A NEW EXCITATION MODEL FOR PROBE-FED PRINTED ANTENNAS ON FINITE SIZE GROUND PLANES

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ABSTRACT – *This paper presents a new excitation model for probe-fed printed antennas on both infinite and finite size ground planes. The model has been developed within the general frame of the mixed potential integral equation (MPIE) and the method of moments (MoM). The technique is based on a delta-gap voltage model, and a special procedure is implemented inside the integral equation to effectively impose a voltage reference plane into a floating metallic plate which is acting as a ground plane. The present technique allows the accurate calculation of the input impedance of printed antennas, and the effects of finite size ground planes can be easily accounted for in the calculations. In addition, an efficient technique is presented for the evaluation of the radiation patterns of printed antennas, taking also into account the presence of finite size ground planes. Comparisons with measured results show that the new derived excitation method is indeed accurate, and can be used for the prediction of the backside radiation and side lobe levels of real life finite ground plane printed antennas.*

Keywords.— Integral equation, excitation models, finite ground plane, backside radiation, printed antennas.

1 INTRODUCTION

During the last decades, printed circuits and antennas have played an important role in many branches of electrical engineering and the field of application is spreading to new technologies and to even higher frequencies. The need for miniaturisation is increasing in many applications e.g., telecommunications and space missions. Obviously, these compact geometries are not adequate for the use of models assuming infinite ground planes.

The need to take into account for finite ground plane dimensions in microstrip antennas modelling arises especially in applications where patches are used as free standing structures and front-to-back ratio must be maximized in order to avoid interference problems [Bokhari *et al.* 1992], or to locate a potential main beam deformation caused by the diffraction from the ground plane edges. Moreover, the need to model the excitation on two floating metallic patches can become inevitable in applications like dual band stacked printed antennas where a first patch acts as ground plane for a second radiating element [Zürcher *et al.* 1999].

To solve this problem a new excitation model and de-embedding technique for the computation of the input impedance of probe-fed printed antennas on finite size ground planes using a Mixed Potential Integral Equation technique (MPIE) [Mosig and Gardiol 1988, Hall and Mosig 1996] has been developed. This approach accounts for the effect of the ground plane dimensions on the input impedance, the mutual coupling, and the radiation characteristics of a single antenna element or a finite array.

As a first step to attain this goal, a new attachment mode for probe-fed printed antennas on infinite ground plane has been developed. The most widely used excitation model for probe-fed antennas is the impressed-current model [Pozar 1982, Hall and Mosig 1989]. This model assumes that a constant impressed current is exciting the antenna and it use the derived distribution of currents on metallic surfaces to compute the voltage at the probe location. This method may lead to accurate results but needs the computation of a surface integral over all the metallic surfaces present in the structure to obtain the input impedance. Contrary to the previous one, the model presented here, as described in Sec. 3, uses a delta gap voltage excitation model (to the authors' knowledge used until now only for microstrip line fed antennas [Davidovitz and Lo 1989, Harokopus and Katehi 1991, Eleftheriades and Mosig 1996]). This model assumes an impressed voltage between the antenna and the ground plane and, once the surface currents have been computed, only a normalisation by the excitation voltage is needed to obtain the input admittance. Another remarkable difference between the two models is the type of special basis functions used in the attachment mode. Considering the case of triangular meshing (the extension to rectangular cells is straightforward), in the impressed current model one (or more) entire basis function with opposite sign of the current on its two halves is used to model the horizontal spreading of the vertical current coming from the coaxial probe. In the present model, one to three half basis functions are introduced for the attachment mode depnding on the location of the feed. This implies that the present excitation model can be used for any probe location inside the patch, including its edge and also for microstrip line fed antennas [Tiezzi *et al.* 1999] without exception.

These excitation models as well as the subsequent technique for computing impedances are implicitly based on the assumption of an infinite ground plane, which according to image theory automatically produces a zero voltage at ground plane level. In Sec. 4 the attachment mode is modified in order to take into account the finiteness of the ground plane. Here, instead of using Green's functions including the ground plane effect through image theory, a specific numerical treatment is applied to the ground plane. To the authors' knowledge, the first approach using an MPIE formulation for the study of finite size ground planes can be found in [Bokhari *et al.* 1992]. This work, however, only represents an approximation of the real finite structure, since the currents induced on the antenna are computed using an infinite ground plane model. Once the induced currents are computed, the finite size nature of the ground plane is taken into account, at a later stage, during the calculation of the scattering problem associated with the computed currents. Hence the results presented in [Bokhari *et al.* 1992] are only accurate, if the ground plane is sufficiently large: it would therefore be desirable to develop a rigorous method, which remains valid even for very small ground planes. The method presented in this paper is a full wave method based on the MPIE technique, and the only approximation introduced is that we use the Green's functions multilayered media formulated in the traditional form of Sommerfeld integrals [Mosig and Gardiol 1988, Mosig 1989]. Therefore the currents induced in the structure are computed taking into account since the beginning the finite size of the ground plane but the second-order effect of dielectric truncation is neglected. This approximation has been introduced to maximize the numerical efficiency and its accuracy is confirmed by our results. In addition to being more rigorous, another advantage of this approach is that the effects on the input impedance of the finite size ground planes can accurately be evaluated and moreover scattering from ground plane edges can be taken into account. Thus full range (including backside scattering) radiation patterns can also be predicted.

