A compact low insertion loss bandpass filter based on meandered self-coupled ring resonator

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Abstract. The traditional structure of dual mode bandpass filter is formed by a circular or rectangular resonator, which causes excessive circuit areas to be occupied. A 5.2 GHz bandpass filter, based on the triple-branch self-coupled ring resonator, is designed to achieve most essential features such as low insertion loss, compact size, and wide stopband. The coupling of each coupled branch introduces the even- and odd-mode perturbation, which produce sharp bandpass and wide stopband responses due to the generation of band-edge transmission zeros. The simulation and measurement in this paper are respectively verified by using IE3D electromagnetic simulator and Agilent's HP8722C network analyzer. Experimental results show that the filter has 19.2% bandwidth centered at 5.2 GHz, 1.2 dB insertion loss, 25 dB stopband rejection from 6 to 9 GHz. Moreover, the circuit size in this paper has been down to $9 \times 6 \text{ mm}^2$ and can be applied to communication of microwave applications.

1 Introduction

The component of passive circuit, such as the filter occupying the largest percentage of communication system structure, is more important to preserve high performance and reduce the electronic circuit size. Many researches of dual-mode bandpass filter have been proposed in papers [1-9].

The bandpass filters based on the dual-mode ring resonator was first analyzed by Wolff [1]. Some papers proposed dual mode ring resonator bandpass filters by using step-impedance or ring resonators with different perturbation schemes [2-4]. The perturbation element is a plus to or cut from the resonator, and it is also viewed as capacitive or inductive [5]. One of the methods is the investigation of designing a compact, low insertion loss and sharp rejection by cascaded dual mode ring resonator [6]. Furthermore, dual-mode microstrip filters are highly attractive for radio frequency systems because their sizes are compact and easy to fabricate [7]. By adding open stubs symmetrically on ring resonator, the dual mode filter is realized [8].

However, there is a novel distributed perturbation by self-coupled segments of the ring resonator, which brings the easier implementation scheme [9]. In this paper, the ring resonator is based on the theory of a self-coupled, simple bandpass filter with low insertion loss and the compact size designed, which showed the easy structure and narrow bandwidth with two transmission zeros near the passband. This filter of 5.2 GHz band is designed. The measurement results show good consensus with simulation.

2 Filter design

The traditional structure of dual mode bandpass filter is shown in Fig. 1. Generally, the filter structure is formed by a circular or rectangular resonator, which causes excessive circuit areas to be occupied. To make up for the disadvantage, the ring resonator was bended in this study to achieve the goal of saving the filter size.

Fig. 1. Structure of traditional dual mode bandpass filter.

2.1 Schematic of proposed bandpass filter

Unlike those studies in the past using the perturbation approach, the ring resonator with coupling method is applied to cause the even and odd mode to be separated, and to achieve resonance effects. Thus, the filter offers an advantage of a simpler circuit layout and is easier to be implemented. The schematic types of the proposed self-coupled bandpass filter with input and output ports are shown in Fig. 2. The proposed bandpass filter is composed of self-coupled ring resonator with parallel-

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coupled input and output networks. The self-coupled ring includes a central coupled branch and two other coupled branches, connected at both ends of the central coupled branch. An intuitive way to understand this self-coupled ring resonator is the meandering of the conventional circular ring resonator such that the ring size can be considerably miniaturized. Different from the lumped perturbation of the conventional ring resonator, the distributed coupling of the proposed ring resonator provides the mode perturbation between even and odd modes.

2.2 Resonance conditions

In Fig. 2, the layout discontinuity effect of both coupled sections are neglected, with the structure of the ring resonator symmetric to line $\overline{AB}$, and the even-odd mode theory applied to analyze the resonance mode characteristics. Fig. 3(a) and (b) show that the equivalent even-mode, odd-mode circuits, the even-mode and odd-mode resonance conditions can be derived from equation (1)-(6) and ignore the coupling support effect on both sides of the bent section. Two coupled sections at ends have the same parameters: the electrical length of θ_1 , the coupling coefficient of c_1 , and the characteristic impedance Z_1 . The central coupled section has the electrical length of θ_2 , the coupling coefficient of c_2 , and the characteristic impedance of Z_2 .

