Dual-Wideband Bandpass Filter Using Short-Circuited Stepped-Impedance Resonators

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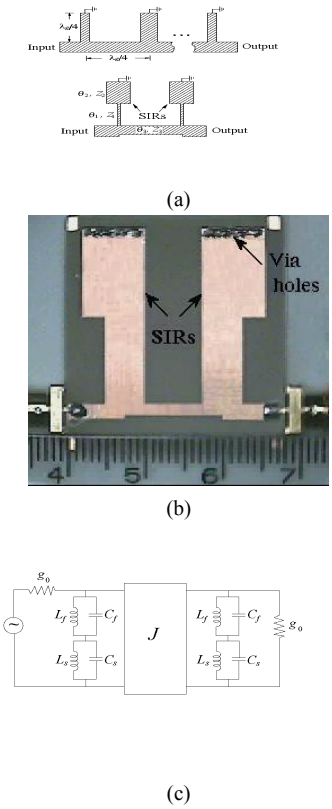
[1] ***Abstract*—A compact dual-wideband bandpass filter featuring two controllable passbands is presented and experimentally validated. The proposed filter uses shorted stepped-impedance resonators to achieve dual-wideband bandpass capability and to minimum size. Design curves are given for the filter design. It is found that the frequency ratio of the second to first band has a wide achievable range. An experimental circuit is fabricated and evaluated to validate the design concept. The measurement results correlated well with the simulation results.**

***Index Terms*—Wideband filter, dual-band filter, stepped-impedance resonator, filter synthesis.**

I. Introduction

CONTINUOUS advances in wireless communication require RF-integrated multiple-band transceivers to enable user access to various services. For instance, the global system for mobile communications (GSM) requires an RF transceiver for receiving and transmitting signals at both 900 and 1800 MHz. Similarly, wireless local area networks (WLANs) have standards such as IEEE 802.11b (2.4 GHz) and IEEE 802.11a (5.2 GHz). Dual-band bandpass filters are being developed for these applications to reduce circuit size and cost [1]-[10]. Applying a two-step frequency-variable transformation to the well-known low-pass prototype can achieve dual passband capability [1]. In [2], a pair of coupled semi-lumped quarter-wave resonators was employed as the basic unit of a dual-band filter. Chen *et al.* [3] integrated two stacked loops into a module for dual-mode dual-band operation. In [4], [5], the synthesis method of using series and parallel stubs as resonators to comply with dual-band performance requirements was proposed. A novel equivalent circuit using lumped-element composite series resonators to obtain dual passbands was presented in [6]. Kuo *et al.* used a rigorous method to design dual-band bandpass filters in parallel-coupled and vertical-stacked configurations with stepped-impedance resonators (SIRs) [7]. Other approaches have also used the spurious frequencies of SIRs to create the second passband [8]-[10]. Most of the aforementioned filters are basically appropriate for the design of narrow band (e.g., less than 10%), and only provide a very limited range of frequency ratio of the two passbands. However, a continuing challenging is designing dual-band filters with wide bandwidth, arbitrary frequency ratio, and a compact size.

This work presents a novel dual-wideband bandpass filter designed to control center frequencies and bandwidths using short-circuited SIRs. Experimental results indicate that the proposed structure provides very wide bandwidth characteristics. The widest measured bandwidth is 54% and the achievable frequency ratio is 1.19-2.9.



(a)

(b)

(c)

Fig. 1 (a) Conventional single-band bandpass filter using quarter-wave short-circuited stubs. (b) Configuration of the proposed dual-wideband bandpass filter with shorted SIRs (order N = 2). (c) Equivalent circuit using composite parallel LC resonators and a J inverter.

II. Design Procedure

According to Fig. 1(a), quarter-wave short-circuited stubs are often used to implement single-band bandpass filters. These quarter-wave stubs have an electrical length of 90° at only a single frequency because their impedance is uniform. However, dual-band performance can be achieved by applying shorted stepped-impedance resonators in filter design. The stepped-impedance resonator can simultaneously provide an

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equivalent electrical length of 90° at two frequencies because of its nonuniform impedance. Figure 1(b) shows the proposed configuration of the dual-wideband bandpass filter with order $N = 2$. Figure 1(c) illustrates a circuit equivalent to that in Fig. 1(b) while the series-connected parallel LC resonator structure is constructed to perform the two resonance frequencies. When the resonators L_fC_f resonate at ω_f (the first center frequency) and L_sC_s resonate at ω_s (the second center frequency), their input impedances approach infinity. Thus, composite LC resonators function like open circuits at ω_f and ω_s, thereby allowing the signal to pass through and achieve the dual-passband response.

