A Wide-Stopband Microstrip Bandpass Filter Using Stepped Impedance Resonators with Open-Circuited Stubs and Asymmetric Coupling Structure

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Abstract
For realizing stable and high-speed communication systems, the spectral of signals need to be purified, thus, an asymmetric third-order bandpass filter, which composed by open-circuited stubs loaded dissimilar stepped impedance resonators, is proposed to realize 20 dB stop-band attenuation up to 20.6$f_0$ (17.85 GHz). Open-circuited stubs are loaded to resonators for scattering harmonics and suppressing the unexpected harmonics. Asymmetric coupling structure has been applied to suppress the harmonics and the inevitable fly-over coupling between resonators. Stepped impedance resonators are designed with extremely low impedance ratio ($R_z$) for enhancing the open-circuited stubs and asymmetric coupling structures’ suppression on harmonics and reducing the circuit size. Resonance frequencies, external quality factor ($Q_e$) and coupling coefficient ($K$) between dissimilar resonators have been synthesized and derived by admittance analysis. The open-circuited stub and $K$’s influence on fundamental resonance, and $Q_e$ have been derived by equations and investigated by simulations respectively. Both the calculations and simulations show that the open-circuited stub and asymmetric coupling structure influence fundamental resonance little while the harmonics significantly.

Keywords: Microstrip Bandpass Filter, Dissimilar Stepped Impedance Resonators, Open-Circuited Stubs, Asymmetric Coupling Structure, Wide Stop-Band Suppression Response

1. Introduction
With various communication systems working in our modern society, the electromagnetic interference (EMI) degrades the system stability and the transmission efficiency. For example, in high power systems (i.e. Radars), some weak signal will be amplified and interfere the operation frequency. For purifying the frequency spectrum, a filter, which has a wide 20 dB reject zone ($20f_0$ until K-band), is need to suppress the harmonics’ power to 1%.

Many filters, which have wide stop-band response, have been designed for eliminating the EMI. The stepped impedance resonator filters whose impedance ratios can be arranged precisely to force the second harmonic to shift to higher band,[1, 2] however, the harmonics distribution would be extremely sophisticated and unpredictable when they beyond $8f_0$. Some works are integrated with electromagnetic band-gap structures [3] to support band-stop response, defected ground structure [4, 5] and capacitively loaded cells [6] to form low-pass response to attenuate the harmonics, however, those structures increase the insertion loss at $f_0$ and need more time for design. By introducing cross-coupled open-circuited stubs,[7] transmission zeros could be introduced and enhance stop-band response, but some unexpected harmonics would be introduced by cross-coupling sections and shorten the reject zone. Moreover, as reported by,[8] the transmission zero could be shifted to extend the stop-band of the filter and suppress harmonics. For avoiding extra pass-band which formed by harmonics couplings, dissimilar resonators are widely applied in both cavity filters and geometrical structures filters.[9–11] But as depicted in reference,[10] harmonics have been staggered by tuning the $R_z$ of dissimilar resonator. Wide stop-band response has been achieved, but unexpected harmonics are generated and make the distribution of harmonics becomes very complicated, moreover, at least 4 dissimilar resonators need to be designed and tuned which mean a lot of time and energy. In references,[12, 13] short-
open-circuited stubs have been loaded to resonators for breaking periodical property. Those works showed very sophisticating theory and designed fifth-order bandpass filters (BPFs). Those BPFs achieve 20 dB reject zone extended to 18f0. Also, some chip elements are employed for minimizing the circuit size. In reference, [14] open-circuited stubs have been loaded to resonators for staggering the harmonics modes to shift to each other’s rejection zone, without taking advantage of Rz, the circuit size is relative large, and chip elements are need to be loaded on open-circuited stubs.