Fig. 2. Structure of proposed coupled-ring bandpass filter.

The equation derived is shown as below. The equivalent even circuit of dual-mode ring resonator is shown in Fig. 3 (a). By calculating the input impedance of the even mode, the even mode resonates when $Z_{Be} = \infty$. On the other hand, the input admittance of the even mode $Y_{Be} = 0$.

$$Y_{Be} = jY_{2e} \frac{Z_{2e} - \frac{1}{2} \left(Z_{1o} \tan \frac{\theta_{1e}}{2} - Z_{1e} \cot \frac{\theta_{1e}}{2} \right) \tan \theta_{2e}}{\frac{1}{2} \left(Z_{1o} \tan \frac{\theta_{1e}}{2} - Z_{1e} \cot \frac{\theta_{1e}}{2} \right) + Z_{2e} \tan \theta_{2e}} = 0 \quad (1)$$

Then

$$Z_{2e} - \frac{1}{2} \left(Z_{1o} \tan \frac{\theta_{1e}}{2} - Z_{1e} \cot \frac{\theta_{1e}}{2} \right) \tan \theta_{2e} = 0 \quad (2)$$

The even mode resonance condition becomes

$$\tan \theta_{2e} = \frac{2 \sqrt{\frac{1+C_2}{1-C_2}}}{Z \left(\sqrt{\frac{1-C_1}{1+C_1}} \tan \frac{\theta_{1e}}{2} - \sqrt{\frac{1+C_1}{1-C_1}} \cot \frac{\theta_{1e}}{2} \right)} \quad (3)$$

Where

$$C_1 = \frac{Z_{1e} - Z_{1o}}{Z_{1e} + Z_{1o}}, C_2 = \frac{Z_{2e} - Z_{2o}}{Z_{2e} + Z_{2o}}, Z_1 = \sqrt{Z_{1e} Z_{1o}}, Z_2 = \sqrt{Z_{2e} Z_{2o}}$$

$$Z = \frac{Z_1}{Z_2}$$

$$\frac{\theta_{2e}}{\theta_2} = \frac{f_{re}}{f_{norm}} = \frac{\theta_{1e}}{\theta_1}; \theta_1 + \theta_2 = \pi$$

Fig. 3. Equivalent even and odd-mode circuit of dual-mode ring resonator.

Similarly, The equivalent odd circuit of dual-mode ring resonator is shown in Fig. 3 (b), and the input impedance vanishes when the odd mode resonates,

$$Z_{Bo} = jZ_{2o} \frac{\left(Z_{1e} \cot \frac{\theta_{1o}}{2} - Z_{1o} \tan \frac{\theta_{1o}}{2} \right) \tan \theta_{2o} + \frac{2Z_{1e}Z_{1o}}{Z_{2o}}}{\left(Z_{1e} \cot \frac{\theta_{1o}}{2} - Z_{1o} \tan \frac{\theta_{1o}}{2} \right) - \frac{2Z_{1e}Z_{1o}}{Z_{2o}} \tan \theta_{2o}} = 0 \quad (4)$$

Then

$$\left(Z_{1e} \cot \frac{\theta_{1o}}{2} - Z_{1o} \tan \frac{\theta_{1o}}{2} \right) \tan \theta_{2o} + \frac{2Z_{1e}Z_{1o}}{Z_{2o}} = 0 \quad (5)$$

The odd mode resonance condition is obtained as

$$\tan \theta_{2o} = \frac{2Z \sqrt{\frac{1+C_2}{1-C_2}}}{\left(\sqrt{\frac{1-C_1}{1+C_1}} \tan \frac{\theta_{1o}}{2} - \sqrt{\frac{1+C_1}{1-C_1}} \cot \frac{\theta_{1o}}{2} \right)} \quad (6)$$

Where

$$C_1 = \frac{Z_{1e} - Z_{1o}}{Z_{1e} + Z_{1o}}, C_2 = \frac{Z_{2e} - Z_{2o}}{Z_{2e} + Z_{2o}}, Z_1 = \sqrt{Z_{1e} Z_{1o}}, Z_2 = \sqrt{Z_{2e} Z_{2o}}$$

$$Z = \frac{Z_1}{Z_2}$$

$$\frac{\theta_{2o}}{\theta_2} = \frac{f_{ro}}{f_{norm}} = \frac{\theta_{1o}}{\theta_1}; \theta_1 + \theta_2 = \pi$$

Under the even-odd mode resonant frequency f_{re} and f_{ro} , the θ_{1e} , θ_{2e} , θ_{1o} and θ_{2o} are the electrical lengths corresponding to the physical lengths ℓ_1 and ℓ_2 respectively. Thus, the resonance frequency condition is obtained as $f_r = \sqrt{f_{re} \cdot f_{ro}}$.

