Analysis of Circular Slots Leaky-Wave Antenna in Cylindrical Waveguide by Wave Concept Iterative Procedure

Z. Houaneb[1], H. Zairi[1], A. Gharsallah[1], and H. Baudrand[2]

[1] Unité Circuit Système électronique HF. Faculté des sciences de Tunis. Tunisie
zied_houaneb@yahoo.fr, Hassen.zairi@ipsi.rnu.tn, ali.gharsallah@fst.rnu.tn

[2] Laboratoire LAPLACE, UPS-INP Toulouse. France
henri.baudrand@yahoo.fr

***Abstract*—** This paper presents the analysis of circular slots leaky-wave antenna by using an iterative method based on the wave concept. The classic method has been reformulated in cylindrical coordinates in order to be adequate for the analysis of the leaky-wave antenna with fast convergence. The proposed leaky-wave antenna can be used to replace a micro-strip patch array. Numerical results are presented to illustrate the advantages of the proposed structure. A good agreement between the new wave concept iterative procedure (WCIP) method results and published data is obtained.

***Index Terms*—** Circular patch, Hankel transform, leaky-waves, multi slits antenna, wave concept.

I. INTRODUCTION

For many years, there has been an increasing interest in microstrip leaky-wave antennas (LWAs) for their several features, which made them attractive candidates for many applications, ranging from mobile communications to phased array radar systems. They are replacing the conventional antennas in electronic scanning applications by frequency steering [1, 7].

Two-dimensional (2-D) microstrip planar antennas constitute a prime candidate for the exploration, separately, in two distinct perpendicular planes. The design of these antennas depends upon understanding the effects of patches excitation, substrate thickness, dielectric constant, and grids spacing on its scan performance. A few years ago, the use of 2-D scanning possibility both in elevation and azimuth planes was realized by additional phase-shifters in one plane and the frequency scanning in the other. This geometry was proposed in order to enhance the capabilities of 1-D uniform radiating structures by frequency changes [8, 10]. In recent years, there has been significant development in planar radiating structures for 2-D scanning features in millimetre-wave range applications, particularly in two dimensional periodic structures.

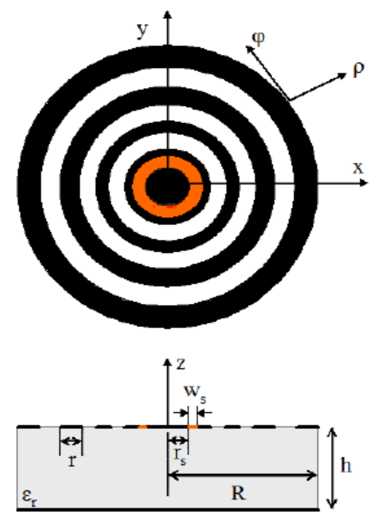

Fig. 1. Multi-slits antenna design.

The aim of this paper is to use the two-dimensional leaky-wave, which propagates outward radially along a planar surface. Among the printed antennas which can satisfy the above conditions is a series of concentric slits around a circular patch in cylindrical waveguide fed by a central ring, which maximizes the excitation of leaky-waves (See Fig. 1). Before exciting the fundamental TM_0 mode of the cylindrical waveguide, an electric source independent of φ is chosen and placed in the first slot. Thus, the various parts of the antenna are successively excited, in order to generalize leaky-wave phenomenon in all the directions of plane with circular wave forms.

To study this structure, a new wave concept iterative procedure (WCIP) in cylindrical coordinates is used. It consists in generating a recursive relationship between a given wave source and reflected waves at the interface containing the circuit [11]. The implementation of the iterative calculation is shown to extract the scattering parameters (S_{ij}) and generate the radiation patterns of the new structure. It consists of generating a recursive relationship between a given source wave and reflected waves at the interface containing the circular circuit. This discontinuity plane is divided into cylindrical cells and characterized by a scattering matrix operator depending on boundary conditions. Then, a Hankel transform is used to pass from spatial to spectral domain for each iteration of the recursive process. The advantage of the use of this method of simulation is to take into account the coupling between the circular slits without making additional calculations.

II. WCIP FORMULATION

Let us consider the shielded circular circuit, assumed to be lossless, presented in Fig. 2. The air-dielectric interface (plane Ω) is divided into cylindrical cells denoted by three subdomains corresponding to metal, dielectric and source domains. The wave concept is introduced by writing the transverse electric field $\vec{E}_i$ and surface tangential current density $\vec{J}_i$ in terms of incident ($\vec{B}_i$) and reflected ($\vec{A}_i$) waves. This leads to the following set of equations [12]:

$$\begin{cases} \vec{A}_i = \dfrac{1}{2\sqrt{Z_{0i}}}\left(\vec{E}_i + Z_{0i}\vec{J}_i\right) & i = 1,2 \\ \vec{B}_i = \dfrac{1}{2\sqrt{Z_{0i}}}\left(\vec{E}_i - Z_{0i}\vec{J}_i\right) & i = 1,2 \end{cases}, \qquad (1)$$

where $Z_{0i} = \sqrt{\mu_0 / \varepsilon_0 \varepsilon_{ri}}$ is the characteristic impedance of region i (i = 1, 2) and ε_{ri} is the relative permittivity of the region i. $\vec{J}_i$ is the surface tangential current density as $\vec{J}_i = \vec{H}_i \times \vec{n}_i$, with $\vec{n}_i$ a unit vector normal to the interface Ω and $\lfloor \times \rfloor$ is the cross product operator.

