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A NOVEL MICROSTRIP ANTENNA ARRAY WITH METAMATERIAL-BASED ELECTRONIC BEAM STEERING AT 2.4 GHz

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Abstract—A metamaterial based electronically controlled microstrip structure, performing as leaky wave (LW) antenna with beam steering capability was synthesized. The structure has the configuration of metamaterial transmission line (TL) composed of cascade composite right-/left-handed (CRLH) unit cells. The direction of maximum radiation is tuned via the variation of varactros' capacities incorporated to the structure. Theoretical analysis and synthesis, based on the metamaterial TL theory, was made and novel methods to realize some of the elements of the units cells are proposed. Results, received via simulation, demonstrate that the LW antenna has steering capability of the direction of maximum radiation in a range of 40°, gain changing a little, 6 dB to 7 dB, during the scanning, whereas small number of cells is enough to obtain this performance.

1. INTRODUCTION

The rapid evolution of mobile communication technology and the ability to offer to users services through cellular systems and local nets, demand the creation of portable electronic telecommunication devices with unique specifications. This type of equipments must support the wireless telecommunication protocols of the networks and simultaneously have the capability of receiving and transmitting signals with sufficient efficiency in manmade environments, where the propagation of the electromagnetic waves is a composite phenomenon, with high possibility of signal loss. The antennas of the portable devices should be capable of adapting at their environment and

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ensuring the optimum signal transmission and reception. Their receiving or transmitting ability, especially in the cases of weak signals, due to high propagation path loss, could be improved by increasing their directive gain. However the gain increment leads to main radiation lobes with small width. This feature suppresses the 'viewing area' of the antenna, making it a less sensitive receiver in an 'mobile' environment in which the direction of arrival of signals (DoA) is a continuously varying parameter. The best, perhaps solution to overcome this trade off and the antenna to have both high directive gain and high adaptivity to the direction of the incoming signals, is the antenna to be properly designed to have beam steering ability, and so to successfully confront the phenomenon of the continuously change of signals' DoA.

The basic category of radiators having the attribute of beam steering are the phased array antennas. This kind of antennas has been widely equipped in both military and civilian applications. However, they suffer some flaws of complex and feeding network, expensive transceiver system, and bulk layout. These features make them not suitable for commercial handset equipment of wide use.

Recently, several concepts of steering antennas based on active devices have been introduced, having used technologies such as electronically tunable impedance surface, electronically controlled transmission line, and microelectromechanical systems (MEMS), to achieve beam-steering [1–5]. By designing the antennas with these techniques we would potentially overcome the disadvantages of the classical phased-array systems. To the above types would be added the category of leaky wave antennas, especially in microstrip or generally in printed form. They are radiating structures which would meet the requirement of beam steering and simultaneously are arrangements of low profile and of simple feeding.

A leaky wave antenna is substantially a waveguide-type structure that leaks out power all along its length due to the existence of a fast travelling wave. Therefore, the structure is treated as a transmission line when terminated on a matched load. Instead, if it is open or short ended, the mechanism of resonance is dominant and the antenna is treated as a radiating element in resonance. In the case that beam steering is required there has to be suitable feeding, that would insert to the current waves a phase shifting from point to point along the array, thus obtaining the adaptation of the beam. An ideal method to obtain this performance is the synthesis of the antenna via the metamaterial concept [6–13].

Metamaterials are artificial structures that can be designed to exhibit specific electromagnetic properties not commonly found in nature. They are commonly termed as left-handed (LH) materials, having simultaneously negative permittivity (ε) and permeability (μ), and have allowed novel applications, concepts, and devices to be developed.

Metamaterials, in printed form, are realized by periodically printed sub-wavelength metallic cells which compose a uniform periodic structure, the period of which defines the operating frequency band. Metamnaterials may be equivalently described in terms of *media* parameters (electric/magnetic susceptibilities, permittivity, permeability), or in terms of *transmission-line* (TL) parameters (inductance/capacitance, impedance/admittance, propagation constant/characteristic impedance). The latter approach has been used in the majority of the practical applications reported to date. Whatever of the above two is the theoretical approach, the key for a metamaterial beam steering array is to find a way to change, in real time, the propagation constant of the wave along the periodic structure, thus causing the suitable phase shifting from cell to cell. This way can be translated as equivalent controlling of the refractive index of the metamaterial structure or change of the equivalent parameters of the transmission line and as a consequence the variation of propagation constant.

