Design and Implementation of Microwave Dual-Band Bandpass Filters Using Microstrip Composite Resonators

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***Abstract* — In this paper, a novel type of dual-band bandpass filter (DBBPF) is proposed by using microstrip composite resonators, and its design method is developed based on the lumped-element equivalent circuit of the filter. Both the midband frequencies and the passband widths of the dual bands of the filter can be controlled separately. Four DBBPFs with different specifications are designed, fabricated and tested, and the measured frequency responses agree favorably with theoretical predictions, validating sufficiently the proposed design theory and filter structures.**

***Key words* — Bandpass filter, dual-band, composite resonators, microstrip line.**

I. Introduction

In recent years, the rapid growth of various wireless communication services has results in strong demands for communication systems and devices capable of operating at multi-mode and multi-bands [1]-[2]. Many reports on dual-band RF and microwave filters [3]-[9] are now available. In [3], a dual-band filter was accomplished by connecting in parallel two bandpass filters (BPFs) with different passbands. In [4], a wideband BPF was cascaded with a narrow-band bandstop filter to get a dual-passband response. As the circuits in both [3] and [4] consisted of two different filters, their configurations were comparatively large. Other papers, like [5] used dual-behavior resonators, and [6] and [7] utilized the dominant and first higher order modes of stepped impedance resonators (SIRs) to realize DBBPFs with separately controllable dual-band center frequencies. Some more recent papers, including [8]-[10] reported dual-band BPFs consisting of dual-band inverters to obtain controllable dual-band response, but no explicit design formulas relating the filter specifications and circuit parameters were provided to facilitate the design of these filters.

In this paper, a novel type of dual-band bandpass filter is proposed by using lumped-element composite series resonators, and a design method is developed based on the circuit analysis. Both the midband frequencies and the passband widths of the dual bands of the filter can be controlled separately. Moreover, circuit transformation is performed to allow easy realization of the filter by using microstrip or other planar transmission lines. Four DBBPFs with different specifications are designed, fabricated and tested, and their measured frequency responses agree favorably with theoretical predictions, validating solidly the proposed design theory and filter structures.

II. Synthesis Theory

Fig. 1 shows an n-degree BPF using conventional shunt LC resonators and J-inverters. The parameters, including the inductor L_{ri}, capacitor C_{ri}, and admittance-inverter $J_{i,i+1}$ are related by the following well-known formulas [11].

$$L_{ri} = \frac{1}{\omega_0^2 C_{ri}} \quad (i=1 \text{ to } n),\ \ \Omega_c=1 \text{ (rad/sec)} \tag{1}$$

$$J_{0,1} = \sqrt{\frac{FBW G_0 \omega_0 C_{r1}}{\Omega_c g_0 g_1}},\ \ J_{n,n+1} = \sqrt{\frac{FBW \omega_0 C_{rn} G_{n+1}}{\Omega_c g_n g_{n+1}}} \tag{2}$$

$$J_{i,i+1} = \frac{FBW \omega_0}{\Omega_c} \sqrt{\frac{C_{ri} C_{ri+1}}{g_i g_{i+1}}} \quad (i=1 \text{ to } n\text{-}1) \tag{3}$$

where ω_0 is the midband angular frequency of the BPF, FBW is the fractional bandwidth, and g_i (i=0,1, …, n+1) is the element values of a Butterworth or Chebyshev prototype lowpass filter.

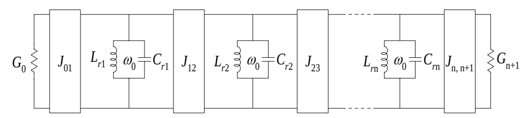
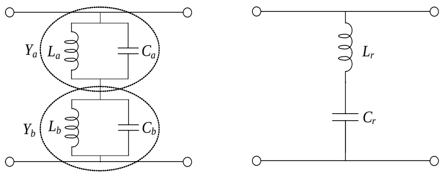


Fig. 1. Conventional BPF with J-inverters and shunt LC resonators

Fig.2(a) shows a composite series resonator. It is a cascade of two shunt LC resonators. These two shunt resonators have susceptance Y_a and Y_b, and resonate at ω_a and ω_b, respectively. Assume $\omega_a < \omega_b$, then when $\omega_a < \omega < \omega_b$, $Y_a>0$ is capacitive, and $Y_b<0$ is inductive. As a consequence, the composite resonator in Fig. 2(a) can be represented by a series LC resonator shown in Fig. 2(b), which resonates at ω_0.

