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MINIATURIZED FORCED-MODE RING RESONATOR WITH CAPACITIVE LOADING

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Abstract—The miniaturization of conventional ring resonators is demonstrated by forcing a voltage minimum at one end of the resonator. In addition, the resonator is loaded with a capacitance to achieve further miniaturization and reducing its sensitivity to substrate thickness tolerance. The final resonator is 73% smaller than a conventional ring resonator and has a tenfold decrease in sensitivity to substrate thickness variations. Using this resonator a 4-pole quasi-elliptic filter is fabricated showing good agreement between simulation and experimental results.

1. INTRODUCTION

Ring resonators have been widely studied in the literature for filter applications [1]. A ring resonator consists of a 360° closed-loop transmission line where a full-wavelength standing wave is excited. Due to the large size of these resonators, different techniques have been suggested in the literature to achieve miniaturization. In [2] a cross-shaped structure is presented where a size reduction of about 30% is accomplished. In [3] two interconnected rings (one inside the other) are used to obtain a size reduction of 50%. Finally, in [4] a voltage maximum is forced by the introduction of a capacitive gap hence attaining 50% size reduction. In the present work a new miniaturization technique is proposed by means of a via to ground that forces the resonant mode to have a voltage minimum at one end of the resonator. By doing this the resonator size reduces 50%. In addition, an interdigital capacitance is added across the resonator. Since one side

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is grounded, it is equivalent to increasing the capacitance to ground, thus further miniaturization is obtained. The conventional method to increase ground capacitance is to widen the resonator transmission lines or to add large patches to the resonator [5]. However, the disadvantage is that the resonant frequency becomes more sensitive to tolerances in substrate thickness, which is undesirable for mass production filters where little or nil trimming is required [6–9]. It will be shown that the proposed interdigital technique reduces resonator size while reducing the sensitivity to substrate thickness. The final resonator is 73% smaller than a conventional ring resonator, and the resonator's sensitivity to substrate thickness decreases to about 1/10 of the original sensitivity.

In this paper, a 4 pole quasi-elliptic filter using the highly miniaturized ring resonator is presented. The filter has a bandwidth of 11% and it is centered at $f_0 = 0.55\,\text{GHz}$. Full design procedure will be exposed and good agreement between simulated and experimental results will be shown.

2. RESONATOR DESIGN

The design starts with a conventional square ring resonator [1] on a substrate of relative permittivity $\varepsilon_r = 10.8$ with a thickness of $h = 1.27\,\text{mm}$ at a center frequency of 1.1 GHz as shown in Figure 1(a). The length of the resonator is 28 mm and the line width is $w = 0.8\,\text{mm}$. Then, a via to ground is added at one end which forces the resonant mode to have a voltage minimum at one end. Forcing voltage maximum in the ring resonator can result in the reduction of the resonant frequency. Introduction of via to ground in the square ring resonator helps in removing the voltage maximum, this affects the resonant frequency of the resonator. In this way, the resonator's size is reduced by 50% to 14 mm for the same resonant frequency (Figure 1(b)). Figure 2 depicts the current distribution at the resonant frequency of the ring resonators with and without via. Furthermore, an interdigital capacitance is added across the resonator as shown in Figure 1(c), where the width of the fingers and the gap between fingers are $x = 0.4\,\text{mm}$. It is well known that the resonant frequency is inversely proportional to the inductance and capacitance. By increasing the capacitance of the resonator, we can reduce the resonant frequency of the resonator.

The equivalent circuit of the latter is shown in Figure 1(c), where two transmission lines with length L are connected in parallel to the series interdigital capacitance C and its spurious capacitance Cs [10]. For this model, lossless components will be considered. This circuit can be expressed by equivalent impedance as (1).

$$Z_{eq} = \frac{Z\tan(\beta L)}{j\left[\omega(C+Cs)Z\tan(\beta L)-2\right]} \tag{1}$$

Since $Z_{eq} = V/I$ and for a standing wave condition we have that $I = 0$, then we can derive (2).