The scope of this paper was the design of a microstrip antenna suitable for portable communication devices with small size and the capability of controlling the direction of maximum radiation via a simple and easy to be realized configuration. The synthesis of the antenna system, was based on the concept of composite right-/left-handed (CRLH) metamaterial theory. The antenna was composed of periodically arranged radiating printed cells which constitute an array that performs as a leaky wave radiating scheme, in which the direction of maximum radiation depends on the propagation constant of the wave which travels on the array. The value of propagation constant is controlled by capacitively loading the antenna with varactors. The controlled variation of the reverse DC varactors' voltage, and the respective changing of their capacity, results to variation of the propagation constant which in turn inserts a progressive phase shift along the array of cells. Thus, the beam steering is achieved. The synthesis and modeling of the operation of the array was based on the basic equivalent circuit theory of the metamaterial transmission lines. The series and parallel capacities and inductances of the model, were realized by interdigital capacitors, varactors, printed short-circuited stubs as well as new equivalent schemes for series connected inductances.

2. THEORETICAL ANALYSIS

The concept of composite right-/left-handed (CRLH) transmission-line metamaterials, describes in a simple manner the fundamentally dual right-handed (RH)/left-handed (LH) nature of metamaterials, and has been widely accepted as an efficient tool for the analysis of metamaterial phenomena and the design of metamaterial arrangements. The general CRLH TL model (Fig. 1) consists of an inductance L_R in series with a capacitance C_L and a shunt capacitance C_R in parallel with an inductance L_L [14]. It comes from the combination of 1) a RH TL model, which is represented as the combination of the per-unit length series inductance L_R and the per-unit length shunt capacitance C_R and 2) a LH TL model, presented as the combination of the per-unit length series capacitance C_L and the per-unit length shunt inductance L_L.

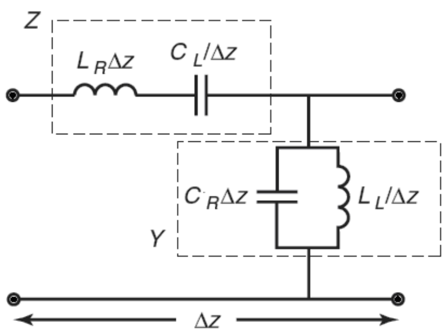

Figure 1. The equivalent circuit model of the CRLH unit cell of the metamaterial transmission line.

Generally, the propagation constant of a TL is given by $\gamma = \alpha + j\beta = \sqrt{Z\gamma}$ where Z and Y are, respectively, the per-unit length impedance and per-unit length admittance. In the particular case of the CRLH TL, Z and Y are defined as

$$\mathrm{Z}(\omega) = j\left(\omega L_R - \frac{1}{\omega C_L}\right) \tag{1a}$$

$$Y(\omega) = j\left(\omega C_R - \frac{1}{\omega L_L}\right) \tag{1b}$$

Therefore, the dispersion relation for a homogenous CRLH TL is

$$\beta(\omega) = s(\omega)\sqrt{\omega^2 L_R C_R + \frac{1}{\omega^2 L_L C_L} - \left(\frac{L_R}{L_L} + \frac{C_R}{C_L}\right)} \tag{2}$$

where

$$s(\omega) = \begin{cases} -1 \text{ if } \omega < \omega_1 = \min\left(\dfrac{1}{L_R C_L}, \dfrac{1}{L_L C_R}\right) \\ +1 \text{ if } \omega > \omega_2 = \max\left(\dfrac{1}{L_R C_L}, \dfrac{1}{L_L C_R}\right) \end{cases} \tag{3}$$

The phase constant β can be purely real or purely imaginary depending on whether the radicand is positive or negative, respectively. In the frequency range where β is purely real, a pass-band is present since $\gamma = j\beta$. In contrast, a stop-band occurs in the frequency range where β is purely imaginary since $\gamma = \alpha$. This stop-band is a unique characteristic of the CRLH TL, which is not present in the RH or the LH cases. Thus a CRLH TL has a hybrid performance. There is a frequency band in which the group velocity $v_g = \frac{\partial\omega}{\partial\beta}$ and the phase velocity $v_g = \frac{\omega}{\beta}$ are antiparallel ($v_g v_p < 0$), thus the line performs as a LH structure. There is also another frequency region in which it performs as a RH structure ($v_g v_p > 0$). The frequency stop band appeares between these LH and RH band pass regions.