Fig. 3 shows a typical frequency response of the composite series resonator. Two passbands around ω_a and ω_b are observed, and they are formed by the resonances of the two LC shunt resonators shown in Fig. 2(a). Meanwhile a transmission zero is produced at ω_0, which is caused by the series resonance shown in Fig. 2(b). Following the analysis procedure described in [12] for a composite shunt resonator, we get:

$$\omega_a = \frac{1}{\sqrt{L_a C_a}}, \quad \omega_b = \frac{1}{\sqrt{L_b C_b}}, \quad \omega_0 = \frac{1}{\sqrt{(L_a + L_b)\dfrac{C_a C_b}{C_a + C_b}}} \tag{4}$$



(a) (b)

Fig. 2. (a) A composite series resonator, and (b) its equivalent LC series resonator.

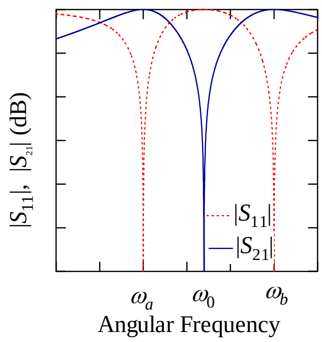


Fig.3. Typical frequency response of a composite series resonator.

As stated above, a composite series resonator exhibits dual passbands around its two resonance frequencies. We use this property of a composite series resonator to build a novel equivalent circuit of dual-band BPF. We replace all the shunt LC resonators in Fig. 1 with the composite series resonator shown in Fig. 2(a), and get a new circuit as shown in Fig. 4. Let all the $L_{ai}C_{ai}$ (i=1,2,…,n) shunt resonators in Fig. 4 resonate at ω_a. Then a passband (namely passband 1) around ω_a is formed by these resonators. Similarly, let all the $L_{bi}C_{bi}$ (i=1,2…,n) shunt resonators resonate at ω_b, then passband 2 around ω_b is formed. Therefore, the circuit in Figure 4 is a filter possessing dual passbands. The midband frequencies, ω_a and ω_b, as well as the passband widths, FBW_a and FBW_b, of the dual bands are determined separately by appropriately choosing L_{ai}, C_{ai}, L_{bi}, and C_{bi} (i=1,2…,n) using (1)-(3). It is important to notice that, as we know, in the conventional BPF, the values of the J-inverters are assumed unvaried with frequency. In our design of the DBBPF, we also assume that the J-inverters are unvaried over the dual passbands. To end this purpose, we can determine $C_{a,i}$ and $C_{b,i}$ (i=1,2…,n) in Fig. 4 by using the following relation:

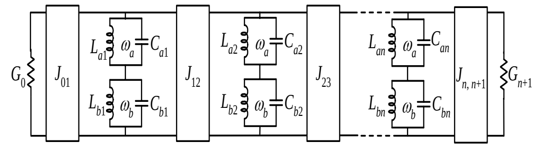


Fig. 4. Equivalent circuit of a DBBPF using composite series resonators.

$$C_{a1} = \frac{\Omega_c g_0 g_1 J_{0,1}{}^2}{FBW_a G_0 \omega_a}, \quad C_{an} = \frac{\Omega_c g_n g_{n+1} J_{n,n+1}{}^2}{FBW_a G_{n+1} \omega_a} \tag{5}$$

$$C_{b1} = \frac{\Omega_c g_0 g_1 J_{0,1}{}^2}{FBW_b G_0 \omega_b}, \quad C_{bn} = \frac{\Omega_c g_n g_{n+1} J_{n,n+1}{}^2}{FBW_b G_{n+1} \omega_b} \tag{6}$$

$$C_{ai+1} = \frac{\Omega_c^2 J_{i,i+1}{}^2 g_i g_{i+1}}{FBW_a^2 \omega_a^2 C_{ai}}, \quad C_{bi+1} = \frac{\Omega_c^2 J_{i,i+1}{}^2 g_i g_{i+1}}{FBW_b^2 \omega_b^2 C_{bi}} \tag{7}$$

