Lumped element balanced multi-band bandstop filter for ultra high frequency applications

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Abstract: An ultra-compact multi-band balanced (differential) bandstop filter (BSF) topology with all lumped elements has been introduced for the first time in this study. The proposed filter is a four-port structure where each symmetrical bisection is a series cascade of circles represent the source(S)/load(L), non-resonating node (NRN) and frequency (RF) and microwave communication systems by ended counterpart [6–8]. So far, many works of literature have been found to deal with balanced bandpass filter [9–14] but the balanced counterpart has not been able to draw much attention in the past applications [6, 15–17]. Another disadvantage is that all these configurations employ either open stub.

1 Introduction

Bandstop filter (BSF) plays an important role in many radio frequency (RF) and microwave communication systems by protecting the receiver from jamming by high power signals. The development of modern communication systems with multi-band services is attracting much attention for notch filters with multiple bandstop filters [1, 2]. For example, dual-band BSFs are often used in high power amplifiers and mixers due to their ability to suppress the double sideband spectrum [3, 4]. Especially, multi-band BSF with high stopband attenuation, low insertion loss, high selectivity and compact physical size is getting popularity in ultra high frequency (UHF) band applications as many RF systems operate in that frequency range [5].

Moreover, the demand of differential (balanced) filter topology is growing rapidly due to its higher immunity to environmental noise and crosstalk, reduced generation of transient noise and higher suppression of second-order non-linearities than its single-ended counterpart [6–8]. So far, many works of literature have been found to deal with balanced bandpass filter [9–14] but the balanced BSF has not been able to draw much attention in the past and current studies. Moreover, the existing limited work on differential notch filter only deals with the single-band response and it is difficult to generalise the developed architectures to high multi-band frequency response simultaneously [6, 15–17]. Another disadvantage is that all these configurations employ either microstrip or waveguide technology for which the circuit size for UHF band applications becomes very large. Moreover, the common mode (CM) response of the differential BSF designs in [15, 17] has very high CM insertion loss for which the CM rejection ratio (CMRR) value is not optimum. Additionally, the design procedures described in [6, 17] are complicated due to the application of substrate integrated waveguide resonators coupled to the opposite polarity transmission lines or double-sided parallel-strip lines. In [18], a proof of concept design of the planar multi-band balanced BSF is reported and it shows remarkable agreement between circuit simulation and experimental results (Fig. 1).

1.5 GHz is designed and fabricated. It offers a rejection level of about 25 dB for each DM stopband and maintains a CMRR value of about 25 dB for both bands. Moreover, it covers an area of only 0.003 λ2, where λ is the guided wavelength at the centre frequency between the two bands.

Fig. 1 Planar multi-band balanced BSF where, white, grey and yellow circles represent the source(S)/load(L), non-resonating node (NRN) and open stub(OS1, OS2, ..., OSK), respectively. Black lines and black circles are used to denote immittance inverters and half wavelength lines (L1, L2, ..., L6), respectively.

2 Filter topology

The topology of the K series cascaded multi-band cells in the proposed lumped element multi-band differential BSF is similar to the one reported in Fig. 1 [18], except it eliminates the use of open stub. The open stubs in Fig. 1 are used for optimum CM noise rejection but they do not affect the DM response of the filter. As the engineered multi-band balanced lumped element circuit offers the maximum CMRR just by possessing the symmetrical property of the circuit, there is no need to load any additional element. For UHF band applications, each immittance inverter used in Fig. 1 can be extrapolated to realise an arbitrary number of DM stopbands; (ii) it does not need any additional component in the symmetry plane to suppress the CM noise; and (iii) it occupies a very small area. For demonstration, a second-order dual-band differential BSF with resonant frequencies 1.15 and 1.36 GHz is designed and fabricated. It offers a rejection level of about 25 dB for each DM stopband and maintains a CMRR value of about 25 dB for both bands. Moreover, it covers an area of only 0.003 λ2, where λ is the guided wavelength at the centre frequency between the two bands.
be replaced by its equivalent lumped element π-circuit and each half-wavelength line can be represented by an LC parallel circuit. Finally, all the capacitors connected between any two nodes of the circuit can be greatly reduced and due to the implementation of the entire circuit as lumped elements, the overall size of the filter is very compact. To achieve satisfactory stopband performance, high-Q components should be used for the circuit implementation.