In general, the series and shunt resonances of the CRLH TL are different. This situation is termed as unbalanced. However, when the series and shunt resonances are equal, namely

$$L_R C_L = L_L C_R \tag{4}$$

the LH and RH contributions balance exactly each other at a specific frequency. At this case of balance, the propagation constant reduces to the simpler expression

$$\beta(\omega) = \beta_R(\omega) + \beta_L(\omega) = \omega\sqrt{L_R C_R} - \frac{1}{\omega\sqrt{L_L C_L}} \tag{5}$$

where the phase constant distinctly splits up into the RH phase constant β_R and the LH phase constant β_L. In accordance to Equation (5) by the progressive increase of frequency, the performance of the TL changes starting from LH dominant mode (at low frequencies) to RH dominant mode at high frequencies. As it is the case of balanced line, there is not stop band and the transition from one mode to the other one occurs at the specific frequency

$$\omega_o = \frac{1}{\omega\sqrt[4]{L_R C_R L_L C_L}} \tag{6}$$

In the case of a balanced CRLH TL, β is zero at ω_o, thus an infinite guided wavelength ($\lambda g = 2\pi/|\beta|$) results. However, due to the non zero value of v_g at ω_o, propagation still occurs and radiation can leak from the structure. In addition, at ω_o ,the phase shift between adjacent cells of length d_c is zero ($|\phi| = \beta d_c = 0$), thus the leaking power of radiation has maximum at the broadside. Phase advance ($\phi > 0$) occurs in the LH frequency range ($\omega < \omega_o$), and phase delay ($\phi < 0$) occurs in the RH frequency range ($\omega > \omega_o$). On the basis of exactly this performance a CRLH transmission line would be used as a leaky wave radiating system with beam scanning ability. Advance or phase delay, depending on the frequency, would create the required condition for phase shifting between adjacent cells, and in this way the steering of the main beam is obtained. In Fig. 2 the general scheme of the metamaterial's TL composed of unit cells is shown. Each cell is considered approximately as one of the N elements of a linear antenna array which, in this way, has array factor, $AF(\theta)$, described by Equation (7).

$$AF(\theta) = \sum_{n=1}^{N} \dot{I}_n e^{j(n-1)k_o d_c \sin(\theta)} \tag{7}$$

where N represents the number of unit cells constituting the structure, $\dot{I}_n$ is the complex 'equivalent current' of the nth cell and k_o is the wavenumber in free space.

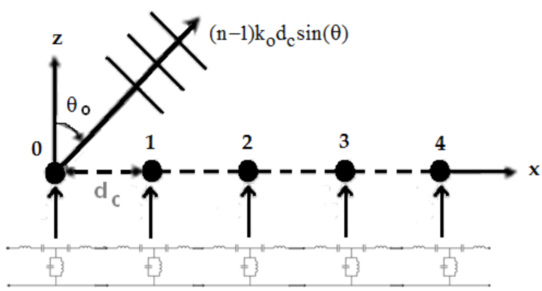

Figure 2. The LW antenna array. The array's elements and the equivalent periodically arranged CRLH unit cells of the metamaterial line.

From the transmission line view the propagation constant γ describes the propagation between the input and the output terminals of each cell, in the way described by Equation (8).

$$\begin{bmatrix} A & B \\ C & D \end{bmatrix} \begin{bmatrix} V_{n+1} \\ I_{n+1} \end{bmatrix} = e^{\pm\gamma d_c} \begin{bmatrix} V_n \\ I_n \end{bmatrix} \tag{8}$$

where d_c is the length of the cell.

In the case of the CRLH balanced line the propagation at all frequencies is $\gamma(\omega) = j\beta(\omega)$ where $\beta(\omega)$ is given by the Equation (5).