For extending the 20 dB reject zone to 20f0, only 2 simple types dissimilar stepped impedance resonators are designed based on microstrip structure to minimize the insertion loss, and improving the design efficiency. Extremely low impedance ratio Rz, which shortens the electrical length of harmonics, is investigated and applied to force the harmonics shift to higher band, reduce the filter circuit size, and enhance the suppression of asymmetric coupling structure and open-circuited stubs. Asymmetric coupling structure has been developed and carefully investigated for suppressing the harmonics above 8f0 systematically. The effect of open-circuited stubs on f0, external quality factor Qe and coupling coefficient K have been carefully investigated by calculations and simulations for predicting and adjusting the response of filter. Moreover, no chip elements or multilayer structures would be need to be employed in this filter.

2. Synthesis of Bandpass Filter with Wide Stop-band Responses

The proposed third-order filter, which is composed by two kinds of stepped impedance resonators, is shown in Fig. 1. One kind of stepped impedance resonator is defined as type (A), which is set on left and right sides in Fig. 1. While, the other one is defined as type (B), which is set on center in Fig. 1. The two coupling structures between type (A)s and (B) are configured as asymmetric coupling structure for suppressing harmonics.

2.1 Synthesis of proposed stepped impedance resonators’ characteristics

As shown in Fig. 1, for achieving compact size of the filter, stepped impedance resonators are need to be clustered close to each other, thus, inevitable coupling between resonators would support some unexpected harmonics response, for scattering and suppressing them, the resonant properties of resonators need to be anticipated.

The middle resonator (B) is designed based on classic stepped impedance resonator with extremely low Rz and shown in Fig. 2. The fundamental resonance f0 and the harmonics of resonators can be obtained by equation (1) when Ym = 0.

\[
Y_m = \frac{Y_2}{f} \frac{R_z - \tan \theta_1 \tan \theta_2}{\tan \theta_1 + R_z \tan \theta_2} 
\]

(1)

\[
R_z = \frac{Y_1}{Y_2} \quad \theta_n = \beta l_n \quad (n = 1, 2, \ldots, 8) 
\]

(2)

where \(\theta_n\) and \(l_n\) represent the electrical length and practical length of admittance section, respectively.

Side resonator (A)s are designed based on stepped impedance resonator with extremely low Rz and open-circuited stub loaded as shown in Fig. 3. The proposed (A) contains four admittance sections, \(Y_2\) is the admittance of

![Fig. 1 Proposed filter.](image_url)

![Fig. 2 Structure of classic stepped impedance resonator of middle stepped impedance resonator.](image_url)

![Fig. 3 The structure of stepped impedance resonator (A) with open-circuited stub loaded.](image_url)
transmission line section with electrical length $\theta_0$, $Y_1$ and $Y_3$ are the low admittance sections with length $\theta_1$ and $\theta_3$ respectively. $Y_2$ is the admittance of open-circuited stub with length $\theta_2$. There is no open-circuited stub loaded to resonator when $\theta_2$ is equal to 0, the structure shown in Fig. 3 is identical to the one shown in Fig. 2. With electrical length compensation, the resonant properties can be obtained by equation (1). Selecting the (A)'s $R_z = 0.13$ without open-circuited stub loaded, and (B)'s $R_z = 0.09$, resonant properties of (A) and (B) are shown in Fig. 4. The calculations in Fig. 4 show the resonant properties with the green line representing the (A)'s resonance and purple line for (B)'s. The third harmonic of (A) is closed to the second of (B). And only two dissimilar open-circuited stubs are need for scattering the harmonics of all three resonators.

With open-circuited stub loaded, choosing the admittance $Y_1 = Y_2 = Y_3$, length $\theta_1 + \theta_2 = \theta_4$ and $R_z = Y_1 / Y_4$ to simplify the calculation, the input admittance can be obtained by equation (4), and when $Y_{in} = 0$, the resonant properties can be obtained.

\[
Y_{in} = \frac{Y_1 \cdot j \cdot \tan \theta_1 \cdot \tan \theta_2 \cdot \tan \theta_3 \cdot \tan \theta_4}{-\tan \theta_4 \cdot \tan \theta_2 \cdot R_z - \tan \theta_4 \cdot \tan \theta_2 \cdot R_z - \tan \theta_4 \cdot \tan \theta_2 \cdot R_z + \tan \theta_4 \cdot \tan \theta_2 \cdot \tan \theta_3 \cdot \tan \theta_4 \cdot R_z}
\]

