Wide-Stopband Microstrip Bandpass Filters Using Dissimilar Quarter-Wavelength Stepped-Impedance Resonators

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***Abstract*—Wide-stopband and compact microstrip bandpass filters (BPFs) are proposed using various dissimilar quarter-wavelength ($\lambda/4$) stepped-impedance resonators (SIRs) for multiple spurious suppression. The use of $\lambda/4$ SIRs is essential in widening the filter stopband and reducing the circuit size. By properly arranging the individual $\lambda/4$ SIR, which has the same fundamental resonance frequency f_0, but has different spurious (harmonic) resonance frequencies, and also carefully misaligning the maximum current density nodes, several higher order spurious resonances may be suppressed so that a BPF with wide stopband may be realized. In this study, the basic concept of multiple spurious suppression is demonstrated by thoroughly investigating the spurious characteristics of the fourth-order interdigital BPFs, which consist of two different types of $\lambda/4$ SIRs. To widen the rejection bandwidth, the fourth-order coupled-resonator BPFs based on three and four different types of $\lambda/4$ SIRs are implemented and carefully examined. Specifically, a very wide-stopband microstrip BPF composed of four dissimilar $\lambda/4$ SIRs is realized and its stopband is extended even up to $11.4f_0$ with a rejection level better than 27.5 dB.**

***Index Terms*—Bandpass filter (BPF), coupled resonator, microstrip, quarter-wavelength stepped-impedance resonator, spurious suppression.**

I. Introduction

PRESELECTED bandpass filters (BPFs) with excellent out-of-band rejection and high selectivity are essential components of wireless communication system. In particular, the wide-stopband bandpass filters are usually needed in association with the nonlinear components (e.g., mixers or power amplifiers) so as to eliminate the undesired interference or noise in the stopband.

The filters composed of uniform distributed-element resonators suffer from the existence of spurious harmonics at multiples of the fundamental resonance frequency (f_0) due to their higher order resonances. The occurrence of spurious responses degrades the filter performance in rejecting the out-of- band interference. Various types of BPFs have been designed and fabricated using different forms of distributed resonators [1]–[10]. The filters based on half-wavelength ($\lambda/2$) uniform-impedance resonators (UIRs) such as coupled-line filters [1] and open-loop coupled-resonator filters [2], [3] have been well established, and these types of filters have the spurious responses around nf_0 ($n = 2, 3, 4, \ldots$). Quarter-wavelength ($\lambda/4$) resonators were also adopted to implement the filters, such as interdigital and combline filters [4]–[8]. Trisection or cascade quadruplet filters may also be realized using $\lambda/4$ resonators. The filters composed of $\lambda/4$ UIRs possess higher order harmonics occurring around $(2n+1)f_0$ ($n = 1, 2, \ldots$). Thus, by making good use of $\lambda/4$ resonators, one may make less effort of implementing a filter with better out-of-band rejection.

Several approaches were proposed to push the spurious responses higher or simply to suppress these unwanted passbands. The most direct way is to cascade a low-pass filter into the designed BPF [9]. However, the use of an additional low-pass filter degrades the insertion loss and increases the size of the filter, thus it is not the best way to eliminate the spurious responses. The first spurious passband may be pushed up to a higher frequency by a deformation of the linewidth using the SIRs [10]–[12]. However, due to the restriction of fabrication process, the SIR may at most push the spurious passband to $5f_0$. Alternatively, by the strip-width perturbation of a microstrip line to implement the wiggly-line filter [13], the first spurious passband of the coupled-line BPF located at $2f_0$ can be suppressed. By employing a different perturbation period in each coupled-line section, intended multiple spurious rejection can also be realized by using the microstrip wiggly-line structures [14].

Other methods available in the literature try to minimize the difference between the even- and odd-mode velocities, or to equalize the modal electrical lengths of microstrip coupled lines. In addition to providing tight coupling, a modified structure for the microstrip line to incorporate a centered slot at the ground plane was proposed to tune the even-/odd-mode phase velocities [15]. The even- and odd-mode phase velocities of parallel coupled-line filters can be equalized by employing the suspended coupled microstrips on a substrate with a proper suspension height [16], but the process of fabrication may thus be complicated. The even-mode velocity may also be speeded up by using a coupled-Schiffman section [17]. The technique of incorporating the over-coupled end stages to the coupled-line filters was proposed to increase the image impedance and, thus, to reduce the difference between modal propagation constants [18]. However, the filters in [15]–[18] based on the technique of modal velocities equalization can only remove the unwanted

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passband at $2f_0$, making the next higher order resonance appear immediately around $3f_0$.

