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Synthesis of X-band Trisection Bandpass Filters Using Hybrid Structures of $\lambda/4$ and $\lambda/2$ Resonators

Muhammad Sufian Anwar¹, Symeon Nikolaou², Hamza Nawaz³, Muhammad Ali Babar Abbasi²

¹Department of Electrical Engineering, University of Engineering and Technology,
Taxila, Pakistan

²School of Electrical Engineering, Frederick University, 1036 Nicosia, Cyprus

³Electronics Engineering Department, Shanghai Jiao Tong University, China

sufian-anwar@hotmail.com, *hamza_nawaz@hotmail.com

Abstract - This paper presents the design of compact cross-coupled microstrip trisection bandpass filters based on the hybrid structures of quarter-wavelength ($\lambda/4$) and half-wavelength ($\lambda/2$) resonators. The desired cross-couplings for transmission zeros (TZ) are realized by parallel coupled transmission line sections (J -inverter) or coupled via sections (K -inverter) of $\lambda/4$ resonators. The main line couplings on the other hand, are formed by mixed-coupled $\lambda/4$ and $\lambda/2$ resonators. In order to validate the proposed hybrid configurations, two X-band trisection filters with 6.5% and 6.2% fractional bandwidth (FBW) were designed, fabricated and measured. The first filter generates a TZ below the passband, while the second one produces a TZ above the passband. Minimum insertion loss of 1.7 dB and 1.5 dB were observed for the two trisection filters, respectively, while the hybrid configurations caused significant size reduction in the overall filter size. Moreover, the first spurious response for the proposed filter topologies was pushed to $3f_0$, resulting in a wide-stopband $-f_0$ is the passband center frequency. The measured and simulated S-parameters show good agreement, which proves the validity and accuracy of the proposed hybrid design methodology.

Key Terms- Bandpass filter, transmission zero, hybrid structure, trisection, coupling matrix

I. INTRODUCTION

Bandpass filters with good selectivity and high out-of-band rejection remain key components in modern RF front-end systems, in order to discard the undesired harmonics. Several recent studies use different forms of microstrip $\lambda/4$ or $\lambda/2$ resonators with cross couplings to achieve this selectivity. One commonly used method for the creation of finite TZs is based on alternative J/K inverters and $\lambda/4$ resonators [1, 2], where the first spurious response is detected at $3f_0$. Trisection filters with quarter-wavelength and half-wavelength topologies were developed in [3] and [4] respectively. A few studies published lately have tried to utilize the hybrid structures involving $\lambda/2$ or $\lambda/4$ resonators to get improved filter performance [5, 6].

The X-band filters reported in recent years generally use substrate integrated waveguide (SIW) structures [7]-[9], but they suffer from fabrication complexities due to the involvement of large via walls. The stopband suppression in SIW filters is limited to frequencies below $2f_0$. Likewise the designs proposed in [10, 11] require a demanding fabrication process as a result of the needed multiple substrate layers. A compact, flexible two pole X-band filter with low insertion loss was presented in [12], but it requires ultra-thin liquid crystal polymer substrate and shows narrow upper stopband characteristics (up to $2f_0$).

Considering the issues mentioned above, this paper proposes the design of considerably simple and fairly compact filters, exploiting the hybrid structures of $\lambda/4$ and $\lambda/2$ resonators in the X-band. The fabrication simplicity makes the proposed methodology very convenient for the X-band and for even higher frequency applications. To the best of our knowledge, the technique of using both $\lambda/4$ and $\lambda/2$ resonators for trisection filters is implemented for the first time in the current manuscript. The conventional trisection filters using only $\lambda/2$ or folded- $\lambda/4$ resonators [3, 4], are difficult to implement at high frequencies because the strip lengths at 10 GHz or higher, are quite small, and from manufacturing point of view it requires small impractical coupling gaps or tiny complex folded shapes. These fabrication limitations are overcome in this work with the introduction of two novel hybrid, 3rd-order filter configurations with asymmetric frequency response. These novel configurations, allow the direct control of TZs through manageable coupling gaps, variability in FBW and stopband selectivity $\geq 3f_0$. More importantly, the upper stopband of proposed filters provide high isolation with the upcoming 5G communication systems having the spectrum around 27.5-28.5 GHz [13]. Furthermore, the standard single layer microstrip technology and only two vias for the $\lambda/4$ strips, results in fairly easy and reliable fabrication when compared to multi-layer half-wavelength [10, 11] or multiple vias SIW filters [7]-[9], respectively.