For periodic leaky wave antennas the direction θ_o of maximum radiation, at the basic mode of operation, is given by

$$\theta_o = \sin^{-1}\left(\frac{\beta(\omega)}{k_o}\right) \tag{9}$$

Therefore, when the leaky wave antenna is a CRLH transmission line the main beam can scan the space via ω.

3. SYNTHESIS OF THE PRINTED CRLH LEAKY WAVE ANTENNA AND RESULTS

In the present work, a CRLH LW antenna array was synthesized on the basis of the transmission line model. It was designed with intend to have the capability of main beam steering at a fixed frequency, instead of scanning via variation of frequency, because this operation is preferable in the majority of applications. In the proposed LW array, the direction of maximum radiation is tuned by controlling $\beta(\omega)$ (Equation (5)) via the variation of the capacitance C_R realized by a varactor diode.

In accordance to the theory of Section 2, the design of the proposed LW antenna was based on the symmetric circuit unit cell shown in Fig. 3(a). The realization of the circuit was made via microstrip elements, a shown in Fig. 3(b). The series capacitance C_L was realized with an interdigital capacitor, the series inductance $L_R/2$ was initially represented by a linear microstrip line with proper length and the parallel induction L_L with a short-circuit ended microstrip stub. The parallel capacitance C_R was realized with a varactor and it was the element by the variation of which the change of $\beta(\omega)$ is done and consequently the beam steering is obtained. The resulting unit cell is

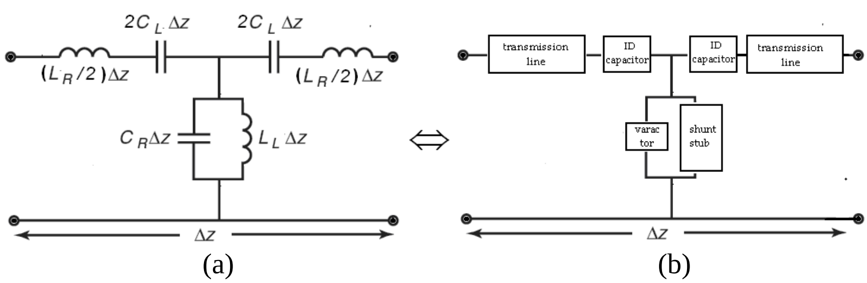

Figure 3. (a) The equivalent symmetric circuit of the CRLH cell used for the design of the LW antenna. (b) The microstrip elements and the varactor by which the symmetric circuit model is realized.

depicted in Fig. 6 and the details of the design are explained in the following.

The used varactor was one of the abrupt tuning varactors of ASI [15], with capacitance changing, at the microwave frequency of 2.4 GHz, in the range of 0.8 pF to 6.8 pF. By this varactor and with intend to obtain steering of the main beam in a range of about 50°, in accordance with the Equations (5), (6) and (9), the values of the parameters were found $L_R/2 = 95\,\text{nH}$, $L_L = 0.1\,\text{nH}$, $\text{C}_L/2$=0.35 pF.

3.1. Calculation of the Values of the Parameters

For the design, the used dielectric substrate was, the FR4 with dielectric constant $\varepsilon_r = 4.4$ and thickness 3.15 mm. The width w_{50} (Fig. 6) of the body of the LW structure was selected equal to 6 mm, which corresponds to an ordinary microstrip line characteristic impedance of 50 Ohms.

3.1.1. Design of L_R

It is well known that in printed structures the parallel connected inductive loads are easy to be implemented in contrast to the series ones. One of the few techniques for easy implementation of a series inductances is the utilization of a microstrip line. In the proposed LW antenna the problem that arises, when following this simple way, is the required large length of the microstrip line, for the realization of $L_R/2 = 95\,\text{nH}$. It is about 0.31 m because the line, of 6 mm width, on the selected substrate, has inductance equal to 306 nH/m. The problem was confronted by replacing the $L_R/2$ inductance with an equivalent circuit as shown in Fig. 4(b). The equivalent inductance L' is much smaller than $L_R/2$, thus can easily be realized via a microstrip line of short length ℓ', as shown in Fig. 4(c). With concern to the jX_p element it can be realized with a short-circuit ended microstrip transmission line.