Special spurious-suppression schemes were also proposed. The employment of uniplanar compact photonic-bandgap (UC-PBG) structure in a microstrip BPF to extend the stopband was proposed, but it can only provide a rejection of around 20 dB [19]. A modified version of the coupled-line filter introducing microstrip gaps to create a transmission zero was reported to suppress the spurious harmonic at $2f_0$, but it can only improve the upper stopband rejection to a level of 15 dB [20]. The rejection bandwidth may be extended by introducing some notch bands exactly at the spurious frequencies. Specifically, by means of proper tappings at both input and output resonators, two independent notches can be created at required frequencies, thereby canceling the spurious passbands [21].

In [22], the n-stage capacitively coupled BPFs with wide stopband characteristic were proposed using different types of dielectric-filled coaxial resonators that have the same fundamental resonance frequency, but with different spurious resonance frequencies. However, the filters proposed in [22] only adopted two different types of resonators, the UIR and SIR, thus the stopband performance is not optimized. Currently, the microstrip filters with less space, weight, and cost become attractive in modern communication system. Thus, the spurious-suppression method proposed in [22] needs to be extended to the microstrip circuits. Recently, a microstrip BPF composed of $\lambda/2$ UIRs and $\lambda/2$ SIRs [23] was proposed using the similar method of multiple spurious suppression as in [22] to push the stopband up to $4.93f_0$ with a rejection level of 20 dB. However, the stopband characteristic is not optimized due to the use of only two different types of resonators.

In this study, the concept of multiple spurious suppression in [22] is extended to design the compact microstrip BPFs with very wide stopband. Specifically, different types of $\lambda/4$ SIRs are suitably selected and arranged so that a very wide-stopband microstrip BPF may be realized. The adoption of $\lambda/4$ SIRs is a key to broaden the stopband and also to compact the circuit size. Basically, the span between adjacent spurious resonance frequencies of a $\lambda/4$ SIR is greater than that of a $\lambda/2$ SIR. Thus, the rejection bandwidth of the filter composed of dissimilar $\lambda/4$ SIRs may easily be made larger than that of the filter consisted of dissimilar $\lambda/2$ SIRs. According to the design procedure for conventional coupled-resonator filters [24], several fourth-order microstrip BPFs using different types of $\lambda/4$ SIRs are designed and examined. By properly arranging the spurious harmonic frequencies of each resonator, a fourth-order microstrip BPF with a very wide stopband even up to $11.4f_0$ is achieved without any degradation of insertion loss in the passband. The proposed filter is also compact because $\lambda/4$ resonators are adopted and no additional components or distributed elements are required.

II. Quarter-Wavelength SIRs

The structure of quarter-wavelength stepped-impedance resonator ($\lambda/4$ SIR) under consideration is shown in Fig. 1. This resonator is composed of two transmission-line sections of different linewidths for different characteristic impedances. The narrower line section of characteristic impedance Z_1 and electrical length θ_1 is connected to the ground through a round via-hole. The wider line section of characteristic impedance Z_2 and electrical length θ_2 is open at one end and connected to the narrower section at the other end.

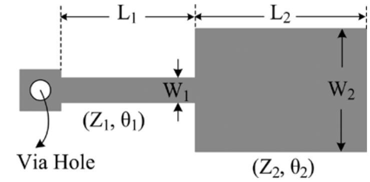


Fig. 1. General microstrip quarter-wavelength stepped-impedance resonator with one end short circuited to ground.

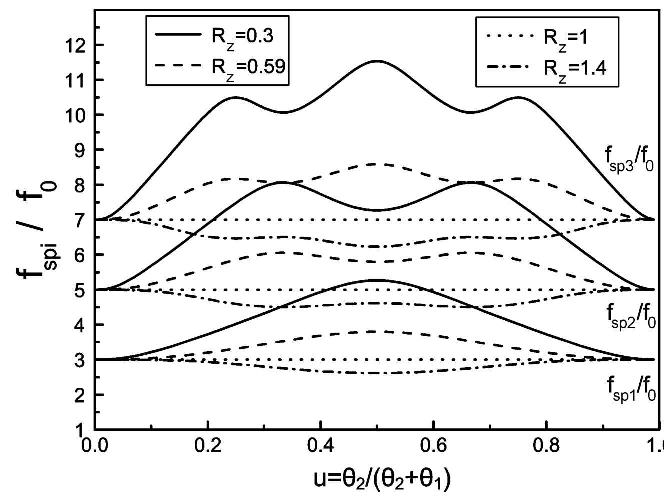


Fig. 2. Spurious (higher order resonance) frequencies $f_{spi} (i = 1, 2, 3)$ of $\lambda/4$ SIRs normalized with respect to their fundamental resonance frequency f_0 with $R_z = 0.3, 0.59, 1$, and 1.4 as parameters.