The proposed lumped element balanced multi-band BSF is presented in Fig. 2a, which is symmetrical to the horizontal central plane. Each symmetrical bisection is designed by cascading \( K \) \( N \)-band sections in series, through \( L_q C_q \), \( π \)-sections. The term ‘\( N \)-band section’ corresponds to the unit of \( N \) LC parallel circuits \((L_{gs}, C_{gs})\) which is responsible for producing \( N \) stopbands in DM operation. When the circuit is excited by a differential signal, the symmetry plane acts as a perfect electric wall, i.e. electrical short-circuit (Fig. 2b). Therefore, it results in a \( K \)th order \( N \)-band BSF response in DM operation (Fig. 2d). It is to be noted that \( L_{gs} \), \( C_{gs} \) values in the original circuit (Fig. 2a) becomes \( 0.5 L_{gs} \), \( 2C_{gs} \) in the DM equivalent circuit (Fig. 2b). The DM multi-band frequency transformation associated with this topology is given by [2]

\[
\Omega_{\text{DM} \text{pass}}(f) = - \sum_{i=1}^{N} \frac{1}{1/\Delta_i ((f/f_i) - (f/f_i))} \tag{1}
\]

where \( \Omega \) is the normalised lowpass frequency, \( N \) is the total number of stopbands, \( f \) is the frequency variable, \( f_i \) and \( \Delta_i \) are the centre frequency and bandwidth scaling factor of the \( i \)th stopband, respectively.

On the other hand, in case of CM operation, the central plane acts as a perfect electric wall, i.e. electric short-circuit (Fig. 2b). Therefore, the two-port bisection in CM operation can no longer produce those \( N \) stopbands as the DM equivalent circuit, rather exhibits a flat passband response (Fig. 2d).

The extraction of all elements in Fig. 2a will be discussed in the following section, starting with an example of dual-band balanced BSF.

### 3 Design example

To illustrate in detail, let us consider a second-order balanced dual-band BSF with the DM specifications: \( f_1 = 1.15 \) GHz, \( f_2 = 1.36 \) GHz and \( \Delta_1 = \Delta_2 = 6\% \). First, applying (1) with \( i = 1 \) and 2, a standard two-pole Chebyshev lowpass ladder network with shunt capacitances is transformed into the DM equivalent circuit which is a second-order dual-band BSF as shown in Fig. 3a [18]. Here, \( \text{Res}_1 \) (characteristic impedance \( Z_{\text{res}} \)) and \( \text{Res}_2 \) (characteristic impedance \( Z_{\text{res}} \)) are short-circuited quarter-wavelength resonators at \( f_1 \) and \( f_2 \), respectively, and \( J_{36}, J_{60}, J_{12} \) are the immittance inverters. These inverter-coupled quarter-wavelength resonators are equivalent to series LC resonators. Therefore, equating the input impedances at resonance, the relationships among different parameters are shown below, where \( g_6 = g_8 \) are the element values of the lowpass filter prototype [19–21].

\[
J_{12} = \frac{Z_0}{\sqrt{g_6 - g_8}} \tag{2}
\]

\[
Z_{\text{res}_1} = \frac{\sqrt{\Delta g_6}}{2\Delta_1} \tag{3}
\]

\[
Z_{\text{res}_2} = \frac{\sqrt{\Delta g_8}}{2\Delta_2} \tag{4}
\]
In Fig. 3b, $J_{21}$, $J_{11}$ and $J_{31}$ are replaced by their equivalent lumped element π-circuits and (9)–(13) are used to calculate them [19, 22, 23]. Usually, to achieve good selectivity in DM operation, the $L_2$ and $C_2$ values should be calculated for $J_{12} = 50\, \Omega$ [19]

$$L_2 = J_{12}/\omega_0$$  \hspace{1cm} (9) \\
$$C_2 = 1/(J_{12} \cdot \omega_0)$$  \hspace{1cm} (10) \\
$$C_{ci} = J_{12}/\omega_0$$  \hspace{1cm} (11) \\
$$C_{cj} = J_{12}/\omega_2$$  \hspace{1cm} (12) \\
$$\omega_0 = (\omega_1 + \omega_2)/2$$  \hspace{1cm} (13)

The negative capacitors will be absorbed into $C_2$, $C_{p1,DM}$ and $C_{p2,DM}$ to give the DM equivalent circuit in Fig. 3d. By setting $L_{p1,DM}$ (or $C_{p1,DM}$) and $L_{p2,DM}$ (or $C_{p2,DM}$) to any physically realisable values, scaling inverters $J_{si}$ and $J_{s2}$ are obtained. Similarly, $L_2$ and $C_2$ are calculated for $Z_0 = 50\, \Omega$. Also, $Z_{res}$ and $Z_{res}$ are generally set to 50. Therefore, the closed form equations of all the elements in Fig. 3d are

$$C_{ci} = \sqrt{\frac{\Delta g_i}{L_{p1,DM}Z_{res}/\omega_0}}$$  \hspace{1cm} (14) \\
$$C_{p1,DM} = \frac{1}{\omega_0 L_{p1,DM}} - C_{ci}$$  \hspace{1cm} (15) \\
$$C_{p2,DM} = \frac{1}{\omega_0 L_{p2,DM}} - C_{cj}$$  \hspace{1cm} (16) \\
$$L_2 = \frac{Z_0}{\omega_0 \Delta g_i}$$  \hspace{1cm} (17) \\
$$C_L = \frac{\Delta g_i}{Z_0/\omega_0} - C_{ci} - C_{cj}$$  \hspace{1cm} (19)