Since the complicated form of equation (1)-(6) is difficult to obtain the analytic solution, we use the graphic expression for analyzing the effect of the even-mode and odd-mode resonant frequencies with the impedance ratio and coupling coefficients.

2.2.1 Impedance ratio effect Z

A case about $c_1=0.2$ and $c_2=0.2$ is plotted in Fig. 4 and it shows the dependence of mode resonant frequencies with respect to the impedance ratio of Z with several characteristics as below.

- 1) When impedance ratio $Z = 1$, the even- and odd-mode resonant frequency are not separated.
- 2) When impedance ratio $Z < 1$, the odd-mode resonant is lower than the even-mode and has the inductive perturbation feature.
- 3) When impedance ratio $Z > 1$, the odd-mode resonant frequency is higher than the even-mode and the type is in a capacitive perturbation state.
- 4) The geometric average resonant frequency, $f_{r,SCR} = \sqrt{f_{re} f_{ro}}$ is apathetical to Z.

Therefore, these arguments are determined to support the operating band and bandwidth of the filter in this paper.

Fig. 4. Resonant frequencies with respect to the impedance ratio of Z.

Fig. 5. Resonant frequencies with respect to the coupling coefficient.

2.2.2 Characteristics of Coupling Coefficient

Fig. 5 (a) and (b) are the relative resonant frequency between coupling coefficient C_1 and C_2 for different impedance ratios under the $\theta_2 = 38^\circ$. There are several characteristics as below.

- 1) When the coupling coefficient increases, the resonant frequency will increase at any impedance ratio.

- 2) When $Z < 1$, the resonant frequency will increase following the incremental impedance ratio in the any coupling coefficient.
- 3) When $Z > 1$, the the resonant frequency will decrease following the incremental impedance ratio in the any coupling coefficient.
- 4) When $Z = 1$, it shows the highest resonant frequency.

2.3 Transmission zeros analysis

The main advantage of the ring resonator filter is generating two transmission zeros near the passband. Because there are two transmission paths between the input and output ports, therefore when the current of the two paths is offset each other at the output, the transmission zeroes on both sides of the passband are generated. In this way, Y-parameter matrix method (Admittance Matrix Method) is used to facilitate the analysis of transmission zero generation.

Ignoring the discontinuous effect of the circuit branch bending, as shown in Fig. 6, the entire resonator can be disassembled into a parallel coupling line with electrical length $\theta_1/2$, impedance Z_1 and a parallel coupling line with electrical length θ_2 , impedance Z_2 . Here $\theta_c = \theta_{io}/2$, $\theta_a = \theta_1/2 - \theta_c$ and θ_{io} as the length of the electrical input between the input and the output port are defined, and $\theta_1 + \theta_2 = \pi$.

Fig. 6. Self-coupling ring resonator two-way transmission path equivalent circuit.

In Fig. 7, the equations (7) to (12) are the formulas for the parallel coupling microstrip Y parameter of the four ports and the two ports. Further, the formulas (7) to (12) are converted into the following formulas (13) to (18) using the relationship between the coupling coefficient and the impedance.