By neglecting the effects of discontinuities and open end, one may express the input impedance seen from the open end as

$$Z_{\text{in}} = jZ_2 \frac{Z_1 \tan\theta_1 + Z_2 \tan\theta_2}{Z_2 - Z_1 \tan\theta_1 \tan\theta_2}. \tag{1}$$

The parallel resonance occurs when $Y_{\text{in}} = 1/Z_{\text{in}} = 0$, from which the condition for odd-mode resonance can be given as

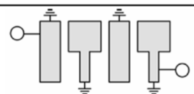
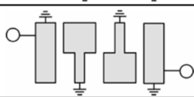
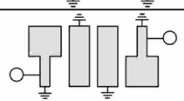
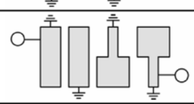
$$R_z = \tan\theta_1 \tan\theta_2 \tag{2}$$

where $R_z = Z_2/Z_1$ is the impedance ratio of the SIR. To simplify the calculation, the parameter $u = \theta_2/(\theta_1 + \theta_2)$ is adopted as in [21].

A good prediction of the required parameters u and R_z to give the desired spurious frequencies for each $\lambda/4$ resonator is essential in designing a wide-stopband filter. Fig. 2 shows the diagram of the spurious frequencies $f_{spi} (i = 1, 2, 3)$, for $\lambda/4$ SIRs with different values of u and R_z, which are calculated from (2) and normalized with respect to their corresponding fundamental resonance frequency f_0. Thus, for the required spurious frequencies of each resonator, one may get the related u and R_z from this diagram.

Compared with the $\lambda/2$ UIR, the $\lambda/4$ UIR with one end short circuited can only resonate at the odd mode. Thus, the spurious frequencies of $\lambda/4$ UIR are roughly at odd multiples of the fundamental frequency, i.e., $(2n + 1)f_0$ $(n = 1, 2, \ldots)$, and the

TABLE I
VARIOUS ARRANGEMENTS OF FOURTH-ORDER FILTERS USING TWO DISSIMILAR $\lambda/4$ SIRs

	Filter Layout	Spurious Rejection (dB)			
		3 f_0	4.09 f_0	4.99 f_0	6.05 f_0
(A)		23.4	14.38	20.74	14.94
(B)		36	4.8	32.27	8.2
(C)		27.17	50	11.85	35.9
(D)		22.56	22.19	21.33	19.44

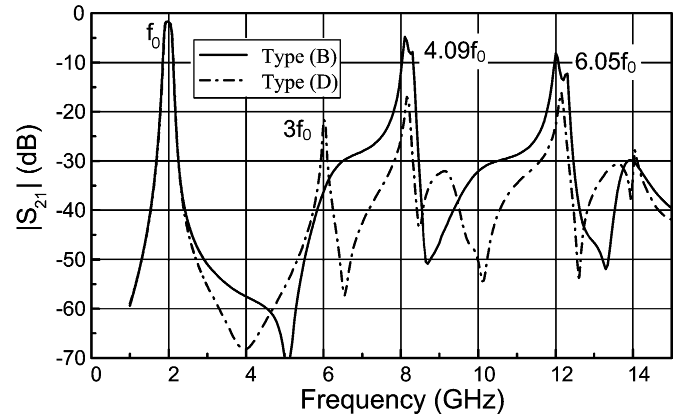


Fig. 3. Simulated frequency responses for type (B) and (D) filters listed in Table I.

frequency span of adjacent higher order resonances is approximately $2f_0$. Taking advantage of the stepped-impedance technique, the range of spurious span can further be made apart. This feature facilitates the goal of widely separating the spurious frequencies, which is required in the design phase.

In addition to the merit of separating spurious frequencies, the utilization of $\lambda/4$ resonators also effectively reduces the filter size. Conventional filters using $\lambda/2$ resonators may occupy circuit size four times that of the filters using $\lambda/4$ resonators. This property is attractive for size reduction in mobile communication.

The design flexibility and advantage of controlling the spurious frequencies available in $\lambda/4$ SIRs make them good candidates for building the coupled-resonator filter for multiple spurious suppression.

III. INTERDIGITAL FILTERS USING TWO DISSIMILAR $\lambda/4$ RESONATORS

The filter composed of identical $\lambda/4$ UIRs exhibits spurious responses at the odd harmonics $(2n+1)f_0$ $(n = 1, 2, \ldots)$ of the resonator. In [22], a method of multiple spurious suppression was proposed and tested for the filters using dielectric-filled coaxial resonators. This method will be extended to the design of microstrip filters.

Here, four possible types of microstrip interdigital coupled-resonator filters using two different types of $\lambda/4$ resonators, as listed in Table I, are carefully examined. All filters are implemented on a Rogers RO4003 substrate of thickness 0.508 mm, dielectric constant 3.38, metal thickness 17 μm, and loss tangent 0.0027. They are all designed for Butterworth response with 3-dB fractional bandwidth (3-dB FBW) of 10% and center frequency f_0 at 2 GHz. The UIR utilized in Table I has the fundamental resonance frequency at 2 GHz and higher order resonances at 6.011 and 9.983 GHz ($3f_0$ and $4.99f_0$), respectively. The SIR is selected to possess the same fundamental frequency with the UIR, but has higher order resonances at 8.18 and 12.1 GHz ($4.09f_0$ and $6.05f_0$). Table I shows the rejection levels at the spurious resonance frequencies of the UIR and SIR, respectively. The resonators, from left to right of each filter in Table I are also indexed with the numbers 1–4 for further reference.

