3. EXPERIMENTAL RESULTS

The proposed aperture-coupled circularly polarized antenna for UHF RFID applications has been fabricated and experimentally analyzed. It should be noticed that a thick high dielectric substrate with a relative permittivity of $\varepsilon_r = 8$ is employed for size reduction. Generally, the impedance of an aperture-coupled antenna can be matched when the aperture size is below resonant size. In this article the aperture size is determined to limit the level of back radiation to below $-20$ dB relative to the main lobe and rectangular stubs are used for effective impedance matching. Figure 4 shows the simulated and measured return loss characteristics. The measured impedance bandwidth (VSWR $\leq 1.5$) spans 60 MHz from 880 to 940 MHz, which covers the entire North American UHF RFID frequency band. In Figure 5, the measured axial ratio in the broadside direction is plotted. It is observed that a $3$ dB axial ratio bandwidth is about $30$ MHz or $3.28\%$ with respect to a center frequency of $915$ MHz. Figure 5 also illustrates the measured right-handed circular polarization (RHCP) gain within the CP bandwidth. A peak gain of $2.69$ dBi and an average gain of $2.58$ dBi are obtained in the frequency band of interest. The measured RHCP gain patterns in the two principle planes of $x$-$z$ and $y$-$z$ are shown in Figure 6.

4. CONCLUSION

In this article, a novel UHF RFID reader antenna for a handheld applications was presented. The prototype of the proposed aperture-coupled antenna has been fabricated and the measured return loss, axial ratio, and CP gain characteristics are excellent in the North American UHF RFID frequency band. The compact size of $100 \text{ mm} \times 100 \text{ mm} \times 9.6 \text{ mm}$ enables the proposed antenna to be easily integrated with portable RFID reader systems.

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1. INTRODUCTION

Metamaterials have been a subject of great interest in the scientific community since the beginning of this millennium, when it was experimentally demonstrated by Smith and coworkers [1, 2]. The existence of left handed wave propagation in an artificial medium composed of metallic wires and split ring resonators (SRRs) [3], and negative refraction in an artificial prism fabricated by means of the same metallic inclusions [2]. This represented the corroboration of the seminal (theoretical) work carried out by Veselago in 1968, where the exotic properties of artificial structures with simultaneous negative permeability and permittivity were anticipated [4].

The study of metamaterials has been very active in recent years. In microwave engineering, applications include the design of compact filters [5] and other microwave components with superior performance and/or exhibiting novel functionalities (see the monographs [6–8] to gain insight on the multiple possibilities of metamaterials in microwave engineering). From the scientific point of view, the demonstration of invisible cloaks operating at microwave frequencies [9] and beyond, and the possibility to extend the metamaterial concepts and applications to the THz and optical regions are hot topics at present [10–12].

To achieve most of the relevant characteristics of metamaterials, the artificial structures must exhibit effective media properties. To this end, the constituent building blocks (SRRs, etc.) must be electrically small, that is, with dimensions significantly smaller than the signal wavelength at the frequencies of interest. In the case of SRRs, this sub-wavelength behavior is achieved through the coupling between the inner and outer rings of the structure [13]. However, besides the subwavelength dimensions, other important requirements in magnetically driven resonators, such as SRRs and other related particles, is the presence of a strong magnetic moment at resonance. Having in mind these considerations, several resonant particles of interest for the synthesis of metamaterials have been already proposed by Marqués and coworkers [14, 15]. The main objective of this work is to demonstrate that by combining two metal layers (at both sides of a dielectric layer) connected by vias with the appropriate topology, it is possible to design new resonant particles with high level of miniaturization. A comparative analysis of the different particles in terms of dimensions and quality factor is pointed out. This analysis reveals that the application of the proposed strategies for area reduction does not represent a severe degradation in the quality factor. The immediate consequence of this fact is that, beyond the applications of these new resonators to the synthesis of metamaterials, they can also be potentially used to reduce the area of planar passive components (the application of conventional SRRs to compact planar filters has been already demonstrated [16]).

