Efficient Analysis for the Design Refinement of Large Multilayered Printed Reflectarrays

Francesca DE VITA[1], *Patrizio DE VITA* [1,2], *Angelo FRENI*[1]

[1] Dept. of Electronics and TLC, University of Florence, Via di S. Marta 3, 50139 Florence, Italy
[2] IDS, Ingegneria Dei Sistemi S.p.A, ViaLivornese 1019, 56010 Pisa, Italy

francesca.devita@gmail.com, p.devita@ids-spa.it, freni@unifi.it

Abstract. *In this paper, we present an efficient numerical technique for the analysis of a reflectarray and its design refinement by the characterization of the "actual" influence of each radiating element when embedded in the antenna structure. The method makes use of the MLayAIM, a fast full-wave formulation suitable for the analysis of electrically large multilayered printed arrays which have one or more planar metallizations and vertical conductors. The low numerical complexity of the analysis method allows the development of a recursive procedure that, starting from the equivalent currents relevant to each cell of the reflectarray when this is immersed in the actual antenna layout, calculates the real phase-shift introduced by each radiating element and corrects its dimensions to better fit the antenna requirements.*

Keywords

Reflectarrays, Method of Moments, Adaptive Integral Method.

1. Introduction

Recently, printed reflectarrays have received increasing attention [1]. Compared to conventional shaped reflectors, the manufacturing processes are much simpler, because they are based on conventional techniques used in multi-layer printed circuits [2]. In addition, when contoured beams are desired, they do not require any specific mould as in shaped reflectors. As a result, increasingly tight requirements have started to be accommodated and this has significantly complicated their structure.

Several approximations are necessary in practical design work-flows to design the antenna with reasonable computational, human and time resources [3]. As a consequence, pre-prototyping refinements of the initial design are very desirable to cut production and/or prototyping costs, especially in view of the complexity of the reflectarray. Full-wave modeling methods play a key role in this scenario for their potential ability to take into account all the physical issues of interest, such as complex multilayered substrates, mutual coupling among cells, finiteness of the radiating aperture, etc. However, to be useful, they must be accurate and at the same time flexible and fast, since the computational electromagnetic task associated to the reflectarray is, in many cases, a challenge [4]. They are electrically large overall, and require a very fine, or non-uniform meshing of the radiating elements because these latter are resonant. The phase response critically depends on the accuracy with which the resonant behavior is captured in the numerical model.

For this purpose, a two steps approach has been used: first, the reflectarray is designed using a computational procedure similar to that described in [3], which has been proven to yield accurate patterns and absolute gains. Specifically, a Physical Optics approach is used where the phase-shift introduced by each radiating element of the reflectarray is supposed equal to the phase of the reflection coefficient of the zero order Floquet mode calculated by assuming the element embedded in a periodic lattice. Then, a full-wave analysis of the entire antenna is performed by using the MultiLayer Adaptive Integral Method (MLayAIM) [5], that is highly suitable for the analysis of multilayered structures, such as reflectarrays, which present a large number of planar metallization and a few vertical conductors. MLayAIM is an extension of the AIM formulation [6] in which the array patches can be of arbitrary shape and orientation and are modeled with sub-domain triangular basis functions. This method makes use of a 2D-FTT/CG scheme, reducing the CPU time per iteration to O(Nlog2N) and the memory requirement to O(N). No restrictions are present on the thickness and the number of dielectric layers that separate each metallization or the vertical conductors can cross.

The full-wave analysis allows us to estimate the equivalent currents relevant to each cell of the reflectarray when this is immersed in the actual antenna layout. Thus, the right mutual coupling effect is taken into account. From these currents, the actual element factor of each cell is computed and compared to the one resulting in the step one, where the radiating element is treated as a cell of a periodic lattice. The results of the analysis are then reused in the design procedure in order to better fit the antenna requirements.