Table I indicates that the type (A) and (D) filters have better spurious rejection than the other two types. The type (A) filter adopts the arrangement proposed in [22] using resonators of different types one after another. However, its rejection levels (14.38 and 14.94 dB) are not satisfactory at $4.09f_0$ and $6.05f_0$. The stopband performance may further be improved by modifying the arrangement of resonators in the type (A) filter. In the type (D) filter, the first pair of resonators is a UIR and the last pair is a SIR. Among the four types of filters listed in Table I, the type (D) filter is the better one to achieve a rejection level of approximately 20 dB. The simulated results in Table I suggests that the only use of two different types of $\lambda/4$ SIRs can at most provide the out-of-band rejection at approximately 20 dB. Shown in Fig. 3 are the simulated frequency responses using Ansoft Designer simulator for the type (B) and (D) filters, in which the type (B) filter has very bad spurious responses at $4.09f_0$ and $6.05f_0$.

Note that the two filters, i.e., types (A) and (D), with a satisfactory rejection level both have two dissimilar $\lambda/4$ SIRs (resonators 2 and 3) placed in the interstage. In synthesizing the design parameters of a fourth-order coupled-resonator filter, the coupling coefficient between resonators 1 and 2 is larger than that between resonators 2 and 3. If a filter had a pair of identical resonators placed in the interstage (e.g., the type (B) filter), the signals associated with the spurious frequencies of resonators 2 and 3 would be resonated within these two interstage resonators so that these strongly resonating signals might be coupled to resonators 1 and 4 and finally to the input and output. Therefore, the filter with identical interstage resonators might have poor stopband rejection at the higher order resonances of these identical resonators. To block or suppress the signals resonating at the spurious frequencies, it is better to make resonators 2 and 3 different and then to separate the higher order resonance frequencies of these two resonators as far as possible. Since the coupling between resonators 2 and 3 is weaker, and their higher order harmonics are separated, these two dissimilar interstage resonators may thereby serve as the spurious-blocking resonators to suppress the unwanted spurious passbands.

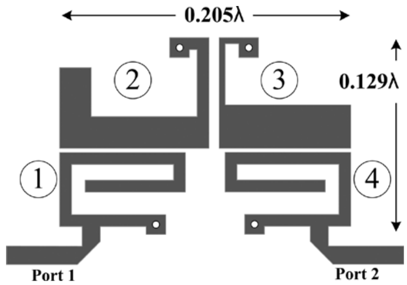


Fig. 4. Layout of the proposed microstrip coupled-resonator BPF using three dissimilar $\lambda/4$ SIRs.

For a fourth-order coupled-resonator filter, which contains four resonators, its spurious suppression is mainly determined by the arrangement and choice of these four resonators. Sagawa *et al.* [22] only implemented the filters with two different types of resonators. It is seen that the use of two dissimilar $\lambda/4$ resonators can at most provide a rejection of approximately 20 dB. Thus, the employment of only two different types of resonators is not the best choice to optimize the out-of-band rejection. In fact, there are four resonators that may be arbitrarily chosen. Below, the four resonators will be made different and properly arranged such that a wide-stopband filter may be implemented to improve the out-of-band rejection.

IV. Coupled-Resonator Filter Using Three Dissimilar $\lambda/4$ Resonators

Here, a wide-stopband filter using three different types of $\lambda/4$ SIRs is implemented on the RO4003. Shown in Fig. 4 is the layout of the proposed fourth-order Butterworth microstrip coupled-resonator BPF composed of three dissimilar $\lambda/4$ SIRs using the Butterworth filter to demonstrate the concept of multiple spurious suppression. All diameters of via-holes have the same size of 0.508 mm.

On designing a filter for multiple spurious suppression, the first step is to widely separate the first spurious frequencies of resonators 2 and 3. Thus, the values of u and R_z for these two resonators may then be determined from Fig. 2 such that their first spurious frequencies may be located at the appropriate frequency points. Having decided the distribution of first spurious frequencies for interstage resonators, one can then go on to locate the spurious frequencies of the other two resonators.