2. PRINCIPLE FOR RESONATOR’S MINIATURIZATION

Most of the subwavelength metamaterial resonant particles are constituted by at least two coupled resonators. Size reduction in these particles is due to the interaction between these individual resonators. From the physical point of view, this interaction means that there is an effective energy interchange between the individual resonators through the electromagnetic field. Depending on whether the energy interchange is mainly produced through the electric or the magnetic field, such coupling is identified as electric or magnetic. It can be shown that the nature of the coupling can be determined by analysing the symmetry of the current distribution in the resonators at the resonances [13, 16].

In general, from the point of view of the equivalent circuit model of two coupled resonators, the individual resonant particles can be modeled by LC tanks and their electric and magnetic coupling by a mutual capacitance and inductance, respectively. As a consequence of the interaction between the individual resonators, the first resonance frequency of the coupled resonators is driven to small values and, hence, the resulting resonant element (formed by the two coupled individual resonators) can be made electrically small [17]. Actually, the coupling produces a split-off of the fundamental frequency of either single resonator, and two new resonance frequencies appear: one of them smaller ($f_{down}$) and the other higher ($f_{up}$) than the resonance frequency of the individual particles. Regardless of the kind of coupling, the coupling coefficient can be estimated by [17]:

$$M = \frac{f_{up} - f_{down}}{f_{up} + f_{down}} \quad (1)$$

As an illustrative example, Figure 1 shows the transmission coefficient of a microstrip line loaded with a NB-SRR (solid line) and loaded with a single resonator (dashed line). In the inset is represented the topology of the NB-SRR and relevant dimensions, where $l = 9.4$ mm, $c = 0.4$ mm, $d = 0.05$ mm (the Rogers RO3010 substrate with thickness $h = 1.27$ mm and dielectric constant $\varepsilon_r = 10.2$ has been considered). The simulation has been performed by means of the Agilent Momentum commercial software.
$f_{\text{down}}$ is driven below the smaller of the two resonance frequencies of the individual resonators, whilst $f_{\text{up}}$ appears above the higher of these resonances [13]. The coupling between the individual resonators and the resulting displacement of $f_{\text{down}}$ towards small values is the fundamental principle used by Pendry et al. for the synthesis of electrically small resonant particles, such as the SRR [3]. In the vicinity of the first resonance, an array of such particles can be considered to form a continuous medium, provided the dimensions of the particles can be made small as compared with the wavelength. Size reduction can also be achieved in particles not consisting on two coupled resonators. This is the case of the spiral resonator (SR) [18–21], where selfcoupling within the particle is responsible for resonator miniaturization. However, in this case we cannot talk in terms of a coupling coefficient.

3. NEW DESIGN STRATEGIES FOR MINIATURIZATION

The use of two different metal layers connected by vias in the design of metamaterial resonators provides clear advantages in different ways. The first evident improvement arises from the possibility of using a second metallic level to increase the length of the particles without expanding their area. On the other hand, the two metal layers and the vias provide further flexibility and the possibility to enhance both interrings or selfcoupling, specially by etching the metallic strips of the elements face-to-face in the substrate (broad-side coupling). Figure 2(a) shows the layout and resonance frequencies of a subwavelength resonant particle based on the SRR, in which the inner and outer rings have been extended through vias to the bottom side of the substrate. The results of this figure show that the reduction ratio is around $R = 25\%$, which is higher than that obtained by considering SRRs with the conventional topology proposed by Pendry et al. (around 19%). We attribute this to the enhancement of edge coupling between the two individual resonators caused by their higher length.