2. MultiLayer Adaptive Integral Method (MLayAIM)

Consider an arbitrarily microstrip structure consisting of N_m metalized layers in a multilayer medium, as that sketched in Fig. 1. Each layer is characterized by relative permittivity ε_{ri}, relative permeability μ_{ri}, and thickness h_i. As well known, the equivalent currents $\vec{J}_s$ relevant to the microstrip structure can be found by solving the following MPIE [7]

$$\hat{n}\times\vec{E}^i = j\omega\hat{n}\times\left\langle \underline{\underline{G}}^A,\vec{J}_s\right\rangle - \frac{1}{j\omega}\hat{n}\times\vec{\nabla}\left\langle G^{\Phi},\vec{\nabla}'\cdot\vec{J}_s\right\rangle, \quad (1)$$

where $\underline{\underline{G}}^A$ and G^{Φ} are the Green's function for the vector and scalar potential, respectively. The notation $\langle\ ,\ \rangle$ is used for integrals of dot products of two functions separated by the comma over their common spatial support. To solve the integral equation (1), one first subdivides the surface S of the metallic structure into small triangular patches, and the unknown current is expanded by using a suitable set of basis functions $\vec{j}_n(\vec{r})$ (e.g. the Rao-Wilton-Glisson (RWG) basis functions [8]), i.e., $\vec{J}_n(\vec{r}) = \sum_{n=1}^{N} I_n\vec{j}_n(\vec{r})$, where I_n are unknown coefficients and N the number of interior edges (unknowns). Then, the MoM solution technique is applied to the surface integral equation (1), and the matrix equation is $\mathbf{Z}I = V$.

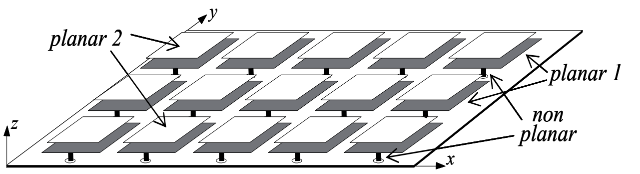

Fig. 1. Geometry of the problem (array of 5x3 elements).

The impedance matrix $\mathbf{Z} = \mathbf{Z}^e + \mathbf{Z}^p$ is then divided into the sum of a non-planar (external) matrix $\mathbf{Z}^e$ and a planar matrix $\mathbf{Z}^p$. The planar matrix represents the interaction relevant to those basis functions that lie on a z-constant plane. In particular, in reflectarray analysis these basis functions represent the majority; hence, the external matrix $\mathbf{Z}^e$ is a very sparse matrix and its storage does not present any problem. Hence, in the following, we will focus our attention on the planar-matrix $\mathbf{Z}^p$ evaluation only.

Let P be the number of metallizations on which the planar part of the structure lies, each element Z^p_{mn} of the planar-matrix $\mathbf{Z}^p$ can be expressed, assuming the Galerkin discretization scheme, as follows

$$\begin{aligned} Z^p_{mn} &= j\omega\iint_S \vec{j}_m(\vec{r})\cdot\iint_{S'} G^a_{pp'}(\vec{r},\vec{r}')\cdot\vec{j}_n(\vec{r}')\,d\vec{r}'d\vec{r}\, + \\ &+\frac{1}{j\omega}\iint_S \vec{\nabla}\cdot\vec{j}_m(\vec{r})\iint_{S'} G^{\Phi}_{pp'}(\vec{r},\vec{r}')\vec{\nabla}'\cdot\vec{j}_n(\vec{r}')\,d\vec{r}'d\vec{r} \end{aligned} \quad (2)$$

where the basis function $\vec{j}_n(\vec{r})$ lies on the p' metallization, while $\vec{j}_m(\vec{r})$ on the p metallization. $G^a_{pp'}$ and $G^{\Phi}_{pp'}$ are the xx-component of the dyadic Green's function and the scalar potential, respectively, due to the interaction between the metallization p and p'. It is worth noting that to solve the planar problem by using the formulation proposed in [9] it is sufficient to evaluate the xx-component of the dyadic Green's function.

