ANALYSIS OF THICK WALL APERTURE COUPLED STRIPLINE RADIATORS

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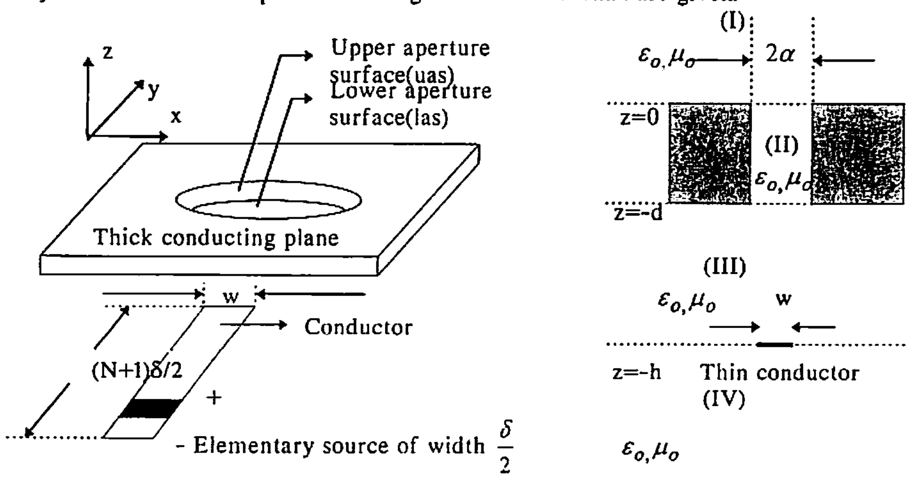
Abstract
A planar stripline thick wall aperture coupled antenna is analysed. As a primary approach a semi-analytical method is developed to compute the electromagnetic interaction between an arbitrarily shaped two dimensional conductor, excited by an external source and a thick wall aperture. Fields in the region of the thick wall and the aperture are expanded in terms of canonical wave solutions. The current distribution on the conductor surfaces is described as a superposition of pulse functions. The boundary conditions on both the aperture surfaces and the conductor are satisfied by using a point-matching technique. In order to reduce numerical cost, analytical techniques are employed to calculate the matrix of the linear system. Taking advantage of the geometrical symmetry, the conductor is placed between two thick wall aperture screens, while the dielectric substrate effect is taken into consideration, employing a mean value surface impedance boundary condition technique.

1. Introduction

The motivation of the present work has been the analysis of novel satellite communications antennas in the microwave and mm-wave frequencies. Especially, the recent plans of developing direct television signal receivers in the mm-wave region, calls for the analysis of "convenient" antennas, in terms of electromagnetic (EM) characteristics (gain, side lobes, bandwidth, G/T), mechanical stability and size. The three-dimensional structure of the conventional wide-used paraboloidal reflector antenna, together with its mechanical sensitivity, has turned the interest of many researchers towards analysing planar two-dimensional small radiators, mainly in the microwave region [1-3].

In the present paper, the proposed structure consists of thin planar aperture coupled stripline antennas. Since the mm-wave region is included in the frequencies of interest, the apertures thickness cannot be neglected. To this end, the EM coupling between an arbitrarily shaped conductor and a circular aperture lying on a finite thickness conducting screen, as shown in Fig.1, is analysed and some concepts for building the overall antenna are given.



a. 3D view b. Cross-section

Fig.1. Circular aperture on a thick conducting plane coupled to a straight surface current

2. Derivation of the integral equation

Assuming an exp(+jωt) time dependence for all the fields quantities and applying the Gauss' theorem $\nabla \cdot \underline{E} = 0$ in the regions I, III, IV, shown in Fig.1, the corresponding electric fields are expressed in terms of their Fourier transform, as

$$\underline{E_I}(x,y,z) = \int_{-\infty}^{+\infty}\int_{-\infty}^{+\infty} dk_x dk_y \cdot e^{jk_x x + jk_y y - \gamma z} \cdot \underline{C_I}(k_x, k_y)$$

$$\underline{E_{III}}(x,y,z) = \int_{-\infty}^{+\infty}\int_{-\infty}^{+\infty} dk_x dk_y \cdot e^{jk_x x + jk_y y} \cdot (\underline{C_{III}^{+}}(k_x,k_y)\cdot e^{-\gamma z} + \underline{C_{III}^{-}}(k_x,k_y)\cdot e^{+\gamma z})$$