In Figure 2(b), we have proceeded in a similar way, but this time trying to etch the individual resonators face-to-face. The reduction ratio is this time $R = 39\%$, which is attributed to broad-side coupling. Obviously, $R$ can be further enhanced by simply using thinner substrates. Although the structures shown in Figures 2(a) and 2(b) are quite similar from the point of view of area, strip width, and even geometry, the results show that the subwavelength property becomes more intense in the structure of Figure 2(b).

At this point, we would like to mention that an important aspect of these particles for their use in the synthesis of metamaterials is their magnetic moment at resonance. To exhibit a strong magnetic behavior, it is important that the resonator topology preserves the cancellation of the different components of current distribution in the particle. Although the analysis of the polarizabilities of the proposed particles is complicated, the inspection of the current distributions inferred from electromagnetic solvers can lead us to useful information concerning the suitability of the particle for its use in the synthesis of metamaterials. This aspect will be considered in Section 5. In the next section, we deal with another relevant aspect which is intimately related to particle losses, that is, the quality factor.

4. LOADED AND UNLOADED QUALITY FACTOR

In the most general definition, the quality factor is the ratio between the total stored energy and the dissipated energy at resonance. This important figure establishes an upper limit for the performance of microwave components based on electromagnetic resonators, and the quality factor is the usual way to have into account the effects of losses. The loaded quality factor ($Q_L$) can be estimated from the following expression:

$$Q_L = \frac{f_0}{\Delta f_{3\text{dB}}}$$

(2)

where $f_0$ is the resonance frequency and $\Delta f_{3\text{dB}}$ is 3 dB the bandwidth. The unloaded quality factor of the resonator can be obtained from $Q_L$ according to:

$$Q_u = \frac{Q_L}{1 - |S_{21}|}$$

(3)

where $S_{21}$ is the magnitude of the transmission coefficient at resonance. The unloaded quality factor, which is the actual figure of merit of the resonators, can be severely influenced by device miniaturization. The reason is that the quality factor depends inversely on the current density in the resonator at the resonance frequency and therefore, since the miniaturization implies in general an increment of the current density, a certain level of degra-
TABLE 1 3D Representations of the New Proposed Metamaterial Resonators and Relevant Characteristics for the Two Considered Substrate Thicknesses

