A Dual-Mode Bandpass Filter With Enhanced Capacitive Perturbation

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***Abstract*—A dual-mode ring bandpass filter with two pairs of capacitors has been designed. The capacitors are used to control the location of the even- and odd-mode frequencies independently, allowing weak coupling for narrow-band filter design with realizable capacitance values. Theoretical expressions have been derived for these frequencies. A 4% bandwidth bandpass filter centered at 1.9 GHz was designed and tested with good agreement between theoretical and measured results.**

***Index Terms*—Bandpass, dual mode, filter, ring.**

I. Introduction

THE microstrip ring resonator has been extensively used in the design of filters, mixers, and couplers in microwave engineering. It has also been used in the measurement of dispersion, phase velocity, and effective dielectric constant. The microstrip ring bandpass filter has received much attention for its simple implementation and robustness, which is highly sought after in mobile and satellite communication systems. Fig. 1 shows a microstrip ring resonator weakly coupled to the feedlines. Two reference planes AA' and BB' are included in Fig. 1 for future reference. Resonance is established when the circumference of the ring is equal to an integral number of the guided wavelength. The theory and application of various ring circuits are well documented in [1].

It is well known that the ring resonator can support two resonance orthogonal modes. When a perturbation is introduced either along AA' or BB', reflections are generated in the two opposing traveling waves propagating along the ring [2]. This results in the generation of two split modes. Depending on the magnitude of the reflected waves, it will influence the level of coupling and bandpass response results. Perturbations in the form of a stub [3] and impedance step [4] along one of the principal diagonals AA' or BB' have been reported. However, in [3], a single pair of perturbation along AA' is only able to influence the even-mode frequency. This fact is later generalized in [5], whereby a single pair of perturbation is only able to control either one of the split modes. Although simultaneous control of the split modes is allowed in [4], we observed that a small impedance ratio might be difficult to realize.

In this paper, we proposed a new perturbation topology. Instead of perturbing in only one of the principal diagonals, we introduced series capacitances at both AA' and BB' planes, as shown in Fig. 2. An arrangement like this is able to control both split-mode frequencies independently, and will be proven below. It offers total control of the split-mode frequencies resulting in a more robust design. The coupling coefficient can also be shown to be a function of the difference of C_1 and C_2. In [5], it is a function of the magnitude of the reactive element. This will relieve the burden to rely on high capacitance for narrow-band filter design. A high-capacitance capacitor has low self-resonant frequency and is not suited for high-frequency operation. Due to the symmetry of the newly proposed ring resonator, it will also be shown later that the design of the bandpass filter is governed simply by the characteristic equation of the ring.

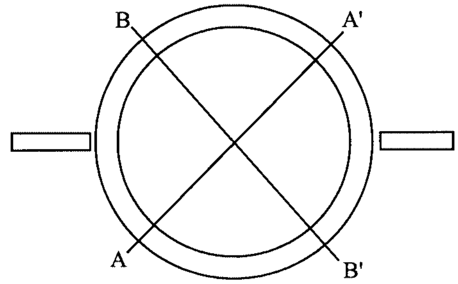

Fig. 1. Weakly coupled microstrip ring resonator.

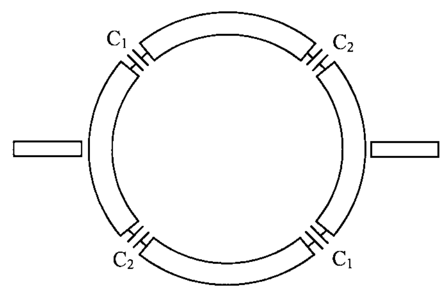

Fig. 2. Proposed dual-mode ring resonator topology.

II. Resonator Analysis

For dual-mode operation, there must be symmetry in all diagonals [3]. Thus, the opposite capacitors are chosen to be the same, namely, C_1 and C_2, as shown in Fig. 2. Fig. 3 shows the proposed resonator with its weakly coupled input and output ports separated spatially by 90°. The structure in Fig. 3 can be analyzed by adopting even–odd-mode analysis [2]. From conventional resonator theory, the ring is resonant when its input

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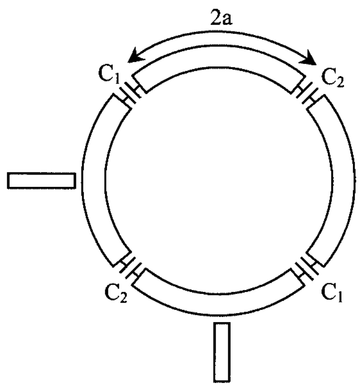


Fig. 3. Newly proposed dual-mode resonator.