(4)

According to equation (4), the fundamental resonance is mainly determined by $\theta_1$, $\theta_2$ and $\theta_4$, the total number of polynomial terms of numerator is six, and the $\theta_2$, which related to open-circuited stub's length $l_2$, only be contained in two terms while $\theta_1$, $\theta_2$ and $\theta_4$ contained in at least three terms. Moreover, $\theta_2$ is always contained in terms $\tan \theta_1 \cdot \tan \theta_2$, which means the influence of open-circuited stubs can be weaken by shortening the $\theta_2$, in other words, the influence of open-circuited stub can be weaken when attached near to the grounded via.

Since the $R_z$ is extremely low and makes second harmonic higher than $f_0$, it is obvious that $\theta_2$ of fundamental resonance is significantly longer than the electrical length of harmonics, the $l_2$ influences the second harmonic much stronger than $f_0$, i.e., when the practical length of open-circuited stub $l_2$ equals 7 mm, the corresponding electrical length $\theta_2$ is $7\pi/108$ at 0.87 GHz and makes the $\tan \theta_2$ lower than 0.0005, $\theta_2$ equals $\pi/2$ at 7 GHz (second harmonic) and makes the value of $\tan \theta_2$ equals to infinity.

Moreover, the lower $R_z$ is, the harmonics would accept more influence of open-circuited stub. Calculated open-circuited stub's influence on second harmonic versus $R_z$ is shown in Fig. 5. According to Fig. 5, the influence of open-circuited stub has been enhanced by the extremely low $R_z$ significantly. When $R_z = 0.1$, with selecting practical length $l_2 = 1$ and 10 mm respectively, the second harmonic would be generated at 8.1 and 3.67 GHz, when $R_z = 0.9$, the harmonic is generated at 2.78 and 2.58 GHz.

Since the $Q_e$ is need to be extracted for evaluating the coupling between input or output feed lines and resonators, a $Q_e$ extraction model has been built as shown in Fig. 6 and can be obtained by equation (5)[15]:

\[
Q_e = \frac{\omega}{2Y_s} \frac{\partial \text{Im}[Y_{inQe}]}{\partial \omega} \bigg|_{\omega = \omega_0}
\]

(5)

where $Y_s$ is the characteristic admittance of input/output

![Fig. 4 Resonant properties of stepped impedance resonators (A) and (B).](image)

![Fig. 5 Theoretical calculated open-circuited stub’s influence on second harmonic versus $R_z$.](image)

![Fig. 6 Schematic structure of tapped-line feeding.](image)
transmission lines which equals to 0.02 S, \( Y_{\text{in}0e} \) is the input admittance, \( \omega_0 \) represents fundamental angular frequency. With choosing \( Y_1 = Y_2 = Y_3 = Y_4 \), the length \( \theta_1 + \theta_2 + \theta_1 = \theta_0 \) and \( R_e = Y_1 / Y_0 \), \( Y_{\text{in}0e} \) can be simplified to a polynomial with eighteen terms as equation (6), and the influence of open-circuited stubs could be analyzed by this equation.

\[
Y_{\text{in}0e} = Y_4 \cdot j \cdot (R_e \tan \theta_1 \tan \theta_2 \tan \theta_3 \tan \theta_4 + \tan \theta_1 \tan \theta_2 \tan \theta_4 \tan \theta_5 + \tan \theta_1 \tan \theta_2 \tan \theta_3 \tan \theta_6 - R_e \tan \theta_1 \tan \theta_2 - R_e \tan \theta_1 \tan \theta_6 - R_e \tan \theta_2 \tan \theta_4 - \tan \theta_1 \tan \theta_5 - \tan \theta_2 \tan \theta_4 - \tan \theta_3 \tan \theta_5 + R_e) / (-\tan \theta_1 \tan \theta_2 \tan \theta_3 \tan \theta_4 \tan \theta_6 + \tan \theta_1 \tan \theta_2 \tan \theta_3 \tan \theta_6 + \tan \theta_2 \tan \theta_3 \tan \theta_5 + R_e \tan \theta_4 - R_e \tan \theta_2 - \tan \theta_3 \tan \theta_5 + R_e) (6)
\]