<table>
<thead>
<tr>
<th>3D DIAGRAM</th>
<th>$h = 1.27$ mm</th>
<th>$h = 0.254$ mm</th>
<th>3D DIAGRAM</th>
<th>$h = 1.27$ mm</th>
<th>$h = 0.254$ mm</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>$f_0 = 0.80$ GHz</td>
<td>$f_0 = 0.82$ GHz</td>
<td>$\lambda_e/9.5$</td>
<td>$\lambda_e/7.6$</td>
<td>$R = 30.2%$</td>
</tr>
<tr>
<td></td>
<td></td>
<td></td>
<td>$M = 0.6$</td>
<td>$M = 0.59$</td>
<td>$Q_u = 150$</td>
</tr>
<tr>
<td>2</td>
<td>$f_0 = 0.85$ GHz</td>
<td>$f_0 = 0.49$ GHz</td>
<td>$\lambda_e/9$</td>
<td>$\lambda_e/16$</td>
<td>$R = 25.6%$</td>
</tr>
<tr>
<td></td>
<td></td>
<td></td>
<td>$M = 0.45$</td>
<td>$M = 0.95$</td>
<td>$Q_u = 129$</td>
</tr>
<tr>
<td>3</td>
<td>$f_0 = 0.43$ GHz</td>
<td>$f_0 = 0.25$ GHz</td>
<td>$\lambda_e/18$</td>
<td>$\lambda_e/32$</td>
<td>$R = 26.4%$</td>
</tr>
<tr>
<td></td>
<td></td>
<td></td>
<td>$M = 0.5$</td>
<td>$M = 0.5$</td>
<td>$Q_u = 130$</td>
</tr>
<tr>
<td>4</td>
<td>$f_0 = 0.18$ GHz</td>
<td>$f_0 = 0.09$ GHz</td>
<td>$\lambda_e/41$</td>
<td>$\lambda_e/83$</td>
<td>$R = 23.7%$</td>
</tr>
<tr>
<td></td>
<td></td>
<td></td>
<td>$M = 0.61$</td>
<td>$M = 0.33$</td>
<td>$Q_u = 135$</td>
</tr>
<tr>
<td>5</td>
<td>$f_0 = 0.35$ GHz</td>
<td>$f_0 = 0.18$ GHz</td>
<td>$\lambda_e/21$</td>
<td>$\lambda_e/43$</td>
<td>$R = 21.2%$</td>
</tr>
<tr>
<td></td>
<td></td>
<td></td>
<td>$M = 0.50$</td>
<td>$M = 0.87$</td>
<td>$Q_u = 114$</td>
</tr>
<tr>
<td>6</td>
<td>$f_0 = 0.84$ GHz</td>
<td>$f_0 = 0.49$ GHz</td>
<td>$\lambda_e/9$</td>
<td>$\lambda_e/15$</td>
<td>$R = 24.6%$</td>
</tr>
<tr>
<td></td>
<td></td>
<td></td>
<td>$M = 0.45$</td>
<td>$M = 0.85$</td>
<td>$Q_u = 136$</td>
</tr>
<tr>
<td>7</td>
<td>$f_0 = 0.31$ GHz</td>
<td>$f_0 = 0.20$ GHz</td>
<td>$\lambda_e/24$</td>
<td>$\lambda_e/39$</td>
<td>$R = 26.4%$</td>
</tr>
<tr>
<td></td>
<td></td>
<td></td>
<td>$M = 0.64$</td>
<td>$M = 0.5$</td>
<td>$Q_u = 130$</td>
</tr>
<tr>
<td>8</td>
<td>$f_0 = 0.34$ GHz</td>
<td>$f_0 = 0.32$ GHz</td>
<td>$\lambda_e/22$</td>
<td>$\lambda_e/24$</td>
<td>$R = 23.7%$</td>
</tr>
<tr>
<td></td>
<td></td>
<td></td>
<td>$M = 0.61$</td>
<td>$M = 0.33$</td>
<td>$Q_u = 135$</td>
</tr>
</tbody>
</table>

Current density distribution (obtained by means of Agilent Momentum) is interesting because it gives an idea of the magnetic behavior of the particles at resonance. No current cancellation is observed in these particles. Therefore these are of interest for the synthesis of metamaterials and, in general, in those applications requiring magnetically coupled resonators. In fact, the different topologies shown in the table have been designed with the aim to avoid the destructive interferences between the magnetic fields produced by the currents in the strips at both sides of the substrate. All the simulations have been performed by considering the Rogers RO3010 substrate with $\varepsilon_r = 10.2$, and two different thicknesses: $h = 0.254$ mm and $1.27$ mm. In all the layouts the values of the indicated geometric parameters are $l = 15.28$ mm, $d = 0.2$ mm, and $c = 0.6$ mm. In the second and third columns are depicted the first resonance frequency (renamed $f_0$), $M$, $Q_u$, and the size of the particle (side) expressed in terms of the guided wavelength, for the two considered substrate thicknesses. $Q_u$ has been obtained from
electromagnetic simulations (by considering ohmic and dielectric losses) through the use of expressions (2) and (3), where the particles have been coupled to two microstrip lines (input and output lines) with via shorts in the extremes, and the particle being placed in between at a fixed distance from the extremes of each line. With this set-up the coupling between the particles and the lines varies (the particles are not necessarily located in those positions where the current in the line is a maximum) and this may produce certain influence on the value of $Q_u$. To properly compare the $Q_u$ of the different particles, the same coupling is convenient, but this is difficult to achieve in practice.

