

# MODELISATION OF PROBE FEED EXCITATION USING ITERATIVE METHOD

H. Zairi\*, A. Gharsallah\*, A. Gharbi\*, H. Baudrand\*\*

\* Department of Physics, Faculty of Sciences of Tunis

2092 El Manar Tunisia, e-mail: [Ali.gharsallah@fst.rnu.tn](mailto:Ali.gharsallah@fst.rnu.tn)

\*\* ENSEEIHT, Toulouse, France

## ABSTRACT

This paper presents an implementation of an iterative method based on the waves concept for analyzing patch antennas fed by coaxial probes. This method includes a two-dimensional fast Fourier transform (FFT-2D) in a wave guide environment. The method has the advantage of simplicity in that it does not involve basis functions and inversion of matrices, as used in other calculation methods. Therefore, it is capable of analyzing larger bodies than other classical techniques. An implementation of the iterative calculation is shown for the extraction of S parameters of microwave components and antennas. The good agreement between simulation results and experimental published data justifies the design procedure and validates the present analysis approach.

**INDEX TERMS** : Iterative method, fast Fourier transform in waveguide environment, probe feed.

## I- INTRODUCTION

Microstrip patch antennas are widely used in wireless communication because of their advantages, such as being low profile, light weight, and conformal. Different numerical electromagnetic analysis techniques such as the method of moments [1], the finite elements method [2], and the finite difference time domain method can be used to accurately simulate the microstrip antenna [3]. In most cases those numerical techniques are not practical to use

directly in CAD software for design and optimisation purposes, due to the enormous amount of computer time required. Circuit simulators on the other hand are very fast. However, models of microwave integrated circuits used in circuit simulators are often inaccurate or even invariable. To overcome these difficulties, the use of the iterative method, which is based on the concept of waves, has been proposed. It consists of generating a recursive relationship between incident waves and reflected waves at the interface containing the circuit which is divided into cells [4]. A high computational speed has been achieved by using 2D fast Fourier Transform in wave-guide environment [5].

In this paper, a general implementation of the iterative method is proposed to treat microstrip patch antennas. The theory as well as its procedure implementation is described. The numerical results are compared to measured data [8] to establish the validity and usefulness of the iterative method given in this study.

## II- GENERAL FORMULATION OF ITERATIVE METHOD

We consider the shielded microstrip circuit, assumed to be loss-less, presented in Fig. 1.

The air-dielectric interface  $\pi$  is divided into cells denoted by three sub-domains corresponding to metal, source and dielectric.

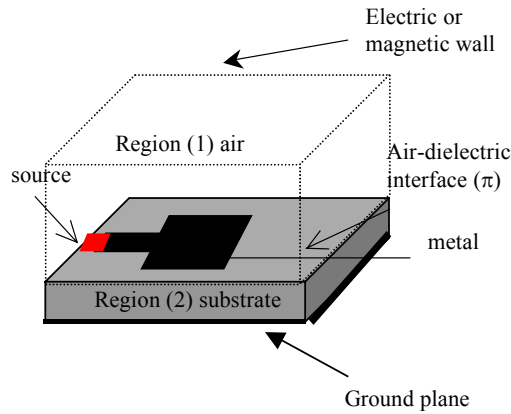


Figure . Typical planar circuit.

The wave concept is introduced by writing the transverse (plane  $\pi$ ) electric field  $E_i$  and current density  $J_i$  in terms of incident and reflected waves [5]. It leads to the following set of equations:

$$\begin{cases} A_i = \frac{1}{2\sqrt{Z_{0i}}} (E_i + Z_{0i}J_i) \\ B_i = \frac{1}{2\sqrt{Z_{0i}}} (E_i - Z_{0i}J_i). \end{cases} \quad (1)$$

$J_i$  is defined as follows:

$$J_i = H_i \times n, \quad n \text{ is oriented as the incident waves } A_i.$$

$Z_{0i}$  is the characteristic impedance of the medium  $i$  ( $i = 1, 2$ ). It is equal to:

$$Z_{0i} = \sqrt{\frac{\mu_0}{\epsilon_0 \epsilon_{ri}}}.$$

In order to generate two waves,  $B_{ix}$  and  $B_{iy}$ , in the space domain, the structure is excited by an electric planar source. The model of the source is an electric field  $E_{oi}$  equivalent to a magnetic current density.