As in the classical Adaptive Integral Method (AIM) [6], the efficiency is achieved by splitting the reaction integral domain of the planar part into a near-interaction (strong) region, and a non-near-interaction (weak) region, based on an appropriate choice of the error of the reaction integral terms. Hence, the planar impedance matrix $\mathbf{Z}^p$ is further divided into the sum of a strong- and a weak-matrix, i.e., $\mathbf{Z}^p = \mathbf{Z}^s + \mathbf{Z}^w$, where $\mathbf{Z}^s$ represents the (strong) near-field interaction and $\mathbf{Z}^w$ the (weak) non-near-field interaction. The former one is calculated $\mathbf{Z}^s = \mathbf{Z}^p - \mathbf{Z}^w$ similarly as in [6], and is the only one that we have to store. Concerning the weak- matrix, we introduce a set of auxiliary basis functions $\vec{\psi}_n(\vec{r})$ and $\psi^d_n(\vec{r})$ which produce a good approximation of the field radiated by a RWG basis function at a large distance. In particular, for each RWG basis function lying on a plane $z = z_p$, we chose a set of $L = (M+1)^2$ point-like current elements (usually $L = 9$) located at the nodes of a regular Cartesian two–dimensional grid, parallel to the x, y plane and located at $z = z_p$, i.e.,

$$\vec{j}_n(\vec{r}) \cong \vec{\psi}_n(\vec{r}) = \sum_{i=1}^{L}\vec{\Lambda}_{n,i}\delta(\vec{r}-\vec{r}_i), \quad (3)$$

$$\nabla\cdot\vec{j}_n(\vec{r}) \cong \psi^d_n(\vec{r}) = \sum_{i=1}^{L}\vec{\Lambda}^d_{n,i}\,\delta(\vec{r}-\vec{r}_i) \quad (4)$$

where $\vec{\Lambda}_{n,i}$, $\Lambda^d_{n,i}$ are the translation coefficients of the expansion and $\vec{r}_i$ is the position vector of the i-th grid node. The translation coefficient are chosen so as to reproduce the first $(M+1)^2$ multipole moments of the original basis function [4], i.e., for $0 \le q_1, q_2 \le M$,

$$\begin{aligned} &\sum_{i=1}^{L}\vec{\Lambda}_{n,i}(x_{n,i}-x_c)^{q_1}(y_{n,i}-y_c)^{q_2} = \\ &= \iint_{T_n}\vec{f}_n(x,y,z_p)(x-x_c)^{q_1}(y-y_c)^{q_2}\,dx\,dy \end{aligned}, \quad (5)$$

$$\begin{aligned} &\sum_{i=1}^{L}\Lambda^d_{n,i}(x_{n,i}-x_c)^{q_1}(y_{n,i}-y_c)^{q_2} = \\ &= \iint_{T_n}\vec{\nabla}\cdot\vec{f}_n(x,y,z_p)(x-x_c)^{q_1}(y-y_c)^{q_2}\,dx\,dy \end{aligned}, \quad (6)$$

where, to minimize the numerical roundoff errors, x_c, y_c, z_p are the center of the RWG basis function support. The set of auxiliary basis functions allows us to approximate each element of the weak matrix as follows

$$\begin{aligned} Z^w_{mn} &= j\omega\sum_{j=1}^{L}\sum_{i=1}^{L}\vec{\Lambda}_{m,j}\cdot\vec{\Lambda}_{n,i}G^a\left(z_p, z_{p'}, \left|\vec{\rho}_i-\vec{\rho}_j\right|\right) + \\ &+\frac{1}{j\omega}\sum_{j=1}^{L}\sum_{i=1}^{L}\Lambda^d_{m,j}\,\Lambda^d_{n,i}G^{\Phi}\left(z_p, z_{p'}, \left|\vec{\rho}_i-\vec{\rho}_j\right|\right). \end{aligned} \quad (7)$$

