A Flat-Passband Microstrip Filter With Nonuniform-Q Dual-Mode Resonators

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***Abstract*—A compact lossy microstrip bandpass filter is realized with flat passband and high selectivity, using the method of nonuniform Q. Two dual-mode resonators are utilized. The Q-factor of odd mode can be properly reduced by loading resistors over the symmetric plane of each resonator, with the even-mode Q unchanged. The geometric and resistance parameters are determined by the derived formulas and optimization. The 0.2 dB bandwidth of the fabricated prototype is improved by 62% when compared with the uniform-Q counterpart. The equivalent Q-factor is as high as 970, which is extracted from measurement.**

***Index Terms*—Dual-mode, flat-passband, lossy bandpass filter, microstrip, (non)uniform Q-factor.**

I. Introduction

MINIATURIZED filters are key elements in various microwave and wireless communication systems and always desired in integrated electronics. However, the resulting lower Q-factors make narrowband filters suffer relatively high in-band insertion loss (IL), degraded passband flatness and frequency selectivity [1]. Recently, a new class of bandpass filters, called lossy filters, has been introduced, and some design techniques have been presented, including predistortion [1], [2], lossy coupling matrix synthesis [1], [3], and nonuniform-Q methods [4]–[7]. These lossy filters have improved frequency selectivity and passband flatness, but at the cost of additional in-band IL. For a receive filter, the IL can be compensated by cascading after a low-noise amplifier and it will have little impact on the overall noise figure of receiver [1].

The nonuniform-Q method was first proposed in [4]. For a uniform-Q filter, the resonators which dominate the response around bandedges will generate more dissipation than those dominating middle band. So the passband may be flattened by increasing the Q-factor of bandedge-dominating resonators or decreasing the Q-factor of middle-dominating resonators. This method has the advantage of easy realization and may be directly applied to the existing topology, without introducing additional resistive couplings [5]. More recently, dual-mode [6] and triple-mode [7] microstrip resonators are exploited to design third-order nonuniform-Q filters for miniaturization.

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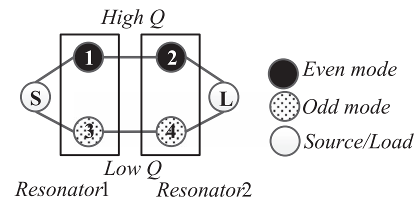

Fig. 1. Coupling topology for the proposed dual-mode filter with nonuniform-Q resonant modes.

In this letter, a new compact fourth-order nonuniform-Q filter is proposed using two dual-mode microstrip resonators. The coupling topology is shown in Fig. 1, which consists of two coupling paths. One is with two high-Q even modes while the other is with two low-Q odd modes. And the Q-factor of odd modes can be tuned by attaching resistors to proper position of the microstrip resonators. Compared with the conventionally designed filter counterpart, a much flatter passband is successfully achieved with the proposed filter.

II. Analysis and Design

Fig. 2(a) shows the structure of the proposed lossy filter, which consists of two dual-mode microstrip resonators positioned symmetrically to the BB' plane. Due to the symmetry of the whole structure, the even- and odd-modes of two resonators are orthogonal to each other, which means they are decoupled. Therefore, there are two independent coupling paths, as shown in Fig. 1. The resonant nodes 1 and 3 correspond to the even- and odd-modes of dual-mode Resonator 1, respectively, and so do the resonant nodes 2 and 4 of Resonator 2. In addition, a resistor R is loaded over the symmetric plane AA' of each resonator to adjust the odd-mode Q-factor but not disturbing the even-mode Q-factor.

The even-/odd-mode analysis is applied to determine the geometry and the value of R. Fig. 2(b) shows the simplified equivalent circuit of the proposed filter when only the odd modes of resonators are considered. The odd-mode of resonator is regarded as a $\lambda/2$ uniform-impedance resonator with its middle point shorted by the electric wall at AA'. The resistors should be positioned close to the AA' plane to reduce the influence of the parasitic capacitance of resistors on the odd modes, due to the large susceptance slope parameter of the resonator seen from the location of resistor. The simplified equivalent circuit when only the even modes of resonators considered is provided in Fig. 2(c). Here, the line widths W_2 and W_3 are properly chosen to satisfy 2 $Z_2 = Z_3$, where Z_2 and Z_3 are the characteristic impedances of the wide and narrow line sections of the stub loaded on the middle of resonators, respectively. The 180° out-of-phase I/O feedings are utilized. For the even-mode resonance, the electric field of upper half of each resonator is symmetric to that of the lower half, while asymmetric for the odd-mode resonance. Thus, the source and load external