2 BACKGROUND AND STATEMENT OF THE PROBLEM

The new excitation model presented in this paper has been developed in the frame of the analysis of multilayered printed circuits and antennas following the MPIE formulation [Mosig and Gardiol 1988]. The generic structure under analysis is presented in Fig. 1. As shown, it is composed by one or more conducting patches embedded on a stratified medium. Either a perfect conductor ground plane or a free space layer extending to $z = -\infty$ can be placed at the bottom of the structure. Each dielectric layer, which may be lossy, is assumed to be homogeneous, isotropic and transversally infinite. The conducting patches are assumed to have finite transverse size, arbitrary shape, negligible thickness and an infinite conductivity, although finite conductivity can easily be taken into account using Leontovich boundary conditions [Mosig and Gardiol 1985].

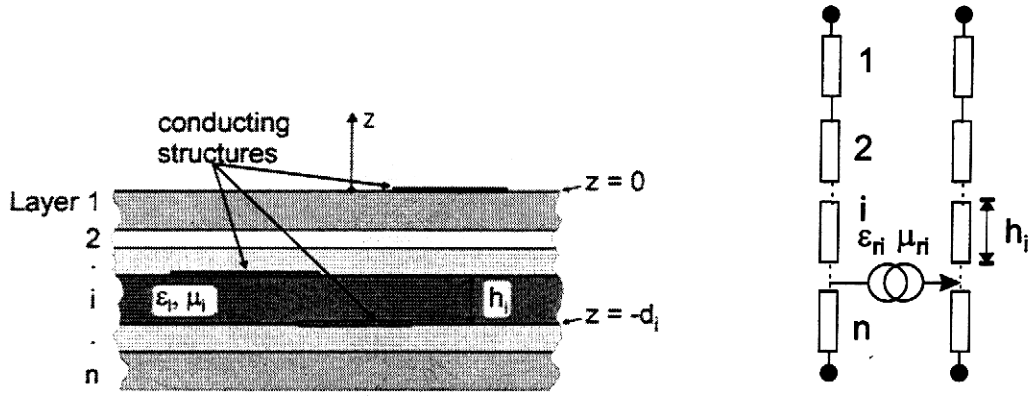
Under these assumptions the boundary condition for the electric field on the surface of the conducting strips is written as

$$\hat{e}_z \times (\vec{E}^e + \vec{E}^s) = 0 \qquad (1)$$

where $\vec{E}^e$ and $\vec{E}^s$ are respectively the excitation and the scattered electric field.

The scattered field is expressed in terms of the vector and scalar potential $\vec{A}$ and V as

$$\vec{E}^s = -j\omega\vec{A} - \nabla V \qquad \vec{H} = \frac{1}{\mu}\nabla \times \vec{A} \qquad (2)$$



(a) Multilayered medium.

(b) Equivalent network.

Figure 1: Generic multilayered structure containing an arbitrary number of finite metallizations.

with the potentials related by the Lorentz gauge [Mosig 1989]

$$j\omega\mu\epsilon V + \nabla \bullet \vec{A} = 0 \tag{3}$$

The vector and scalar potentials $\vec{A}, V$ can in turn be expressed in terms of superposition integrals of the corresponding Green's functions $\overline{\overline{G}}_A, G_V$ weighted by the unknown distribution of surface current and electric charge $\vec{J}_s, \rho_s$ as

$$\begin{aligned} \vec{A} &= \int_S \overline{\overline{G}}_A(r|r') \bullet \vec{J}_s(r')\, dS' \\ V &= \int_S G_V(r|r')\, \rho_s(r')\, dS' \end{aligned} \tag{4}$$

and finally, using the continuity equation to express the electric charge in terms of current, the boundary condition in equation (1) becomes

$$\begin{aligned} \hat{e}_z \times \vec{E}^e &= \hat{e}_z \times \left(j\omega \int_S \overline{\overline{G}}_A(r|r') \bullet \vec{J}_s(r')\, dS' \right. \\ &+ \left. \frac{1}{j\omega} \nabla \int_S G_V(r|r')\, \nabla \bullet \vec{J}_s(r')\, dS' \right) \end{aligned} \tag{5}$$

which is the basic integral equation to be solved to find the unknown distribution of surface currents.

The multilayered media Green's functions appearing in equation (5) are derived, in the spectral domain, from the equivalent transmission line circuit shown in Fig. 1(b), as described in [Mosig and Gardiol 1988, Michalski and Mosig 1997]. Furthermore, these Green's functions are calculated in the spatial domain using special numerical methods for the evaluation of the Sommerfeld integral, as extensively described in [Mosig 1989, Alvarez-Melcon and Mosig 1996].