II. NETWORK TOPOLOGY:

The general topology of asymmetric trisection filters with one cross coupling is depicted in Fig 1(a). The $\lambda/4$ and $\lambda/2$ (hair-pin) uniform impedance resonators (UIRs) used in this paper are shown in Figs. 1(b) and 1(c) respectively.

Trisection filters with asymmetric frequency selectivity can realize one TZ, either below or above the passband due to cross coupling. Specifically, inductive cross coupling creates a TZ below the passband while capacitive cross coupling produces a TZ above the passband [3, 4]. The other two couplings in these third order filters are equal magnitude mixed couplings [14]. Generally, the filters consisting of $\lambda/4$ UIRs when compared to $\lambda/2$ UIR filters exhibit smaller size due to smaller physical length, higher insertion loss due to vias and greater FBW due to the wide coupling spaces between the strips. The spurious responses for $\lambda/4$ resonators occur at the odd multiples of f_0 i.e. $f=(2n-1)f_0$, while for the $\lambda/2$ UIRs they occur at the integer multiples of f_0 i.e. $f=nf_0$ [14], where $n \in \{2,3,\dots\}$. The hybrid configurations proposed in this paper combine the advantages of both types of resonators, while keeping their related individual drawbacks to a minimum. The resultant structures are consequently made compact, low-loss and their first spurious response is shifted to $3f_0$ or beyond thus improving the filter's upper stopband performance.

The dielectric substrate chosen for the fabricated filters is Roger's RT/Duroid 5870 with $\epsilon_r = 2.33$, $\tan\delta = 0.0012$ and thickness 0.254 mm. Resonator widths in Figs. 1(b) and 1(c) were uniformly selected as $W_1=0.71$ mm ($Z_0=50 \Omega$) and $W_2=0.3$ mm ($Z_0=80 \Omega$) respectively, where Z_0 is the corresponding characteristic impedance of the microstrip transmission line. Another important parameter is the diameter (D) of the via-hole drilled through the one end of the $\lambda/4$ resonators (Fig. 1(b)). By varying D , the inter-resonator couplings as well as the resonant frequencies of $\lambda/4$ resonators can be changed [1]. In addition the external quality factor (Q_e) also depends on the via- diameter [1]. For the presented filters, $D=0.4$ mm was used everywhere.

III. DESIGN AND IMPLEMENTATION OF TRISECTION FILTERS:

A. Trisection Filters Design

Fig. 2 shows the proposed structure of a trisection filter with a TZ below the passband. It can be seen that inductive cross coupling (K-inverter) [2, 3], between $\lambda/4$ UIRs is formed and controlled by changing the spacing 'd1'. Mixed couplings are created by coupling the side arms of the hair-pin resonator with the open ends of the $\lambda/4$ UIRs as shown in Fig. 3(b). The first step in filter synthesis is the calculation of ' $N+2$ ' $\times$ ' $N+2$ ' coupling matrix [14, 15], containing the couplings between source-load terminations and resonator nodes, where ' N ' is the order of the filter. The next step is the calculation of filter dimensions, which are calculated by using LineCalc tool and full wave EM simulations in Advanced Design System (ADS). The coupling coefficients obtained from the coupling matrix can be estimated by carrying momentum simulations in ADS.

Figure 2 Proposed trisection filter with TZ below the passband

Fig. 3 plots the variations of the coupling coefficients with respect to the distances between the corresponding resonators for the filter presented in Fig. 2. These plots are generated by varying the distance between the two resonators under consideration, and by calculating the coupling coefficients ($M_{i,j}$) using the formula given in (1), where indexes 'i' and 'j' represent the resonator number [14].

$$M_{i,j} = \pm \frac{f_{p2}^2 - f_{p1}^2}{f_{p2}^2 + f_{p1}^2} \quad (1)$$

The frequencies f_{p1} and f_{p2} are the characteristic frequencies of the coupled resonators that correspond to the two resonant peaks [14], attained as a result of EM simulations.