By sorting out the RWG basis functions with respect to the P planar metallizations, it is possible to rewrite the weak matrix $\mathbf{Z}^w=\{\mathbf{Z}^w_{pp'}\}$ as a sum of $P\times P$ sub-matrices

$$\begin{aligned}\mathbf{Z}^w_{pp'} &= j\omega\left[\boldsymbol{\Lambda}^x_p\right]^T \mathbf{G}^a_{pp'}\boldsymbol{\Lambda}^x_{p'} + \\ &+ j\omega\left[\boldsymbol{\Lambda}^y_p\right]^T \mathbf{G}^a_{pp'}\boldsymbol{\Lambda}^y_{p'} + \frac{1}{j\omega}\left[\boldsymbol{\Lambda}^d_p\right]^T \mathbf{G}^\Phi_{pp'}\boldsymbol{\Lambda}^d_{p'}\end{aligned} \tag{8}$$

where $\boldsymbol{\Lambda}^x_p$, $\boldsymbol{\Lambda}^y_p$ and $\boldsymbol{\Lambda}^d_p$ are sparse matrices with each row containing only L nonzero elements. Considering the unknown vector relevant to the weak part composed by P vectors as $\underline{I}^w=\left[\underline{I}^w_1;\underline{I}^w_2;\ldots;\underline{I}^w_P\right]$, and by employing a conjugate gradient (CG) method as the iterative solver, one can write the matrix-vector multiplication $\underline{V}^w=\left[\underline{V}^w_1;\underline{V}^w_2;\ldots;\underline{V}^w_P\right]=\mathbf{Z}^w\underline{I}^w$ as follows

$$\begin{aligned}\underline{V}^w_p &= j\omega\left[\boldsymbol{\Lambda}^x_p\right]^T \sum_{p'=1}^{N_p}\mathbf{G}^a_{pp'}\boldsymbol{\Lambda}^x_{p'}\underline{I}_{p'} + \\ &+ j\omega\left[\boldsymbol{\Lambda}^y_p\right]^T \sum_{p'=1}^{N_p}\mathbf{G}^a_{pp'}\boldsymbol{\Lambda}^y_{p'}\underline{I}_{p'} + \frac{1}{j\omega}\left[\boldsymbol{\Lambda}^d_p\right]^T \sum_{p'=1}^{N_p}\mathbf{G}^\Phi_{pp'}\boldsymbol{\Lambda}^d_{p'}\underline{I}_{p'}\end{aligned}. \tag{9}$$

It is worth noting that $\mathbf{G}^a_{pp'}$ and $\mathbf{G}^\Phi_{pp'}$ are Toeplitz block matrices. The matrix-vector products in (9) are evaluated using the 2D-FFT pair. In case that vertical basis functions are not present, this gives rise to a complexity of the order $O(N\log_2 N)$. When vertical basis functions are present, the asymptotic behavior $O(N\log_2 N)$ continues to be prevalent only if their number N_v is a few hundred. Differently, for structures where the number of vertical basis functions is proportional to the area of the array, we have an asymptotic behavior $O(NN_v)$, where $N_v = N/\beta$ and β is the average density ratio between the number of unknowns and the number of vertical basis functions.

3. Element Factor of Each Reflectarray Cell

In the design procedure, each cell of the reflectarray is considered as a radiating element able to add a specific phase difference to the field incident on it. In order to calculate this phase difference for a specific frequency and direction of the incident field (supposed a plane wave) each element is considered embedded in a periodic lattice. The latter assumption, although approximate, allows us to easily obtain a reflectarray layout.

Then, the full-wave analysis of the entire antenna is performed and the actual contribution of each single radiating element is extract. In fact, among the outputs of the full-wave analysis we can easily identify the equivalent currents flowing on each reflectarray element and find the far-field radiated in any direction of interest. To this field we have also to add the influence of the multilayered structure on which the radiating element is printed. As a matter of fact, in the numerical simulation the multilayered structure is considered infinite, but now is relevant only to a single cell of the reflectarray. The influence of the cell finiteness can be conveniently considered by using the physics optics approximation in evaluating the scattering from a portion of the multilayer substrate, having the same geometry and position of the cell containing the radiating element. Thus, for each element of the reflectarray on which a locally plane wave is impinging along $\hat{\imath}$, we have to add the electric field $\vec{E}^s$ scattered by the element cell of dimension L_x, L_y to that radiated by the equivalent currents coming from the full-wave formulation. Namely,