The previous integral equation (5) is solved by the Method of Moments. The conducting patches are segmented into triangular cells and triangular rooftops [Rao *et al.* 1982] are used as basis and test functions, applying a Galerkin method. If coaxial excitation is used, modified basis functions are introduced at the coaxial pin location in order to model the spread on the patch of the current flowing on the vertical pin.

3 A NEW ATTACHMENT MODE

A special set of basis functions, called the attachment mode, is used to ensure the continuity of the current between the coaxial probe and the antenna. In the present approach the attachment mode is derived directly from the delta-gap voltage excitation model used on microstrip lines [Eleftheriades and Mosig 1996]. As shown in fig. 2 an efficient excitation model is obtained for the microstrip case applying a voltage source of magnitude V_m between an infinitesimally small gap of length $\delta \to 0$ across the feeding line and the ground plane. The flow of induced currents through the edge of the microstrip line is modeled introducing one or more half subsectional basis functions (half rooftop in the present case) as shown in Figs. 2a, 2c, and 2d [Eleftheriades and Mosig 1996].

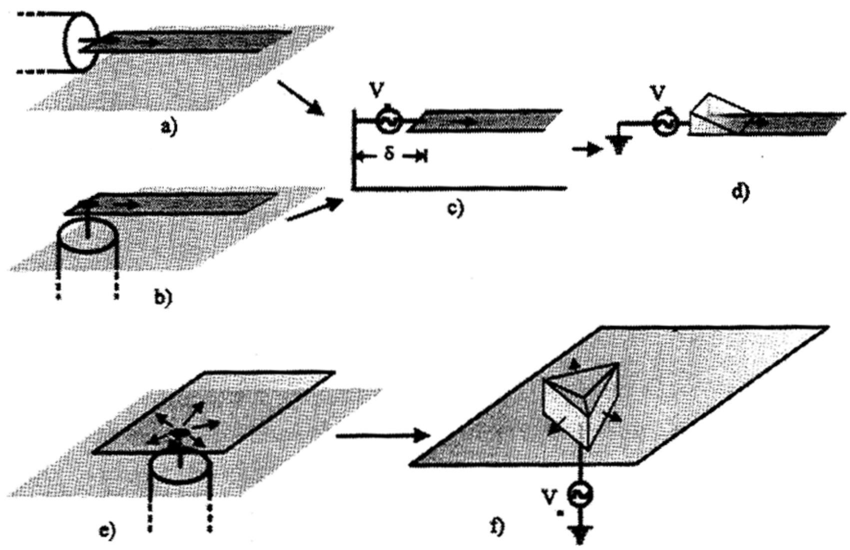


Figure 2: A delta-gap voltage source exciting printed circuits. a) Colinear transition between a coaxial probe and a microstrip line. b) perpendicular transition between a coaxial probe and a microstrip line. c) Delta-Gap voltage model applyed to a coaxial probe-fed microstrip line. d) Associated MoM description of the excitation model. e) Coaxial probe-fed patch antenna. f) Associated MoM description of the of the excitation of a probe-fed patch antenna

It is well known that at least for electrically thin dielectrics, no difference in the measurement can be noticed when the microstrip line is fed by a vertical coaxial probe (Fig. 2b), so it can be affirmed that the previous delta-gap excitation model is still valid in this case. The next step is to apply the same method to a point located inside the patch (see Fig. 2e) having in mind that current can spread in any direction. This behaviour can be obtained introducing 3 (or less if the feed is close to the edge) new half rooftops, one for each edge of the triangle containing the feeding point, which are superimposed to the halves of the standard rooftops already attached to the triangle (see Fig. 2f). It must be stressed that at this point six half rooftops (one couple for each side) are present in the triangle, but only three of them are involved in the attachment mode and they are attached to three virtual vertical half rooftops, while the other three are connected to the halves located in the adjacent triangles to form standard "planar" basis function. It is also important to point out that to reach a good model of the physical excitation, the area of the triangle with the attachment mode must be reasonably small, the lower limit being imposed by the section of the internal conductor of the coaxial cable.

The application of the Method of Moments (MoM) to solve the integral equation (5) leads to a system of linear equations that can be shortly expressed as

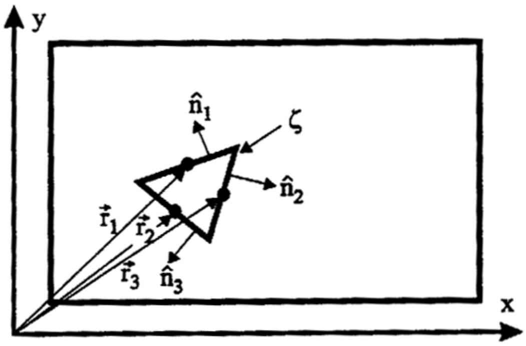


Figure 3: Basic geometry of a probe-fed printed antenna used in the formulation of the excitation model.