Fig. 7. parallel coupling transmission line diagram (a) four port (b) two port.

$$\begin{bmatrix} Y_{11} & Y_{12} & Y_{13} & Y_{14} \\ Y_{21} & Y_{22} & Y_{23} & Y_{24} \\ Y_{31} & Y_{32} & Y_{33} & Y_{34} \\ Y_{41} & Y_{42} & Y_{43} & Y_{44} \end{bmatrix} = \begin{bmatrix} Y_{11} & Y_{12} & Y_{13} & Y_{14} \\ Y_{12} & Y_{11} & Y_{14} & Y_{13} \\ Y_{13} & Y_{14} & Y_{11} & Y_{12} \\ Y_{14} & Y_{13} & Y_{12} & Y_{11} \end{bmatrix}$$

$$Y_{11} = (-\frac{j}{2})[Y_e + Y_o] \cot \theta \quad (7)$$

$$Y_{12} = (-\frac{j}{2})[Y_e - Y_o] \cot \theta \quad (8)$$

$$Y_{13} = (-\frac{j}{2})[Y_e - Y_o] \csc \theta \quad (9)$$

$$Y_{14} = (-\frac{j}{2})[Y_e + Y_o] \csc \theta \quad (10)$$

$$\begin{bmatrix} Y_{11}' & Y_{12}' \\ Y_{21}' & Y_{22}' \end{bmatrix} = \begin{bmatrix} Y_{11}' & Y_{12}' \\ Y_{12}' & Y_{11}' \end{bmatrix}$$

$$Y_{11}' = \frac{j}{2}[Y_e \tan \theta - Y_o \cot \theta] \quad (11)$$

$$Y_{12}' = \frac{j}{2}[Y_e \tan \theta + Y_o \cot \theta] \quad (12)$$

$$Y_{11} = Y \frac{-j}{\sqrt{1-c^2}} \cot \theta \quad (13)$$

$$Y_{12} = Y \frac{-jc}{\sqrt{1-c^2}} \cot \theta \quad (14)$$

$$Y_{13} = Y \frac{j}{\sqrt{1-c^2}} \csc \theta \quad (15)$$

$$Y_{14} = Y \frac{-jc}{\sqrt{1-c^2}} \csc \theta \quad (16)$$

$$Y_{11}' = Y \frac{-j}{\sqrt{1-c^2}} \cot \theta \quad (17)$$

$$Y_{12}' = Y \frac{-j}{\sqrt{1-c^2} \sin 2\theta} (c + \cos 2\theta) \quad (18)$$

By using the formulas (13) to (18), we can solve the position of transmission zero in Fig. 6. The transitive formula is shown in equation (19), and when Y_{21} is zero, the transmission zero is generated.

$$Y_{21} = Q_d + \frac{2S_d T_d (G_d + P_u) - (H_d + Q_u)(S_d^2 + T_d^2)}{(H_d + Q_u)^2 - (G_d + P_u)^2} \quad (19)$$

And within the formula (19),

$$\begin{aligned}
 P_u &= -\frac{jY_2}{\sqrt{(1-c_2^2)}\sin 2\theta_2}(c_2 + \cos 2\theta_2), \\
 Q_u &= \frac{jY_2}{\sqrt{(1-c_2^2)}\sin 2\theta_2}(1 + c_2 \cos 2\theta_2), \\
 S_d &= Q_{ac} - \frac{S_{ac}G_{ac}}{2H_{ac}}, T_d = -\frac{S_{ac}G_{ac}}{2H_{ac}}, \\
 G_d &= T_{ac} - \frac{G_{ac}^2}{2H_{ac}}, H_d = -\frac{G_{ac}^2}{2H_{ac}}, Q_d = -\frac{Q_{ac}^2}{2H_{ac}}, \\
 Q_{ac} &= T_c - S_c \frac{(Q_a + Q_c)}{P_a + P_c}, \\
 S_{ac} &= S_c - T_c \frac{(Q_a + Q_c)}{P_a + P_c}, \\
 T_{ac} &= P_c - \frac{S_c^2}{P_a + P_c}, \\
 G_{ac} &= Q_c - \frac{S_c T_c}{P_a + P_c}, H_{ac} = P_c - \frac{T_c^2}{P_a + P_c}, \\
 P_c &= Y_1 \frac{-j}{\sqrt{1-c_1^2}} \cot \theta_c, \quad Q_c = Y_1 \frac{j c_1}{\sqrt{1-c_1^2}} \cot \theta_c, \\
 S_c &= Y_1 \frac{j}{\sqrt{1-c_1^2}} \csc \theta_c, \quad T_c = Y_1 \frac{-j c_1}{\sqrt{1-c_1^2}} \csc \theta_c.
 \end{aligned}$$

The complicated form of (19) makes it difficult to obtain the analytic solutions in calculation. Then we turn to the graphic expression of Y_{21} for analyzing the effects of the impedance ratio Z and input-output separation length θ_{10} [9].