The Fast Fourier Transform (FFT-2D) in waveguide environment is then used to obtain two spectral waves  $B_i^{TE}$  and  $B_i^{TM}$  in each region. Then, these spectral waves are reflected in the spectral domain of the Region(1) and Region(2) as described in Fig. 2.

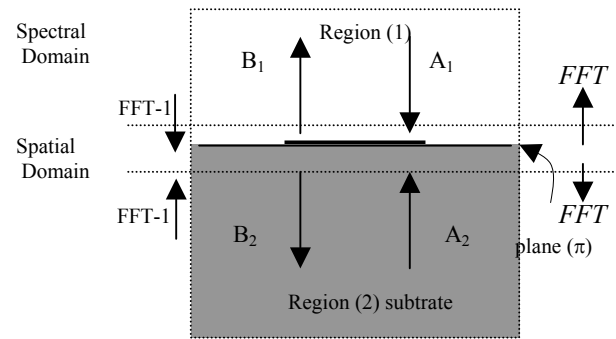


Figure 2. Definition of incident and diffracted waves.

The travelling part of these waves is then stored in memory, whereas the evanescent part constitutes the incident waves for the second iteration. The implementation of the iterative process consists of establishing a recursive relationship between waves (incident and reflected) at the  $q$  and  $q-1$  iteration.

Initially, the interface circuit (plane  $\pi$ ) on which the boundary conditions have to be satisfied (spatial domain) must be meshed.

Let us note  $H_d$  and  $H_m$  the indicator functions of respectively the dielectric and metal. These are equal to one in the considered domain and zero elsewhere. Due to the continuity relationship ( $E_{t1}=E_{t2}$  and  $J_{1+} + J_{2-}=0$  on the dielectric,  $E_{t1}=E_{t2}=0$  on the metal) in each point of the discontinuity plane, it is easy to deduce from the equation (1) and (2) the following system:

$$\begin{bmatrix} B_1 \\ B_2 \end{bmatrix} = \begin{bmatrix} -H_m \frac{1-N_2}{1+N_2} H_d & \frac{2N}{1+N_2} H_d \\ \frac{2N}{1+N_2} H_d & -H_m \frac{1-N_2}{1+N_2} H_d \end{bmatrix} \begin{bmatrix} A_1 \\ A_2 \end{bmatrix}. \quad (3)$$

Then the scattering matrix corresponding to the metal and dielectric domains can be expressed as follows:

$$\Gamma\pi = \begin{bmatrix} -H_m - \frac{1-N^2}{1+N^2} H_d & \frac{2N}{1+N^2} H_d \\ \frac{2N}{1+N^2} H_d & -H_m - \frac{1-N^2}{1+N^2} H_d \end{bmatrix}$$

$H_m=1$  on the source and 0 elsewhere.

$H_d=1$  on the source and 0 elsewhere.

$$N = \sqrt{\frac{Z_{o1}}{Z_{o2}}}$$

- Sub-domain of the source

There are numerous possibilities for choosing the source. The most simple consist in a realistic description of the excitation by a microstrip line (Fig.1).

This source generates two waves on both sides of the interface. The boundary condition on the source can be written as follows:

$$\begin{bmatrix} B_1 \\ B_2 \end{bmatrix} = \begin{bmatrix} -1 & 0 \\ 0 & -1 \end{bmatrix} \begin{bmatrix} A_1 \\ A_2 \end{bmatrix} + \begin{bmatrix} \frac{E_0}{\sqrt{Z_{o1}}} \\ \frac{E_0}{\sqrt{Z_{o2}}} \end{bmatrix}$$

$$E_1 = E_2 = E_0 - Z_0(J_1 + J_2).$$

Then the scattering matrix on the sub-domain source is expressed as

The drawback of this source is that we must use a box with electric walls to connect the source to the ground plane.

Another technique of spatial excitation is described in Section (III).