Figure 3 VSWR and Return-loss of proposed Antenna Array

Fig. 4 shows the proposed structure of a second trisection filter with a TZ above the passband. Capacitive cross coupling (J-inverter) [1, 2] for this filter is attained by coupling the strip portions of the $\lambda/4$ UIRs opposite to vias, as shown in Fig. 5(a) whereas mixed couplings (Fig. 5(b)) are realized by coupling the center strip of the hair-pin resonator with the quarter-wavelength strips. Fig. 5 plots the coupling coefficients of the filter in Fig. 4 vs the inter-resonator distances using (1).

Figure 4 Proposed trisection filter with TZ above the passband

Figure 5 (a) Capacitive coupling coefficient ($M_{1,3}$) vs microstrip line lengths & distance between resonators (L_1 , L_2 & d). (b) Mixed coupling coefficient ($M_{1,2}=M_{2,3}$) vs distance between resonators, d .

B. External Quality Factor (Q_e) Calculation

Another important step in filter design is the calculation of the external quality factor (Q_e). Input and output (I/O) resonators are connected by a tapped microstrip line to external 50Ω terminations as can be seen in both Fig. 2 and Fig. 4. As a consequence, the width of the tapped strip (W), its length (L) as well as the feed position (d) (Fig. 6), they all play a significant role in determining Q_e . For simplicity, $W=0.4$ mm, is uniformly set for the filters presented in this paper. The other two variables are estimated from Fig. 6, which plots Q_e against ' d ' for two different values of ' L '.

Figure 6 External Quality Factor (Qe) vs the tapped microstrip line width, length and position (W, L & d).

The Q_e in Fig. 6 is extracted from the frequency response of a singly loaded I/O resonator [14]. Momentum simulations of the tapped resonator depicted in Fig. 6 are carried out in ADS. The phase response of the resulting reflection coefficient (S11) provides the data for Q_e calculation using the formulas given in (2) and (3).

$$Q_e = \frac{\omega_o}{\Delta\omega_{\pm 90^\circ}} \quad (2)$$

$$\Delta\omega_{\pm 90^\circ} = \Delta\omega_+ - \Delta\omega_- \quad (3)$$

ω_o in (2) corresponds to center frequency, whereas $\Delta\omega_+$ and $\Delta\omega_-$ in (3) correspond to $+90^\circ$ and -90° shifts from the absolute phase at ω_o in the S11 phase response, respectively.

C. Implementation of Trisection Filter with TZ Below the Passband

In this section the synthesis method for the proposed trisection filter with a TZ below the passband (Fig. 2) is applied to design a filter at 10.0 GHz. An ' $N+2$ ' $\times$ ' $N+2$ ' normalized coupling matrix for a filter with $f_o=10.0$ GHz, FBW=6.5%, return loss RL=20 dB and TZ at 9.23 GHz (-2.45j) is synthesized [14, 15] and given in (4).

$$.m1 = \begin{bmatrix} 0 & 1.17 & 0 & 0 & 0 \\ 1.17 & -0.11 & 0.95 & -0.45 & 0 \\ 0 & 0.95 & 0.44 & 0.95 & 0 \\ 0 & -0.45 & 0.95 & -0.11 & 1.17 \\ 0 & 0 & 0 & 1.17 & 0 \end{bmatrix} \quad (4)$$

The coupling coefficients $M_{i,j}$ and the external quality factor Q_e are derived from the coupling matrix (m) using (5) and (6) [14, 15].

$$M_{i,j} = m_{i,j} \times FBW \quad (5)$$

$$Q_e = \frac{m_{S,i}}{FBW} = \frac{m_{L,i}}{FBW} \quad (6)$$

where, $m_{i,j}$, $m_{S,i}$ and $m_{L,i}$ are the normalized inter-resonator, source-resonator and load-resonator coupling values given in matrix ‘m’ respectively, while indexes ‘i’ and ‘j’ represent the resonator numbers. The filter design parameters are summarized in Table I. The inductive dominant coupling $M_{1,3}$, calculated using (5), creates a TZ below the passband. Similarly, using (5) and (6), the mixed couplings $M_{1,2} = M_{2,3}$ and Q_e are computed and are also presented in Table I.