$$\underline{E_{IV}}(x,y,z) = \int_{-\infty}^{+\infty}\int_{-\infty}^{+\infty} dk_x dk_y \cdot e^{jk_x x + jk_y y + \gamma z} \cdot \underline{C_{IV}}(k_x, k_y)$$

where $\underline{C_I}(k_x,k_y), \underline{C_{III}^{+}}(k_x,k_y), \underline{C_{III}^{-}}(k_x,k_y), \underline{C_{IV}}(k_x,k_y)$ are the vector coefficients to be determined, $\gamma = \sqrt{k_x^2 + k_y^2 - k_o^2}$ is the propagation constant and $k_o = \omega\sqrt{\varepsilon_o \mu_o}$ is the propagation constant in the free space. It is notable that, in order to satisfy the radiation conditions, it is required that Real(γ)>0 and due to the exp(+jωt) time dependence, it is also required that Imag(γ)>0.

In region II, the electric field is expressed in terms of the circular waveguide eigenfunctions $\underline{e_{TE}^{mn}}$, $\underline{e_{TM}^{mn}}$ [4], as

$$\underline{E_{II}}(\rho,\phi,z) = \sum_{m=-\infty}^{+\infty}\sum_{n=1}^{+\infty} (A_{mn}^{+}\cdot e^{-\gamma'_{mn} z} + A_{mn}^{-}\cdot e^{\gamma'_{mn} z})\cdot \underline{e_{TE}^{mn}} + (B_{mn}^{+}\cdot e^{-\gamma_{mn} z} + B_{mn}^{-}\cdot e^{\gamma_{mn} z})\cdot \underline{e_{TM}^{mn}}$$

where $\underline{e_{TE}^{mn}} = \left(\frac{\omega\mu_o m a^2}{\rho_{mn}'^2 \rho} J_m\left(\frac{\rho'_{mn}}{a}\rho\right).e^{jm\phi}, \quad \frac{j\omega\mu_o a}{\rho'_{mn}} J_m'\left(\frac{\rho'_{mn}}{a}\rho\right).e^{jm\phi}, \quad 0\right)$

$$\underline{e_{TM}^{mn}} = \left(\frac{-\gamma_{mn} a}{\rho_{mn}} J_m'\left(\frac{\rho_{mn}}{a}\rho\right).e^{jm\phi}, \quad \frac{-j\gamma_{mn} a^2 m}{\rho_{mn}^2 \rho} J_m\left(\frac{\rho_{mn}}{a}\rho\right).e^{jm\phi}, \quad J_m\left(\frac{\rho_{mn}}{a}\rho\right).e^{jm\phi}\right)$$

$\gamma_{mn} = +\sqrt{\frac{\rho_{mn}^2}{a^2} - k_o^2}$, $\gamma'_{mn} = \sqrt{\frac{\rho_{mn}'^2}{a^2} - k_o^2}$, ρ'_{mn} and ρ_{mn} being the n-th roots of the m-th order Cylindrical Bessel function and its derivative respectively. Applying the Maxwell's equation $\nabla \times \underline{E} = -j\omega\mu_o \underline{H}$, the H-fields are obtained in each region.

Imposing the boundary conditions on the z=0 and z=-d surfaces (i.e. continuity of the tangential components of the electric and magnetic fields on the upper (S_{uas}) and on the lower aperture surface (S_{las})) and on the z=-h surface (i.e. continuity of the tangential components of the E-field and discontinuity of the tangential components of the H-field on the conductor surface due to the presence of the excitation), a system of three coupled two-dimensional integral equations is derived in terms of the unknown transverse electric field components $\underline{E_{at}^{u}}(x,y)$, $\underline{E_{at}^{l}}(x,y)$ on the upper(uas) and lower aperture surface(las) respectively and in terms of the conductivity currents $\underline{J_s}(x,y)$ which flow on the conductor surface, as

$$\iint_{S_{apert}} dx'dy' \cdot [\overline{\underline{K}}_{uu}(x,y|x',y')\cdot \underline{E_{at}^{u}}(x',y') + \overline{\underline{K}}_{ul}(x,y|x',y')\cdot \underline{E_{at}^{l}}(x',y')] = \underline{0} \quad x,y \in S_{uas}$$