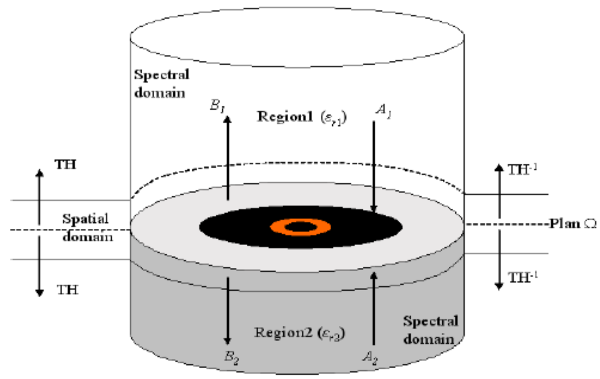

Fig. 2. The ideal WCIP structure.

$\vec{B}_i$ and $\vec{A}_i$ are incident and reflected waves in region i associated with the discontinuity interface Ω. The iterative process consists in establishing a recurrence relationship between waves $\vec{B}_i$ and $\vec{A}_i$. In order to generate incident waves $\vec{B}_i$ in the space domain, the circular circuit is excited by an electric planar source. Supposing that the space above the patch is equipped with cylindrical symmetry, it will be possible to consider the circular patch as a discontinuity between two half spaces that form two cylindrical guides of infinity radius.

The decomposition of the incident wave vector $B_i(\rho,\varphi)$ on the basis of mode TE and TM of the cylindrical guide, leads us to obtain:

$$B_i(\rho,\phi) = \sum_{m,n}\left(B^{\mathrm{TE}}_{i;mn}\left|f^{\mathrm{TE}}_{mn}(\rho,\phi)\right\rangle + B^{\mathrm{TM}}_{i;mn}\left|f^{\mathrm{TM}}_{mn}(\rho,\phi)\right\rangle\right), \quad (2)$$

where f^{α}_{mn} are mode functions of cylindrical guide with α={TE,TM}. By choosing an electric excitation source with a shape of ring with thickness w_s on the patch, the expressions of the cylindrical modes can be reduced due to fact that the electric and magnetic fields in the plane Ω are independent of φ. Thus, the radial component of electromagnetic TM mode is only excited.

By replacing f^{TM}_{mn} by their expressions of mode, we obtain the magnitude of TM mode:

$$B^{TM}_{i,mn} = -\frac{1}{\sqrt{\pi}} k_\rho \left\langle J'_n(k_\rho \rho) \middle| B_{i,\rho}(m,n) \right\rangle, \quad (3)$$

where k_ρ is the radial wave number and J'_n is the derivative of the Bessel function.

The scalar product in (3) becomes then:

$$\left\langle J_n^{'}\left(k_\rho \rho\right) \middle| B_{i,\rho}\left(m,n\right)\right\rangle = \int_0^R B_{i,\rho}\left(m,n\right) J_n^{'}\left(k_\rho \rho\right)\rho d\rho. \quad (4)$$

By using the recurrent relation of Bessel functions of integer order, the integral in equation (4) can be written as a Hankel transform [13, 14]. This transform enables us to move from space domain to the spectral domain.

As far as separable geometry is concerned, the set of functions associated with both TE and TM transverse electric field provides a complete set of orthogonal basis functions suitable to expand electric fields in the boxed structure as [15]:

$$E_T(\rho,\phi) = \sum_{\alpha,m,n} e_{mn}^{\alpha} f_{mn}^{\alpha}(\rho,\phi). \quad (5)$$

The tangential current density is expressed as:

$$J_T(\rho,\phi) = \sum_{\alpha,m,n} e_{mn}^{\alpha} Y_{mn}^{\alpha} f_{mn}^{\alpha}(\rho,\phi). \quad (6)$$

The expressions (5) and (6) support the expansions in the spectral domain of the integral operator $\hat{Y}$ defined as:

$$\vec{J} = \hat{Y}\vec{E}$$
$$\hat{Y} = \sum_{\alpha,m,n} \left| f_{mn}^{\alpha} \right\rangle Y_{mn}^{\alpha} \left\langle f_{mn}^{\alpha} \right|. \quad (7)$$

