DESIGN OF MICROSTRIP LINE FOR DUAL BAND PASS FILTER

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ABSTRACT: Dual-band bandpass filters with controllable fractional bandwidths (FBWs) are constructed by cascading the multiple λ/2 stepped-impedance resonators (SIRs) through the distributed parallel-coupled microstrip lines (PCMLs). By suitably choosing the aspect ratio of two strip widths or impedances in the SIR, the first two resonant frequencies are allocated to 2.4 and 5.2 GHz for dual-band filter application. The prototype of the bandpass filter achieved insertion loss of 1.25 and 1.87 dB, 11 of 29 and 40 dB, and bandwidth of 21% and 12.7% at 2.4 and 5.2 GHz, respectively.

I. INTRODUCTION

Multiband devices, such as multiband antenna, multiband filter and multiband low-noise amplifier, have been recently receiving a tremendously increasing application in exploring many advanced wireless systems with simultaneous operations at multiple frequency bands [1]. As pointed out [2], multiband passive circuits basically determine the multiband operation quality, overall size and fabrication cost of a RF and wireless module using various integration technologies. Of them, bandpass filters with multiple passbands are considered as one of the most key components in these multiband systems. Due, to the shortage of matured design procedure, it becomes the most challenging issue for one to design the multiband filters with good passband performances, including a dual-band case.

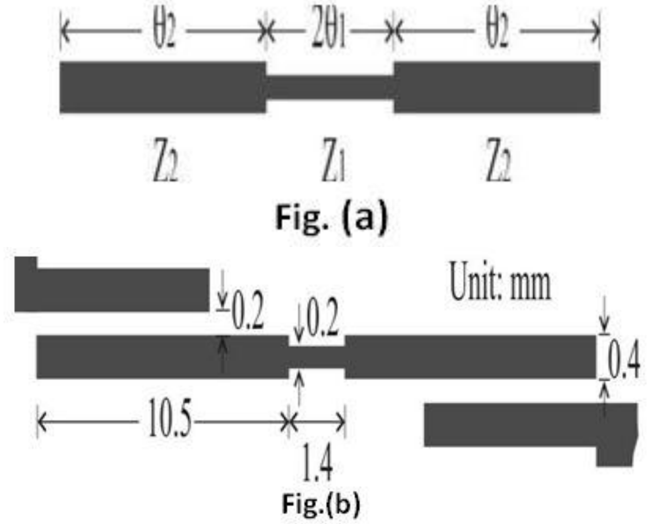


Fig. 1 Geometrical diagrams of the stepped-impedance resonator (SIR) and its constituted dual-band bandpass filter. (a) SIR (RZ < 1); (b) one-stage SIR bandpass filter (coupling length:LC1).

DUE to the explosive growth of various wireless communication services, today's microwave communication systems often require multi-band operations. In such systems, it is essential to have simple and low-cost, yet high-performance multi-band filters such as those in [1]–[3]. This letter presents simple yet effective planar dual-band filters. Since the structures are based on the well-known parallel coupled line filters, a complete set of design equations is derived based on the standard filter synthesis technique that can be applied to an arbitrary number of stages. Also, the proposed filter is perfectly compatible with planar circuitry and well suited for conventional 50- systems. Experimental results for a 0.1-dB ripple Chebyshev-type dual-band filter with passband center frequencies at 2.4 GHz and 5.2 GHz are in excellent agreement with the simulated results.

II. RESONATOR PROPERTIES OF SIR and PCML

A. Coupling Dispersion of PCML

Fig. 2(a) depicts the geometry of the PCML with the overlapped coupling length of LC1. Its equivalent circuit may be expressed as a J-inverter susceptance (J) and the two equal electrical lengths (θ / 2) with characteristic admittanceY0 , as shown in Fig. 2(b). As the characteristic impedances (Z0e, Z0o) and phase constants (β e, βo) of the even and odd dominant modes in the uniform PCML are modeled, the two-port impedance matrix can be deduced with the four elements,

Under the network equivalence, the two J-inverter network parameters can be obtained in terms of the two independent normalized susceptances,

where B11 = B11 / Y0 , B12 = B12 / Y0 , n is an integer number.