In this proposed filter, three dissimilar resonators are adopted, and resonators 1 and 4 are kept identical for symmetrical input/output tappings. The values of u and R_z for resonator 2 are chosen as 0.696 and 0.532, respectively, and this brings the first three higher order resonances located at 7.25, 12.75, and 18.9 GHz ($3.625f_0$, $6.375f_0$, and $9.45f_0$), respectively. The values of u and R_z for resonator 3 are 0.3 and 0.62 and this makes the first three higher order resonances located at 9.4, 15.05, and 19.4 GHz ($4.7f_0$, $7.525f_0$, and $9.7f_0$), respectively. In order to make the spurious frequencies of resonator 1 different from those of resonators 2 and 3, here the UIR is used for resonators 1 and 4, i.e., $u = 0$ and $R_z = 1$, and the spurious frequencies of these two resonators are roughly at odd multiples of fundamental frequency. The related dimensions and parameters are tabulated in Table II.

TABLE II
Physical Dimensions and Parameters for Each Resonator in Fig. 4

	Resonator 1	Resonator 2	Resonator 3	Resonator 4
Z_2 (Ω)	64.64	34.37	27.12	64.64
Z_1 (Ω)	64.64	64.64	90.83	64.64
R_z	1	0.532	0.299	1
W_2 (mm)	0.76	2.03	2.79	0.76
W_1 (mm)	0.76	0.76	0.38	0.76
L_2 (mm)	0	12.8	8.51	0
L_1 (mm)	24.13	5.59	5.2	24.13
u	0	0.696	0.62	0

Shown in Fig. 5(c) is the distribution of spurious frequencies of each resonator up to 20 GHz $(10f_0)$. Note that the distribution of spurious frequencies for each resonator has been carefully arranged.

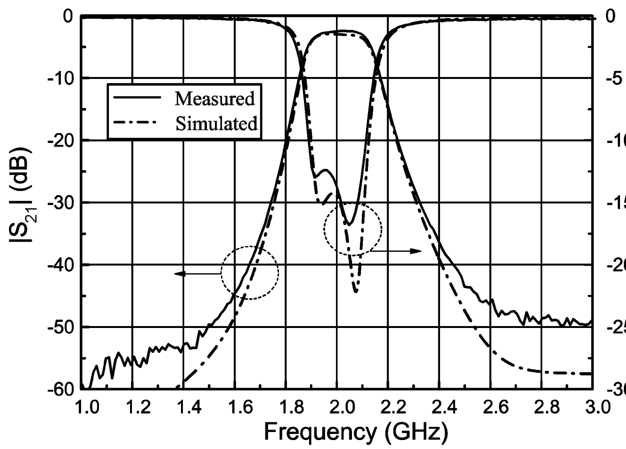
The design parameters of this proposed fourth-order filter are determined for the Butterworth response with center frequency at 2 GHz and 3-dB FBW of 10.2%. The external quality factor and coupling matrix $[M]$ are given as follows:

$$Q_{ei} = Q_{eo} = 7.504.$$
$$[M] = \begin{bmatrix} 0 & 0.0858 & 0 & 0 \\ 0.0858 & 0 & 0.055 & 0.0858 \\ 0 & 0.055 & 0 & 0 \\ 0 & 0 & 0.0858 & 0 \end{bmatrix}. \quad (3)$$
.

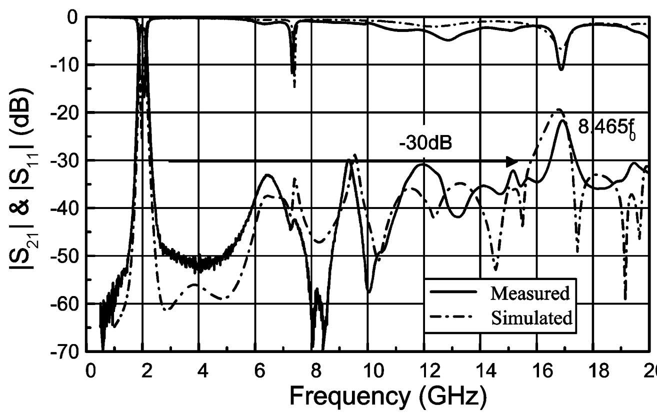
Fig. 5(a) and (b) shows the measured and simulated frequency responses of the proposed filter in Fig. 4. The measured center frequency is at 2.01 GHz, the measured FBW is approximately 12.64%, the minimum insertion loss is 2.48 dB at 2.01 GHz, and the return loss at 2.58 GHz is 13.7 dB. The fabricated filter is compact and its size is only approximately $0.205\lambda \times 0.129\lambda$ (18.82 mm $\times$ 11.84 mm), where λ is the guided wavelength at the center frequency. Fig. 5(b) shows the wide-band frequency responses of the filter in Fig. 4 ranging from 0.5 to 20 GHz. Note that this proposed filter has significantly pushed the stopband up to 16.478 GHz $(8.2f_0)$ with a rejection level around 30 dB.

From Fig. 5(c), it is observed that some spurious resonances are nearly coincident with each other at some specific frequencies, e.g., the spurious frequencies of resonators 1 and 3 around 9.5 GHz and the spurious frequencies of resonators 1 and 2 around 13 GHz. However, the spurious frequencies of resonators 2 and 3 have been separated wide enough to block these spurious harmonics around 9.5 and 13 GHz so that the filter still shows good rejection level around these frequency points.