Hence, from definition (1), the waves can be expanded on the same set of basis functions of the tangential fields and the $\hat{\Gamma}_i$ operator such that:

$$\vec{A}_i^{\alpha} = \hat{\Gamma}_i^{\alpha} \vec{B}_i^{\alpha}, \quad (8)$$

where i = 1,2 refers to the sides of interface Ω. Thus, $\hat{\Gamma}_i^{\alpha}$ has the general form:

$$\hat{\Gamma}_i^{\alpha} = \sum_{m,n} \left| f_{mn} \right\rangle \Gamma_{mn}^{\alpha} \left\langle f_{mn} \right|. \quad (9)$$

Applying $\hat{\Gamma}$ simply consists in multiplying the modal amplitude of the waves by the corresponding numbers $\hat{\Gamma}_i^{TE}$ and $\hat{\Gamma}_i^{TM}$ in (11), such that:

$$\hat{\Gamma}_i = \frac{1 - Z_{0i} Y_{i,mn}^{\alpha}}{1 + Z_{0i} Y_{i,mn}^{\alpha}}, \quad (10)$$

where $Y_{i,mn}^{\alpha}$ is defined in Table 1, $k_r^2 = \omega^2 \varepsilon_0 \varepsilon_r \mu_0$, $\gamma^2 = k_\rho^2 - k_r^2$ and $Y_r = \sqrt{\varepsilon_0 \varepsilon_{ri} / \mu_0}$ is the admittance of each domain.

Table 1: Mode admittance expressions of a uniform waveguide

	TE mode	TM mode
Infinite guide	$-jY_r\gamma / k_r$	$-jY_r k_r / \gamma$
Guide short circuited at distance h	$-jY_r\gamma \coth(\gamma h)/k_r$	$-jY_r k_r \coth(\gamma h)/\gamma$

The spectral wave $\vec{B}_i^{\mathrm{TM}}$ is reflected in each domain by the $\hat{\Gamma}$ operator to give the reflected wave $\vec{A}_i^{\mathrm{TM}}$:

$$\vec{A}_i^{\mathrm{TM}} = \hat{\Gamma}_i \vec{B}_i^{\mathrm{TM}}. \quad (11)$$

To return to the spatial domain, an inverse Hankel transform must be used:

$$\vec{A}_{i,\rho} = \mathbf{HT}^{-1}\left\{\vec{A}_i^{\mathbf{TM}}\right\}. \quad (12)$$

The wave $\vec{A}_{i,\rho}$ is scattered in interface plane Ω by Γ_Ω factor [16, 17] in order to constitute the incident wave $\vec{B}_i$ for the next iteration:

$$\vec{B}_i = \left[\Gamma_\Omega\right]\vec{A}_i + \vec{B}_0, \quad (13)$$

where $\vec{B}_0$ is source excitation.

The implementation of the iterative procedure consists in establishing a recurrent relationship between each side of the interface (discontinuity). By using the boundary conditions in spatial domain (13) and reflection in the spectral domain (11) the following relationships can be obtained:

$$\vec{B}_i^{(k)} = \Gamma_\Omega \vec{A}_i^{(k-1)} + \vec{B}_i^{(0)} \quad (14)$$
$$\vec{A}_i^{(k)} = \hat{\Gamma}_i \vec{B}_i^{(k)}, \quad (15)$$

where i is the index media, k is the number of iterations. The iterative process is stopped when the electric field and the current density converge. So, the main characteristics of the circuit can be extracted. Once convergence is achieved, the $\vec{B}_i$ and $\vec{A}_i$ waves are expressed in spatial domain and the electric field and current density can be determined at the interface plane Ω. It is done using the equations in (1).

$$\vec{E}_i = \sqrt{Z_{0i}}\left(\vec{A}_i + \vec{B}_i\right) \quad (16)$$

$$\vec{J}_i = \frac{1}{\sqrt{Z_{0i}}}\left(\vec{A}_i - \vec{B}_i\right). \quad (17)$$

The algorithm of the iterative process is shown by Fig. 3. In order to implement this process, the FORTRAN language is used.

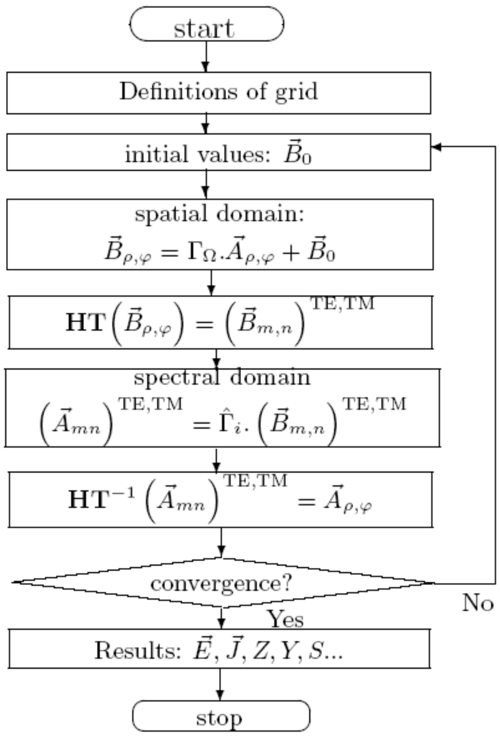

Fig. 3. The WCIP algorithm.