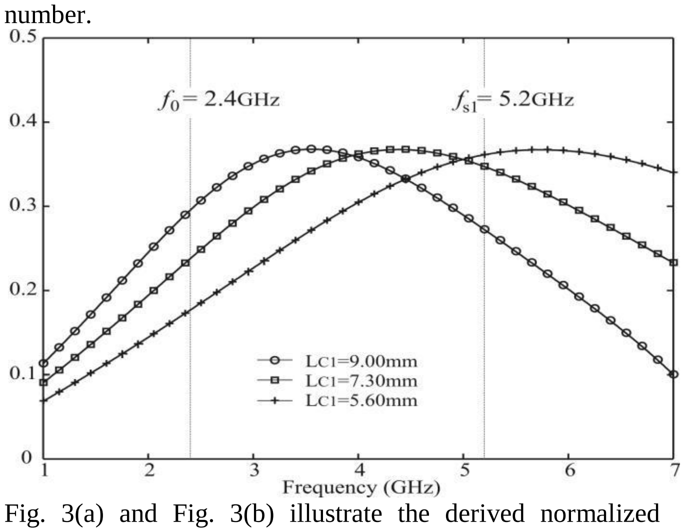
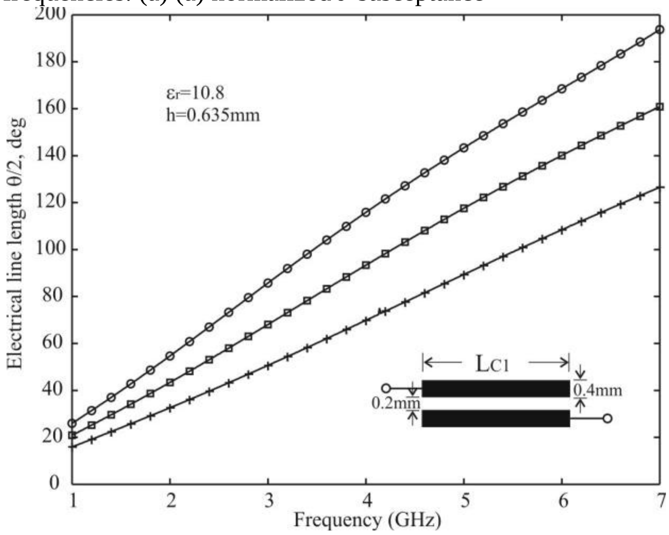


Fig. 3(a) and Fig. 3(b) illustrate the derived normalized Jinverter susceptance (J) and equivalent electrical length (θ / 2) under different lengths (LC1). The parameter J is seen to increase, reach to the maxima and then decrease as a nonmonotonic function of the frequency over 1.0 to 7.0 GHz for all the three cases. Further, as the length (LC1) is stretched, the band and stronger at the second operating band. In other words, the FBWs of both bands may be freely tuned by adjusting the coupling length LC1. Fig. 3(b) shows the quasilinearly increased electrical lengths versus frequency in relation to determinant of the dual resonant frequencies. (a) (a) normalized J-susceptance



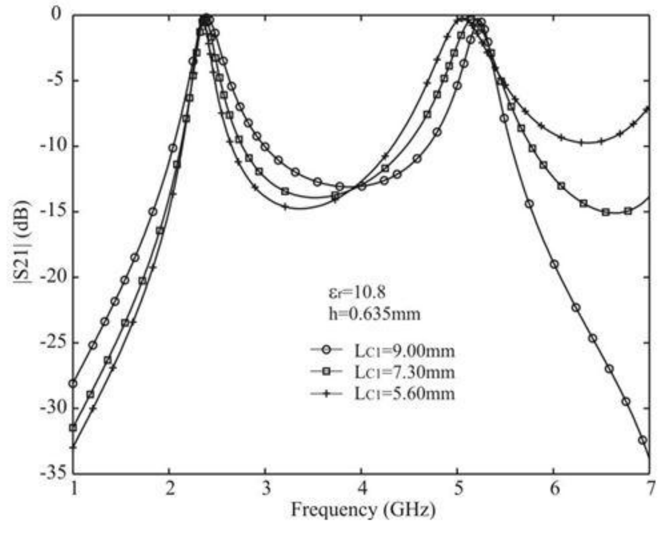
(b)

Fig. 2 Parallel-coupled microstrip line (PCML) to be characterized. geometry; (b) equivalent J-inverter network. normalized J-susceptance electrical lengths

Fig. 3 Extracted J-inverter network parameters of two-port PCML.