In order for the equivalence between the circuits of Figs. 4(a) and 4(b) to be valid, their ABCD transfer matrices have to be equal, namely

$$\begin{bmatrix} A & B \\ C & D \end{bmatrix} = \begin{bmatrix} A' & B' \\ C' & D' \end{bmatrix} \tag{10}$$

The above matrices are simply calculated as

$$\begin{bmatrix} A & B \\ C & D \end{bmatrix} = \begin{bmatrix} 1 & j\omega(L_R/2) \\ 0 & 1 \end{bmatrix} \tag{11}$$

$$\begin{bmatrix} A' & B' \\ C' & D' \end{bmatrix} = \begin{bmatrix} 1 & j\omega L' \\ 0 & 1 \end{bmatrix} \begin{bmatrix} 1 & 0 \\ \frac{1}{jX_p} & 1 \end{bmatrix} \begin{bmatrix} 1 & j\omega L' \\ 0 & 1 \end{bmatrix} \tag{12}$$

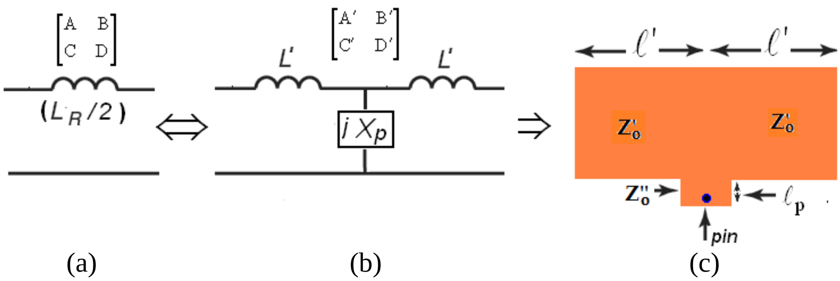

Figure 4. (a) The series inductance of the unit cell. (b) A T-circuit equivalent to the series inductance. (c) Realization of the T-circuit via a microstrip structure.

When the circuit of Fig. 4(b) is implemented with transmission line parts (Fig. 4(c)), the respective matrix is calculated as

$$\begin{bmatrix} A' & B' \\ C' & D' \end{bmatrix} = \begin{bmatrix} \cos(k\ell') & jZ_o'\sin(k\ell') \\ jY_o'\sin(k\ell') & \cos(k\ell') \end{bmatrix} \begin{bmatrix} 1 & 0 \\ \frac{1}{jZ_o''\tanh(k\ell_p)} & 1 \end{bmatrix} \begin{bmatrix} \cos(k\ell') & jZ_o'\sin(k\ell') \\ jY_o'\sin(k\ell') & \cos(k\ell') \end{bmatrix} \tag{13}$$

In agreement to Equation (13), the values of the parameters are

$$\begin{aligned} A' &= \cos(k\ell')^2 + jZ_o'\sin(k\ell')\left[\frac{\cos(k\ell')}{jZ_o''\tanh(k\ell_p)} + jY_o'\sin(k\ell')\right] \\ B' &= jZ_o'\cos\left(k\ell'\right)\sin(k\ell') + jZ_o'\sin\left(k\ell'\right)\left[\frac{Z_o'\sin(k\ell')}{Z_o''\tanh(k\ell_p)} + \cos\left(k\ell'\right)\right] \\ C' &= jY_o'\cos\left(k\ell'\right)\sin(k\ell') + \cos(k\ell')\left[\frac{\cos(k\ell')}{jZ_o''\tanh(k\ell_p)} + jY_o'\sin(k\ell')\right] \\ D' &= -\sin(k\ell')^2 + \cos(k\ell')\left[\frac{Z_o'\sin(k\ell')}{Z_o''\tanh(k\ell_p)} + \cos(k\ell')\right] \end{aligned} \tag{14}$$

By Equation (14) and under the condition that A', B', C', D' are equal to A, B, C, D, (Equation (11)), the values of the variables Z_o', ℓ_p, ℓ', Z_o'' are calculated.