With reference to Fig. 2, the input admittances of the shorted SIR and its equivalent circuit are given by

$$Y_{SIR} = j\frac{Z_2\tan\theta_1\tan\theta_2 - Z_1}{Z_1(Z_1\tan\theta_1 + Z_2\tan\theta_2)} \tag{1a}$$

$$Y_c = \frac{jC_fC_s(\omega^2-\omega_f^2)(\omega^2-\omega_s^2)}{\omega[C_s(\omega^2-\omega_s^2)+C_f(\omega^2-\omega_f^2)]} \tag{1b}$$

where $C_f = g_i/\omega_f\Delta_f$, $C_s = g_i/\omega_s\Delta_s$, $\Delta_{f,\ s}$ are the fractional bandwidths of the first and second passband, respectively, and g_i is the low-pass prototype element value. Notably, the resonant frequencies of ω_f and ω_s can be determined by using the resonant condition $Y_{SIR} = 0$. Further, the susceptance slope parameter [11] can be used to ensure that the proposed SIR and its equivalent circuit have the same bandwidth. Based on the susceptance slope and the resonant condition at ω_f and ω_s, four simultaneous equations are formed

$$\frac{Z_2\tan\theta_1\tan\theta_2 - Z_1}{Z_1(Z_1\tan\theta_1 + Z_2\tan\theta_2)} = 0 \tag{2a}$$

$$\frac{Z_2\tan(r_f\theta_1)\tan(r_f\theta_2) - Z_1}{Z_1[Z_1\tan(r_f\theta_1) + Z_2\tan(r_f\theta_2)]} = 0 \tag{2b}$$

$$\frac{Z_1^2\theta_1\cos^2\theta_2 + Z_2^2\theta_1\sin^2\theta_2 + Z_1Z_2\theta_2}{2Z_1(Z_1\sin\theta_1\cos\theta_2 + Z_2\cos\theta_1\sin\theta_2)^2} = \frac{g_i}{\Delta_f} \tag{2c}$$

$$\frac{r_f[Z_1^2\theta_1\cos^2(r_f\theta_2) + Z_2^2\theta_1\sin^2(r_f\theta_2) + Z_1Z_2\theta_2]}{2Z_1[Z_1\sin(r_f\theta_1)\cos(r_f\theta_2) + Z_2\cos(r_f\theta_1)\sin(r_f\theta_2)]^2} = \frac{g_i}{\Delta_s} \tag{2d}$$

where $r_f = \omega_s/\omega_f$ denotes the frequency ratio. Notably, θ_1 and θ_2 in (2) are defined at ω_f. The center frequencies and bandwidths of the proposed filter are controllable via impedance changes and SIR length as shown in (2). By eliminating Z_1 and Z_2, the four equations (2a)-(2d) are further simplified to two equations

$$r_fr_b[\theta_2\cos^2(r_f\theta_1)\csc^2(r_f\theta_2) + \theta_1\cot(r_f\theta_1)\cot(r_f\theta_2)] = \theta_2\cos^2\theta_1\csc^2\theta_2 + \theta_1\cot\theta_1\cot\theta_2 \tag{3a}$$

$$\tan\theta_1\tan\theta_2 = \tan(r_f\theta_1)\tan(r_f\theta_2) \tag{3b}$$

where r_b is the bandwidth ratio Δ_s to Δ_f. Solving for θ_1 and θ_2 simultaneously by using a simple root-searching program and substituting into (2), one can obtain the values of Z_1 and Z_2.

The calculated results in Fig. 3(a) show the electrical lengths θ_1 and θ_2 versus the frequency ratio r_f. The range of r_f is chosen from 1.1 to 2.9, which covers most of the design objectives. In Fig. 3(a), there are three cases in which product $r_f \cdot r_b$ = 0.8, 1, and 1.2. It is found that $\theta_1 > \theta_2$ for $r_f \cdot r_b < 1$ and $\theta_1 = \theta_2$ for $r_f \cdot r_b$

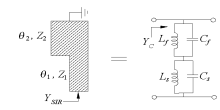
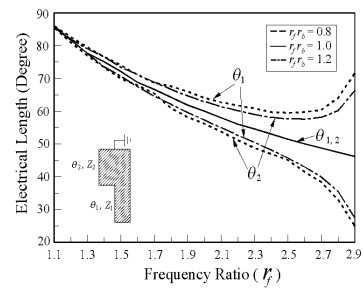


Fig. 2 Equivalence of shorted SIR and series-connected parallel LC resonators.



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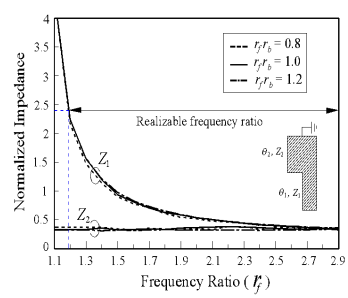


(b)

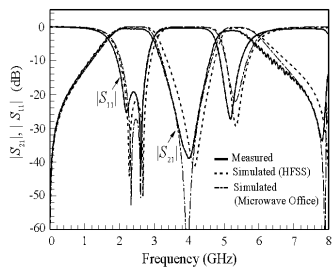
Fig. 3 (a) Electrical lengths θ_1 and θ_2 of shorted SIR versus r_f. (b) Normalized impedances Z_1 and Z_2 versus r_f.