The \( Q_e \) is mainly determined by \( \theta_1, \theta_3, \theta_4, \) and \( \theta_5, \) as the \( \theta_2 \) is contained in four terms in numerator and two in denominator while the others \( \theta \) contained in at least nine terms, i.e., \( \theta_0 \) is contained in six terms in numerator and three in denominator. Same as the analysis of the resonant properties, the \( \tan \theta_n \) always shows itself with \( \tan \theta_n \), thus the influence can be weaken by shortening \( \theta_n \) and the \( \theta_2 \) can be changed with little influence on the \( Q_e \).

All analysis above strongly suggest that the length changing of open-circuited stub has little influence on fundamental resonance and \( Q_e \), while the harmonics could be shifted as well, and it is very flexible in practical circuit tuning by changing the microstrip line length \( l_2 \) to scatter the harmonics.

### 2.2 Synthesis of coupling coefficient constant \( K \) of proposed coupling structure

Since the size of filter circuit is compact, the distance between the resonators is shorter than 1 mm. Inevitable fly-over coupling would support some harmonics' pass-band, so the intensity and type of coupling are need to be evaluated and anticipated.

A circuit model of coupling structure between dissimilar stepped impedance resonators is shown in Fig. 7. Treating parallel coupled lines of coupling structure as a four ports circuit, the coupling coefficient constant \( k_{12} \) between resonators (A) and (B) could be obtained by deriving the input admittance of parallel coupled lines as (7).[16]

\[
k_{12} = \frac{\text{Im}[Y_{\text{in}12}]}{\sqrt{b_A \cdot b_B}} (7)
\]

where

\[
b_A = \frac{\omega}{2} \frac{\partial \text{Im}[Y_{\text{in}A}]}{\partial \omega} \bigg|_{\omega=\omega_0},
\]

and

\[
b_B = \frac{\omega}{2} \frac{\partial \text{Im}[Y_{\text{in}B}]}{\partial \omega} \bigg|_{\omega=\omega_0}, (8)
\]

\( b_A \) is slope parameter of port A, and \( b_B \) for port B,

\[
Y_{\text{in}A} = Y_{\text{in}1} + Y_{\phi 11},
\]

\[
Y_{\text{in}B} = Y_{\text{in}6} + Y_{\phi 22},
\]

\( Y_{\phi 11} \) and \( Y_{\phi 22} \) are the input admittance of port A and B which including admittance \( Y_{\text{in}} \) and \( Y_{\text{sum}} \) of short- and open-circuited stubs as shown in Fig. 7,

\[
Y_{\phi 11} = Y_{11} - \frac{Y_{32} \cdot Y_{13}}{Y_{33} + Y_{\text{in}}},
\]

\[
Y_{\phi 22} = Y_{22} - \frac{Y_{23} \cdot Y_{23}}{Y_{33} + Y_{\text{in}}},
\]

\[
Y_{\phi 32} = Y_{\phi 21} = Y_{12} - \frac{Y_{32} \cdot Y_{13}}{Y_{33} + Y_{\text{in}}},
\]

\( Y_{11}, Y_{12}, Y_{13}, Y_{22}, Y_{23}, \) and \( Y_{33} \) are Y-parameters of coupled line.

Just as analyzed in Section 2.1 the open-circuited stubs' influence on \( K \) is need to be evaluated for ensuring the pass-band width of proposed filter, however, the equation (7) is very complex, the denominator is a polynomial which contains with more than 1.6 billion terms. The theoretical analysis is very difficult, but owing to the extremely low \( R_e \), electrical length of fundamental resonance is significantly longer than the harmonics', it is easy to find that the asymmetric coupling structure has little influence on the fundamental resonance’s \( K \), and this would be verified in Section 3.3 by simulations.

