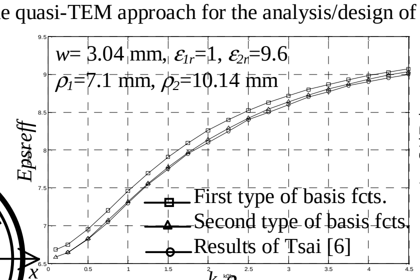

Fig. 2. The calculated ε_{reff} as a function of $k_0\rho_2$.

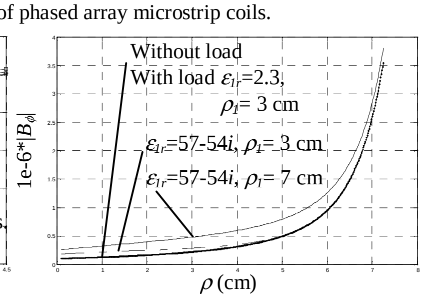

Fig. 3. Normalized $|B_\phi|$ versus the radial distance ρ for various sample loads.