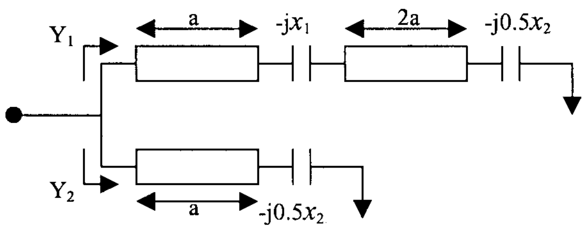


Fig. 4. Odd-mode equivalent circuit of ring resonator.

admittance is zero. For both even and odd modes, their respective resonant conditions are shown as

$$Y_1^{\mathrm{odd,even}} + Y_2^{\mathrm{odd,even}} = 0 \tag{1}$$

where Y_1 and Y_2 are the upper and lower arm input admittance, respectively.

A. Odd Mode

Fig. 4 shows the equivalent-circuit representation of half the ring resonator when an electric wall is applied along AA', as is the case for the odd mode. The value a is an eighth of the ring circumference and x_1 and x_2 are the normalized reactance of the perturbation elements.

The odd-mode input admittance y_1^o and y_2^o of the upper and lower arms, respectively, have been derived as

$$Y_1^o = j\frac{2+x_2\tan 2\theta+\left[(2x_1+x_2)+(x_1x_2-2)\tan 2\theta\right]\tan\theta}{(2x_1+x_2)+(x_1x_2-2)\tan 2\theta-(2+x_2\tan 2\theta)\tan\theta} \tag{2a}$$

$$Y_2^o = j\frac{2+x_2\tan\theta}{x_2-2\tan\theta} \tag{2b}$$

where $\theta = \beta^o a$ and β^o is the odd-mode phase constant.

Hence, the odd-mode resonant frequency is given as

$$\tan(2\beta^o a) + \frac{2}{x_2} = 0. \tag{3}$$

B. Even Mode

Fig. 5 shows the equivalent-circuit representation of half the ring resonator when a magnetic wall is applied along AA', as is the case for the even mode.

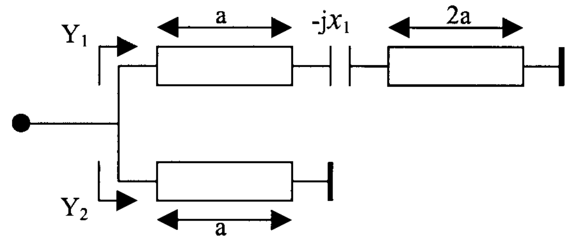


Fig. 5. Even-mode equivalent circuit of ring resonator.

Similar to the odd-mode case, the even-mode input admittance Y_1^e and Y_2^e of the upper and lower arms, respectively, are

$$Y_1^e = j\frac{\tan 2\theta + x_1\tan 2\theta\tan\theta + \tan\theta}{x_1\tan 2\theta + 1 - \tan 2\theta\tan\theta} \tag{4a}$$

$$Y_2^e = j\tan\theta \tag{4b}$$

where $\theta = \beta^e a$ and β^e is the even-mode phase constant.

The even-mode resonant frequency is given as

$$\tan(2\beta^e a) + \frac{2}{x_1} = 0. \tag{5}$$

Two important observations were made from the above even-odd mode analysis of this resonator. The first observation is that the split-mode frequencies can be controlled independently. The odd-mode frequency is controlled by C_2, while the even-mode frequency is controlled by C_1. The second observation is that both resonant conditions, as shown in (3) and (5), have the same mathematical form. If we replaced all four capacitors by a common capacitance value, it will reduce to a single equation depicting the resonant frequency of the natural state of the modified semilumped ring. Hence, the modified semilumped ring's natural frequency, as well as its even-and odd-mode frequencies, all shares the same characteristic equation as follows:

$$f_r = \frac{c_0}{\pi^2 r\sqrt{\varepsilon_{\mathrm{reff}}}}\left[\pi - \tan^{-1}(4\pi f_r C Z_R)\right] \tag{6}$$

where

- f_r even-mode, odd-mode, or modified semilumped ring's natural frequency;
- r average radius of ring;
- C capacitance of the four common capacitors;
- Z_R characteristic impedance of the ring.

III. Bandpass Filter Analysis

The center frequency of a bandpass filter can be approximated from the average of the even- and odd-mode frequencies as follows:

$$f_c = \frac{1}{2}(f_e + f_o) = \frac{c}{4a\sqrt{\varepsilon_{\mathrm{reff}}}}\left[1 - \frac{1}{2\pi}\left(\tan^{-1}\frac{2}{x_1} + \tan^{-1}\frac{2}{x_2}\right)\right]. \tag{7}$$

The coupling between the two degenerate modes is characterized by the coupling coefficient k [6], which can be computed