3.1.2. Design of C_R

The series capacitance C_L was realized with an inter-digital capacitor. The general configuration of this load is shown in Fig. 5(a). For the present work in order to obtain the capacitance $C_L/2 = 0.35\,\text{pF}$ two fingers were used (Fig. 5(b)). The calculation of the values of

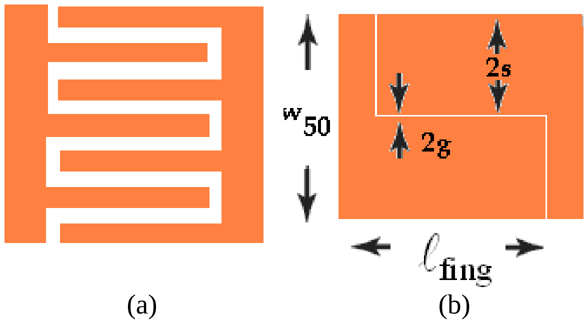

Figure 5. (a) The ordinary configuration of an interdigital capacitor. (b) The interdigital capacitor of two fingers designed for the realization of $C_L/2$ capacitance of the CRLH unit cell.

parameters were based on [16] and the mathematical formulas for two fingers are those of Equations (15) to (19).

$$C_{id} = 21_{fing}\varepsilon_o\varepsilon_{efing}\frac{K(k'_{id})}{K(k_{id})} \tag{15}$$

$$\varepsilon_{efing} = 1+\frac{\varepsilon_r-1}{2}\frac{K(k'_{id1})}{K(k_{id1})}\frac{K(k_{id2})}{K(k'_{id2})} \tag{16}$$

$$k_{id2} = \frac{\sinh\left(\frac{\pi g}{2h}\right)}{\sinh\left(\frac{\pi(g+2s)}{2h}\right)} \tag{17}$$

$$k_{id1} = \frac{g}{g+2s} \tag{18}$$

$$k'_{id} = \sqrt{1-k_{id}^2},\quad k'_{id1} = \sqrt{1-k_{id1}^2},\quad k'_{id2} = \sqrt{1-k_{id2}^2} \tag{19}$$

where ℓ_{fing} is the length of the fingers, and K (*) are elliptic integrals of the first kind.

3.1.3. Varactor and L_L Induction

The used varactor was one of abrupt tuning varactors of the ASI with capacity changing, at 2.4 GHz, from 0.8 pF to 6.8 pF. It was attached to the body of the LW antenna via a transmission line of length $\lambda g/2$, as shown in Fig. 6. The inductive impedance ωL_L was realized via short-circuit ended microstrip line, of length ℓ_L , shown in Fig. 6.

In accordance to the aforementioned analysis, the layout of the entire unit cell is depicted in Fig. 6. An array of cascaded N unit cells

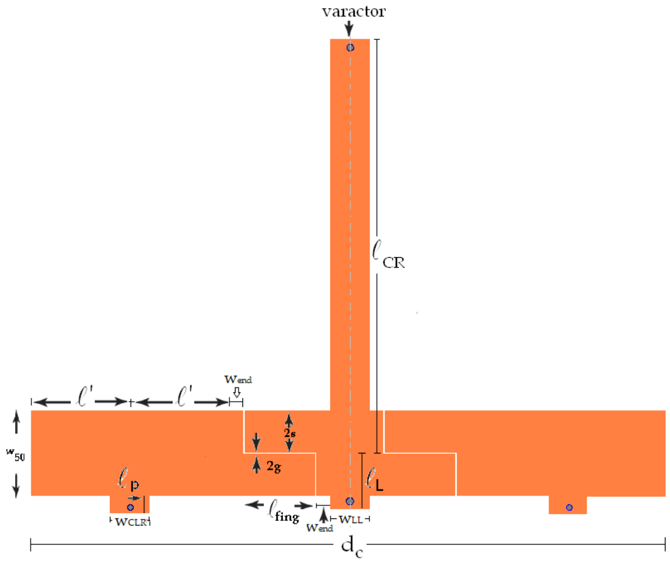

Figure 6. The realized microstrip configuration of the metamaterial CRLH unit cell of the LW antenna.