III. VALIDATION OF WCIP METHOD

The presented formulation was implemented in FORTRAN code. To demonstrate the effectiveness of the method, circular patch antenna has been considered (See Fig. 4).

Figure 5 shows the S_{11} and real and imaginary parts of Z_{in} convergence at 6.6 GHz for the structure given by Fig. 4 for a radius patch equal to 7.0 *mm*. Figure 6 shows the simulated resonant frequency obtained for the space wave against radius R compared to the published data [18]. Thus, a good agreement between simulated and published data is observed.

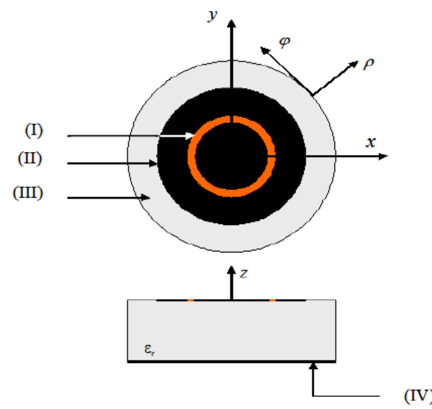

Fig. 4. Configuration of the circular patch: (I) excitation ring, (II) circular patch, (III) dielectric, (IV) ground plane.

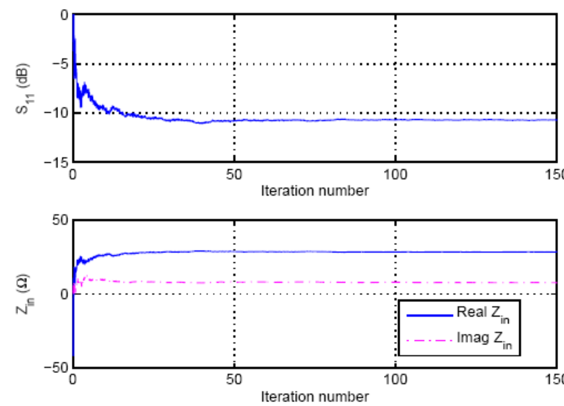

Fig. 5. S_{11} and Z_{in} convergence versus iterations number at 6.6GHz.

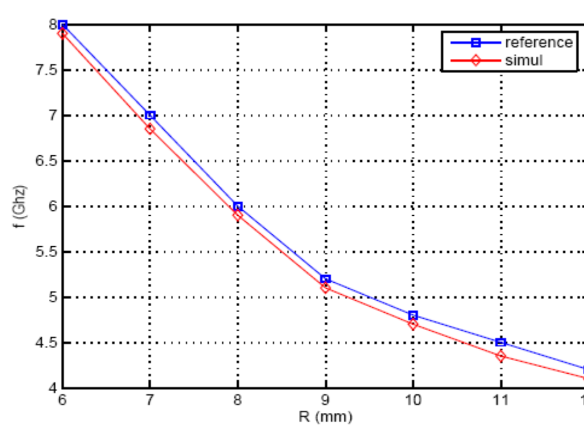

Fig. 6. The variation of resonance frequency f simulated for space wave against R compared to those published.

To validate the WCIP in cylindrical coordinate, the Q-factor of surface wave has been simulated versus the height of substrate (Fig. 7) and the radius of the patch (Fig. 8) for two different values of ε_r and compared to published data [18]. The accuracy of these figures explains that the addition of the tangential electric field and density of current given by magnetic and electric walls is necessary. We denote by E_{ma} and E_{el} the fields given by magnetic and electric walls situated at large value of r approximately equal to $5\lambda g$. The total field is given by:

$$\begin{aligned} E_{tot} &= E_{ma} - jE_{el} \\ &= \cos(k_0 r) - j\sin(k_0 r) \\ &= \exp(jk_0 r). \end{aligned} \quad (18)$$

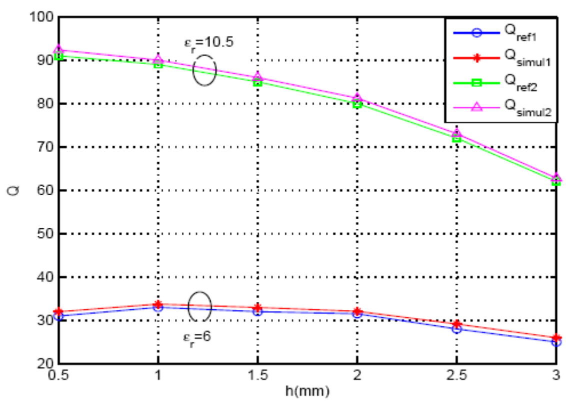

Fig. 7. The variation of Q-factor of surface wave with h for two different ε_r and $R = 7mm$.