TABLE I
FIG. 2 TRISECTION FILTER DESIGN PARAMETERS

f_o (GHz)	FBW (%)	RL (dB)	TZs	Coupling Coefficients
10.0	6.5	20	$\Omega = -2.45j$ (9.23 GHz)	$M_{1,3} = -0.029$ $M_{1,2} = M_{2,3} = 0.062$
				$Q_e = 18$

TABLE II
OPTIMIZED PHYSICAL DIMENSIONS OF FILTER IN FIG. 2 (UNITS: MM)

Filter Size (without tapped feed lines) = $(0.17\lambda \times 0.31\lambda)$			
L_1	L_2	L_3	L_4
5.66	4.11	3.66	6.35
d_1	d_2	d_3	h_1
1.00	0.30	1.50	0.74

The S-parameter plots for the matrix ‘m1’ are given in Fig. 7, which verify the design parameters in terms of f_o , FBW, RL and TZ (Table I). An overall size of “ $0.17\lambda \times 0.31\lambda$ ” (Table II) verifies the claim for a very compact filter. In this case, λ is the microstrip guided wavelength on the used substrate at f_o . Table II illustrates the optimized physical dimensions of the fabricated filter. These dimensions are acquired from the plots in Fig. 3 and Fig. 6 while the resonator strip widths, via-hole diameter and tapped microstrip line width (W) are described in section II and section III-B. ADS Linecalc was used to initially compute the lengths of $\lambda/4$ and $\lambda/2$ resonators (L_1 , L_2 & L_3) at f_o , which were further adjusted using momentum simulations. Similarly, using Fig. 3 and momentum simulations, the coupling gaps d_1 , d_2 and h_1 were decided. Likewise, Fig. 6 and momentum simulations resulted in d_3 and L_4 .

Figure 7 S-parameter plots for Coupling Matrix 'm1'

Fig. 8 (a) shows a photograph of the filter portrayed in Fig. 2, which was fabricated using a standard chemical-etching procedure. Measured and simulated S-parameters are compared in Fig. 8(b). The reason for selecting a substrate with low height was that it allowed better through-hole plating for the via-hole resonators, which in turn improved the filter performance especially at higher frequencies.

(a)

(b)

Figure 8 (a) Fabricated trisection filter with TZ below the passband. (b) Measured and simulated S-parameters

All the measurements reported in this paper were taken using Keysight microwave network analyzer, PNA-N5224A. A minimum insertion loss of 1.7 dB is noted in the passband while at f_o its value is 2 dB. Fig. 8 (b) also incorporates a magnified plot displaying the passband zone for a more detailed inspection. The first meaningful spurious response in measured results appears at $3f_o$, where the S21 rejection is 7.3 dB (29.12 GHz), while below $3f_o$, a minimum rejection of 15.3 dB is detected at 19.6 GHz, as indicated in Fig. 8 (b). This extended out-of-band suppression, justifies the proposed synthesis approach, however it is further improved in the second filter design presented in the subsequent section. A slight increase in bandwidth, as well as a small frequency shift in TZ, can be observed in measured response compared to the simulated results. The measured FBW is 7% while the TZ appears at 9.06 GHz. These deviations can be caused by several unforeseen parasitic effects such as the fabrication tolerance, the PTH (through-hole plating) quality, or even a minor discrepancy in the exact ϵ_r value used for the momentum simulations. Overall the measured results are in good agreement with the simulated ones, something that validates the presented synthesis technique.