The waves  $B_i^\alpha$  are reflected on the upper and lower parts of the structure. Consequently, in spectral domain the relation between waves become:

$$\begin{bmatrix} A_i^{TE} \\ A_i^{TM} \end{bmatrix} = \begin{bmatrix} \rho_i^{TE} & 0 \\ 0 & \rho_i^{TM} \end{bmatrix} \begin{bmatrix} B_i^{TE} \\ B_i^{TM} \end{bmatrix} \quad (4)$$

where the reflection coefficient in the dielectric substrate is:

$$\rho_i^\alpha = \frac{1 - Z_{oi}^\alpha Y_{mn,i}^\alpha \text{Coth}(\gamma_{m,n,i} h_i)}{1 + Z_{oi}^\alpha Y_{mn,i}^\alpha \text{Coth}(\gamma_{m,n,i} h_i)} \quad (i=2)$$

and the reflection coefficient in the free space (region 1) is:

$$\rho_i^\alpha = \frac{1 - Z_{oi}^\alpha Y_{mn,i}^\alpha}{1 + Z_{oi}^\alpha Y_{mn,i}^\alpha} \quad (i=1)$$

( $\alpha$  is TE or TM mode) with

$$Y^{TE} = \frac{\gamma_{m,n,i}}{j\omega\mu_0}; \quad Y^{TM} = \frac{j\omega\epsilon_0\epsilon_r}{\gamma_{m,n,i}}$$

where  $k_0$  is the space wave number.

$$\gamma_{m,n,i}^2 = \left[ \frac{m\pi}{a} \right]^2 + \left[ \frac{n\pi}{b} \right]^2 - k_0^2 \epsilon_r$$

### III- MODELISATION OF PROBE FEED

The coaxial current probe offers a more realistic method to excite currents on a patch antenna [9]-[10]. The coaxial probe is connected through the ground plane with the centre conductor embedded vertically and terminated on the patch surface, where the outer conductor of the coax is connected to the ground plane. Figure 3 shows the attachment of the coaxial probe to the patch antenna surface.

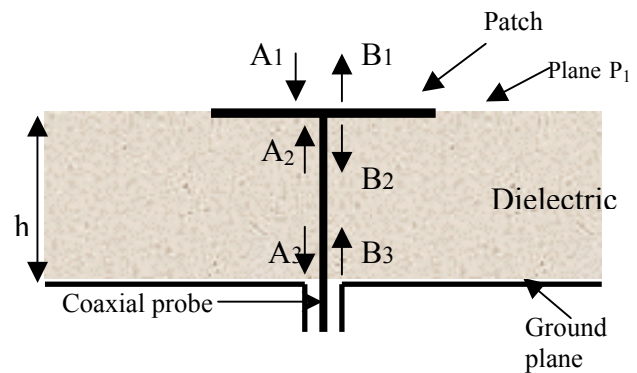


Figure 3. Microstrip patch antenna excited by coaxial probe feed.

The purpose is to determine the relationship between  $B_1$ ,  $B_2$  and  $A_1$ ,  $A_2$  on the sub-domain source.

We suppose that there are four metallic cells on the discontinuity plane ( $P_1$ ) which are connected to the via hole. In this case the current density distribution is illustrated in Fig.4.

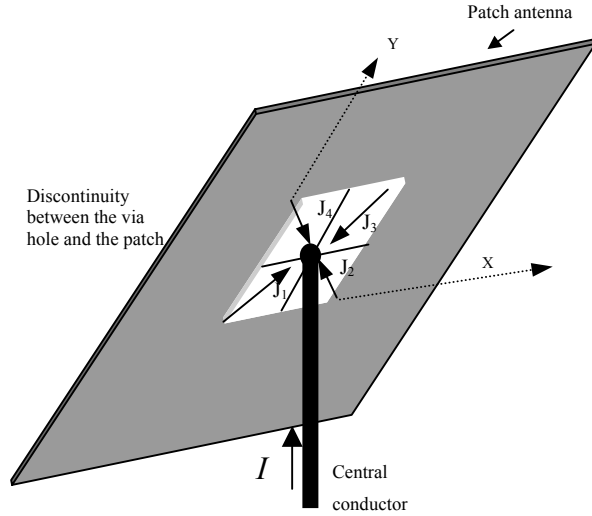


Figure 4. Bottom view of the patch antenna: Distribution of magnetic field on the discontinuity patch-via hole.

Let us note that  $I$  is the current excitation that verifies the following relationship:

$$J \times h = I . \quad (5)$$

According to Fig.4, it is possible to establish the following relationship:

$$J_1 + J_2 + J_3 + J_4 = J . \quad (6)$$

The vector  $J$  which characterizes the current distribution on the pixels of the discontinuity (via hole-patch) is given by:

$$J = \frac{1}{\sqrt{8}} (1, 1, -1, 1, -1, -1, 1, -1)_{(x,y)}^T . \quad (7)$$

In order to modelize the discontinuity ( via hole –plane  $P_1$ ), we assume that only fundamental mode can be propagated in the via hole and the other modes are evanescent. The passage from four cells characterizing the current density  $J$  to modes and vice versa can be considered as multi-port network depicted in Fig. 5.