$$e_i = \sum_{k=1}^{N_f} \alpha_k \, P_{i,k} \,, \qquad i = 1, 2, \cdots, N_f \tag{6}$$

where $P_{i,k}$ is the i,k-th term of the moments matrix, α_k is the k-th term of the unknown electric current density vector, N_f is the total number of basis functions and e_i is the i-th term of the excitations vector. The latter is defined as

$$e_i = \int_S \vec{E}^e \bullet \vec{f}_i(\vec{r}) \, ds \tag{7}$$

where $\vec{E}^e$ represents the impressed electric field, and $f_i(\vec{r})$ is the subsectional testing functions of the MoM. The unknowns electric currents can now be expanded as

$$\vec{J}_s = \sum_{k=1}^{N_f} \alpha_k \, \vec{f}_k(\vec{r}\,') \qquad (8)$$

where N_f is the total number of basis functions, and α_k are the unknown coefficients in the expansion.
With reference to the port geometry shown in Fig. 3, we apply the delta-gap model only to the three "half" basis functions of the attachment mode, which allows us to write the excitation field created by the voltage source as

$$\vec{E}^e = V_m \sum_{p=1}^{3} \delta(\vec{r} - \vec{r}_p)\, \hat{n}_p \qquad (9)$$

where $\vec{r}_p, \quad p = 1, 2, 3$, denotes the position vector of the three edge associated to the port. Substituting equation (9) in equation (7) we obtain

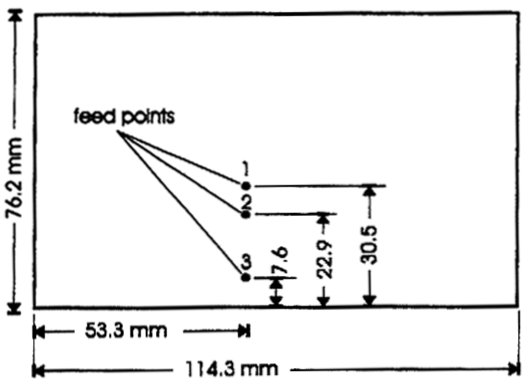


Figure 4: Probe-fed patch antenna on an infinite ground plane. Substrate: REXOLITE 2200, h = 1.59 mm, ε_r = 2.62, $tan\delta \approx 0.002$.

$$e_i = V_m \int_S \left[\sum_{p=1}^{3} \delta(\vec{r} - \vec{r}_p)\, \hat{n}_p \bullet \vec{f}_i(\vec{r}) \right] ds \qquad (10)$$

Using the integration properties of the Dirac delta function and defining $\vec{f}_i^{\hat{n}_p}(\vec{r}) = \hat{n}_p \bullet \vec{f}_i(\vec{r})$ as the component of the basis function perpendicular to p-th triangle's edge, equation (10) reduces to

$$e_i = V_m \oint_{\zeta} \left[\sum_{p=1}^{3} f_i^{\hat{n}_p}(\vec{r}_p) \right] dl \qquad (11)$$

where ζ is the perimeter of the triangle with the attachment mode (see Fig. 3). Defining now

$$\gamma_i = \oint_{\zeta} \left[\sum_{p=1}^{3} f_i^{\hat{n}_p}(\vec{r}_p) \right] dl \qquad (12)$$

which is an integral with an easily obtained analytical solution, we can introduce (11) in (6) and obtain the following system of linear equations

$$V_m \, \gamma_i = \sum_{k=1}^{N_f} \alpha_k \, P_{i,k} \, , \qquad i = 1, 2, \cdots, N_f \qquad (13)$$

The solution of this system of linear equations gives the

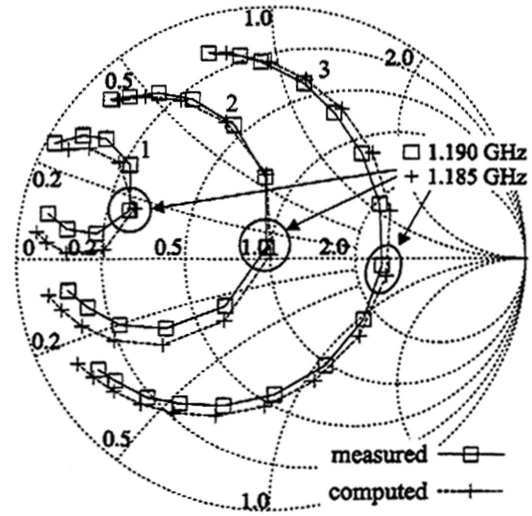


Figure 5: Comparison of measured and computed results of the input impedance of the antenna in Fig. 4. □ measure, + theory. (increment 5 MHz clockwise, measurement reproduced from [James and Hall 1989])

values of the unknown coefficients α_k. These can then be used to compute the current I_m flowing through the port as follows

$$\begin{aligned} I_m &= \oint_{\zeta} \left[\sum_{p=1}^{3} \vec{J}_s(\vec{r}_p) \bullet (\hat{n}_p) \right] dl \\ &= \sum_{k=1}^{N_f} \alpha_k \oint_{\zeta} \left[\sum_{p=1}^{3} f_k^{\hat{n}_p}(\vec{r}_p) \right] dl \\ &= \sum_{k=1}^{N_f} \alpha_k \, \gamma_k \qquad (14) \end{aligned}$$