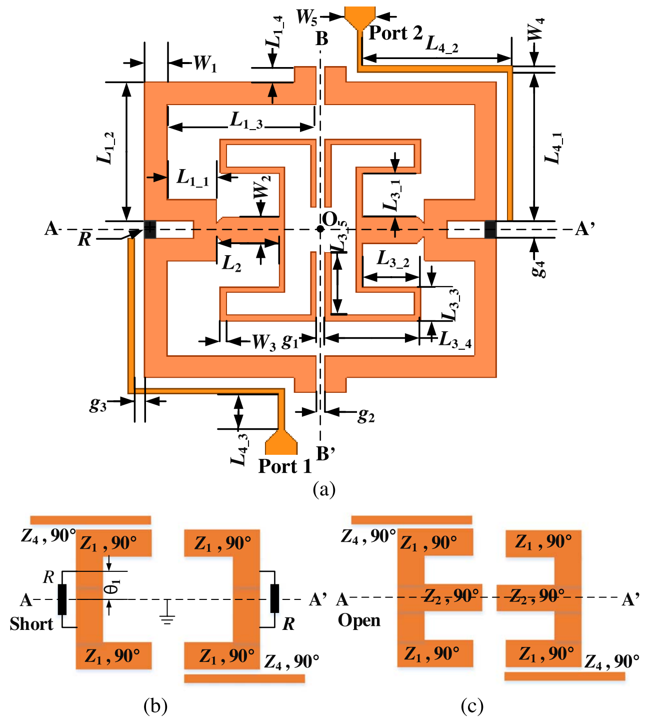

Fig. 2. (a) Layout of the proposed filter; (b) the simplified equivalent circuit when only the odd modes of resonators are considered; and (c) the simplified equivalent circuit when the even modes of resonators are considered.

coupling coefficients of two odd-mode nodes are opposite to each other, while the two external coupling coefficients of even modes have the same value.

The coupling matrix of the proposed dual-mode lossy filter is

$$\mathbf{M}=\begin{bmatrix} 0 & m_{S1} & 0 & m_{S2} & 0 & 0 \\ m_{S1} & -j\delta_{11} & -m_{12} & 0 & 0 & 0 \\ 0 & -m_{12} & -j\delta_{22} & 0 & 0 & m_{S1} \\ m_{S2} & 0 & 0 & -j\delta_{33} & -m_{34} & 0 \\ 0 & 0 & 0 & -m_{34} & -j\delta_{44} & -m_{S2} \\ 0 & 0 & m_{S1} & 0 & -m_{S2} & 0 \end{bmatrix}. \tag{1}$$

The elements of matrix except-$j\delta_{ii}$ ($i = 1$ to 4) can be obtained using the conventional synthesis method [2], according to design specifications. The values $\delta_{ii} = 1/(Q_{ui}\text{FBW})$ of the diagonal elements are resulted from the losses in the resonators, where FBW is the fractional equal-ripple bandwidth. Without loading resistors, it is almost a uniform-Q filter. When loaded with resistors, the resonant modes are grouped by Q-factors, i.e., the higher-Q even modes ($Q_{u1} = Q_{u2} = Q_{ue}$) and the lower-Q odd modes ($Q_{u3} = Q_{u4} = Q_{uo}$). In order to achieve a response with flat passband, the required ratio $k_Q = Q_{ue}/Q_{uo}$ of higher Q to lower Q should be assigned by the nonuniform-Q method in [5].

With a loading resistor, the modified Q-factor of the odd mode can be obtained by

$$\frac{1}{Q_{uo}} = \frac{1}{Q_{uo0}} + \frac{1}{Q_R} \tag{2}$$

where Q_{uo0} is the odd-mode Q-factor without loading resistor, approximately equal to Q_{ue}, and Q_R is the Q-factor due to the loss of resistor, given by

$$W_o = 2 \times \frac{1}{2}\int_0^{\frac{\pi}{2}} \frac{U_0^2 \sin^2\theta}{Z_1\omega_0} d\theta = \frac{U_0^2}{\omega_0 Z_1}\frac{\pi}{4} \tag{3}$$

$$Q_R = \frac{\omega_0 W_o}{\frac{(U_0 \sin\theta_1)^2}{\left(\frac{R}{2}\right)}} = \frac{\pi R}{8 Z_1 \sin^2\theta_1} \tag{4}$$

where W_o is the stored energy of odd mode, ω_0 is the resonant angular frequency, Z_1 is the characteristic impedance of the microstrip line with the width of W_1, and U_0 is the maximum voltage in the Z_1-sections, and the electrical length θ_1 at the central frequency corresponds to the position of the resistor, defined in Fig. 2(b). Then, we can derive from (2) and (4) that

$$R = \frac{8 Z_1 \sin^2\theta_1 Q_{ue}}{\pi(k_Q - 1)}. \tag{5}$$

The value of R will be further optimized to take the discontinuities of the filter into account.