$$\iint_{S_{apert}} dx'dy' \cdot [\overline{\underline{K}}_{lu}(x,y|x',y')\cdot \underline{E_{at}^{u}}(x',y') + \overline{\underline{K}}_{ll}(x,y|x',y')\cdot \underline{E_{at}^{l}}(x',y') +$$

$$+ \iint_{S_{cond}} dx'dy' \cdot [\overline{\underline{K}}_{ls}(x,y|x',y')\cdot \underline{J_s}(x',y') = \underline{0} \quad x,y \in S_{las}$$

$$\iint_{S_{apert}} dx'dy' \cdot \underline{\overline{K}}_{sl}(x,y|x',y') \cdot \underline{E^l_{at}}(x',y') + \iint_{S_{cond}} dx'dy' \cdot \underline{\overline{K}}_{ss}(x,y|x',y') \cdot \underline{J_s}(x',y') = \underline{R}(x,y) \quad x,y \in S_{cond}$$

where $\underline{\overline{K}}_{ij}(x,y|x',y'), (i = u,l,s / j = u,l,s)$ are kernel matrix functions and the right hand vector $\underline{R}(x,y)$ describes the excitation. Special attention is required for the transformation of the field components from orthonormal coordinates to polar ones for the two aperture surfaces and to transverse-longitudinal ones for the conductor.

3. Solution of the integral equation

In order to proceed with the solution of the system of integral equations, the unknown quantities $\underline{E^u_{at}}(x,y)$, $\underline{E^l_{at}}(x,y)$ are expressed as

$$\underline{E^u_{at}}(\rho,\phi) = \sum_{m=-\infty}^{+\infty} \sum_{n=1}^{+\infty} A_{\tilde{m}\tilde{n}} \cdot \underline{e^{\tilde{m}\tilde{n}}_{TE}} + B_{\tilde{m}\tilde{n}} \cdot \underline{e^{\tilde{m}\tilde{n}}_{TM}} \ , \ \underline{E^l_{at}}(\rho,\phi) = \sum_{m=-\infty}^{+\infty} \sum_{n=1}^{+\infty} C_{\tilde{m}\tilde{n}} \cdot \underline{e^{\tilde{m}\tilde{n}}_{TE}} + D_{\tilde{m}\tilde{n}} \cdot \underline{e^{\tilde{m}\tilde{n}}_{TM}}$$

where $A_{\tilde{m}\tilde{n}}, B_{\tilde{m}\tilde{n}}, C_{\tilde{m}\tilde{n}}, D_{\tilde{m}\tilde{n}}$ are the unknown coefficients to be determined. Accordingly, the conductor surface S_{cond} is divided into N+1 transverse elementary segments of equal length $\frac{\delta}{2}$ with N points, in each of which the conductivity current density is given by the piece-wise trigonometric function

$$\Pi(y) = \begin{cases} \sin\frac{2\pi}{\lambda}(\frac{\delta}{2} - |y|), & |y| \le \frac{\delta}{2} \\ 0, & |y| \ge \frac{\delta}{2} \end{cases}.$$

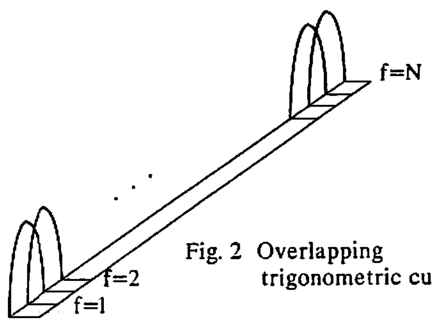


Fig. 2 Overlapping trigonometric currents

Then the overall surface current is expressed as a summation $J_s(y) = \hat{y} \cdot \sum_{f=1}^{N} t_f \cdot \Pi(y - y_f)$, where y_f corresponds to the f-th segment center. It should be noted that the last expansion implies constant dependence of the transverse x-variable and zeroth value of the transverse J_x current component. In addition, the choice of the employed piece-wise function definition intervals leads to overlapping trigonometric distribution currents, which satisfy the edge conditions at the segment interfaces, as shown in Fig.2.