$$\vec{E}^s = \frac{jkL_xL_y}{4\pi}\frac{e^{-jkr}}{r}\frac{\sin(u)}{u}\frac{\sin(v)}{v}\hat{r}\times\left[\vec{M}_s+\zeta_0\hat{r}\times\vec{J}_s\right] \tag{10}$$

where $u = L_xk_0\left(\hat{\imath}+\hat{r}\right)\cdot\hat{x}/2$, $v = L_yk_0\left(\hat{\imath}+\hat{r}\right)\cdot\hat{y}/2$, while

$$\vec{J}_s e^{-jk\hat{\imath}\cdot\vec{r}'} = \frac{\vec{E}^i_e}{\zeta}\left(1-\Gamma_e\right)\left(\hat{z}\times\hat{n}_\perp\right) - \frac{\vec{E}^i_h}{\zeta}\left(1-\Gamma_h\right)\left(\hat{z}\cdot\hat{\imath}\right)\hat{n}_\perp, \tag{11}$$

and

$$\vec{M}_s e^{-jk\hat{\imath}\cdot\vec{r}'} = \vec{E}^i_e\left(1+\Gamma_e\right)\left(\hat{n}\times\hat{z}\right) - \vec{E}^i_h\left(1+\Gamma_h\right)\left(\hat{z}\times\hat{n}_\perp\right). \tag{12}$$

In the previous equations $\hat{z}$ is the unit normal pointing outward the reflectarray, $\hat{n}_\perp = \hat{n}_s\times\hat{\imath}$, $\hat{n}=\hat{n}_\perp\times\hat{\imath}$, Γ_h and Γ_e are the TE_z and the TM_z reflection coefficients, respectively, of the portion of the multilayer substrate under observation, and $\vec{E}^i_h, \vec{E}^i_e$ the relevant components of the incident wave.

Acting in this way, we can regard each radiating element as a single element in the array, but its element factor now takes into account the mutual coupling between the elements of a structure that is aperiodic.

4. Results

As an example of application we have designed (by using a procedure similar to that described in [3]), and manufactured, the reflectarray shown in Fig. 2: it consists in 35×35 modified Malta Cross elements printed on a $h=1.6\,\mathrm{mm}$ grounded substrate ($\varepsilon_r=3$, $\tan(\delta)=0.003$). The element grid spacing is $11.5\,\mathrm{mm}$ so that the reflectarray results $402.5\times402.5\,\mathrm{mm}^2$ (about $16\lambda\times16\lambda$ at the central frequency of $11.7\,\mathrm{GHz}$). The modified Malta Cross element [10] presents two degrees of freedom (the element size and the slots length) which are used to guarantee that each element in the array not only provides the right phase delay at a specific frequency, but also that this phase delay varies in such a way as to compensate for the frequency variation of the incident field phase. The planar reflecting surface lays in the (x,y) plane and its centre has been chosen as the origin of the Cartesian coordinate system. The RA is illuminated by a standard horn (SIVERS IMA Philips PM7320X/01) located in $x=0.1955\,\mathrm{m}$, $y=0$, $z=0.73\,\mathrm{m}$, and radiates the maximum field in the direction $\theta=-15°$, $\phi=0°$.

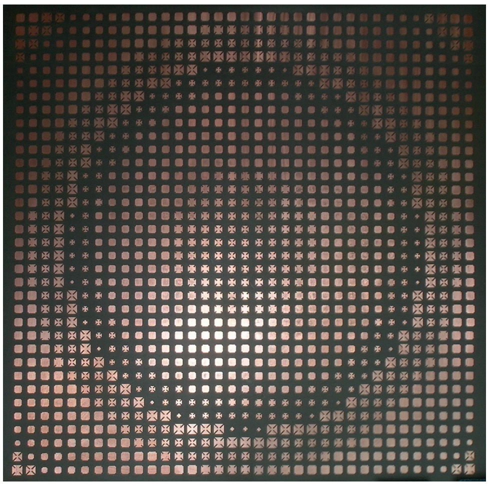

Fig. 2. Photograph of the reflectarray prototype.