D. Implementation of Trisection Filter with TZ Above the Passband

This section describes the synthesis of the second proposed trisection filter with a TZ above the passband that was already portrayed in Fig. 4. A normalized source-load coupling matrix with $f_o=9.75$ GHz, FBW=6.2%, RL=20 dB, TZ=10.65 GHz (2.89j) is computed and revealed in (7).

$$m_2 = \begin{bmatrix} 0 & 1.17 & 0 & 0 & 0 \\ 1.17 & 0.09 & 0.98 & 0.38 & 0 \\ 0 & 0.98 & -0.37 & 0.98 & 0 \\ 0 & 0.38 & 0.98 & 0.09 & 1.17 \\ 0 & 0 & 0 & 1.17 & 0 \end{bmatrix} \quad (7)$$

TABLE III
FIG. 4 TRISECTION FILTER DESIGN PARAMETERS

f_o (GHz)	FBW (%)	RL (dB)	TZs	Coupling Coefficients
9.75	6.2	20	$\Omega=2.89j$ (10.65GHz)	$M_{1,3}=0.024$
				$M_{1,2}=M_{2,3}=0.061$
				$Q_e=18.9$

TABLE IV
OPTIMIZED PHYSICAL DIMENSIONS OF FILTER IN FIG. 4 (UNITS: MM)

Filter Size (without tapped feed lines) = $(0.45\lambda \times 0.25\lambda)$				
L_1	L_2	L_3	L_4	L_5
4.70	1.32	3.33	4.62	4.98

L_6	d_1	d_2	d_3	
2.42	0.46	0.25	1.54	

Table III provides the coupling coefficients extracted from ‘m2’ using (5) and (6). The capacitive coupling $M_{1,3}$ generates a TZ at 10.65 GHz, while $M_{1,2}=M_{2,3}$ and Q_e are the desired mixed coupling values and the external quality factor respectively. Fig. 9 shows the S-parameters for the coupling matrix ‘m2’, which correspond with the selected design parameters given in Table III. Table IV displays the optimized dimensions for the filter shown in Fig. 4 along with its overall size of “ $0.45\lambda \times 0.25\lambda$ ”. ADS Linecalc and momentum tuning were employed to get the strip lengths L_1 , L_2 , L_3 and L_4 at $f_0=9.75$ GHz, whereas Fig. 5 and Fig. 6 provide the parameters d_1 , d_2 , d_3 , L_5 and L_6 . Using these measurements, the filter was fabricated and its picture is shown in Fig. 10 (a). The measured and simulated results are compared in Fig. 10 (b), which also includes the zoomed image of the passband response. At f_0 , the value of insertion loss is 1.6 dB, while a minimum loss of 1.5 dB is observed. This filter has further improved stopband performance as compared to the design in the previous section. A rejection value of 30.8 dB is noted near $2f_0$, however the first meaningful spurious peak occurs beyond $3f_0$ at 33.1 GHz, where the rejection is 16.2 dB, as can be seen in Fig. 10 (b). The measured FBW is 6.3% and the TZ is detected at 10.43 GHz, which shows a downward shift by 220 MHz. A similar shift is witnessed in the filter’s center frequency as depicted in the magnified plot. The frequency downshift can be explained by the under-etching effect, which leaves the actual trace dimensions slightly larger than the designed dimensions. However, overall, the simulated and measured S-parameters show substantial agreement, which verifies the proposed synthesis method.

Figure 9 S-parameter plots for Coupling Matrix ‘m2’

Figure 10 (a) Fabricated trisection filter with TZ above the passband. (b) Measured and simulated S-parameters

IV. CONCLUSION

In this paper, a novel synthesis approach is presented to design two trisection X-band filters with asymmetric frequency selectivity, exploiting the hybrid structures of $\lambda/4$ and $\lambda/2$ resonators. The hybrid assembly makes it possible to develop filters exhibiting an appealing combination of compact size, low insertion loss and most importantly an extended upper stopband ($>3f_0=30\text{GHz}$) which provides high isolation with 5G applications (27.5-28.5 GHz). The conducted S-parameter measurements up to 40 GHz, verify the intended performance. It is important to stress that the presented design methods can be directly applied to implement higher order cascaded triplet filters and to design filters for even higher frequencies.

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