2.3.1 Effect of the impedance ratio Z

Fig. 8 is the coupling coefficient with C_1 and C_2 being equal to 0.2, $\theta_1 = 140^\circ$, $\theta_2 = 40^\circ$ and Z ratio being 0.5, 1.0, 1.8 in the normalization of the frequency of the Y_{21} relationship diagram when the angle between the input and output is 80 degree. As shown in the Fig. 8, it can be found that there is no transmission zero on both sides of the fundamental frequency passband when the inductive coupling $Z < 1$. When $Z > 1$ is capacitively coupled, two transmission zeros are generated on both sides of the passband and a transmission zero occurs near the second passband. When $Z = 1$, the transmission zero will occur near the base band, destroying the passband characteristics.

2.3.2 Effect of angle between input and output ports

Fig.9 is the normalized frequency of Y_{21} relationship diagram showing when the coupling coefficient C_1 and C_2 are equal to 0.2, the Z ratio is 1.8, with the angle

between input port and output being 60° , 70° , 80° , 90° , 100° . It can be seen from the Fig. 9 that when the angle between the input and output port is $70^\circ \sim 80^\circ$, the passband on both sides of the transmission zero points shows the most symmetry.

Fig. 8. Z ratio of the normalization frequency of Y_{21} relationship diagram.

Fig. 9. shows the Y_{21} relationship diagram difference between the input and output ports at the normalized frequency.

3 Experimental results

Following the theoretical analysis in section 2, in this paper, the coupling coefficients C_1 and C_2 are all used at 0.2, the electrical length $\theta_1 = 142^\circ$, $\theta_2 = 38^\circ$, the characteristic impedance Z_1 is 109Ω , Z_2 is 61.5Ω , and the impedance ratio $Z = 1.77$. Moreover, the relative dimensions of the resonator is equivalent to $W_2 = 1.2$ mm, $W_1 = 0.3$ mm, $W_t = 0.2$ mm, $L_1 = 8.5$ mm, $L_2 = 4.5$ mm, $g_2 = 0.6$ mm, $g_1 = 0.6$ mm, and $g_t = 0.2$ mm.

This filter is fabricated on the substrate with a relative dielectric constant of 3.38, with loss tangent being 0.0025 and thickness being 0.762 mm. The actual photograph of 5.2GHz self-coupled bandpass filter is shown in Fig. 10. The size of self-coupled bandpass filter is 9×6 mm². On the other hand, the circuit size is relatively 54% of the size to the traditional uniform ring case.

Fig. 10. Photograph of 5.2 GHz bandpass filter.

Fig. 11. shows the measurement and simulation results of bandpass filter.

The simulation and measurement are respectively verified by using IE3D electromagnetic simulator and Agilent's HP8722C network analyzer [10]. Fig. 11

shows the simulated and measured results. Finally, there is very good consensus between the measured and simulated results in this paper that within 5-5.4 GHz, the measurement return loss is greater than 22 dB and measurement insertion loss is less than 1.2 dB and this condition improves the response of passband. The measured 3-dB bandwidth is about 19.2% at the center frequency of 5.2 GHz. The filter has three transmission zeros close to the passband at 4.39 GHz, 6.42 GHz and 7.58 GHz, which produces sharp bandpass and wide stopband response.

4 Conclusion

A 5.2 GHz dual-mode bandpass filter has been proposed in this study by using self-coupled ring resonator for small size, low insertion loss, wide stopband and it is easy to implement. The proposed filter size is $9 \times 6 \text{ mm}^2$, three transmission zeros are close to the passband at 4.39 GHz, 6.42 GHz and 7.58 GHz, and the 3-dB bandwidth is about 19.2% from 4.75 GHz to 5.73 GHz, and insertion loss is less than 1.2 dB at center frequency 5.2 GHz. A good consensus has been reached in this study by simulation and measurement showing highly valuable application of proposed bandpass filter in wireless communications.

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