### 2.3 Harmonics suppression by asymmetric coupling structure

It is difficult to scatter the modes of harmonics and guarantee 20 dB attenuation in stop-band since the electromagnetic far field harmonics of the proposed filter are
very complicated. For stopping the energy of harmonics propagating within the resonators, asymmetric electromagnetic coupling structure has been proposed and investigated. By adjusting the length of $\theta_6$ and $\theta_7$, the coupling structure could be arranged asymmetrically, and the coupling type could be changed from negative to positive (or conversely as well). Treating the positive coupling as magnetic coupling and the negative as electric, it is obvious that the electromagnetic energy which coupled by mutual inductors (magnetic coupling) can hardly be coupled by mutual capacitors (electric coupling), so the harmonics could be suppressed. To verify this method, an ideal schematic model, which shown in Fig. 8, has been built by Keysight Advanced Design System 2014. This schematic in Fig. 8 contains RLC elements to represent resonators and ideal transmission lines to represent coupling structures between resonators. The TL2 between the resonators is set as $-90$ degree to realize the electric coupling while the TL3 is set as the 90 degree for magnetic coupling, and the impedance Z12 and Z23 are set to realize the coupling intensity. Simulation results are shown in Fig. 9.

The pure electric or magnetic coupling can not attenuate the second harmonic, while the mixed coupling method can cause a 7 dB attenuation. This strongly suggests that asymmetric coupling structure performs well in harmonic suppression. Comparing to the asymmetric coupling structure’s coupled line electrical length $\theta_{cp}$, the fundamental resonance has a longer electrical length as $\theta_1 + \theta_2 + \theta_4 + \theta_{cp}$ or $\theta_6 + \theta_7 + \theta_8 + \theta_{cp}$, the tuning of asymmetric coupling structure would have little influence on fundamental coupling coefficient $K_0$ in practical design, so the bandwidth of pass-band would not be influenced.

It needs to be mentioned that, since this filter has a very compact size, side (A)s are fly-over coupling with each other inevitably, the asymmetric structure can also weaken this coupling by staggering the relative position between side (A)s.

3. Theoretical Calculations and Simulations of Proposed Stepped Impedance Resonators and Coupling Structure

3.1 Resonant properties of dissimilar stepped impedance resonators

The dissimilar stepped impedance resonators’ resonant properties $f_{(A)}$ and $f_{(B)}$ are calculated by Maple 2015 and shown in Table 1.

The resonator (B) is designed with extremely low $R_z$ around 0.086 which forced the second harmonic mode shift to 8.7 GHz (10f0). With no open-circuited stubs loaded to (A), relative higher $R_z$ of the (A) has been set as 0.133 to make second harmonic to shift to 6.97 GHz (8f0) which has been positioned in rejection zones of (B), so the second harmonics of (A) and (B) could be scattered and suppressed, however, the second harmonic of (B) is very close to the third of (A), open-circuited stubs are needed to...
With the practical length $l_2$ changed from 0 to 10 mm, the resonances have been theoretical calculated and shown in Table 2. The rate-of-change of fundamental resonance is only 0.8%, and the others’ at least 14.91%, just as the theoretical analysis in Section 2.1, the open-circuited stub has nearly no influence on fundamental resonance, while the second, third, fourth, fifth and sixth harmonics could be shifted dramatically. According to Table 2, the $l_2$ should be selected as 4–7 mm to scatter the harmonics of (A) and (B).

Moreover, the rejection bands of (B) between the first and second, third and fourth harmonics have bandwidth wider than 8 GHz which is quite flexible for inserting and scattering (A)’s fourth and fifth harmonics into them.