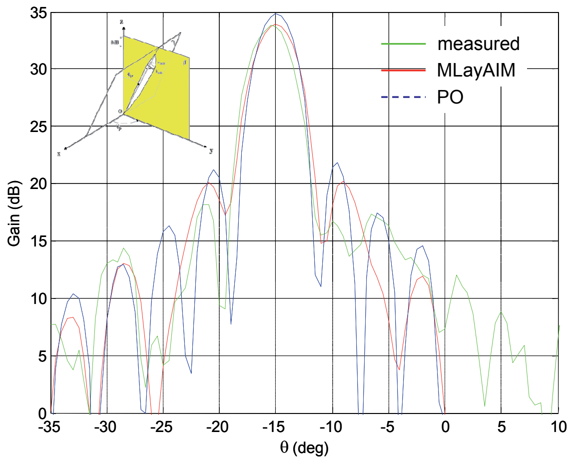

Fig. 3. Measured and calculated gain in the plane *xz* at 11.7 Ghz (vertical polarization).

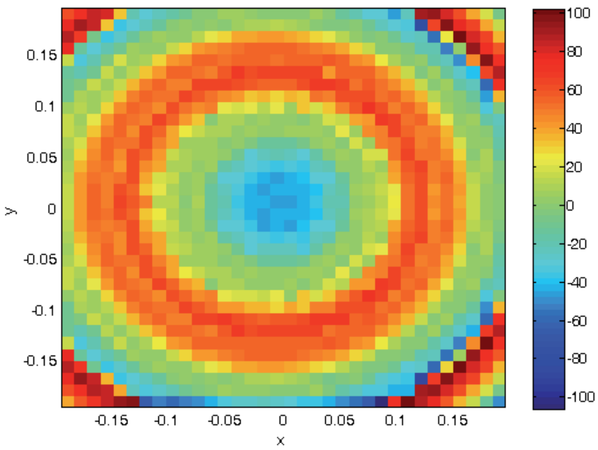

Fig. 4. Map of the absolute phase error between the ideal phase shift that each radiating element should introduce on the reflectarray plane and the one reconstructed by the MLayAIM formulation.

Fig. 3 shows the measured (solid green curve) and computed (solid red curve) gain pattern in the xz plane at 11.7 GHz. In the same figure is also shown the pattern we expected from the design [10] (dashed blue curve), where the characteristics of each radiating element have been evaluated supposing it in a periodic lattice with a periodicity of 11.5 mm both along x and y. We can note that the curve relevant to the MLayAIM solution better agrees with the measurements in comparison with the one expected from the design procedure. This disagreement can be explained by the fact that the design procedure does not take into account the mutual coupling between adjacent cells. If one look at the map of Fig. 4, where the absolute phase difference between the theoretical phase expected from the design and that calculated by using the full wave approach is shown, it is evident that the actual aperiodic configuration introduces a phase perturbation up to 80°, that slightly modifies the theoretically expected radiation characteristics of the reflectarrays.

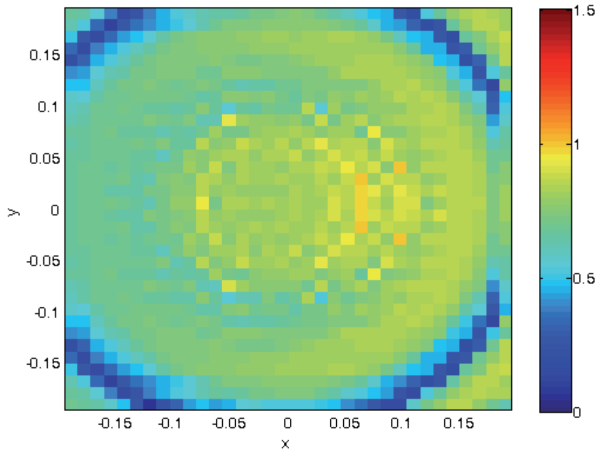

Fig. 5. Amplitude of the equivalent reflection coefficient reconstructed by the MLayAIM formulation.