The first and the second rows of Table 1 are dedicated to the conventional SRR and the BC-SRR. It is interesting to observe the significant reduction in the electrical size for the thinnest substrate in the case of the BC-SRR. This is obviously explained by the enhancement of coupling that can be achieved between metals etched face to face in a narrow substrate. This size reduction is also visible in other particles of Table 1 where broadside coupling is also present.

Particles 3 and 4 are significantly different from the other particles presented in the table since their subwavelength property does not arise from the interaction between two resonators, but from selfcoupling. The structure of these particles is based on the SR, which has been extended through vias to the bottom side of the substrate in such a way that the magnetic field created by the current flow in the strips does not cancel at resonance. The difference between the results obtained for $h = 1.27$ and 0.254 mm points out the importance of the broad-side capacitance between the strips etched at the two metallic levels. Obviously, for these-particles we have not represented $M$ and $R$, since these magnitudes have significance only if the particle is composed of two individual resonators.

The particles 5 and 6 are composed of two coupled resonators. In both cases the geometry has been specially designed to magnify the broadside coupling. The increment in the coupling coefficient that is obtained by reducing the substrate thickness varies from $M = 0.50$ to 0.87 for the Particle 5 and from $M = 0.45$ to 0.85 for the Particle 6. This significant dependence of the coupling coefficient with the substrate thickness proves that in both cases the coupling is mainly governed by the broadside capacitance. This is obvious in Particle 6, but it is not evident in Particle 5, where different metallic strips share the same metal level and are very close. On the other hand, the presence of face-to-face metallic strips does not necessary imply the existence of significant broadside coupling. Particle 8 is a clear example of this situation. Although the geometry of this particle contains metallic strips aligned at both sides of the substrate, it forces the current to close through an edge capacitance at the first resonance. Thus, by the effect of the geometry, the edge capacitance becomes the dominant coupling mechanism at the first resonance. As a consequence, the reduction of substrate width does not substantially decrease the electrical size of the particle.

The different examples collected in Table 1 marshal a representative sample of the miniaturization possibilities of metamaterial resonators by the inclusion of vias and the use of a second metallic layer. Concerning $Q_u$, it can be appreciated that the SRR provides the best result. However, we do not obtain significant degradation in this figure for the other resonator topologies. Nevertheless, we must mention that the technique to obtain $Q_u$ is not
very accurate due to the strong sensitivity of \( Q_u \) with the modulus of the transmission coefficient at resonance (Expression 3) and to the different coupling between the resonators and the input and output lines (as has been mentioned before). Hence, the results provided in the table are estimative, rather than exact values of \( Q_u \).

6. ILLUSTRATIVE EXAMPLES OF APPLICATION TO THE DESIGN OF LEFT HANDED STRUCTURES

The proposed structures can replace the conventional SRRs in many applications, like, for instance, the design of left handed transmission lines. In Figures 3 and 4, the frequency responses, including the dispersion relation, of one-cell left handed microstrip structures based on the conventional SRR (see Fig. 3) and on the resonator labeled 7 in Table 1 (see Fig. 4), are depicted (the characteristics of the Rogers RO3010 substrate have been considered, with dielectric constant \( \varepsilon_r = 10.2 \) and thickness \( h = 1.27 \) mm). The \( \beta l \) diagram, which has been inferred from the phase of the transmission coefficient, \( \phi_T \), according to

\[
\cos(\beta l) = \frac{\cos(\phi_T)}{|S_{21}|}
\]