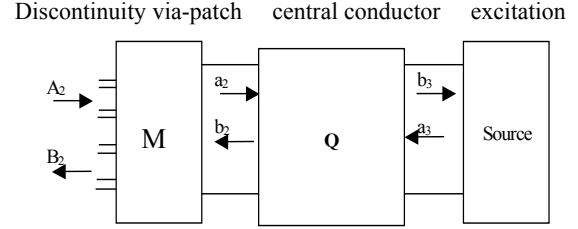


Figure 5. Coupling of Cells-Modes, Modes-Cells in the discontinuities between the via hole and the planes  $P_1, P_2$ .

$Q$  is a two-port network which characterizes the via hole, using the theory of transmission lines.

The scattering matrix of this two-port network is given by:

$$[S] = \begin{bmatrix} 0 & e^{j\theta} \\ e^{j\theta} & 0 \end{bmatrix}, \theta = k_0 h ; k_0 = \frac{2\pi}{\lambda_0} .$$

$\lambda_0$ : length wave in the free space.

Consequently, the relationship between waves  $a_2, a_3, b_2,$  and  $b_3,$  can be deduced:

$$\begin{bmatrix} b_2 \\ b_3 \end{bmatrix} = [S] \begin{bmatrix} a_2 \\ a_3 \end{bmatrix} . \quad (8)$$

The expression of the matrix  $M$  characterizing the multi-port network is demonstrated in reference [6]. It is given by the following equation:

$$[M] = \begin{bmatrix} JJ^T & J \\ J^T & 0 \end{bmatrix} .$$

Therefore, the relationship between incident and reflected waves is:

$$\begin{bmatrix} B_2 \\ a_2 \end{bmatrix} = \begin{bmatrix} JJ^T - 1 & J \\ J^T & 0 \end{bmatrix} \begin{bmatrix} A_2 \\ b_2 \end{bmatrix} . \quad (9)$$

The source can be modeled by the equivalent circuit illustrated in Fig. 6.

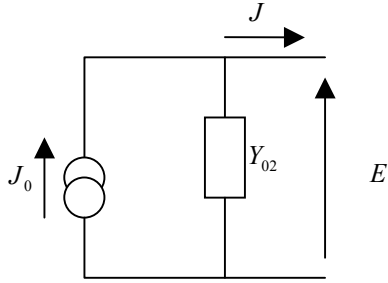


Figure 6. Equivalent circuit of the source.

As is known from the equivalent circuit given in fig.6, it is possible to deduce:

$$J = J_0 - Y_{02}E, \quad (10)$$

where  $Y_{02}$  is the admittance of Region (2),

$$Z_{02} = Y_{02}^{-1}.$$

Let us assume that the source excites the mode  $a_3$ . However the magnitude of  $b_3 = 0$ . Using the equations (5) and (6) we deduce:

$$B_2 = (JJ^T - 1)A_2 + e^{j\theta} J a_3. \quad (11)$$

According to equations (7) and (11) we deduce:

$$a_3 = \frac{1}{2h} \sqrt{Z_{02}} I_0. \quad (12)$$

In the region (1), if we suppose that we have a metal domain, it is possible to establish:  $B_1 = -A_1$ .

According to the equations (11) and (12), we can deduce the relationship between incident and diffracted waves on the sub-domain source:

$$\begin{bmatrix} B_1 \\ B_2 \end{bmatrix} = \begin{bmatrix} -1 & 0 \\ 0 & JJ^T - 1 \end{bmatrix} \begin{bmatrix} A_1 \\ A_2 \end{bmatrix} + \begin{bmatrix} 0 \\ \frac{1}{2} \sqrt{Z_{02}} J_0 e^{j\theta} J \end{bmatrix}. \quad (13)$$