The stored energy of the even mode can be calculated by

$$W_e = \frac{U_0^2}{\omega_0 Z_1}\frac{\pi}{4} + \frac{U_0^2 Z_3}{\omega_0 Z_1^2}\frac{\pi}{4}. \tag{6}$$

Since the even and odd modes share the same I/O feedings, the ratio of the corresponding external Q-factors (i.e., Q_{ee} and Q_{eo}) is calculated by

$$\frac{Q_{ee}}{Q_{eo}} = \frac{W_e}{W_o} = 1 + \frac{Z_3}{Z_1} = \frac{m_{S2}^2}{m_{S1}^2}. \tag{7}$$

Therefore, we have

$$Z_3 = \left(m_{S2}^2/m_{S1}^2 - 1\right) Z_1. \tag{8}$$

A filter prototype with a fourth-order generalized Chebyshev response is designed. The in-band return loss (RL) is 20 dB. A pair of transmission zeros are located at $\pm 2j$. The passband is centered at 2.00 GHz with an FBW of 5.36%. The substrate is a 40-mil-thick Taconic RF-35A2, with $\varepsilon_r = 3.5$ and $\tan\delta = 0.0015$, leading to $Q_{ue} = 180$. The coupling matrix is synthesized, where $m_{S1} = 0.525$, $m_{S2} = 0.879$, $m_{12} = 1.278$, and $m_{34} = 0.690$. The calculated value of $k_Q = 3.75$, leading to $Q_{uo} = 48$. Here, the characteristic impedance Z_3 is set to 112 Ω with $W_3 = 0.4$ mm. Then, we have $Z_2 = 56$ Ω with $W_2 = 1.87$ mm, $Z_1 = 62\,\Omega$ with $W_1 = 1.56$ mm from (8), and $R = 334\,\Omega$ from (5) when $\theta_1 = 10°$.

The line width W_4 for external coupling is not unique. The smaller the W_4 becomes the larger the gap width g_3 becomes when a fixed external coupling is required. Here, W_4 is set equal to W_3. The g_3 is then optimized using EM simulator, ANSYS HFSS, to meet the required value of m_{S1}. The gap width g_2 is determined by m_{34}, and then the width g_1 is determined by m_{12}, when the other geometry parameters are fixed. The coupling element m_{S2} can be realized simultaneously if (8) is satisfied. Full-wave simulations are performed for the final optimal dimensions, as listed in Table I. Note that the calculated values of W_1, W_2, W_3, and W_4 are unchanged during the optimization.

III. Results and Discussion

The lossy filter is fabricated. A chip resistor of $R = 330\ \Omega$ is mounted on each dual-mode microstrip resonator. Excluding

TABLE I
DIMENSIONS OF THE HYBRID NONUNIFORM-Q FILTER PROTOTYPE (UNIT: mm)

W_1	L_{1_1}	L_{1_2}	L_{1_3}	L_{1_4}	W_2	L_2	W_3
1.56	3.50	9.88	10.55	1.05	1.87	4.40	0.40
L_{3_1}	L_{3_2}	L_{3_3}	L_{3_4}	L_{3_5}	W_4	L_{4_1}	L_{4_2}
3.50	4.10	2.40	6.72	4.43	0.40	11.02	10.55
L_{4_3}	W_5	g_1	g_2	g_3	g_4		
2.93	2.20	0.62	0.56	0.74	1.20		

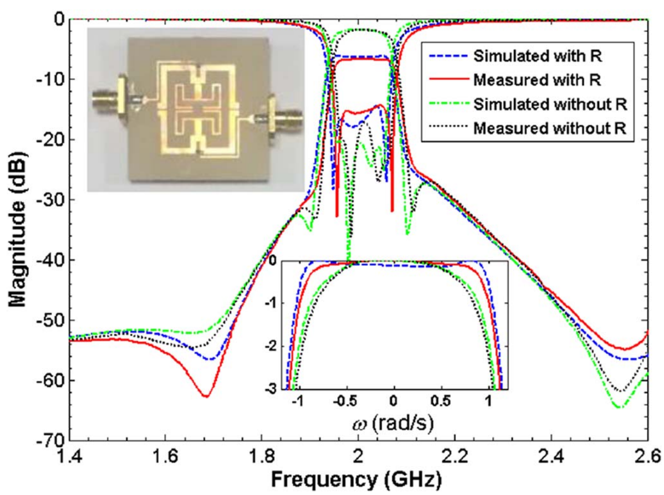

Fig. 3. Simulated and measured results of the proposed filter prototypes with and without loading resistors. One inset is the photo of prototype and the other is the enlarged view of normalized S_{21} in the normal frequency domain.