where \( |S_{21}| \) is the magnitude of the transmission coefficient, indicates that the phase and group velocities are antiparallel in the region of interest (to derive Expression 4, it has been assumed that the attenuation constant is null in the transmission band). The central frequency of the left handed band is centered in the vicinity of 2.3 GHz in both cases. However the structure of Figure 4 is much smaller for the reasons that have been explained before. The significant insertion losses in Figure 4 are attributed to the smaller mutual inductance between the line and the particle. It is expected that particle excitation is enhanced by using coplanar waveguide (CPW) configurations. These CPW-based cells, which were proposed previously [23], consist on a CPW line with thin metal wires that connect the central strip to the ground planes (acting as shunt inductances) and two coupled resonators. A three-dimensional (3D) representation of this structure, using BC-SRRs (Particle 2 in Table 1) and its frequency response and dispersion are depicted in Figure 5. The dimensions for the CPW and the shunt inductances are as follows: CPW line width \( W = 7 \) mm, slots width \( G = 1.35 \) mm, and shunt inductances width \( l_0 = 0.4 \) mm. The geometrical dimensions for the resonators are: side of the outer ring \( l = 7.8 \) mm, strip width \( c = 0.6 \) mm, and separation between resonators \( s = 0.55 \) mm. Simulations have been performed using the Agilent Momentum simulator. The considered substrate has a dielectric constant of \( \varepsilon_r = 10.2 \) and the three metal levels are separated a distance of 0.635 mm. The characteristics of the frequency response and dispersion of the structure of Figure 5 are similar to those of Figures 3 and 4.

Figure 5 Left-handed cell implemented by using BC-SRRs (particle number 2 of Table 1) coupled to a CPW with shunt inductances. (a) 3D representation; (b) S-parameters and \( \beta l \)

Figure 6 Fabricated prototype of a left handed cell corresponding to a CPW loaded with resonators labeled as number 3 in Table 1 (a), and measured frequency response (b). [Color figure can be viewed in the online issue, which is available at www.interscience.wiley.com]
Two additional left handed cells in CPW technology have been designed and fabricated, but using two of the new resonators. One of them contains the resonator number 3 (see Fig. 6) and the other one contains the resonator number 5 (see Fig. 7). The measured frequency responses (obtained by means of the Agilent HP8720ET vector network analyzer) are also depicted in these figures. The left handed cells have been fabricated using the Rogers RO3010 substrate with relative permittivity $\varepsilon_r = 10.2$ and thickness $h = 0.635$ mm. Actually, we have used two different substrates. In one of them we have etched the rings (at both sides) and we have drilled the holes for vias metallization. In the other, the CPW structure and the inductive wires have been etched. After aligning both substrates, they have been soldered and subjected to pressure to minimize the presence of air in the contacting region. CPW dimensions are $W = 5$ mm, $G = 1.16$ mm and the wire inductance has a width of $I_w = 0.4$ mm in both cells. The geometrical parameters of the resonator number 3 are as follows: the side length is $l = 4.8$ mm, the strip width $c = 0.6$ mm and separation between the resonators $s = 1.36$ mm. For the cell with the resonators number 5 the dimensions are $l = 5$ mm, $c = 0.6$ mm, the separation between adjacent strips $d = 0.2$ mm and $s = 1.16$ mm. In view of the measured frequency responses of these left handed cells, insertion losses in the allowed band are within acceptable limits (a significant improvement as compared to Figure 4 has been obtained by using a CPW configuration). The small dimensions of the resonators make the cell to be electrically very small. The cell size is actually given by the side length of the resonators, and this is as small as $\lambda_g/24$ and $\lambda_g/23$ for the resonators of Figures 6 and 7, respectively. In future works, the application of these miniaturized left handed cells to the design of microwave components is expected. Their application to the design of reflectors is another potential application.

7. CONCLUSIONS

As a conclusion, the use of two metal levels connected by vias in the design of subwavelength resonant particles provides the possibility to introduce broadside coupling in the designs and hence to enhance the miniaturization. The geometry and the topology of the particles play a very important role in size reduction since the mere presence of face-to-face metal strips does not guarantee broadside coupling. Several merit figures have been pointed out as numerical indicators of particle performance. The analysis of the proposed particles shows that by enhancing the coupling between the constituent elements of the resonators or self coupling, it is possible to enhance miniaturization without severe quality factor degradation. Some of the proposed resonant particles have been used to design several compact left handed cells in microstrip and CPW technology. The frequency responses are indicative of better performance in CPW-loaded structures. Work is in progress to apply the proposed particles to the design of planar passive microwave devices, where miniaturization is the key point.