We find that the composite series resonator shown in Fig. 2(a) is difficult to realize by using microstrip or other planar transmission lines. So we transform the composite series resonator into a composite shunt resonator by introducing a J-inverter, as shown in Fig. 5. The composite shunt resonator consists of two series LC resonators. The element values after the circuit transformation is as follows:

$$C_a' = L_a J^2, L_a' = \frac{C_a}{J^2}, \; C_b' = L_b J^2, \; L_b' = \frac{C_b}{J^2} \tag{8}$$

Here, J is the characteristic admittance of the J-inverter introduced in the transformation. It can take theoretically arbitrary values. But, in practical design of filters, if the J-inverter is o be realized by using a piece of transmission line, then the value of J is limited by the impedance of the transmission line possible for easy fabrication.

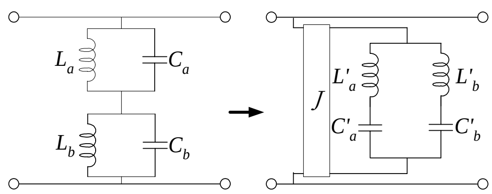


Fig. 5. Transforming a composite series resonator into a composite shunt resonator by introducing a J-inverter.

After all the composite series resonators in Fig. 4 are transformed by introducing J-inverts, as is shown in Fig. 5, the filter shown in Fig. 4 is then transformed to the circuit shown in Fig. 6. As will be explained below, both the LC series resonators and the J-inverters in Fig. 6 can be readily realized by using microstrip open stubs.

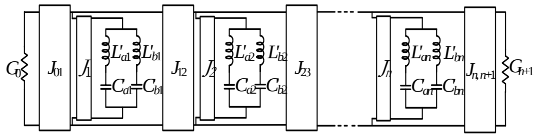


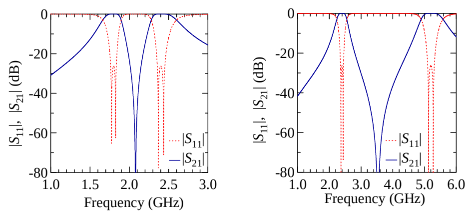
Fig. 6. Equivalent circuit of a DBBPF using *LC* series resonators and *J*-inverters.

III. Design Examples

To verify the above-described design theory, four DBBPFs with different specification, as shown in Table 1, are designed. The element values in the lumped-element circuit in Fig. 4 or Fig. 6 are obtained by using formulas described above, and the frequency responses of the filters are computed by using circuit-based simulators, like the well-know ADS of Agilent Co., or Designer of Ansoft Co.. Because of the space limit, only the response of filter A and D are given in Fig. 7. It is seen that the design specifications, including the bandwidths and center frequencies of the dual-bands of both filters, are satisfied well.

TABLE I Filter Specifications

Filter	A	B	C	D
FBW_a (%)	4	2	4	2
FBW_b (%)	4	4	4	4
f_a (GHz)	1.8	1.8	2.4	1.8
f_b (GHz)	2.4	2.4	5.2	2.4
degree	2	2	2	3



(a) (b)

Fig.7. Frequency response of (a) filter A, and (b) filter C.

IV. Compact Microstrip DBBPFs

The DBBPF shown Fig. 6 consists of *LC* series resonators and *J*-inverters only, so it can be easily accomplished by using microstrip or other planar transmission lines. As described in detail in [12], both an *LC* series resonator and a *J*-inverter can be realized by using quarter-wavelength open stubs. Design formulas relating the circuit elements with the dimensions of the microstrip open stubs were derived in [12], and are omitted here because of the space limit. A substrate with a dielectric constant of 11.3, a thickness of 0.635mm, and a loss tangent of 0.003 is used. The dimensions of the filter are obtained by using an electromagnetic simulator, Sonnet em [13].

Fig. 8 illustrates the configuration and dimensions of filter A designed above in Section III, and Fig. 9 shows a comparison of its frequency responses. The solid lines in Fig. 9 are simulated results of the microstrip filter shown in Fig. 8 by using Sonnet em. Lossless substrate and conductors are assumed in the simulation. On the other hand, the broken lines are computed from the equivalent circuit shown in Fig. 4. The agreement between the solid and broken lines is good at frequencies around the passbands. At frequencies in the stopband, the transmission zeros associated with the distributed microstrip filter results in sharp and large attenuations.