= 1. When $r_f \cdot r_b > 1$, the design curves of θ_1 and θ_2 become reversed ($\theta_1 < \theta_2$) to the solid curve $r_f \cdot r_b = 1$. Figure 3(b) plots the normalized impedances Z_1 and Z_2 versus r_f with the element value of g_1 = 1.4142 (second-order maximally flat low-pass prototype) and Δ_f = 60%. The value of Z_1 increases exponentially as r_f decreases while Z_2 remains almost constant. Neither Z_1 or Z_2 are sensitive to the product $r_f \cdot r_b$. Since Z_1 always exceeds Z_2, the upper section of the proposed SIR should have a wider line width than the lower section to

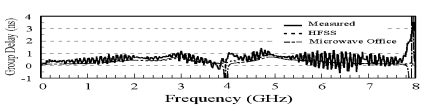
achieve dual-passband performance. When r_f approaches three, Fig. 3 reveals the tendency of $Z_1 \approx Z_2 \approx 0.33$ and $\theta_1 + \theta_2 \approx 90^o$ in which the proposed SIR becomes the conventional $\lambda_g/4$ uniform stub.

Due to the highest practical limit of impedance realization (120 Ω typically), as shown in Fig. 3(b), the frequency ratio is realizable for nearly the entire range except $r_f < 1.19$. This wide achievable range is very helpful to construct dual-band filters with arbitrary frequency ratio. Notably, it is also found that the solutions of Z_1 and Z_2 don't exist or become impractical if $\Delta_{f,\ s}$ are less than 30%. The reason is that the ratio of Z_1/Z_2 is reduced to one as the bandwidth reduces, which conflicts with dual band requirement (requires nonuniform impedance). Thus, the proposed structure is particularly appropriate for wide passband applications.

The admittance inverter between SIRs can be calculated by using $J_{i,i+1} = g_0 g_1 / \sqrt{g_i g_{i+1}}$, and the simplified dual-band structure presented in [12] is adopted to realize the J inverter for the proposed filter.



(a)



(b)

Fig. 4 (a) $|S_{21}|$ and $|S_{11}|$ responses of the proposed dual-wideband bandpass filter. (b) Group delay. $Z_1 = 23.35\ \Omega$, $\theta_1 = 55.63°$, $Z_2 = 16\ \Omega$, $\theta_2 = 57.87°$, $Z_3 = 59.73\ \Omega$, and $\theta_3 = 56.84°$.

III. Simulation And Measurement results

Validity of the proposed structure was tested experimentally by constructing a second-order maximally flat dual-wideband bandpass filter on 25N substrates with $\varepsilon_r = 3.38$, $\tan\delta = 0.0025$, and a thickness of 0.762 mm. The central frequencies of the dual passbands were designed at 2.4 GHz and 5.2 GHz with $\Delta_f = 60\%$ and $\Delta_s = 30\%$, respectively, to match the WLAN specifications. Solving (2) and (3) with $g_1 = 1.4142$, $r_f = 2.17$, and $r_b = 0.5$ revealed that the design parameters for SIR are $Z_1 = 23.35\ \Omega$, $\theta_1 = 55.63°$, $Z_2 = 16\ \Omega$, and $\theta_2 = 57.87°$. The design parameters of the J inverter are $Z_3 = 59.73\ \Omega$ and $\theta_3 = 56.84°$. The calculated circuit dimensions and simulated responses of the experimental filter were then obtained by using the full-wave EM simulator HFSS. Simulated results from the circuit simulator Microwave Office using the chosen ideal transmission line parameters were also plotted for comparison.

Figure 4(a) shows the simulated and measured S-parameters. The dual-band bandpass response was obtained, and there is good agreement between simulated and measured responses. Measured central frequencies were 2.3 and 5.25 GHz with bandwidths of 54% and 20%, respectively. The bandwidths were decreased mainly because the open stubs of the dual-band inverter were ignored. Notably, the bandwidths are wider than those of dual-band bandpass filters presented in [2], [4], [5], [8], [10]. In-band return loss and insertion loss of this filter were better than 20 dB and 0.8 dB, respectively. As Fig. 4(b) shows, the measured group delay in the passband was small and ranged from 0.5 to 1 ns. Figure 1(b) depicts a photograph of the fabricated filter, and the circuit area was about 24 × 24 mm^2. In our further studies, the dual bandwidths of the proposed filter can up to 60% and 67% centered at 3.7 GHz and 9 GHz, respectively, which is not shown here for conciseness.

IV. Conclusions

This work presents a novel dual-wideband bandpass filter using short-circuited stepped-impedance resonators. The circuit possesses two wide passbands with controllable center frequencies and bandwidths. Both dual-band performance and compact size are achieved by using dual-band SIRs instead of conventional single-band quarter-wavelength shorted stubs. The synthesis method was developed and its effectiveness was demonstrated by synthesizing an experimental filter.

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