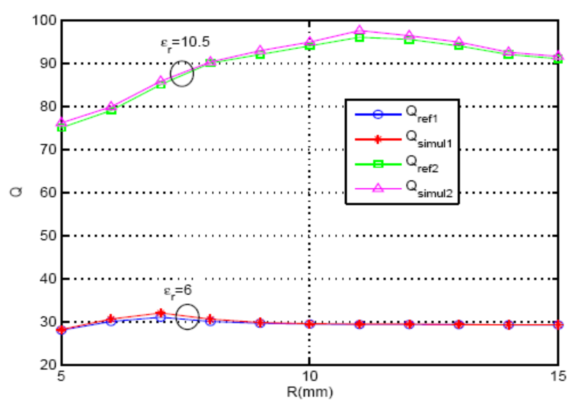

Fig. 8. The variation of Q-factor of surface wave with R for two different ε_r and $h = 1.27mm$.

IV. ANTENNA DESIGN

The structure proposed in this work is an antenna of low profile. Fig. 1 illustrates the circular microstrip patch antenna, which incorporates annular slits. This new design is fed by a circular ring. The circular patch of radius R is mounted on a substrate of thickness h and with a dielectric constant ε_r. The feed ring of width w_s, is located at a distance r_s from the centre of the patch. The annular slits are situated concentrically around the central circular patch. The distance r between the slits is chosen when the derivative Bessel function $J_0^{'}$ is in its maximum. The interface Ω is sampling in 340*40 polar pixels (radial direction by 340 and azimuth direction by 40).

When this antenna is excited by an electric source, that is independent of the variable φ, the radial component of TM_0 mode is only excited. In each slit exists an electric field which is created by the leaky-wave. These fields give rise to a radiated field. This one is identical to an array antenna patch when it is excited by a feeder network. Thus, each patch creates a far field in space. The advantage of our antenna is the use of only one source instead of a network. Also, the condition of spacing between the slits is not necessarily $\lambda_g/2$. Thus, the shape of the proposed antenna is more compact.

First, we study an antenna with a central ring (that represents the source) and an annular slit to validate the WCIP method of simulation. This antenna is mounted on a substrate of thickness h = 2.0 *mm* and with a permittivity ε_r = 4.25. This one is fed by the central ring having an interior radius r_0 = 0.5 *cm* and a width w_s = 1.0 *mm*. The annular slit has a width w = 1.0 *mm* and an interior radius 1.0 *cm*. The distance between the central ring and the slit that surrounds it is r = 2.0 *mm*. The r distance is chosen in a way that the derivative Bessel function $J_0^{'}$ reaches its first maximum. Both the boundary of the domain that is a perfectly cylindrical conductor and the radius of patch have 6.0 *cm* of radius.

In order to demonstrate the validity and the advantages of the iterative approach, a program implanted with a symbolic calculation with FORTRAN language is developed. Figure 9 represents the distribution of the electric field on the plane Ω at the resonance frequency. We notice the appearance of the electric field in the slit, which is due to the leaky-wave.

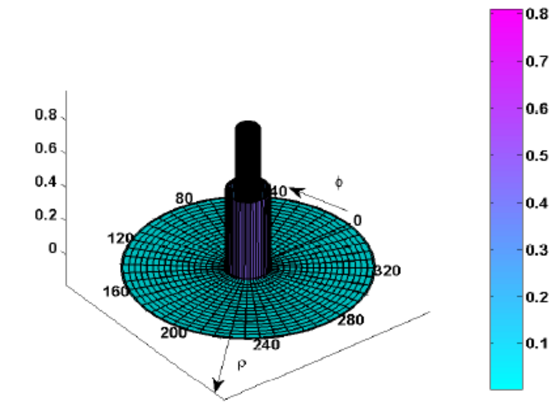

Fig. 9. The electric field magnitude as a function of ρ-φ.

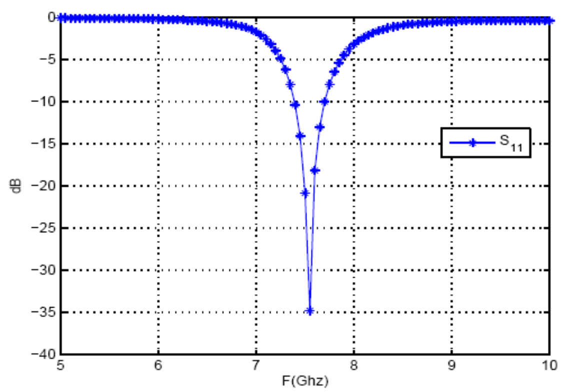

Fig. 10. Return loss as function of frequency.