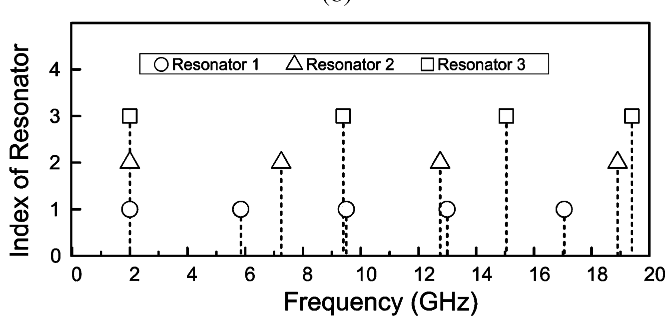
Even though the harmonic frequencies of each resonator have been properly distributed, the rejection response becomes poor around the fourth harmonic frequency 16.8 GHz $(8.465f_0)$ of resonator 1. To explain this poor rejection at $8.465f_0$, the amplitude distribution of current density over each resonator is plotted in Fig. 6(a). Though the higher order harmonic frequencies of resonators 2 and 3 are separated from the fourth harmonic of resonator 1 (16.8 GHz), the maximum current nodes over the adjacent resonators, at 16.8 GHz, are well aligned, thereby weakening the ability of spurious blocking. More precisely, the coupled signal between resonators 2 and 3 is not negligible and degrades the rejection level at that frequency. The overall current distribution over the proposed filter (Fig. 4) is also depicted



(a)



(b)

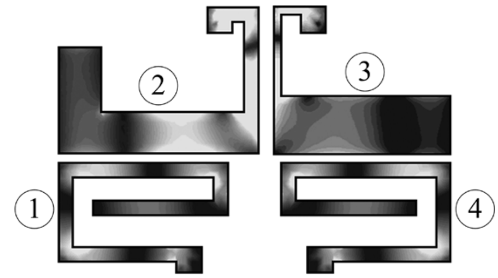


(c)

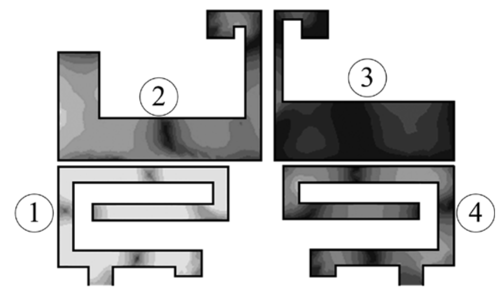
Fig. 5. (a) Narrow- and (b) wide-band measured and simulated responses of the filter in Fig. 4. (c) The fundamental and spurious frequencies of each resonator for the proposed filter in Fig. 4.

in Fig. 6(b), which shows that considerable coupling from resonators 1 to 2 to 3 and then to 4 is observed and gives a rejection of only 20 dB around the frequency of 16.8 GHz.

Note that the only information from the distribution of spurious frequencies [e.g., Fig. 5(c)] is not sufficient to predict the frequency response of a filter. It is the combined information both from the distribution of spurious frequencies and the amplitude distribution of current density that may better predict the overall frequency response of a filter. For the sake of preventing current node alignment between adjacent resonators, each resonator should be carefully designed to possess appropriate structure parameters so that the current nodes become misaligned.



(a)



(b)

Fig. 6. (a) Simulated current distribution over each resonator of the proposed filter in Fig. 4 excited isolatedly at 16.8 GHz. (b) Overall simulated current distribution over the proposed filter in Fig. 4 at 16.8 GHz. (The lighter the color, the stronger the current density, and vice versa.)

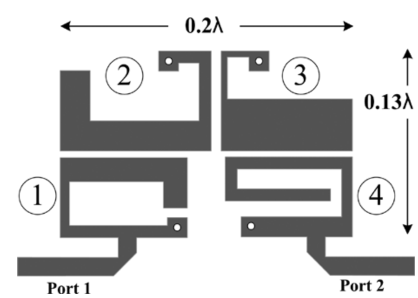


Fig. 7. Layout of the proposed microstrip coupled-resonator BPF using four dissimilar $\lambda/4$ SIRs.

V. Coupled-Resonator Filter Using Four Dissimilar $\lambda/4$ Resonators

Though the stopband of the previous filter is significantly pushed up to $8.2f_0$, the spurious passband again appears at $8.465f_0$. With the flexibility in choosing the four resonators, an intuitive method to remove this unwanted passband is the use of four dissimilar $\lambda/4$ resonators. By means of making resonators 1 and 4 different, one can prevent these two resonators from resonating at the same spurious frequency of the same order and move these nearby spurious frequencies to some points at which the coupling between resonators 2 and 3 becomes less significant. Most important of all, resonator 3 should be appropriately deformed to avoid its current nodes align with those of resonators 2 and/or 4 so that the spurious-blocking mechanism may be enhanced.