Fig. 5 shows the amplitude of the ratio between the field radiated in the direction $\theta = -15°$, $\phi = 0°$ by each cell of the reflectarray and the field that should be radiated in the same direction when the power incident on the cell is scattered isotropically. It is worth noting that a few cells present a ratio greater than one. We can also note that on the periphery of the reflectarray several cells are poorly radiating in the main lobe direction (i.e., the ratio between the scattered and the incident field is less that 0.3). This phenomenon is explainable by considering that a strong mutual coupling between the radiating elements is present and each cell is about half wavelength. Hence, the contribution of each cell is not isotropic and, depending of their position in the reflectarray, the cells show different directivity values in the direction $\theta = -15°$, $\phi = 0°$.

The map of the absolute phase error in Fig. 4 allows a refinement of the antenna design. Namely, the radiating elements dimensions are modified in such a way as to reduce the error between the ideal phase shift and the one reconstructed by using the MLayAIM formulation. To operate this correction, the relation between the phase shift introduced by the element and its dimensions calculated by considering the element embedded in a periodic lattice is still used. As a matter of fact the phase error we would like to compensate is quite small and we can assume that the periodic approximation is enough accurate to model the small difference required in the element dimensions. A damping factor is, however, introduced to assure the convergence.

The design so obtained is once again analyzed by using the MLayAIM formulation, a new map of the absolute phase error calculated, and the procedure repeated recursively. Each step requires about 19 minutes on a Centrino 1.8 GHZ, 2 GB RAM.

After a few steps the procedure converges to the result shown in Fig. 6 where it can be noted an increase of about 1dB with respect to the first design. If we further repeat the refinement procedure the improvements is still present but it is almost unnoticeable.

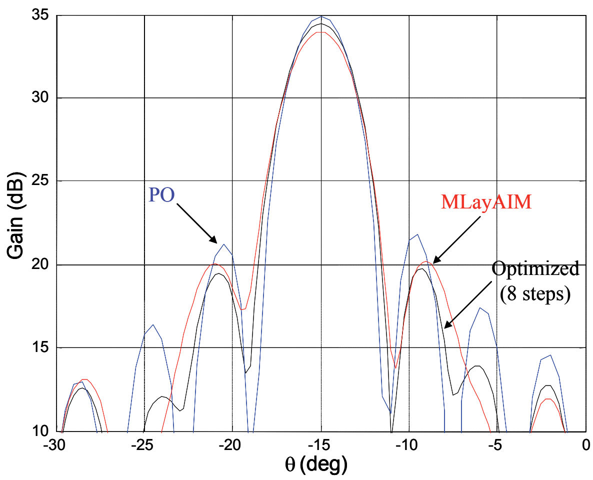

Fig. 6. Gain in the vertical plane before and after the optimization process.

5. Conclusions

In this paper we have presented a method that allows the full-wave analysis of large printed reflectarray. The method maintains the accuracy of the standard MoM but drastically improves the computation efficiency in terms of both memory and computational time requirements. The results obtained by using the proposed full-wave technique are then reused in the design procedure in order to better take into account of the mutual coupling between the radiating elements of a structure that is actually aperiodic. A simple recursive procedure has been described to better fit the antenna requirements.

Acknowledgements

This work has been partially supported by Thales Alenia Space France and by the Italian Ministry of Education, University and Research under MIUR-PRIN grant 2008N3B2LP_003.

References

[1] HUANG, J., ENCINAR, J. A. *Reflectarray Antennas*. IEEE Press, John Wiley, 2008.

[2] POZAR, D. M., TARGONSKI, S. D., POKULS, R. A shaped-beam microstrip patch reflectarray. *IEEE Trans. Antennas Propagat*, 1999, vol. 47, no. 7, p. 1167-1173.