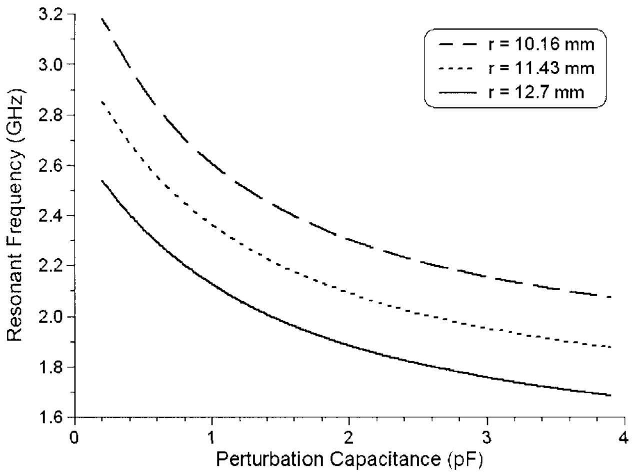


Fig. 6. Graphical representation of the characteristic equation.

from the knowledge of even- and odd-mode frequencies as follows:

$$\begin{aligned} k &= \frac{\Delta f}{f_c} \\ &= \frac{2|f^o - f^e|}{f^o + f^e} \\ &= \frac{\frac{1}{\pi}\left|\tan^{-1}\frac{2}{x_1} - \tan^{-1}\frac{2}{x_2}\right|}{1 - \frac{1}{2\pi}\left(\tan^{-1}\frac{2}{x_1} + \tan^{-1}\frac{2}{x_2}\right)}. \end{aligned} \tag{8a}$$

From (8a), we see that k can be made numerically small, which is needed in narrow-band filter design. Due to the differential term in the numerator, it is noted that weak coupling can be realized easily with large x (or low capacitance), as compared in the case of Type 2 in [5], which required a high capacitance value. In order to gain further insight into the coupling, (8a) can be further approximated for positive weak coupling. It is assumed that $1/x_1 + 1/x_2 \ll 1/\pi$. Equation (8a) is then reduced to

$$k \approx \frac{\frac{1}{\pi}\left(\frac{2}{x_1} - \frac{2}{x_2}\right)}{1 - \frac{1}{2\pi}\left(\frac{2}{x_1} + \frac{2}{x_2}\right)} \approx 4Z_R f_c(C_1 - C_2). \tag{8b}$$

From (8b), we can see that the coupling coefficient is now linearly proportional to the difference of the perturbation capacitor values. Thus, such filters can take advantage of low capacitor value, which has a better quality factor and self-resonant frequency for the required weak coupling. From circuit simulation, it was also noted that the condition for the existence of attenuation poles is $C_1 > C_2$.

The main features of the proposed ring resonator can be described by its characteristic equation. Therefore, it is intuitive to present a graphical plot, as shown in Fig. 6, displaying the relationship between various resonant frequencies and perturbation capacitors. The width of the ring resonator is chosen to be 1.524 mm, resulting in an effective dielectric constant ε_{reff} of 7.5 and Z_R of 29.4 Ω. The resonator is fabricated on RT6010 with a thickness of 0.635 mm and a relative dielectric constant of 10.2.

TABLE I
SUMMARY OF DESIGN PARAMETERS

Centre Frequency	1.90 GHz	
Fractional Bandwidth ω	0.04	
Coupling Coefficient k	0.0283	
f_o	1.93 GHz	C_2 = 1.7 pF
f_e	1.87 GHz	C_1 = 2.1 pF

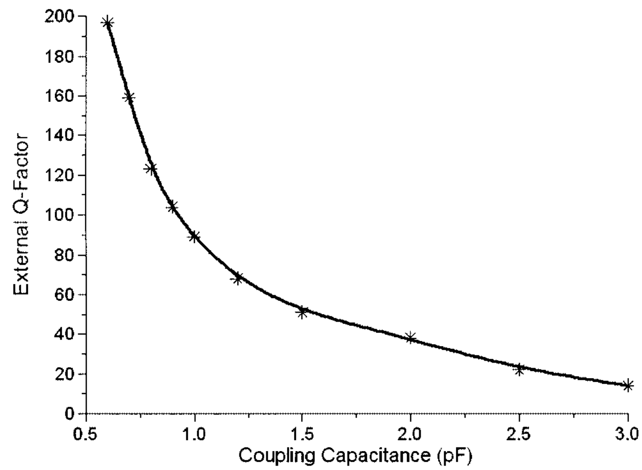


Fig. 7. External Q factor Q_E against coupling capacitance.

IV. Filter Design

A bandpass filter of center frequency 1.9 GHz with a fractional bandwidth ω of 4% will be designed and fabricated using the proposed ring resonator outlined in Section III. A two-stage Butterworth filter [7] has been selected with its low-pass prototype elements as $g_0 = 1.0$, $g_1 = 1.4142$, $g_2 = 1.4142$, and $g_3 = 1.0$. The inter-stage coupling of the filter can be determined by using

$$k = \frac{\omega}{\sqrt{g_1 g_2}}. \tag{9}$$

By equating (9) and (8a), two perturbation capacitors C_1 and C_2 were chosen to provide the necessary coupling from Fig. 6 and are tabulated in Table I.