From equation (14) the input impedance of the circuit is directly obtained by dividing both the terms of the equation by the exciting voltage V_m, and then by inverting the resulting input admittance, namely:

$$Z_{in} = \frac{1}{Y_{in}} \qquad Y_{in} = \frac{I_m}{V_m} = \sum_{k=1}^{N_f} \frac{\alpha_k \, \gamma_i}{V_m} \qquad (15)$$

To verify the validity of the derived model we have analysed the basic probe-fed printed patch antenna presented in [James and Hall 1989]. For simplicity the geometry of the antenna is reported in Fig. 4. The input impedance of the antenna has been measured for the fundamental (TM_{10}) mode and for three different placements of the feed (see Fig. 4). The comparison between the measurement and the computed results, presented in fig. 5, show the accuracy achieved with the present model.

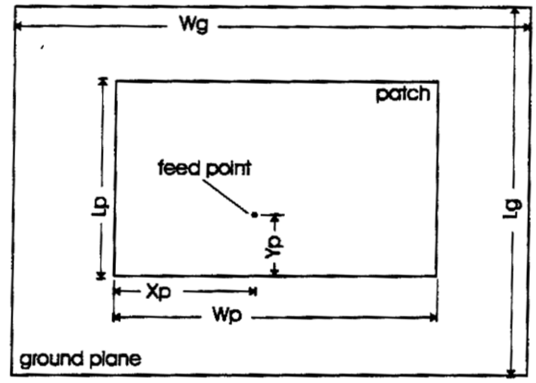


Figure 6: Probe-fed patch antenna on an finite size ground plane.

4 ANALYSIS OF PROBE-FED PATCH ANTENNAS ON FINITE SIZE GROUND PLANES

In this section we describe how the excitation model presented in the previous section must be modified in order to take into account the finiteness of the ground plane. The study is presented for a simple printed patch antenna, but the extension to more complicated structure is straightforward. An important difference between the analysis presented in the present paper and traditional analysis like the one performed in the previous section (see also [Bunger and Arndt 1997]), is that in the present case the Green's functions derived do not take into account infinite ground planes, and therefore, all metallizations are considered to be finite. The main difficulty in doing this is that the condition of null potential at the ground plane is not automatically imposed by the Green's functions. As a consequence, now the finite ground plane must be introduced inside the integral equation to enforce the proper boundary conditions on it, and the currents induced on this reference ground plane must also be computed. Also, a new excitation model and de-embedding technique must be derived to be able to extract the actual input impedance of the antenna when such floating grounds are considered as references. This is mainly due to the fact that the ground plane is no longer acting as an automatic reference plane for the generator, so that the reference condition of the finite ground plane must be introduced explicitly in the model.

The advantages of such finite ground plane models are clear. First, the effects of a finite size ground plane on the input impedance of antennas can be accurately taken into account. Secondly, the diffraction of the radiated field on the edges of finite size ground planes can also be studied. This will give an idea of the back-radiation of microstrip antennas, including the side-lobe levels which might be expected in their radiation patterns. Both elements are of key importance in the design of antennas, and up to now they could only be evaluated through measurements, or with lengthy numerical calculations using techniques such as the finite elements or the finite differences [Ciampolini *et al.* 1996].

Let us now consider the basic microstrip antenna with finite size ground plane represented in Fig. 6. Opposite to the case of an infinite ground plane, where the excitation is injected only through the patch while the ground plane is included in the Green's functions, the model must be modified in the finite ground case so that the finite ground plane is connected to the generator and surface currents must be free to flow through this connection. This is obtained by using a "mirror" attachment model in the ground plane with the sign of the currents reversed. Also, the potential of the ground plane is set to zero by means of a numerical treatment acting on the MPIE formulation. Fig. 7 presents the basic idea of the extended attachment mode. If we take again the case of a transition from a coaxial cable to a microstrip line, but where the size of the microstrip's ground plane is now finite (Fig. 7a), the equivalent excitation model can be represented with a voltage generator connected to the microstrip line as in the previous case, but with the grounded terminal now connected to the physical ground plane (Fig. 7b). As depicted in the figure, the currents flowing through the two terminals of the generator must be the same. Therefore the same "spreading" behavior of the current must be imposed in both the microstrip patch and the ground plane. This behaviour can be obtained in the MoM implementation by introducing one half basis function on the ground