$$C + Cs = \frac{2}{\omega Z\tan(\beta L)} \tag{2}$$

For an interdigital capacitance, the values of C and Cs can be calculated using (3) and (4) [10],

$$C = (\varepsilon_r + 1)\, L\left[(N-3)\, A_1 + A_2\right] \tag{3}$$

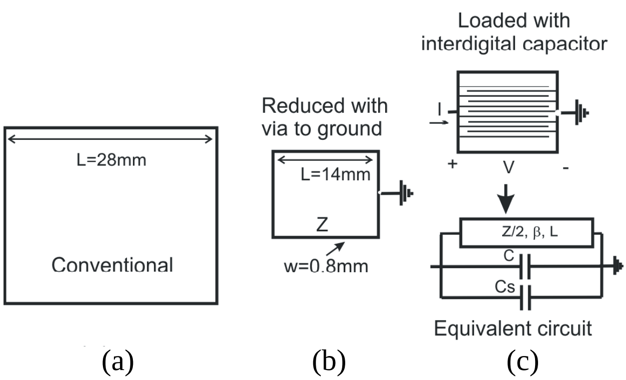

Figure 1. (a) Conventional ring resonator. (b) Miniaturization by forcing mode with via to ground. (c) Loading with capacitor and its equivalent circuit.

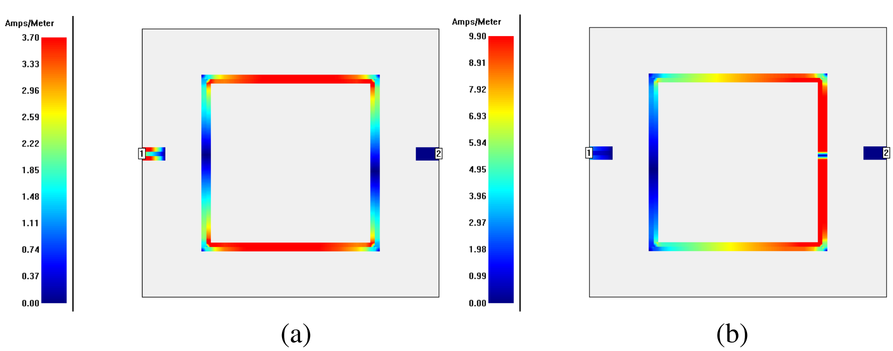

Figure 2. Current distributions on a square ring resonator (a) without and (b) with via to ground.

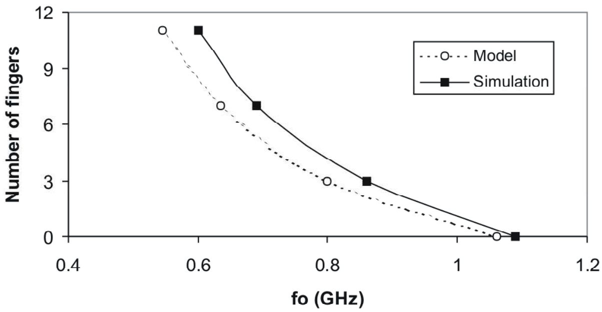

Figure 3. Resonant frequency of the via-hole resonator with different number of capacitor fingers.

where, $A_1 = 0.04\left[0.33 - 0.15\left(\frac{x}{h}\right)\right]^2$; $A_2 = 0.04\left[0.5 - 0.23\left(\frac{x}{h}\right)\right]^2$

$$Cs = \frac{10^{-3}}{2cZ_0 v_p} \tag{4}$$

where N is the number of fingers, c the speed of light, and v_p the phase velocity for the respective substrate. By solving Equation (2) with the values of Cs and C from Equations (3) and (4), the center frequency of the resonator can be calculated for different number of fingers. Figure 3 shows the calculated and simulated [11] center frequency (f_0) of the resonator for different numbers of fingers. In the simulation, lossless materials are assumed. As expected, for a larger capacitance value, the resonant frequency decreases therefore, miniaturizing the structure. For $N = 0$, the calculated resonant frequency is $f_{01} = 1.1$ GHz and $f_{01} = 1.12$ GHz for the simulation. When $N = 11$, the modeled value becomes $f_{02} = 0.45$ GHz, while the simulation shows $f_{02} = 0.6$ GHz. However, there is a slight mismatch between the model and simulation values, especially, as we increase the number of fingers. This could be attributed to higher parasitic effects as the capacitance value increases [12].