The loaded Q factor Q_L of the modified ring is then noted from the 3-dB bandwidth using a general circuit simulator such as Agilent's Advanced Design System. The unloaded Q factor Q_O can then be estimated using [1]

$$Q_O = \frac{Q_L}{1 - 10^{((-L)/20)}} \tag{10}$$

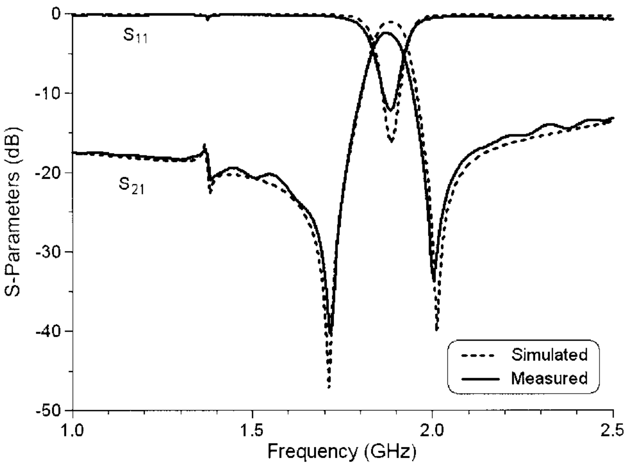
where L is the insertion loss in decibel.

The external Q factor Q_E is then computed [1] using (11), and a plot of Q_E against the coupling capacitance is shown in Fig. 7 as

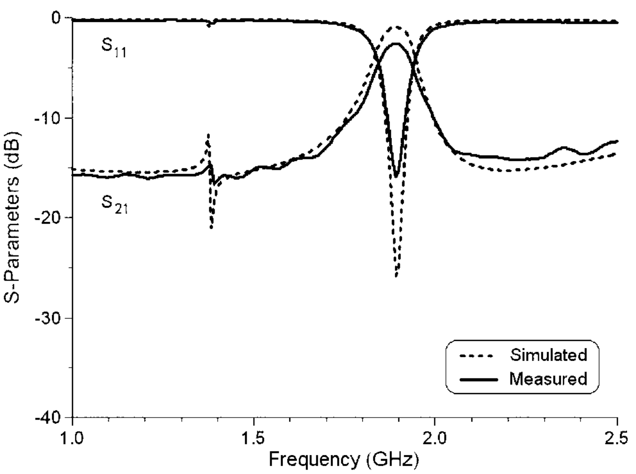
$$Q_E = \frac{2Q_L Q_o}{Q_o - Q_L}. \tag{11}$$

Assuming the same input and output impedance, the required external Q factor can be estimated [7] as

$$Q_E = \frac{g_0 g_1}{\omega}. \tag{12}$$



(a)



(b)

Fig. 8. Response of the designed filter at 1.9 GHz for the case of: (a) $C_1 > C_2$ and (b) $C_2 > C_1$.

The external Q factor was computed to be 35 and, with the aid of Fig. 7, the required coupling capacitance was found to be 2.25 pF. However, a capacitor of this value may have significant series resistance, as well as low self-resonant frequency. Moreover, dc isolation is inherent in this filter due to the perturbation capacitors. As such, a direct tap is applied instead. Two similar filters based on the above specifications have been fabricated and measured using RT6010 with a thickness of 0.635 mm and a relative dielectric constant of 10.2. Fig. 8(a) shows the comparison between simulated and measured results for the case of C_1 greater than C_2. The measured bandwidth is approximately 2.8% with a minimum insertion loss of 2.1 dB in the center of the passband. The measurement in Fig. 8(b) revealed a bandwidth of 2.5% and a minimum insertion loss of 2.5 dB. It is noted here that the condition for the existence of attenuation poles is C_1 greater than C_2.

It was also observed from Fig. 8 that the insertion loss in the passband is higher than that simulated. This is because of the finite quality factor, as well as the equivalent series resistance of the capacitors used.

V. Conclusion

A new coupling method using two sets of orthogonal positioned capacitors in a ring resonator has been described. The self-resonant frequency, as well as the even- and odd-mode frequencies, can be described by a single characteristic equation. By making use of the difference of the capacitances, narrow-band bandpass filter can be designed with such a topology. A filter has been demonstrated with this new perturbation. The simulated and measured results have shown good agreement.

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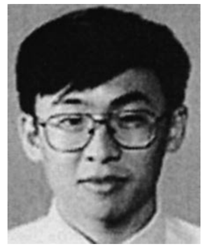

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