[3] POZAR, D. M., TARGONSKI, S. D., SYRIGOS, H. D. Design of millimeter wave microstrip reflectarrays. *IEEE Trans. Antennas Propagat.*, 1997, vol. AP–45, no. 2, p. 287–296.

[4] ENCINAR, J. A., DATASHVILI, L. SH., ZORNOZA, J. A., ARREBOLA, M., SIERRA-CASTANER, M., BESADA- SANMARTIN, J. L., BAIER, H., LEGAY, H. Dual-polarization dual-coverage reflectarray for space applications. *IEEE Trans. Antennas Propag.*, 2006, vol. 54, no. 10, p. 2827–2837.

[5] DE VITA, P., DE VITA, F., DI MARIA, A., FRENI, A. An efficient technique for the analysis of large multilayered printed arrays. *IEEE Antennas and Wireless Propagation Letters*, 2009, vol. 8, p. 104–107.

[6] BLESZYNSKI, E., BLESZYNSKI, M., JAROSZEWICZ, T., AIM: Adaptive integral method for solving large–scale electromagnetic scattering and radiation problems, *Radio Science*, 1996, vol. 5, pp. 1225–1251.

[7] MOSIG, J. R. Arbitrarily shaped microstrip structures and their analysis with a mixed potential integral equation. *IEEE Trans. Microwave Theory Tech.*, Feb. 1988, vol. 36, no. 2, p. 314–323.

[8] RAO, S. M., WILTON, D. R., GLISSON, A. W. Electromagnetic scattering by surfaces of arbitrary shape. *IEEE Trans. Antennas Propag.*, 1982, vol. 30, p. 409–418.

[9] MICHALSKI, K. A., MOSIG, J. R. Multilayered media Green's functions in integral equation formulations. *IEEE Trans. Antennas and Propagat.*, 1997, vol. AP–45, p. 508–519.

[10] DE VITA, P., FRENI, A., DASSANO, G. L., PIRINOLI, P., ZICH, R. E. Broadband element for high-gain single-layer printed reflectarray antenna. *Electronics Letters*, 2007, vol. 43, n. 23, p. 1247–1249.

About Authors ...

Francesca DE VITA received the Laurea degree in Electronic Engineering and the Ph.D. degree in Informatics and Telecommunication Engineering from the University of Florence, Florence, Italy, in 1999 and 2004, respectively. Then, she was a Research Assistant in the Applied Electromagnetic Laboratory at the University of Florence. Her research interests include electromagnetic theory, and numerical methods with applications for array antennas and multilayered structures.

Patrizio DE VITA received the M.S. degree in Electronic Engineering and the Ph.D. degree in Information and Telecommunication Engineering from the University of Florence, Florence, Italy, in 1999 and 2004, respectively. From 1999 to 2008 he has served as a Research Assistant at the Electronics and Communications Department of University of Florence. Since 2009 he is with the computational electromagnetics section of IDS, Ingegneria dei Sistemi S.p.A. His main research interests are on numerical techniques for electromagnetic radiation and scattering problems.

Angelo FRENI received the Laurea (Doctors) degree in Electronics Engineering from the University of Florence in 1987. Since 1990 he is with the Department of Electronic Engineering of the University of Florence, Italy, first as Assistant Professor and from 2002 as Associate Professor of electromagnetism. From 1995 to 1999 he has also been an Adjunct Professor at the University of Pisa, Italy and in 2010 Visiting Professor at the TU Delft University of Technology, Delft, The Netherlands. During 1994, he was involved in research at the Engineering Department of the University of Cambridge, UK, concerning the extension and the application of the finite element method to the electromagnetic scattering from periodic structures. Between 2009 and 2010 he also spent one year as a researcher at the TNO Defence, Security and Safety, The Hague, The Netherlands, working on the electromagnetic modeling of kinetic inductance devices and their coupling with array of slots in THz range. His research interests include meteorological radar systems, radiowave propagation, numerical and asymptotic methods in electromagnetic scattering and antenna problems, electromagnetic interaction with moving media and remote sensing. In particular, part of his research concerned numerical techniques based on the integral-equation, with focus on domain-decomposition and fast solution methods.