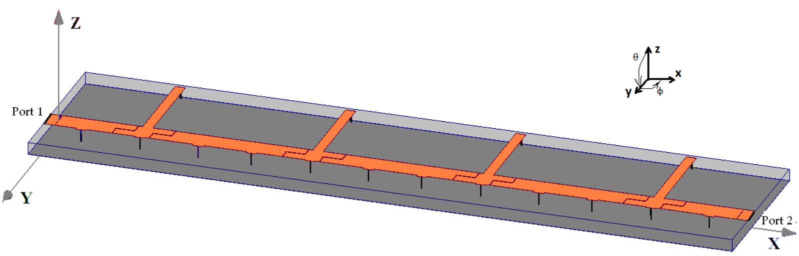

Figure 7. The synthesized LW antenna composed of four CRLH unit cells. The entire length is equal to 140.84 mm.

is theoretically expected to perform as a LW antenna. Fig. 7 depicts the layout of such a LW array with $N = 4$. Its operation was simulated via commercial high frequency electromagnetic software.

It is pointed out that the initial values of the lengths of the various parts of each cell, which were calculated by the theory, were perturbed, the target being the entire array to exhibit the desired performance. The perturbation was necessary because in the theory and the mathematical formulas used, it had not been taken into account the interaction among the individual parts of each cell as well as among the cells. The final values of the parameters are those of Table 1.

In Fig. 8 the scattering coefficient S_{11}, versus frequency, at port 1, for various values of the varactor' s capacitance is depicted. Due to the symmetry of the antenna, the S_{22} at port 2 is similar.

In Fig. 9, the gain patterns, on-xz-plane, of the antenna at 2.4 GHz

Table 1.

variables	Size [mm]
w_{50}	6.06
d_c	35.21
ℓ_{CR}	27.75
ℓ'	7.05
ℓ_p	2.80
w_{CLR} ($Z_o = 75$ Ohms)	0.71
w_{LL} ($Z_o = 75$ Ohms)	2.80
ℓ_L	4.00
ℓ_{fing}	5.00
$2s$	2.599
$2g$	0.15
w_{end} r	1.00 0.2 mm
h	3.15 mm
ε_r	4.4

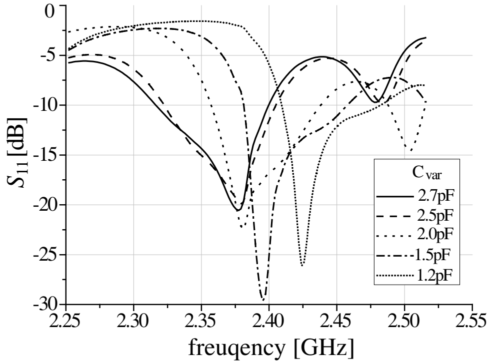

Figure 8. Scattering coefficient of the signal at the input of the LW antenna, versus frequency for various capacitances of the varactors.

and for various values of the capacitance of the varactor are depicted. Beam steering in a range of about 40 degrees was obtained. The angle of maximum radiation versus the varactor's capacity is plotted at Fig. 10. The gain variation versus the capacitance of the varactors, at 2.4 GHz, is depicted in Fig. 11. It is shown that the gain varies about ±0.5 dB, around the value of 6.5 dB, except in the cases of small capacitances.

Concluding and on the basis of the beam steering potentials of the designed LW antenna, if it is incorporated with a proper