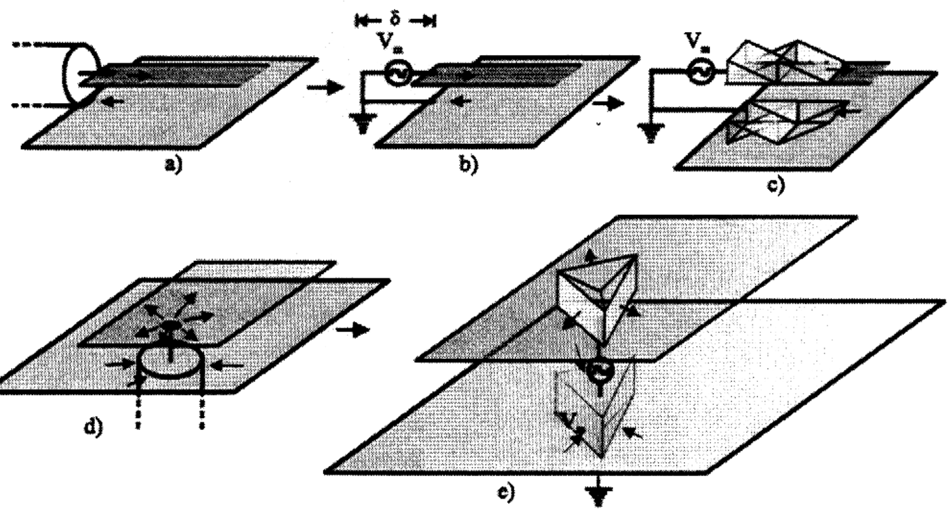


Figure 7: Attachment mode for patch antennas on finite ground planes. a) Colinear transition between a coaxial probe and a microstrip line. b) Delta-Gap voltage model applyed to a coaxial probe-fed microstrip line. c) Associated MoM description of the excitation model. d) Coaxial probe-fed patch antenna. e) Associated MoM description of the of the excitation of a probe-fed patch antenna

plane for each of these present in the microstrip and linking the two halves together to form an entire basis function (see Fig. 7c), i.e. only one unknown term for each couple is present in the MoM matrix [Tiezzi *et al.* 1999]. This implies that the free edges of the two half basis function must have the same length. Applying now the same scheme to the probe-fed patch antenna represented in Fig. 7d, starting from the attachment mode sketched in Fig. 2e, we obtain the new attachment mode composed by three (or less) half basis functions on the patch and the same number of half basis functions with opposite sign on the ground plane.

To demonstrate the effectiveness of the derived model, the antenna in Fig. 4 has been simulated with a ground plane of width $W_g = 214$ mm and length $L_g = 214$ mm for again three position of the coaxial excitation. The agreement between theory and measurement (Fig. 8)is rather good. Indeed our model can work for any size of ground plane from the completely unbalanced antenna (infinite ground plane) to a perfectly balanced antenna (ground plane having the patch's size). The latter case has been tested for an antenna on a RT/DUROID 5870 substrate with thickness $h = 1.57$ mm and relative dielectric constant $\varepsilon_r = 2.33$. With respect to Fig. 6 the dimensions of the antenna are $W_p = W_g = 120.1$ mm, $L_p = L_g = 79.5$ mm, $X_p = 60$ mm, $Y_p = 29$ mm. The results are presented in Fig. 9. The agreement between measured and computed results is ex-

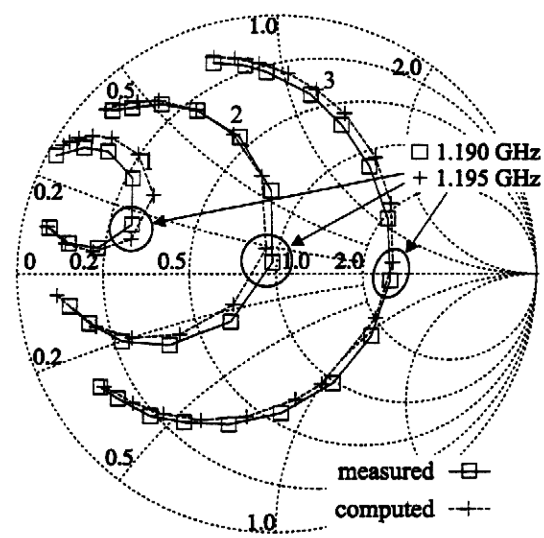


Figure 8: Measured versus simulated results for the patch antenna shown in Fig. 4, when the new excitation model is used: W_g = 214 mm, L_g = 214 mm. (increment 5 MHz clockwise, measurement reproduced from [James and Hall 1989])

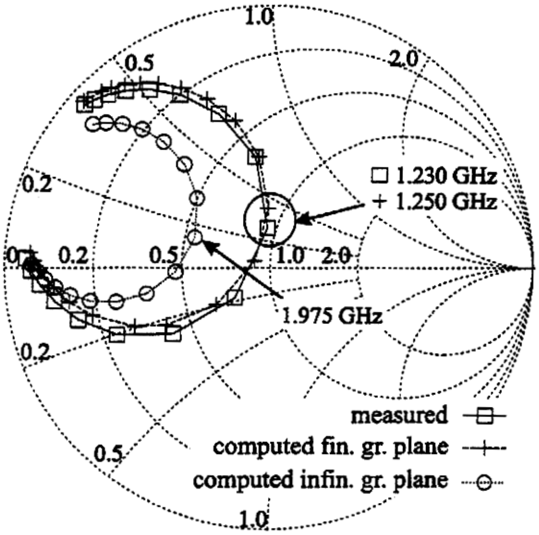


Figure 9: Measured versus simulated results for a perfectly balanced patch antenna: $W_p = W_g = 120.1$ mm, $L_p = L_g = 79.5$ mm, $L_p = L_g = 80$ mm, $X_p = 60$ mm, $Y_p = 29$ mm. Substrate: RT/DUROID 5870, $h = 1.57$ mm, $\varepsilon_r = 2.33$, $tan\delta = 0.0012$. (increment 2.5 MHz clockwise)

cellent. As a matter of comparison, the result obtained using the infinite ground plane model has also been included and it show that in this extreme case the infinite ground plane approximation is definitely too rough.