The electromagnetic simulation models of resonators (A) and (B) built by Ansoft HFSS 12.9 are shown in Fig. 10 (a) and (b) respectively. The simulation results are shown in Fig. 11 with $f_{0}$ to represent the fundamental resonance and $f_h$ for the harmonics. According to Fig. 11, some unexpected harmonics are generated (mainly caused by the low $R_z$), and the resonant properties become very complicated, with the open-circuited stub loaded to the resonator (A), the harmonics can be shifted dramatically, and the fundamental frequency is almost unchanged just the same as the theoretical calculation shows, moreover, when $l_2$ changed from 3 to 7 mm the third and fourth harmonics of side (A) were attenuated significantly.

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### Table 1 The theoretical calculations of stepped impedance resonators’ resonant properties without open-circuited stub loaded.

| Resonance frequency (GHz) | 1st | 2nd | 3rd | 4th | 5th |
|---------------------------|-----|-----|-----|-----|-----|
| $f_{0}(A)$                | 0.875 | 6.97 | 8.72 | 14.83 | 16.58 |
| $f_{0}(B)$                | 0.873 | 8.68 | 10.42 | 18.23 | 19.98 |

### Table 2 The theoretical calculations of stepped impedance resonator (A) resonant properties with open-circuited stub loaded.

| Resonance frequency (GHz) | 1st | 2nd | 3rd | 4th | 5th | 6th |
|---------------------------|-----|-----|-----|-----|-----|-----|
| $l_2$ (mm)                |     |     |     |     |     |     |
| 0                         | 0.875 | 6.97 | 8.72 | 14.83 | 16.58 | 22.68 |
| 1                         | 0.874 | 6.71 | 8.43 | 14.48 | 16.42 | 22.43 |
| 2                         | 0.874 | 6.39 | 8.21 | 14.04 | 16.26 | 21.03 |
| 3                         | 0.873 | 6.01 | 8.06 | 13.35 | 16.02 | 18.43 |
| 4                         | 0.872 | 5.61 | 7.93 | 12.41 | 15.64 | 17.15 |
| 5                         | 0.872 | 5.2 | 7.83 | 11.46 | 15.25 | 16.72 |
| 6                         | 0.871 | 4.81 | 7.75 | 10.69 | 14.93 | 16.52 |
| 7                         | 0.87 | 4.44 | 7.67 | 10.1 | 14.64 | 16.38 |
| 8                         | 0.87 | 4.12 | 7.59 | 9.62 | 14.32 | 16.22 |
| 9                         | 0.869 | 3.82 | 7.51 | 9.24 | 13.91 | 16 |
| 10                        | 0.868 | 3.56 | 7.42 | 8.93 | 13.39 | 15.65 |

Rate-of-change | 0.8% | 48.92% | 14.91% | 39.78% | 19.24% | 31% |

Fig. 10 The 3D models of (A) Side stepped impedance resonator (A) and (B) Middle stepped impedance resonator (B).
3.2 Open-circuited stubs’ influence on $Q_e$

The external quality factor $Q_e$ extraction model is shown in Fig. 12. With the $l_2$ changed from 0 to 7 mm, the theoretical calculation shows that the $Q_e$ is nearly unchanged around 5.82, and the simulations show the $Q_e$ changes from 6.1 to 6.36 as shown in Fig. 13, both of them strongly suggest that the changing of $l_2$ has little influence on $Q_e$.

3.3 Theoretical calculated and simulated $K$ of fundamental resonance

To extract the coupling structure precisely, an HFSS model, which shown in Fig. 14, has been built with grounded (GND) vias and perfect conductor (PEC) wall set between resonators (A) and (B) for shielding the coupling between the high admittance transmission line sections to realize the same coupling structure shown in Fig. 6.

According to Fig. 15, the calculations and the simulations show agreement in variation tendency, owing to the compact circuit shape, some inevitable fly-over coupling make the simulations’ $K$ tighter than the calculations’ about 0.015. According to the phase of $S_{21}$, shifting the (A) up or downwards (to realize the asymmetric coupling structure), the coupling type of fundamental resonance would keep negative. With selecting $l_6 = 4$ mm, the fundamental $K_0$ obtained by simulation is $-0.109$, and $-0.17$ without grounded vias and PEC wall.