ACKNOWLEDGMENT

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Figure 7  Fabricated prototype of a left-handed cell corresponding to a CPW loaded with resonators labeled as number 5 in Table 1 (a), and measured frequency response (b). [Color figure can be viewed in the online issue, which is available at www.interscience.wiley.com]
MINIATURIZED MICROSTRIP CROSS-COUPLED BANDPASS FILTER USING NOVEL STEPPED IMPEDANCE RESONATORS WITH A DESIRABLE UPPER STOPBAND

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ABSTRACT: A miniaturized microstrip cross-coupling bandpass filter using new stepped impedance resonators (SIRs) with simultaneous suppression of the 2nd harmonic passband response is presented. Four improved and miniaturized SIRs constitute the compact filter, which exhibits a sharp transition band due to two transmission zeros at both sides of the passband. Two quarter-wavelength open-ended stub resonators are applied to provide extra transmission zeros for suppressing the 2nd harmonic passband. A bandpass filter with a center frequency at 2.5 GHz was designed and fabricated, of which experimental results validated the proposed filter design.

Key words: microstrip bandpass filter; cross-coupling; harmonic suppression; miniaturization; stepped-impedance resonator

1. INTRODUCTION

In modern wireless communication systems, high-selectivity small-size microstrip bandpass filters with excellent out-of-band rejections are required to enhance the system performance and reduce the fabrication cost. To obtain a sharp transition band, one effective method is to introduce two transmission zeros and locate them at either side of the passband, respectively. Cross-coupled configurations, which exhibit elliptic function-like responses [1], are suitable and commonly applied.

Conventional microstrip bandpass filters are designed with half-wavelength or quarter-wavelength resonators. To reduce filter size, cross-coupled filters using folded half-wavelength hairpin resonators were developed [2]. Stepped impedance resonators have attracted much attention due to their controllable first spurious passband [3, 4]. With modified stepped impedance hairpin resonators, those microstrip filters were more compact and had better stopband extensions [5].

The planar bandpass filters made of resonators inherently have the spurious passbands at multiple of the center frequency, which limit the upper stopbands. Many methods have been proposed to solve the problem, such as providing different electric length for even and odd modes [6], using uniplanar compact electric bandgap (EBG) structure [7] or defected ground structure (DGS) [8], creating transmission zeros by open-ended stub resonators [9], and so on.

In this letter, we propose a novel miniaturized microstrip filter structure that is suitable for realizing the high selectivity and suppressing the 2nd harmonic passband. In the design, four improved folded half-wavelength hairpin SIRs are placed as a two-by-two array to achieve the cross-coupling. Therefore, the cross-coupled capacitance is introduced directly between two adjacent resonators with two transmission zeros introduced. In addition, two quarter-wavelength open-ended stubs are attached at the edges of the resonators in the main-coupling path. One extra transmission zero is introduced at the first harmonic frequency. A compact high-selectivity bandpass filter at 2.5 GHz with harmonic suppression is optimally designed, fabricated, and measured.

2. RESONATOR AND FILTER DESIGN

Figure 1 shows layout of the proposed filter. Four symmetric basic resonators are placed oppositely to each other and form a four-square contour. Tapped-line input/output ports are connected to two resonators, in which the distance between the feed point and the resonator center is \( L_0 \). The resonator consists of two triangular patches connected to both ends of a high impedance microstrip line section, which is bent three times to reduce circuit area. The resonator is an improved variation of the SIR with square-shape as well. The main coupling principle is identical to the hairpin SIR’s. The SIR section is used in each element, of which the size is adjustable with the relevant impedance ratio [10].

Figure 2 shows the equivalent circuit of the proposed filter. In each resonator, the distributed inductance \( L_1 \) and \( L_2 \) are formed by

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