4.1 RADIATION PATTERNS

Another interesting aspect of the excitation model derived in this paper is the prediction of the back radiation and the side lobe levels of microstrip printed antennas. In the present work the far field radiated by the structure has been computed with the aid of asymptotic expressions for the multilayered media Green's functions, valid for large values of source-observer distances. These asymptotic expressions are based on the use of the saddle point method, which allows the analytical evaluation of a Fourier integral by just considering the contribution of the function at the saddle point [Mosig and Gardiol 1982]. It is important to have in mind that in a multilayered medium, horizontal currents can in general produce both horizontal and longitudinal (along z) components of the electromagnetic fields. This comes from the fact that the dyad associated with the magnetic vector potential is not a diagonal dyad, but it rather contains off diagonal elements. For instance, if the so called Sommerfeld choice is selected, then the whole magnetic vector potential dyad can be written, for only horizontal currents, as [Mosig and Gardiol 1985, Mosig 1989]

$$\begin{aligned} \overline{\overline{G}}_A &= \left(\hat{e}_x\, G_A^{xx} + \hat{e}_z\, G_A^{zx}\right)\hat{e}_x \\ &\quad + \left(\hat{e}_y\, G_A^{yy} + \hat{e}_z\, G_A^{zy}\right)\hat{e}_y \end{aligned} \tag{16}$$

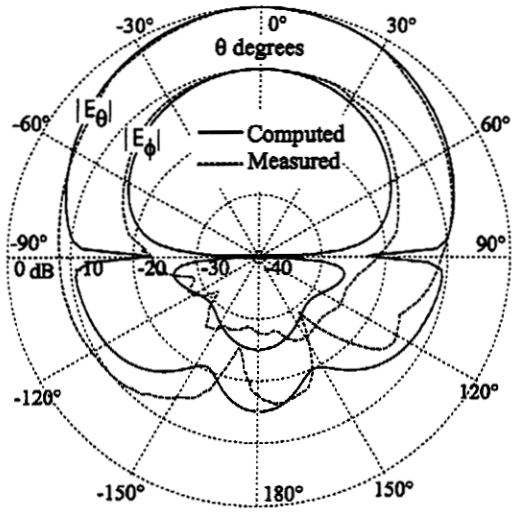
where, as already said, the spectral domain Green's functions appearing in equation (16) are derived from voltages and currents computed in the equivalent transmission line network of Fig. 1(b), as described in [Mosig and Gardiol 1988, Michalski and Mosig 1997]. For the Green's functions of interest in (16) one obtains [Michalski and Mosig 1997]

$$\begin{aligned} \tilde{G}_A^{xx} &= \tilde{G}_A^{yy} = \frac{V_J^{TE}}{j\,\omega}, \\ \tilde{G}_A^{zx} &= -\frac{\mu}{k_\rho^2} j\, k_x \left(I_j^{TM} - I_J^{TE}\right), \\ \tilde{G}_A^{zy} &= -\frac{\mu}{k_\rho^2} j\, k_y \left(I_j^{TM} - I_J^{TE}\right), \end{aligned} \tag{17}$$

where TE, TM denotes transverse electric and transverse magnetic (with respect to the z-axis) waves, and the transverse wavenumbers are given by [Mosig and Gardiol 1982]: $k_\rho = k_0 \sin\theta$, $k_x = -k_0 \sin\theta \cos\varphi$, $k_y = -k_0 \sin\theta \sin\varphi$.
The main difficulty is then reduced to the calculation of these Green's functions in the spatial domain. For this purpose the inverse Fourier integral is evaluated with the saddle point technique, and, as shown in [Mosig and Gardiol 1982], one finally obtains in the spatial domain the following simple relation