the feed lines, the area of the filter is 23.06×24.78 mm^2 ($0.26 \times 0.27\ \lambda_g^2$), where λ_g represents the guided wavelength of a 50-Ω microstrip line on the used substrate at 2.00 GHz.

The measured and simulated S-parameters are plotted in Fig. 3. The responses with and without R are both provided for comparison. Two insets are inserted into Fig. 3 to show the photo of the fabricated prototype and the enlarged view of the normalized S_{21}. Two transmission zeros are located at 1.91 and 2.10 GHz, as expected. There are two additional transmission zeros at 1.66 and 2.55 GHz, owing to the 180° out-of-phase I/O feedings [9]. The stopband rejection is improved. For the filter without R, the measured passband is centered at 2.01 GHz with a 0.2 dB bandwidth of 61 MHz, while its central frequency and 0.2 dB bandwidth are 2.00 GHz and 62 MHz in simulation, respectively. The measured minimum IL is 2.0 dB and the in-band RL is better than 17 dB. The additional 0.1 dB IL over the simulated result is mainly caused by the SMA connectors.

For the filter with R, the measured central frequency is 2.01 GHz with a 0.2 dB bandwidth of 99 MHz, while the simulated central frequency is 2.00 GHz with a 0.2 dB bandwidth of 111 MHz. The minimum IL is 6.7 dB and the in-band RL is better than 14 dB in measurement. The 0.2 dB bandwidth of the filter with R is significantly broadened by 62% when compared with the case without R, equivalent to a uniform-Q of 970, at the cost of additional IL of 4.7 dB. Compared with the simulated response, the measured one has a slight frequency shift of 10 MHz, an extra IL of 0.4 dB, and the shrunk 0.2 dB bandwidth by 12 MHz, mainly due to the tolerances of the fabrication, especially the small gap width between two resonators.

Table II compares the sizes and performance of the reported microstrip filters designed using the nonuniform-Q method [8].

TABLE II
COMPARISON BETWEEN THE REPORTED MICROSTRIP FILTERS DESIGNED USING THE NONUNIFORM-Q METHOD

	f_0 GHz	FBW %	Q:EQ	IL dB	RL dB	RJ dB	Size λ_g^2
[4]	0.96	5.7	250/80:750	7.2	11	46	0.65×0.79
[6]	2.48	8	280/84:540*	1.5	13	18	0.42×0.39
[7]	2.03	8	130/44:650	3.4	10	22	0.19×0.31
[8]	3.80	21	95/57/35:240	2.2	18	4	0.57×0.89
ours	2.01	5.4	180/48:970	6.7	14	32	0.26×0.27

EQ: Equivalent Q-factor; **RJ**: Out-of-band rejection, defined within the range from 2FBW*f_0 to 6FBW*f_0 away from the central frequencies (f_0).
* The equivalent Q-factor of [6] is estimated with the synthesized results.

The minimum IL is mainly determined by the lower Q, FBW and filter order. The IL of this filter is higher than the smallest one in [6] due to the higher order and lower Q's. By using the analytically derived k_Q and properly positioning resistors, our design shows the highest equivalent Q-factor in measurement, evaluated for passband flatness. The out-of-band rejection of our work are better than the others except for [4], due to the fourth-order response with four transmission zeros. Although the sixth-order single-mode filter in [4] shows a high selectivity, its relative area is about seven times that of our design with two dual-mode microstrip resonators. The occupied area of this work is only 20% larger than that of the third-order triple-mode filter in [7], but the out-of-band rejection is enhanced by 10 dB.

IV. CONCLUSION

A planar miniaturized flat-passband filter is designed with the nonuniform-Q method and realized with two dual-mode microstrip resonators. By loading resistors on the resonators properly, the odd-mode Q-factor can be controlled, with the even-mode one unchanged. The formulas are derived for the critical dimensions and resistance. The passband flatness is significantly improved with the equivalent Q of 970. The increased IL can be compensated by low-noise pre-amplifier.

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