By comparing the simulation responses, one can see that $f_{02} = 0.55\ f_{01}$. Therefore the overall resonator miniaturization is about 73% compared to a conventional ring.

Moreover, the addition of the interdigital capacitance also reduces the sensitivity (S) of the resonant frequency to changes in substrate thickness. To prove this, the height of the substrate h was increased by $\Delta h = 2$ mm to $h_b = 3.27$ mm. Then, the different structures were simulated in [11]. To calculate the sensitivity to substrate thickness

Equation (5) is utilized.

$$S(\%) = \frac{f_{0b} - f_0}{f_0} \cdot 100 \tag{5}$$

where f_{0b} is the resonant frequency when $h_b = 3.27\,\text{mm}$ and f_0 when $h = 1.27\,\text{mm}$. When $N = 0$, the sensitivity is $S = 6\%$, whereas when $N = 11$, it is $S = 0.6\%$, a tenfold decrease in sensitivity.

3. FILTER DESIGN

In [13] the design procedure for quasi-elliptic filters is presented. For $\Omega_a = 1.8$, a 4-pole bandpass filter with a fractional bandwidth of 10% has the following low pass g values: $g_1 = g_4 = 0.95974$ and $g_2 = g_3 = 1.42192$: To achieve a quasi-elliptic response, the second and third resonators must be coupled with a negative coupling $J_2 = -0.21083$; and the first and fourth resonators should have a positive coupling $J_1 = 1.11769$. From these values, the coupling coefficients (k) between resonators and the external $Q(Q_{ext})$ can be calculated from (6–9).

$$Q_{ext} = \frac{g_1}{FBW} = 9.5974 \tag{6}$$

$$k_{1,2} = k_{3,4} = \frac{FBW}{\sqrt{g_1 g_2}} = 0.0856 \tag{7}$$

$$k_{2,3} = \frac{FBW \cdot J_2}{g_2} = 0.0786 \tag{8}$$

$$k_{1,4} = \frac{FBW \cdot J_1}{g_1} = -0.02196 \tag{9}$$

To achieve the required positive coupling between resonators 2 and 3, the configuration of Figure 4(a) is utilized, where the magnetic coupling is dominant. For the negative coupling between resonators 1 and 4 the configuration of Figure 4(b) is used which has an electrical coupling. For resonators 1–2 and 3–4, the configuration of Figure 4(c) is used with mixed coupling. In order to extract the filter parameters from the coupling coefficients, a simulation in [11] is performed by weakly coupling the resonators to the feed lines as shown in Figure 5. Then, distance d between resonators is varied and the k is computed from (10).

$$k = \pm \frac{f_2^2 - f_1^2}{f_2^2 + f_1^2} \tag{10}$$

where f_1 and f_2 are the resonant frequencies.

The final results are shown in Figure 4. For the required filter characteristics, $d_1 = 0.41\,\text{mm}$, $d_2 = 0.32\,\text{mm}$ and $d_3 = 0.5\,\text{mm}$.

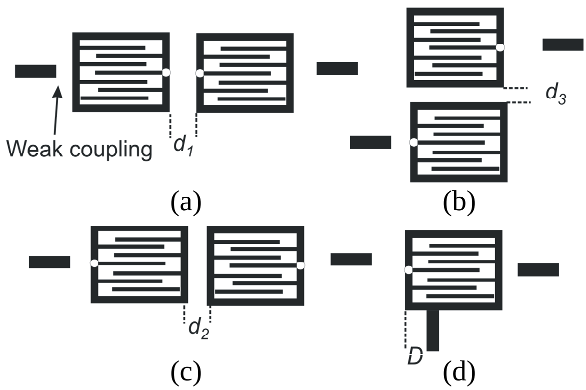

Figure 4. Different coupling configurations. (a) Magnetic coupling. (b) Electric coupling. (c) Mixed coupling. (d) External coupling.