The complete scattering matrix can be expressed as follows

$$\begin{cases} \begin{bmatrix} B_1 \\ B_2 \end{bmatrix} = \begin{bmatrix} -Hm - \frac{1-N2}{1+N2} Hd & \frac{2N}{1+N2} Hd \\ \frac{2N}{1+N2} Hd & -Hm - \frac{1-N2}{1+N2} Hd \end{bmatrix} \begin{bmatrix} A_1 \\ A_2 \end{bmatrix} & \text{On sub-domains metal and dielectric} \\ \begin{bmatrix} B_1 \\ B_2 \end{bmatrix} = \begin{bmatrix} -1 & 0 \\ 0 & JJ^T - 1 \end{bmatrix} \begin{bmatrix} A_1 \\ A_2 \end{bmatrix} + \begin{bmatrix} 0 \\ \frac{1}{2} \sqrt{Z_{02}} J_0 e^{j\theta} J \end{bmatrix} & \text{On sub-domain source} \end{cases}$$

At  $K^{\text{th}}$  iteration, it is possible to calculate the electric field and current density at interface plane:

$$J_i^k = \frac{1}{\sqrt{Z_{0i}}} (A_i^k - B_i^k) \quad (15)$$

$$E_i^k = \sqrt{Z_{0i}} (A_i^k + B_i^k). \quad (16)$$

#### IV- RESULTS AND DESIGN EXAMPLES

As an application, we are interested in characterizing two different shapes of patch antennas. We have developed a program to calculate the input impedance and the reflection coefficient of each antenna.

##### 1. Rectangular patch antenna:

The first example is a rectangular patch antenna deposited on substrate with relative dielectric constant  $\epsilon_r = 2.2$ . The coaxial probe is attached to the patch antenna at  $(x_c, y_c)$ .

In this case, the circuit plane is meshed with 64x64 square cells.

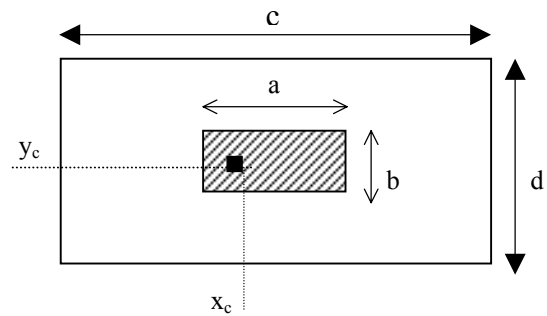


Figure 7. Rectangular patch antenna structure  $c=100\text{mm}$ ,  $d=76\text{mm}$ ,  $a=25\text{mm}$ ,  $b=19\text{mm}$ ,  $x_c=45,5\text{mm}$ ,  $y_c=34,5\text{mm}$ ,  $\epsilon_r=2.2$ .

First, the convergence of parameters is tested. In Fig. 8, The real part of  $Z_{in}$  is illustrated as function of the iteration number. It is seen that convergence is achieved for 160 iterations. The iterative process is terminated when the convergence is reached.

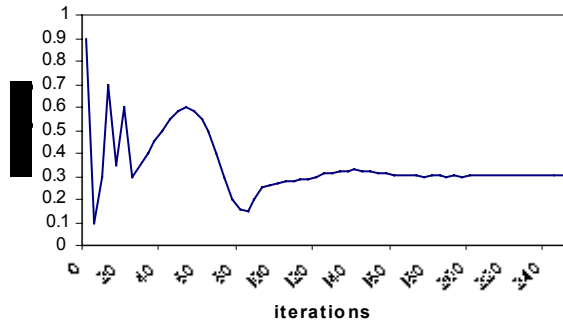


Figure 8: Real part of  $Z_{in}$  as function of the iteration number.

Figure 9. show the comparison between simulation and measured data reference [7]. It is seen that the error between them is 5 to 7%. The error can be minimized by meshing the structure with high resolution (128x128 pixels).

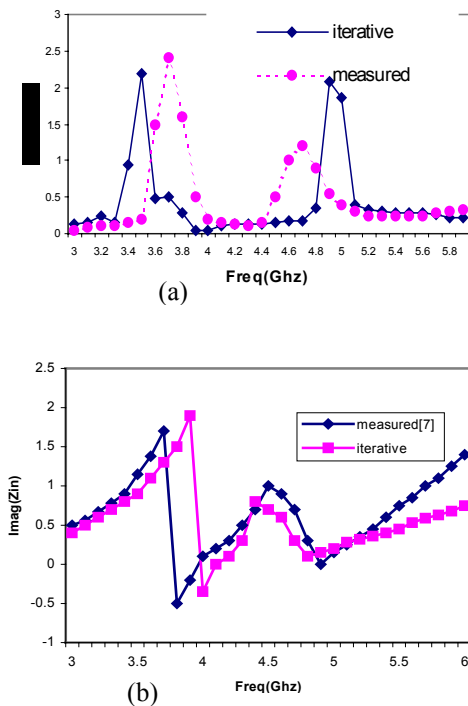


Figure 9. Input impedance of a rectangular patch antenna as a function of frequency, (a) real part and (b) imaginary part.