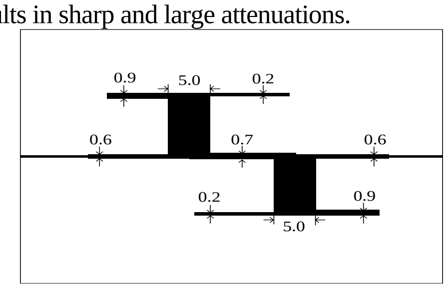


Fig. 8. Configuration and dimensions of the 2-degree filter A.

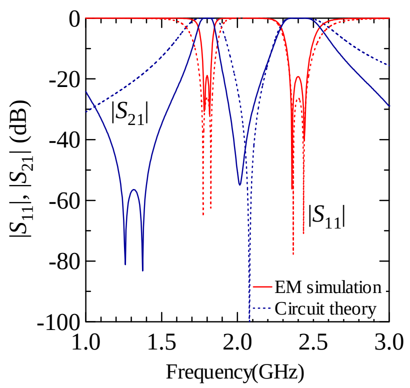


Fig. 9. Comparison of the frequency response of filter A.

The microstrip filter is fabricated and then measured using a network analyzer HP8510C. Fig. 10 is a photograph of the fabricated filter A. In Fig. 11, the measured data are drawn in solid lines, and are compared with the broken lines simulated by Sonnet em when taking into account of the dielectric and copper (σ=58x10^6 S/m) losses. Over both the passband and stopbands, an excellent agreement is observed between the solid and broken lines.

Fig. 12 is a comparison of the measured and simulated response of filter C. Again, a good agreement is observed. Because of the space limit, results for filter B and D are omitted here, but similar conclusions are obtained.

V. Conclusion

A novel class of DBBPF is proposed by using composite resonators, and its design method is developed based on the lumped-element equivalent circuit of the filter. Four DBBPFs with different specifications are designed and fabricated in microstrip form. The measured frequency responses agree favorably with theoretical predictions. The theory and design examples prove that the dual-band responses of the filter can be controlled separately.

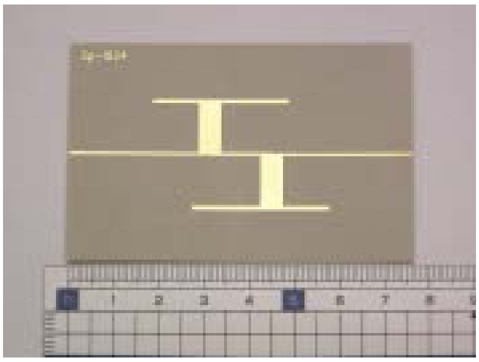


Fig. 10. Photograph of the fabricated filter A.

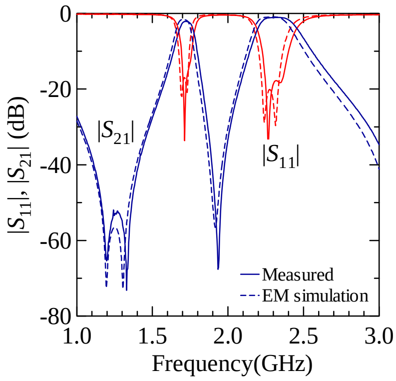


Fig. 11. Comparison of the measured and simulated frequency responses of filter A.

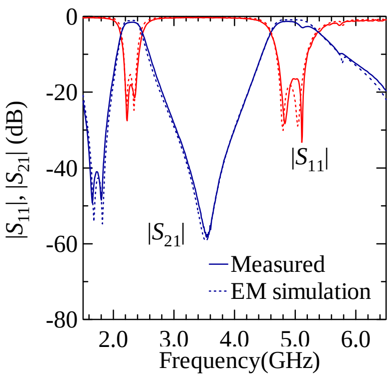


Fig. 12. Comparison of the measured and simulated frequency responses of filter C.

ACKNOWLEDGEMENT

This paper is supported in part by the Grant-in-Aids for Scientific Research (KAKENHI 17560303) from the Japan Society for the Promotion of Science, in part by the High-Tech Research Center Project from the Ministry of Education, Culture, Sports, Science and Technology, Japan, and in part by the Murata Science Foundation.

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