The return loss parameter is extracted in Fig. 10. The resonance frequency of this antenna is located at 7.55 GHz. For each frequency, the WCIP process consumes 13 seconds but the ADS software consumes 19 seconds and the HFSS software 20 seconds. All theses tests are done with the same computer characteristics. Moreover the time of calculation is less than the one using a conventional technique.

In Fig. 11, the real and imaginary parts of input impedance of this antenna with two slits are represented. The real part of impedance is 50 ohm and the imaginary part is around 0 at the resonance frequency, this shows that our antenna is resonating.

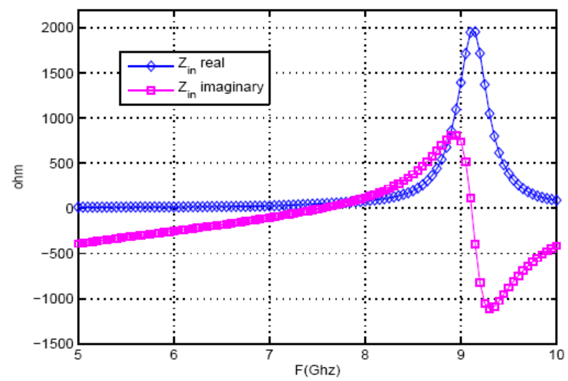

Fig. 11. Real and imaginary parts of input impedance of antenna with two slits.

V. RADIATION PATTERNS

To calculate the radiated field by multi-slits antenna, each slit can be represented as a magnetic loop of current [19]. Each loop creates a far field E_{rad}^{k} where k designs the index of slits. The total field radiated by our antenna (E_{rad}^{t}) is the sum of the fields radiated by each loop of current:

$$E_{rad}^{t} = \sum_{k=1}^{N_s} E_{rad}^{k} \, , \tag{19}$$

where N_s is the number of slits. The centre of the slits defines the origin of the reference mark. As the antenna has symmetry of revolution, the magnetic current is expressed with Fourier series:

$$\vec{M}^{k}(\rho,\varphi) = \sum_{n_\varphi=-\infty}^{+\infty} \left(M_{\rho}^{k}(\rho; n_\varphi)\vec{\rho} + M_{\varphi}^{k}(\rho; n_\varphi)\vec{\varphi} \right) e^{jn_\varphi \varphi} \, . \tag{20}$$

It is the same for the far electric field, given by:

$$E_{\theta}^{k}(\theta,\varphi) = -k\Psi(r_k) \sum_{n=-\infty}^{+\infty} j^{n} e^{jn\varphi} \left[jC_{\theta,\rho}^{k}(\theta;n) + C_{\theta,\varphi}^{k}(\theta;n) \right] \tag{21}$$

$$E_{\varphi}^{k}(\theta,\varphi) = k\Psi(r_k)\cos(\theta) \sum_{n=-\infty}^{+\infty} j^{n} e^{jn\varphi} \left[C_{\varphi,\rho}^{k}(\theta;n) - jC_{\varphi,\varphi}^{k}(\theta;n) \right] \tag{22}$$

with

$$\Psi(r_k) = \frac{e^{-jkr_{ki}}}{r_k}$$

$$C_{\theta,\rho}^{k}(\theta;n) = \int_{r_k - \frac{w_k}{2}}^{r_k + \frac{w_k}{2}} M_{\varphi}^{k}(\rho;n) \frac{nJ_n(k\rho\sin\theta)}{k\rho\sin\theta} \rho d\rho$$

$$C_{\theta,\varphi}^{k}(\theta;n) = \int_{r_k - \frac{w_k}{2}}^{r_k + \frac{w_k}{2}} M_{\rho}^{k}(\rho;n) J_n^{'}(k\rho\sin\theta) \rho d\rho$$

$$C_{\varphi,\rho}^{k}(\theta;n)=\int_{r_k-\frac{w_k}{2}}^{r_k+\frac{w_k}{2}} M_{\varphi}^{k}(\rho;n)J_n^{'}(k\rho\sin\theta)\rho d\rho \quad (23)$$

$$C_{\theta,\varphi}^{k}(\theta;n)=\int_{r_k-\frac{w_k}{2}}^{r_k+\frac{w_{ki}}{2}} M_{\rho}^{k}(\rho;n)\frac{nJ_n(k\rho\sin\theta)}{k\rho\sin\theta}\rho d\rho \quad .$$

The simulated radiation patterns at the resonance frequency are given by Fig. 12.