Figure 10 shows that the second resonance frequency is eliminated when the position of probe feed excitation is at the center of the patch antenna

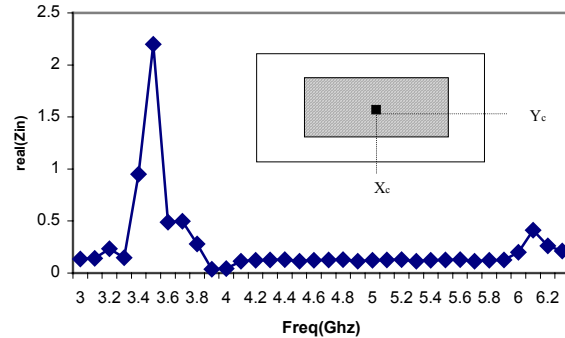


Figure 10. Input impedance of rectangular patch antenna  $c=100\text{mm}$ ,  $d=76\text{mm}$ ,  $a=25\text{mm}$ ,  $b=19\text{mm}$ ,  $x_c=0$ ,  $y_c=0$ ,  $\epsilon_r=2.2$ .

### 2. Patch Inverted-F Antenna (PIFA):

The antenna was printed on a thin, flexible Mylar substrate. The end of the ground plane is wrapped around the former, reducing the overall length of the antenna by the height without affecting the antenna performance.

Figure 11 shows the geometries and dimensions of the studied structure.

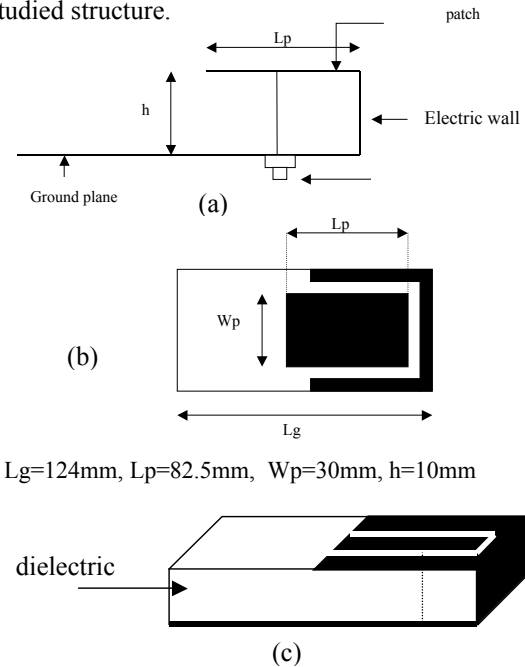


Figure 11. (a) Side view of the antenna (b) top view of the antenna (c) schematic view on the antenna.

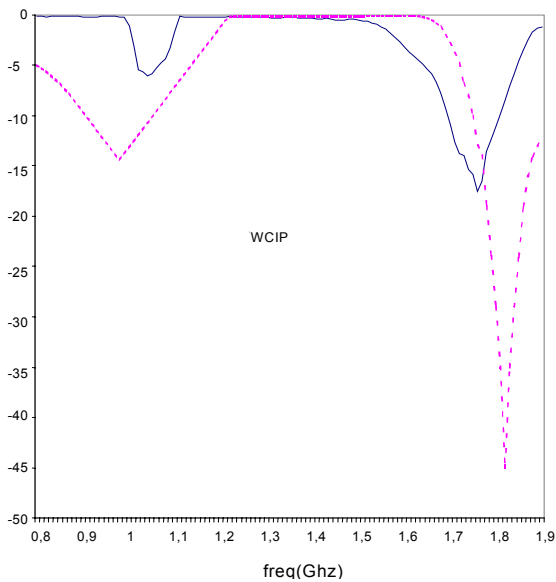


Figure 12. Reflection coefficient against frequency for the pifa antenna WCIP [8].