Using the inverse Fourier transform of the fundamental formula which gives the spherical wave propagation

$$\frac{e^{-jk_0|r-r'|}}{4\pi \cdot |r-r'|} = \frac{1}{(2\pi)^3} \cdot \int_0^{+\infty} k dk \int_0^{2\pi} d\phi_k \int_{-\infty}^{+\infty} dp_z \frac{e^{jkr\cos(\phi-\phi_k)} \cdot e^{jkr'\cos(\phi'-\phi_k)} \cdot e^{jp_z(z-z')}}{k^2 + p_z^2 - k_o^2 + j\varepsilon}, z > z', \varepsilon \to 0^+,$$

applying the Cauchy's 'integral-residuals' theorem and using the 'addition' theorem ([5]) to expand the function $\frac{e^{-jk_0|r-r'|}}{4\pi \cdot |r-r'|}$ in products of cylinder functions of the half order (spherical Bessel functions) and Legendre polynomials, the following linear system is obtained

$$\begin{bmatrix} \underline{\overline{A_1}} & \underline{\overline{A_4}} \\ \underline{\overline{A_2}} & \underline{\overline{A_5}} \\ \underline{\overline{A_3}} & \underline{\overline{A_6}} \end{bmatrix} \cdot \begin{bmatrix} \underline{\overline{AB}_{\tilde{m}\tilde{n}}} \\ \underline{\overline{CD}_{\tilde{m}\tilde{n}}} \\ \underline{\overline{T_f}} \end{bmatrix} = \begin{bmatrix} \underline{\overline{0}} \\ \underline{\overline{0}} \\ \underline{\overline{R_o}} \end{bmatrix},$$

where sub-matrix $\underline{\overline{A_1}}$ expresses the effect of the uas and las on the uas, $\overline{\underline{A_2}}$ the effect of uas and las on the las, $\overline{\underline{A_3}}$ the effect of las on the surface current (sc), $\overline{\underline{A_4}} = \overline{\underline{0}}$ the non effect of sc on the uas, $\overline{\underline{A_5}}$ the effect of sc on the las and $\overline{\underline{A_6}}$ the effect of sc on itself. The sub-matrix $\overline{\underline{R_o}}$ corresponds to the excitation and the sub-matrices $\overline{\underline{AB}}_{\tilde{m}\tilde{n}}, \overline{\underline{CD}}_{\tilde{m}\tilde{n}}, \overline{\underline{T}}_f$ denote the unknown coefficients. Giving point pairs (r_i, ϕ_i) on the uas and las and points y_i on the sc, the system is solved and macroscopic quantities such as radiation pattern are easily computed, independently of the choice of the points (Point Matching Technique).

4. Extension to a stripline antenna structure - Discussion

Once the EM coupling between the arbitrarily shaped conductor and the thick wall aperture is determined, the present method can be easily expanded to the study of the coupling between two co-centric circular apertures, included in two different thick parallel conducting planes and a conductor of any shape, arbitrarily placed between the two apertures. This is easily accomplished due to horizontal symmetry, though the lower aperture has to be of larger radius, if a broadband antenna is required. As expected, a thin dielectric substrate is needed to support the conductor (usually Teflon with $\varepsilon_r = 2.1$). The impact of this substrate is approximately taken into consideration by demanding a surface impedance flow (jz) instead of a zeroth value of the transverse electric field on the conductor surface. Therefore, a new boundary condition is imposed, taking the transverse electric field mean value, as $\frac{\underline{E}_t(z=-h^-) + \underline{E}_t(z=-h^+)}{2} = jz\underline{H}_t^n$, where $_t$, ^{n} denote the transverse and normal components respectively. A third conducting plane can be used as a reflector, completing the design of a planar stripline antenna.

Furthermore, this antenna can be used as individual element in constructing a two-dimensional array with significant gain/directivity requirements, suitable for satellite communications applications. To this end, the second-order phenomena associated to the adjacent apertures coupling, which affects the side lobes behaviour, have to be taken into consideration. This, together with the dielectric substrate detailed consideration are the topics of our current work.

REFERENCES

[1] E. Rammos, "Suspended Substrate Line Antenna Fits 12 GHz Satellite Applications", MSN, pp. 110-126, March 1984.

[2] G.H. Millard, "A 12 GHz Domestic Receiving Aerial for Satellite Broadcasting", EBU Review, pp. 208-213, October 1972.

[3] G. Dubost and C. Vinatier, "Reseaux de Doublets Replies Symmetriques en Plaques a Large Bande autour de 12 GHz", Onde Electrique, V61, pp. 34-41, Avril 1981.

[4] D. S. Jones, *The Theory of Electromagnetism*, London: Pergamon Press, 1964.

[5] I. S. Gradstein and I. M. Ryshik, *Tables of Integrals, Series and Products*, Moscow 1971.