Fig. 7 shows the layout of another proposed filter using four different types of $\lambda/4$ SIRs and still fabricated on RO4003. This filter is designed for Butterworth response with the center frequency at 2 GHz and 3-dB FBW of 7%. It is composed of four dissimilar resonators with same fundamental frequency, but

TABLE III
PHYSICAL DIMENSIONS AND PARAMETERS FOR EACH RESONATOR IN FIG. 7

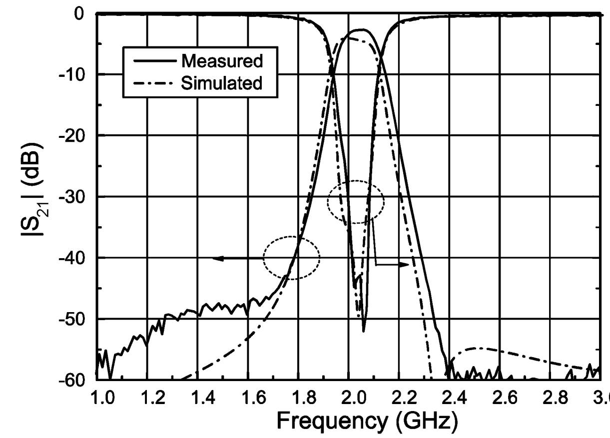
	Resonator 1	Resonator 2	Resonator 3	Resonator 4
Z_2 (Ω)	42.33	36	23.81	64.64
Z_1 (Ω)	64.64	64.64	90.83	47.63
R_z	0.655	0.557	0.262	1.357
W_2 (mm)	1.52	1.9	3.3	0.76
W_1 (mm)	0.76	0.76	0.38	1.27
L_2 (mm)	9.8	12.85	8.38	19.18
L_1 (mm)	9.65	5.77	4.45	5.84
u	0.504	0.69	0.653	0.77

with appropriately separated spurious frequencies. Table III lists the related dimensions and parameters of the filter in Fig. 7. The measured and simulated responses of the filter in Fig. 7 are shown in Fig. 8(a) and (b). The measured center frequency is at 2.044 GHz, the minimum insertion loss is 2.6 dB at center frequency, the measured 3-dB FBW is approximately 7.6%, and the return loss at center frequency is 21 dB. The filter is compact and has a size $0.2\lambda \times 0.13\lambda$ (18.7 mm × 11.87 mm).

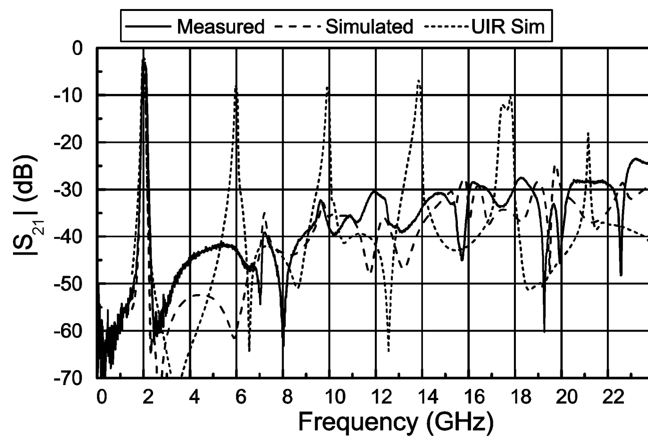
The fundamental and spurious frequencies of each resonator for the proposed filter in Fig. 7 are depicted in Fig. 8(c). Though the higher order resonances of resonators 1 and 2 are located almost at the same frequencies, the signals associated with these resonances may effectively be suppressed by the spurious-blocking mechanism provided by resonators 2 and 3. In this proposed arrangement, resonator 3 plays the primary role in blocking the spurious harmonics. With the impedance and length ratios properly chosen, resonator 3 has its first and second higher order spurious frequencies distributed greatly different from those of the other three resonators. Thus, even around the frequency of 12 GHz, where three higher order resonances of resonators 1, 2, and 4 nearly overlap, the filter still shows an acceptable rejection level of 30 dB around that frequency.

Another region that might result in a problem is the frequency band around 16.45 GHz over which the harmonic frequencies of resonators 1 and 4 are nearly equal. Observing the current distribution of each resonator at 16.45 GHz shown in Fig. 9(a), the maximum current nodes of resonator 3 are misaligned with those of resonators 2 and 4. It is this misalignment of maximum current nodes that effectively improves the rejection to a level of 30 dB, as depicted in Fig. 8(b). The overall filter's current distribution is also shown in Fig. 9(b). Notably, the signal is blocked between resonators 2 and 3 due to the misalignment of corresponding current distribution so that the rejection around 16.45 GHz gets improved.