As shown in Fig. 16, with open-circuited stubs’ length $l_2$ changed from 4 to 7 mm while $l_6$ set as 4.5 mm, the $K$ of fundamental resonance almost keeps unchanged just as...
analyzed in Section 2.2, it strongly suggests that the open-circuited stubs have little influence on $K_0$.

3.4 Theoretical calculated and simulated $K$ of asymmetric coupling structure

Without open-circuited stub loaded, the theoretical calculations of asymmetric coupling structure’s $K$ are shown in Table 3 and Fig. 17.

Choosing $X.5\ GHz$ ($X$ is integer) as the typical frequency to estimate the $K$ variation tendency of the band $X \sim X + 1\ GHz$ roughly (i.e. $7.5\ GHz$ for band $7$ to $8\ GHz$), and dividing the $K$ of harmonics ($K_{hi}$) which has same variation tendency into same group, 4 groups can be obtained, the $K$ of corresponding frequencies bands can be anticipated for designing the asymmetric coupling structure for suppressing.

According to Fig. 17, with selecting $l_6 > 2.5\ mm$, the $K$ of the fundamental resonance is almost unchanged and much bigger than the harmonics’. Moreover, the $K_{hi}$ of harmonics is changing dramatically as they have much shorter electrical length.

According to Fig. 17 (a), to 7.5, 8.5, and 9.5 GHz, they have a same variation tendency of changing from tight to loose coupling as calculated by equation (7). To 7.5 and 8.5 GHz, with $l_6$ set as 3 and 4.5 mm respectively, $K$ of 7.5 GHz are $-0.018$ and $0.05$, and $K$ of 8.5 GHz are $-0.05$ and $0.01$, so asymmetric coupling structure could be realized to suppress the harmonics just as analyzed in Section 2.3. And, from 7 to 10 GHz, $l_6$ should be set as 4.5 mm to realize relative looser coupling between resonators (A) and (B).

According to Fig. 17 (b), to 10.5 and 11.5 GHz, all of their $K_{hi}$ is below 0.026 which means a relative looser coupling between resonators than other harmonics, and $l_6$ should be set as 3.5~4.5 mm to make $K_{hi}$ below 0.015.

According to Fig. 17 (c), to 12.5, 13.5, 14.5, and 15.5 GHz, their $K_{hi}$ change from positive to negative. Asymmetric coupling structure can be realized, however, the $K_{hi}$ of 13.5 and 14.5 GHz is stronger than 0.0562, which mean considerable tighter coupling and no harmonic should be arranged into these bands, and $l_6$ should be set as 4.5 mm to weaken the $K_{hi}$ below 0.03.

According to Fig. 17 (d), to 16.5 and 17.5 GHz, $K_{hi}$ is changing from negative to positive and back to negative
coupling. Since their $K_H$ could be tighter than 0.04 as $l_6$ set as 3.5 mm, and tighter than -0.06 as $l_6$ set as 4.0 to 4.5 mm, asymmetric coupling structure could be realized.

As shown in Fig. 18, two types of HFSS models have been built to evaluate the influence of asymmetric coupling structure, and their simulations have been shown in Fig. 19.

In practical design, the PEC walls are very difficult to realize, thus, models without PEC walls and GND vias are need to be investigated for evaluating the influence of proposed asymmetric coupling structure, their response are shown in Fig. 20. Both of Figs. 19 and 20 show that the pass-band of symmetric and asymmetric filters are identically overlapping with each other, which strongly suggests that the asymmetric coupling structure almost has no influence on the pass-band width, moreover, with asymmetric coupling structure, the attenuation of 2 to 9 GHz has been increased more than 3 dB, the harmonics around 9.4 GHz is about 6.5 dB weaker. It is obvious that all the harmonics' pass-bands are sabotaged and only some harmonics' spurs remaining. With changing the length of...
open-circuited stub $l_2$, these spurs could be shifted to resonator’s reject zones and attenuated.