Finally the reflection coefficient is extracted and compared in Fig. 12 to the measured response given in reference [8], an agreement between them is observed.

#### IV- CONCLUSION

An iterative technique based on the concept of waves has been used for the simulation of the input impedance of rectangular patch antenna and the reflection coefficient of PIFA antenna. Thanks to its simplicity, the presented method does not involve bases functions and inversion of matrix. The good agreement between computed and published results justifies the design procedure and validates the present analysis approach. Consequently, the present approach will be investigated for further new applications such as air bridges, diodes, active elements.

#### REFERENCES

- [1] R. F. Harrington, "Field Computation by Moment Methods", New York, Macmillan, 1968.
- [2] O. Axelsson and V.A. Barker, "Finite Element Solution of boundary value problems", *Theory and Computation, New York, Academic Press*, 1984.
- [3] D. M. Sheen, S. M. D. Abouzahra, and J. A. Kong, "Application of the three-dimensional finite-difference time-domain method to the analysis of

planar micro-strip circuit", *IEEE, MTT*, vol.38, pp.849-857, July 1990.

- [4] M. Azizi, Boussouis, H. Aubert, H. Baudrand, "Three dimensional analysis of planar discontinuities by an iterative method". *Microwave and Optical Techniques Letters* 1996, 20:372-376.
- [5] A. Gharsallah, A. Gharbi, L. Desclos and H. Baudrand, "Analysis of interdigital capacitor and quasi-lumped miniaturized filters using iterative method," *Int. J. Numer. Model*, 2002. 15: 169-179. (DOI: 10,1002/jnm, 448)
- [6] A. Mami, H. Zairi, A. Gharsallah and H. Baudrand. "Analysis of micro-strip spiral inductor by using iterative method", *Microwave and optical technology Letters*. vol. 35, no. 4 , November 2002.
- [7] I. Proust, thesis, "Contribution à l'étude des discontinuités en guides d'ondes coaxiaux. Application aux antennes planaires", *INP Toulouse, France*, 1995.
- [8] K. F. Lee, Y. X. Guo, J. A. Hawkins, R. Chair and K. M. Luk. "Theory and experiment on microstrip patch antennas with shorting walls". *IEE Proc-Microwave Antennas propag.* vol. 147, no. 6, December 2000.
- [9] J. P. Damiano, A. Papiernik, "Survey of analytical and numerical models for probe-fed microstrip antennas", *IEE Proc. Microwave Antennas propag.*, vol. 141, no. 1, pp. 15-22, February 1994.
- [10] E. Lier, "Improved formulas for input impedance of coax-fed microstrip patch antennas" *IEE Proceeding*, vol. 129, no. 4, pp. 161-164, August 1992.



**Hassen Zairi** received his master's degree in physics from the Faculty of Sciences, Tunis, Tunisia, in 2002. He is currently working toward his PhD degree at the Faculty of Sciences, Tunis. His research interests include antenna

for wireless communication and microwave integrated circuits.



**Ali Gharsallah** received the degrees in radio-electrical engineering from the Ecole Supérieure de Télécommunication de Tunis in 1986 and the PhD degree in 1994 from the Ecole Nationale d'Ingénieurs de Tunis.

Since 1991, he was with the Department of Physics at the Faculty of Sciences, Tunis. His current research interests include antennas, multilayered structures and microwave integrated circuits.



**Gharbi Abdelhafidh** Professor in the Faculty of Sciences of Tunis, his current research interests are in the areas of dielectric and electromagnetic fields.



**Baudrand Henri** professor of Electromagnetism and microwave circuits at the Ecole Nationale Supérieure d'Electronique Electrotechnique Informatique of Toulouse, France. Baudrand is specialized in modelization of

passive and active circuits and antennas. He is the author and co-author of three books:

- Introduction au calcul des éléments de circuits microondes
- Optimisation des circuits non linéaires
- Calcul des circuits microondes par les schémas équivalents-exercices CEPADUES Editions). He co-signed over 100 publications in journals and 250 communication in international conferences.
- Dr Baudrand is a member of 'Electromagnetism Academy', senior member of IEE societies. President of URSI France commission B for 6 years (1993-1999), member and past president of IEEE-MTT-ED French chapter. He is a awarded officier des Palmes académiques, and Director Honoris causa of Iasi University, he is the Director of GREMO (Groupe de Recherche en Electromagnétisme et Microondes de Toulouse).