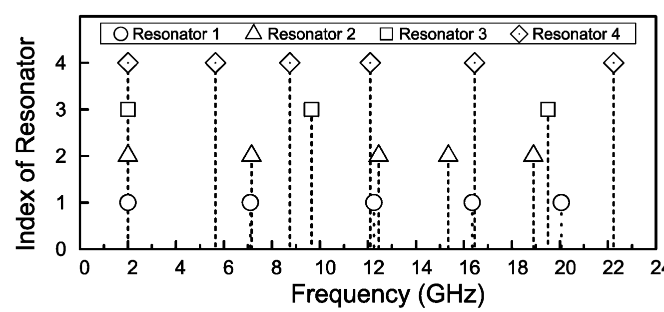
Shown in Fig. 8(b) are the wide-band frequency responses ranging from 0.01 to 24 GHz. Remarkably, the spurious passband around $8f_0$ has been pull down to an acceptable rejection level, thereby pushing the stopband even up to 22.8 GHz $(11.4f_0)$ with a rejection level better than 27.5 dB. For comparison, the frequency response of the microstrip interdigital filter, using all identical $\lambda/4$ UIRs and realized on the same substrate with the same specification, is also included in Fig. 8(b). Compared with this conventional interdigital filter, the proposed filter using four dissimilar $\lambda/4$ SIRs has effectively suppressed the spurious responses occurring at $3f_0$, $5f_0$, $7f_0$, $9f_0$, and $11f_0$, which would be associated with the interdigital filter composed of four identical $\lambda/4$ SIRs.



(a)



(b)



(c)

Fig. 8. (a) Narrow- and (b) wide-band measured and simulated frequency responses of the proposed filter in Fig. 7. The "UIR Sim" curve depicts the simulated response of conventional interdigital filter. (c) The fundamental and spurious frequencies of each resonator for the proposed filter in Fig. 7.

Some guidelines for realizing a wide-stopband filter are worthy of mention. In this study, two methods have been proposed to facilitate the design procedure. The first one is to properly distribute the harmonic frequencies of each resonator and, more importantly, to widely separate the harmonic frequencies of the interstage resonators (resonators 2 and 3). Note that the method of separating the resonator's harmonic frequencies provides an initial design of a filter whose rejection may be not wide and its level may be not acceptable (such as the type (D) filter shown in Table I). To realize a satisfactory wide-stopband filter, the second method, which misaligns the maximum current nodes of adjacent resonators, should be

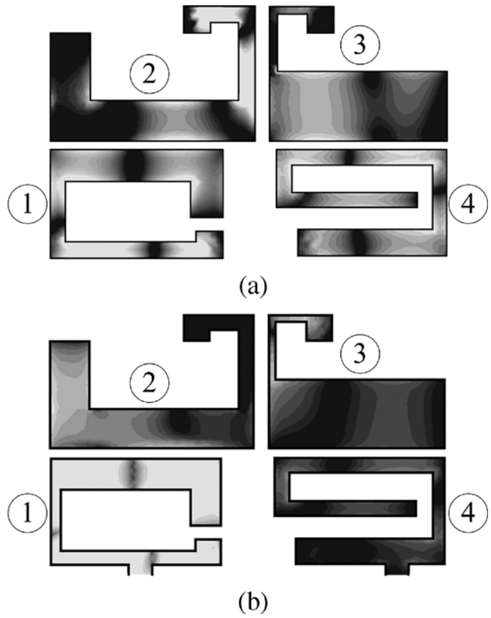


Fig. 9. (a) Simulated current distribution over each resonator of the proposed filter in Fig. 7 excited isolatedly at 16.45 GHz. (b) Overall simulated current distribution over the proposed filter in Fig. 7 at 16.45 GHz.

incorporated with the harmonic frequencies separation method in the design phase so that a very wide-stopband filter (Fig. 7) with an acceptable rejection level may be realized.

VI. Conclusion

Wide-stopband microstrip BPFs have been proposed using different types of $\lambda/4$ SIRs to multiply suppress the spurious harmonic resonances. The use of the $\lambda/4$ SIR, which possesses a wider span between adjacent spurious resonance frequencies, is essential in extending the rejection bandwidth and also reducing the filter size. By properly distributing the spurious frequencies of each resonator and also misaligning the maximum current nodes , a very wide-stopband fourth-order microstrip BPF composed of four dissimilar $\lambda/4$ SIRs has been implemented to suppress all of the spurious harmonics under $11f_0$, thereby pushing the stopband even up to $11.4f_0$ with a rejection level better than 27.5 dB. These wide-stopband filters are useful in a modern communication system for suppressing the undesired interference in the upper stopband.

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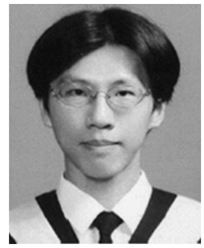

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