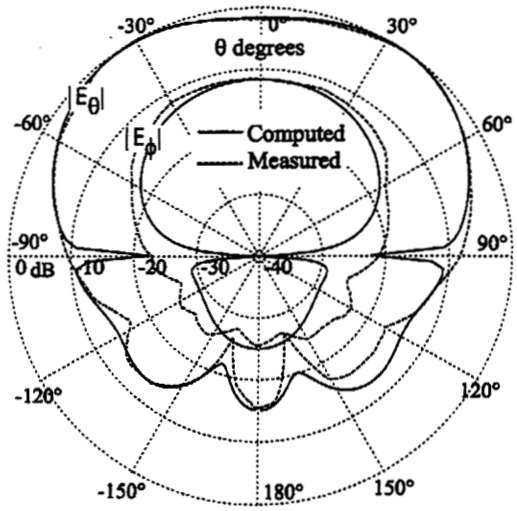
$$G_A^{st} = j\, k_0 \cos(\theta)\, \tilde{G}_A^{st}\, \frac{\exp(-j\, k_0\, R)}{R} \tag{18}$$

where $s, t = x, y, z$, and R is the source-observer distance. It is important to remark that for the derivation of equation (18) the spectral domain Green's functions are assumed to have a free space dependence of the type: $\exp(-j\,\beta\, z)$. The main implication of this is that the voltages and currents in equation (17) must be computed at the first air-dielectric interface for: $0 < \theta < \pi/2$, and they must be computed at the last air-dielectric interface for: $\pi/2 < \theta < \pi$. Having all these computational details in mind, an accurate evaluation of the radiation patterns of microstrip antennas printed on finite size ground planes has been carried out. Figs. 10, 11 and 12. present the measured and computed results



(a) E-plane.

Figure 10: Radiation patterns of the printed patch antenna shown in Fig. 4. Ground plane size: $W_g = 60$ mm, $L_g = 60$ mm. Frequency is 5.020 GHz. (measurement reproduced from [Bokhari *et al.* 1992])



(a) E-plane.

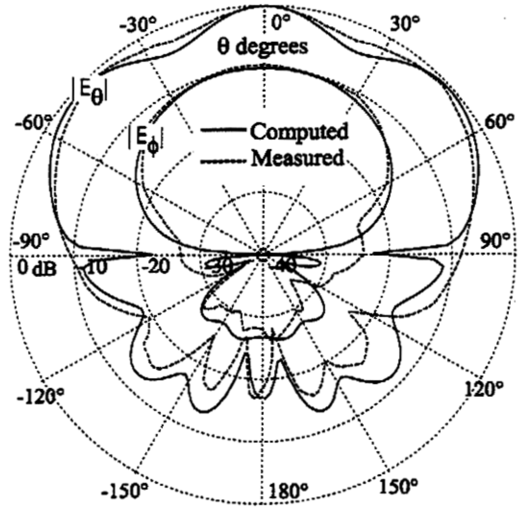
Figure 11: Radiation patterns of the printed patch antenna shown in Fig. 4. Ground plane size: $W_g = 90$ mm, $L_g = 90$ mm. Frequency is 5.020 GHz. (measurement reproduced from [Bokhari *et al.* 1992])

for the E and H-plane radiation patterns of the antenna shown in Fig. 4 with ground plane size: $W_g = 60\ mm$, $L_g = 60\ mm$, $W_g = 90\ mm$ $L_g = 90\ mm$ and $W_g = 180\ mm$ $L_g = 180\ mm$ (respectively $\lambda_0 \times \lambda_0$, $1.5\lambda_0 \times 1.5\lambda_0$ and $3\lambda_0 \times 3\lambda_0$ at 5.02 GHz). The results presented indicate that the agreement is good, and in particular the predicted level of back radiation is approximately the measured one. It is important to mention that a model using an infinite ground plane gives no information concerning the level of back radiation of the antenna, which is assumed to be zero. On the contrary, with the new excitation model derived in this paper, an accurate estimation of the back radiation level can be obtained. It must be also pointed out that the present model still uses layered Green's functions and doesn't include neither the radiation of the probe itself nor the effect of the dielectric layer finiteness.

These two aspects of the problem could also be included in the model by means of respectively, vertical conduction and polarisation currents and work towards this goal is in progress. However the results of figures 10-12 shows clearly that the only noticeable improvement would be the filling of the deep nulls at $\pm 90°$ and that except for this minor correction, our model in its current status follows closely the measured values, while still retaining a reasonable simplicity which would be lost if the aforementioned effects are included.

5 CONCLUSION

A new excitation model for coaxially fed printed microstrip antennas, developed in the frame of the mixed potential integral equation (MPIE) and the method of moments (MoM), has been presented. Moreover, a modified version of this model allows the analysis of these antennas on finite size ground planes. This model has been successfully applied to the prediction of input impedances for patches above ground planes whose size ranges from the patch size to infinity. With this approach, scattering from ground plane edges can be taken into account and full range (including backside scattering) radiation patterns can also be predicted. The paper has first presented the theoretical basis of the new derived excitation method, including the numerical details needed for a correct far field computation. Theoretical results have been compared with measurements, for both the input impedance and the radiation patterns. Comparisons have revealed that the accuracy achieved with the new excitation method is very satisfactory, and in particular the backside radiation and side lobe levels of real life printed antennas can accurately be predicted.



(a) E-plane.

Figure 12: Radiation patterns of the printed patch antenna shown in Fig. 4. Ground plane size: $W_g = 180$ mm, $L_g = 180$ mm. Frequency is 5.020 GHz. (measurement reproduced from [Bokhari *et al.* 1992])

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