B. PCML-Excited Dual-band SIR

Using the above-modeled PCML section, a dual-band SIR circuit is constructed as shown in Fig. 1(b). Its S21-magnitude under the three different coupling lengths is calculated by the Momentum simulator [11] and plotted in Fig. 4 to initially demonstrate the controllable dual-band pass bandwidths at 2.4 and 5.2 GHz. In this aspect, the FBW observes to be widened from 6.8%, 8.2% to 12% at 2.4 GHz-band and narrowed from 10.5%, 8.2% to 5.2% at 5.2 GHz-band, respectively. Of importance, these results illustrate that the bandwidths of the dual passbands can be freely adjusted to well realize the three distinctive dual-band cases, which have not yet been reported so far.



III. DUAL BAND REDUCED COUPLED LINES

the proposed dual-band reduced-length parallel coupled line. For dual-band operations, the proposed coupled line must have the same electrical performance at the two frequencies f1 & mf1, and , as that of the conventional parallel coupled line in Fig. 1(b) at . The conditions for such can be obtained by analyzing its even- and odd-mode equivalent circuits. The conditions are where k is an integer such that ≤ ≤ . These conditions guarantee the exact same performance at the two frequencies f1, and mf1 . Thus, when a dual-band filter is developed by cascading the structure in Fig. 1(a), the resulting will be two passbands at f1and mf1 with the same absolute bandwidths.

Jf1 = Yj/2 (C/D – E/F)
…………….(2.1)

Bf1 = YB/2 (C/D + E/F) – 2Yo1(G/H + K/L)-1
……………..(2.2)

Where,

C = (1 +Rz1) tanθ1
D = 1 - Rz1tan2 θ1
E = tanθ1 – Rz1cot θ1
F = 1 + Rz1
G = (1 + Rz2) tan θo1
H = Rz2 – tan2 θo1
K = Rz2 tan θo1 - cot θo1
L = 1+Rz2

And

Yj = Yt1
…………..(2.3)

YB = Yt1 + Yg1
…………..(2.4)

For the structure in fig (a)

YJ = (Y0o1 – Y0e1) / 2
…………..(2.5)

YB = (Y0o1 + Y0e1) / 2
………….(2.6)

For dual band operation at f1 and mf1 the J inverter values and the susceptance must satisfy the following conditions

Jf1 = Jmf1
…………(2.7)

Bf1 = 0 and Bmf1 = 0

IV. DESIGN EQUATIONS

In the previous section, a dual-band reduced-length parallel coupled line is proposed and the conditions for dual-band operations are derived. Then a dual-band filter is developed by cascading the proposed coupled lines. First, a conventional single band filter in Fig. 2(b) is designed with a center frequency at f1 . Then based on this design, the design parameters for a dual-band filter in Fig. 2(a) with center frequencies at and can be obtained. The following are the complete design equations for the proposed-order dual-band filter:

$$z_2 = z_1 / \tan\theta_{2_{f_1}}$$

$$z_3 = 0.5 z_2 \tan^2(2\theta_{2_{f_1}})$$

$$z_4 = \frac{z_1}{\sin(\theta_{4_{f_1}})}$$

$$z_5 = z_4 \tan(\theta_{5_{f_1}}).\tan(\theta_{4_{f_1}})$$

f2 / f1 = R frequency ratio (resonant frequencies) $\theta1 = \theta2 = \theta3 = \theta4 = \theta5 = \theta = \pi / (R+1)$ for equal electrical length of line. Although the proposed filter has no theoretical limitations on the frequency separation between the two passbands,

V. CONCLUSION

In this work, a new dual-band filter with controllable fractional bandwidths is proposed and constructed by using the distributed PCML coupled-line and dual-resonance SIR. After coupling dispersion of the PCML with varied lengths is studied in terms of explicit J-inverter susceptance, the three SIR circuits are modeled to exhibit the two tunable passbands. Further, the three SIR dual-band filters with two SIRs are optimally designed with varied dual-band FBWs as well verified by our experiment. This simple but effective design procedure provides us a powerful capacity in exploring the dual-band filter without needing any external dual-band matching network , thus miniaturizing the overall size and reducing the design complexity. This letter demonstrates a planar dual-band filter based on reduced-length parallel coupled lines.

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