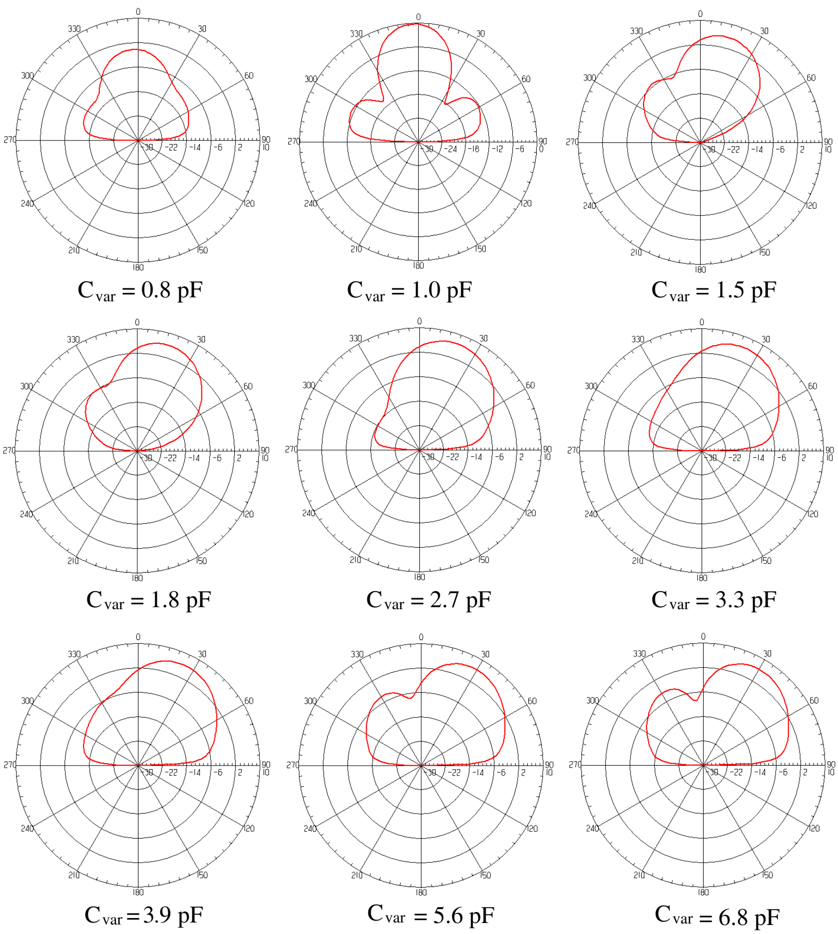

Figure 9. Gain patterns of the antenna on the-xz-planed at 2.4 GHz for various varactors' capacitances.

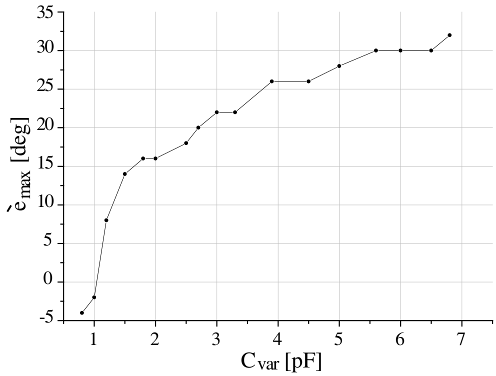

Figure 10. Directions of maximum radiation, with respect the broadside, at 2.4 GHz.

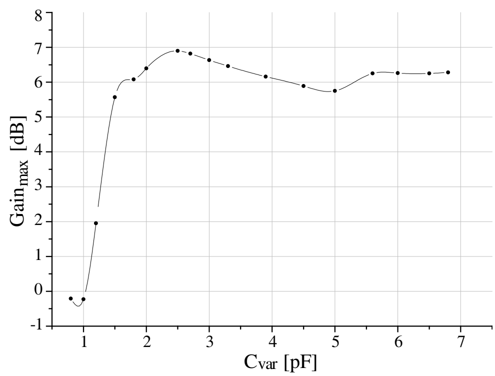

Figure 11. Variation of the Gain versus varactors' capacitances at 2.4 GHz.

microcontroller, when the received signal weakens, the antenna would steer the main lobe until it finds another direction from which the signal is stronger.

4. CONCLUSIONS

A LW antenna was designed on the basis of metamaterial transmission line concept. The structure is composed of cascaded CRLH unit cells and performs as a leaky wave antenna, with beam steering capability which results from the variation of the wave propagation constant at 2.4 GHz. This variation is controlled by the capacity of varactors incorporated into the transmission line's cells. The structure was synthesized by utilization of the equivalent circuit model of the TL. For the synthesis, new methods to realize some of the elements of CRLH unit cell are introduced. Tuning of the maximum radiation angle in a range of 40° was obtained. During the tuning, the gain varies a little, between 6 dB and 7 dB, whereas the scattering coefficient of the signal at the input of the antennas remains less than −10 dB. It was found that only four cells , resulting in antenna length equal to 140.84 mm, are enough for the above performance to be achieved. The obtained features make the antenna, if incorporated with a microcontroller for the tuning of the varactors' voltages, capable of adapting to its environment, searching, in real time, for directions from which strong signals would be received.

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