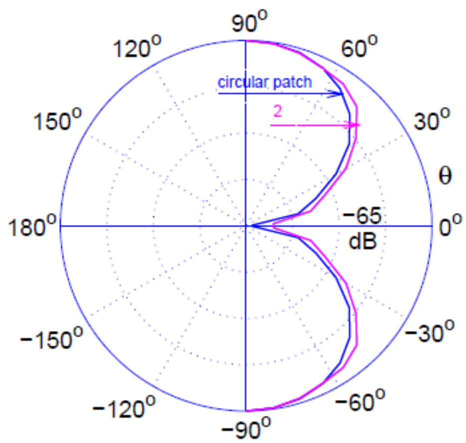

Fig. 12. The radiation patterns at frequency 7.55 GHz. (The numbers in the figure represent the number of slits on the patch.)

The directivity of the proposed antenna is improved by addition of one slit in the circular disc. The beam width to -3dB is reduced by 5 degrees if we add one slit in this circular patch. That is, due to the surface waves which were excited by the source that an electric field in the second slit taking part in the radiation can be seen. This is equivalent to an antenna array with two patches. To improve the directivity in a considerable way, it is necessary to increase the number of slits. These slits are outdistanced by *r* distance between them such as the derivative Bessel function $J_0^{'}$ which reaches its maximums on each slit. The radiated field at $\phi = 0$ is usually equal to zero for any number of slits. This is due to the symmetry of this antenna. The study of coupling between the slits is not necessary since the new WCIP method as electromagnetic simulator is used. The advantage of this method is to take into account the coupling between the cells, which constitutes the interface plane Ω where the circular antenna with slits is put.

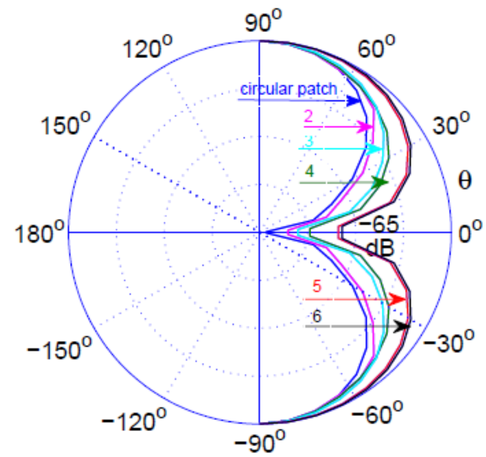

Fig. 13. The radiation patterns at frequency 7.55GHz of different number of slits. (The numbers in the figure represent the number of slits on the patch.)

According to Fig. 13, the 3dB radiation ranges from 90 degree of circular patch to between 30 and 60 degree for a structure with slits. Thus, the band increases with the number of slits. In Table 2, the reduction of the beam width to -3dB of the antenna with slits compared with circular patch is shown.

Table 2: Beam width reduction

Number of slits	Beam width reduction (degree)
2 slits	5°
3 slits	23°
4 slits	31°
5 slits	36°
6 slits	38°

We notice that if we have more than six slits, the reduction of the beam width will not be significant any more. So, we can limit ourselves to an antenna with six slits.

VI. CONCLUSION

In this paper, the formulation and the implementation of a new iterative method based on the concept of waves is presented to study a new leaky-wave antenna with circular shape containing a number of concentric slits around a circular patch, which can be used for broadcast applications. It is demonstrated that this antenna can play the part of an array antenna if we increase the number of slits in the structure. The advantage of this new antenna is its compact form and facility of realization. Numerical results have been obtained with reduced CPU time because of the Hankel transformation used in the iterative method. A good agreement between our results and published data is obtained.

REFERENCES

[1] C. Lin and C. Tzuang, "A dual beam micro CPW leaky-mode antenna," *IEEE Trans. Antenna And propag.,* vol. 48, pp. 310-316, 2000.

[2] C. Caloz, S. Lim, and T. Itoh, "A novel leaky-wave retrodirective reflector using short/matched terminations," *33rd European Microwave Conference,* pp. 1071-1074, October 2003.

[3] T. Zhao, D. Jackson, J. Williams, D. Yang, and A. Oliner, "2-D periodic leaky-wave antennas- Part I: Metal patch design," *IEEE Trans. Anten. and Propag.,* vol. 53, pp. 3505-3514. 2005.

[4] T. Zhao, D. Jackson, and J. Williams, "2-D periodic leaky-wave antennas- Part II: Slot design," *IEEE Trans. Anten. and Propag.,* vol. 53, pp. 3515-3524. 2005.

[5] T. Zhao, D. Jackson, J. Williams, and A. Oliner, "General formulas for 2-D leaky-wave antenna," *IEEE Trans. Anten. and Propag.,* vol. 53, pp. 3525-3532, 2005.

[6] D. Killips, J. Radclife, L. Kempel, and S. Schneider, "Radiation by a linear array of half-width leaky-wave antennas," *ACES Journal*, vol. 21, no. 3, pp. 248-255, 2006.

[7] J. R. James and G. Andrasic, "Analysis and computation of leaky-wave hyperthermia applicator," *ACES Journal*, vol. 7, no. 2, pp. 72-84, 1992.