4. Design Example

Based on all above theoretical calculations and simulations, the fundamental resonance, its $Q_e$ and $K$ would not be influenced by adjusting the open-circuited stubs, so the design flowchart of proposed filter could be designed by conventional synthesis as shown in Fig. 21.

To verify the methods used for suppressing the harmonics, a third-order Chebyshev filter was fabricated on a Panasonic MEGTRON7(N) substrate with a relative dielectric constant of 3.4, a thickness of 0.5 mm, and loss tangent for 0.001. The specifications of filter are shown as follows.

1) Operation frequency: 0.87 GHz
2) Pass-band: 0.81 to 0.93 GHz
3) Ripple: 0.004 dB

The fabricated filter is shown in Fig. 22, and the broad and narrow band frequency response, which measured by vector network analyzer ZVB 20 (Rohde & Schwarz GmbH & Co. KG), are shown in Figs. 23 and 24, respec-

Fig. 21 Flowchart of design procedure.

Fig. 22 Fabricated filter.

Fig. 23 Comparison of measured and simulated wide band response of proposed filter.

Fig. 24 Comparison of measured and simulated narrow band response of proposed filter.
tively. The measured operation frequency of proposed filter is 866 MHz, the 3-dB fractional bandwidth is 14.1% from 0.794 to 0.916 GHz, in-band return loss $|S_{11}|$ is better than 17.2 dB, and the minimum insertion loss $|S_{21}|$ is 1.52 dB, transmission zero, which is caused by fly-over coupling between side resonators, is generated at 1.288 GHz. The rejection level of 0.962 to 17.85 GHz is better than 20 dB except a very weak harmonic generated at 11.32 GHz. The size is $39 \times 17.6 \text{ mm}^2$, approximately $0.177 \times 0.082 \lambda_g^2$, where $\lambda_g$ represents the guide wavelength on the substrate at operation frequency. Due to the comparison, the proposed filter has a much wider reject zone than the conventional uniform impedance resonator filter. Moreover, for investigating the temperature dependency of proposed filter, the responses by changing resistances of resonators in equivalent circuit for third-order BPF have been simulated, the simulations show the reject zone response of this filter accepted little influence while the insertion loss changed a little bit more, measurements will be carried on for verifying this.

Below 2 GHz the measurement and the simulation show a great agreement with each other. According to the data-sheet of MEGTRON7 (N), the relative dielectric constant becomes lower than 3.4 when the frequency above 2 GHz (frequency dependence by relative dielectric constant), this makes the harmonics shift to higher band as the measurement shows, and make the simulations and calculations become very sophisticated to predict the response of practical filter. Comparisons of the proposed filter and previous works have been illustrated in Table 4 and indicate that proposed filter has both compact size and wide stop-band response.

It needs to be mentioned that, the suppression methods used in this theme can be also applied to 5G microstrip filters (operation frequency 3.53 GHz, bandwidth: 200 MHz, 20 dB reject zone up to 65.95 GHz), similar response had been achieved by simulations, which verified these methods can be applied to higher band applications.

5. Conclusion

Proposed BPF is achieved the characteristics of 20 dB stop-band attenuation up to $20f_0$. The resonant properties of proposed stepped impedance resonators, the influence of open-circuited stub and asymmetric coupling structures have been analyzed theoretically and verified by simulations. With extremely low $R_z$, the suppression of open-circuited stubs and asymmetric coupling structure on the harmonics have been enhanced, and the circuit size is reduced to $0.177 \times 0.082 \lambda_g^2$. As the theoretical calculated, simulated and measured results show, with extremely low $R_z$, open-circuited stubs have nearly no influence on fundamental resonance, its $Q_e$ and $K_0$, so the filter can be designed by conventional synthesis methods. As verified by simulations and measurement, the asymmetric coupling structure can be applied to attenuate the far field sophisticated harmonics, moreover, as the simulations show, these methods could be used to suppress the harmonics in higher band.

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