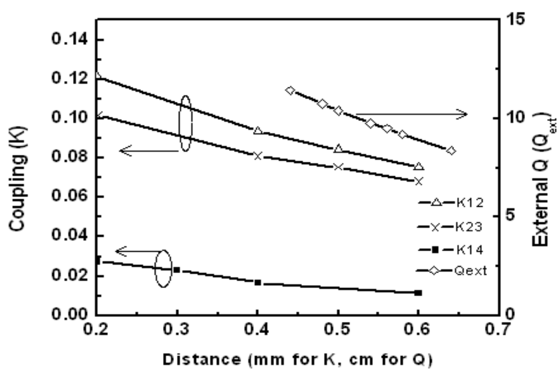

Figure 5. Values of k for different separation distances (d) and external coupling values (Q_{ext}) for different values of D.

To extract the external coupling, Figure 4(d) is simulated for different values of D and (11) is computed and plotted in Figure 5. From the figure, the required dimension is D =0.55 cm.

$$Q_e = \frac{f_0}{\Delta f_{3\,\mathrm{dB}}} \tag{11}$$

4. EXPERIMENTAL AND SIMULATION RESULTS

The final filter was simulated in [11] including material losses. Then, it was fabricated on a Rogers Duroid-6010 substrate with $\varepsilon_r = 10.8$ and

$h = 1.27$ mm. Figure 6 shows the photograph of the fabricated circuit.

Figure 7 depicts the simulated and measured results. The simulated response shows a center frequency of 0.55 GHz with a fractional bandwidth of 11%. The insertion losses in the filter passband are better than 1.77 dB, and the reflection losses are better than 21 dB. The measured values show a center frequency of 0.51 GHz with a fractional bandwidth of 11.3%. The insertion losses are better than 1.88 dB, whereas the reflection losses are better than 11 dB. The differences are attributed to inaccuracies in the fabrication process and material tolerances.

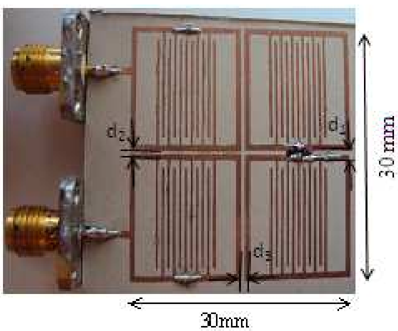

Figure 6. Photograph of the quasi elliptic filter using ring resonator with interdigital capacitor and via hole. ($d_1 = 0.41$ mm, $d_2 = 0.32$ mm and $d_3 = 0.5$ mm).

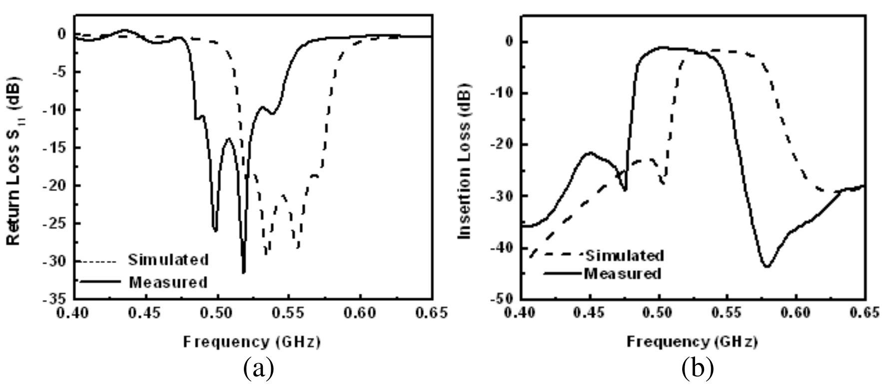

Figure 7. Measured and simulated. (a) Return loss. (b) Insertion loss.

5. CONCLUSIONS

In this work, the miniaturization of conventional ring resonators was demonstrated by forcing a voltage minimum at one end of the resonator. In addition, the resonator was loaded with a capacitance to achieving further miniaturization and reducing its sensitivity to substrate thickness tolerance. The final resonator is 73% smaller than a conventional ring resonator and has a tenfold decrease in sensitivity to substrate thickness variations. Finally, a 4-pole quasi-elliptic filter was demonstrated at a center frequency of 0.55 GHz showing good agreement between simulation and experimental results.

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