[8] P. Liao and R. York, "A new phase-shifters beam scanning technique using arrays of couple oscillator," *IEEE Trans. Microwave Theor. Techn.,* vol. 41, pp. 1810-1815, 1993.

[9] K. Chen, Y. Qian, C. Tzuang, and T. Itoh, "A periodic microstrip radial antenna array with a conical beam," *IEEE Trans. Anten. Propag.,* vol. 51, no. 4, pp. 756-765, 2003.

[10] C. Hu, C. Jou, and J. Wu, "An aperture coupled linear microstrip leaky-wave antenna array with two dimensional dual beam scanning capability," *IEEE Trans. Anten. Propag.,* vol. 48, pp. 909-913, 2000.

[11] R. S. N'gongo and H. Baudrand, "A new approach for microstrip active antennas using modal FFT algorithm," *IEEE AP-S Int symp Ind USNC/URSI Natl Radio Sci Mtg,* Orlando, FL, July 1999.

[12] H. Zairi, A. Gharsallah, A. Gharbi, and H. Baudrand, "Modelling of coupled microstrip antennas integrated with EBG structure using an iterative method," *ACES Journal,* vol. 23, no. 4, pp. 357-362, Dec. 2008.

[13] M. Guizar-Sicairos and J. C. Gutierrez-Vega, "Computation of quasi-discrete Hankel transforms of integer order for propagating optical wave fields," *J. Opt. Soc. Am.,* vol. 21, no. 1, pp. 53-58, 2004.

[14] Z. A. Delecki, "Eigenvalue computation for application of the finite Hankel transform in coaxial regions," *ACES Journal,* vol. 4, no. 1, pp. 41-56, Mar. 1989.

[15] H. Baudrand, "The wave concept in electromagnetic problem: application in integral methods," *Asia Pacific Microwave Conference*, New Delhi, December 1996.

[16] A. Gharsallah, A. Gharbi, and H. Baudrand, "Analysis of interdigital capacitor and quasi-lumped miniaturized filters using iterative method," Int *J Num Modelling: Electronic Networks, devices and Fields,* vol. 15, issue 2, pp. 169-179, Apr. 2002.

[17] A. Gharsallah, R. Garcia, A. Gharbi, and H. Baudrand, "Wave concept iterative process merges with modal fast Fourier transformation to analyze microstrip filters," *ACES Journal*, vol. 16, no. 1, pp. 61-67, Mar. 2001.

[18] D. El-Kouhen, H. Aubert, M. Ghomi, and H. Baudrand, "Q-factor computation of radiation loss corresponding to surface wave in a patch circular resonator," *Electronics Letters*, vol. 32, no. 22, pp. 2039-2041, October 1996.

[19] S. J. Orfanidis, "Electromagnetic Waves and Antennas," Copyright (c) 1996-2008 by Sophocles J. Orfanidis, www.ece.rutgers.edu/ orfanidi/ewa.

Zied Houaneb was born in Tunisia. He received his master's degree in telecommunication from Ecole Supérieure des Communications de Tunis, Tunisia in 2002. He is currently working toward his Ph.D. degree at the Faculty of Sciences, Tunis, Tunisia. His research interests include circular devices and antennas for wireless communication and microwave integrated circuits.

Hassen Zairi received his master's degree in physics from the Faculty of Sciences, Tunis, Tunisia, in 2002 and the Ph.D. degree in 2005. He is currently an Assistant Professor with the Department of Electrical Engineering, Ecole Supérieure de Technologie et d'informatique, Tunis, Tunisia. His main research interests are in numerical methods for the analysis of microwave and millimeter-wave circuits.

Ali. Gharsallah received the degrees in radio-electrical engineering from the Ecole Supérieure de Télécommunication de Tunis in 1986 and the Ph.D. 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.

Henry Baudrand, Professor Emeritus at the Ecole Supérieure d'Electronique Electrotechnique Informatique, ENSEEIHT, of National Polytechnic Institute of Toulouse, France is specialized in Modelling of Passive and Active Circuits and Antennas. He is the author and co-author of three books:

- Introduction au calcul des éléments de circuits micoondes
- Optimisation des circuits non linéaires. (in collaboration with M.C.E. Yagoub)
- Calcul des circuits microondes par les schémas équivalents. (in collaboration with H.Aubert)

These books are published by CEPADUES Editions Toulouse.

He co-signed over 130 publications in international journals, four chapters in scientific books and 250 communications in international conferences.

He is a Fellow Member of IEEE Society, a Fellow Member of the Electromagnetism Academy and a Senior Member of the IEE society. He was President of URSI France commission B for 6 years (1993-1999), President of IEEE-MTT-ED French chapter (1996-1998), and President of International Comity of O.H.D. (Hertzian Optics and Dielectrics) between 2000 and 2004. He is awarded "Officier des Palmes académiques" and Doctor Honoris causa of lasi University (1996).