For the given specifications, the initial optimisation parameters of the DM equivalent circuit are calculated using (14)–(19) and the ideal differential circuit is shown in Fig. 4a. It should be noted that $C_{pi}$ and $C_{p2}$ values for this circuit are half of $C_{p1,DM}$ and $C_{p2,DM}$, respectively. Similarly, $L_1$ and $L_2$ values in this balanced circuit are twice of $L_{p1,DM}$ and $L_{p2,DM}$, respectively. Fig. 4b shows that DM stopband rejection (Sdd21) mainly depends upon the resonator $Q$ value whereas CM passband response (Scc21) remains unaltered for different $Q$. Here, as the resonator $Q$ changes from 50 to 200, the lower band rejection improves from 16.5 to 34 dB whereas the upper band rejection lies in the range of 20–38 dB. In comparison to that, 'Q's of both $C_{ci}$ and $C_{cj}$ have very low impact on the Sdd21 plot. Fig. 4c demonstrates that the DM stopband rejection for each band is only improved by 5 dB even when 'Q-values of these coupling capacitors are increased by four times. Similarly, the $Q$ of

$$L_{p1,DM} = \frac{\Delta g_i}{\omega_0 Z_{res}/\omega_0}$$  \hspace{1cm} (5) \\
$$C_{p1,DM} = \frac{Z_{res}/\omega_0}{\Delta g_i}$$  \hspace{1cm} (6) \\
$$L_{p2,DM} = \frac{\Delta g_i}{\omega_0 Z_{res}/\omega_0}$$  \hspace{1cm} (7) \\
$$C_{p2,DM} = \frac{Z_{res}/\omega_0}{\Delta g_i}$$  \hspace{1cm} (8)

Fig. 4  Dual-band differential BSF and effect of component $Q$ on frequency response  
(a) Initial optimisation parameters using the equations, (b) Effect of resonator ($L_{p1}$, $C_{p1}$ & $L_{p2}$, $C_{p2}$) Q, (c) Effect of coupling capacitor($C_{s1}$ and $C_{s2}$) Q, (d) Effect of π-circuit capacitor ($C_i$) Q

In Fig. 3b, Res1 and Res2 are replaced by their equivalent LC parallel resonators. The equations for calculating the elements of these parallel resonant circuits are [22]
C_L has almost no effect on both DM and CM responses of the proposed design (Fig. 4d).

Now, (14)–(19) can be generalised for extracting the components of the multi-band differential BSF structure in Fig. 2a as follows, where \( i = \{1, 2, \ldots, N\} \) and \( j = \{1, 2, \ldots, K\} \):

\[
L_d = \frac{Z_0}{\alpha 0 \sqrt{8 g k + 1}}
\]

(20)

\[
C_{L_d} = \begin{cases} \frac{\Delta L_d}{0.5 L_0^p Z_{res}^0} & \text{if } i \text{ is odd} \\ \frac{\Delta L_d}{0.5 L_0^p Z_{res}^0} & \text{if } i \text{ is even} \end{cases}
\]

(21)

\[
C_{Li} = \begin{cases} \frac{2 \sqrt{8 g k + 1}}{Z_{p0}} \sum_{j=1}^{N} C_{Li} & \text{if } j = 1, K \\ \frac{2 \sqrt{8 g k + 1}}{Z_{p0}} \sum_{j=1}^{N} C_{Li} & \text{if } j \neq 1, K \end{cases}
\]

(22)

\[
C_{Cl} = \frac{0.5 (10.5 \alpha \beta Z_{p0} - C_{cl})}{0.5 Z_{p0} Z_{res}^0}
\]

(23)

4 Results and discussion

For demonstration, the abovementioned dual-band balanced tunable BSF is built on Rogers RO4003 with a substrate thickness of 1.52 mm, the dielectric constant of 3.38 and loss tangent of 0.0027 @ 10 GHz (Fig. 5). This implemented circuit uses 0402 series wirewound inductors from Johanson Technology, which have a \( Q > 55 \) at 1 GHz [24]. For the capacitors, ‘Johanson 0805 S-series’ is employed which exhibits a \( Q > 900 \) at 1 GHz [25]. The design is simulated in NI/AWR Microwave office and the fabricated model is characterised using N5224 4-port PNA. Fig. 6 demonstrates that the simulation and the measured results agree to a great extent. As expected, the centre frequencies of the DM stopbands are found to be 1.15 and 1.36 GHz, respectively. The stopband rejection for both bands in DM operation (Sdd21) is close to 25 dB whereas Scc21 plot is a flat 0 dB passband response. The FBW of the lower band is measured to be about 6.3% whereas FBW of the upper band is 6.1%. On the other hand, the passband portion of the DM response shows an insertion loss of <1 dB. This insertion loss mainly depends upon the resonator \( Q \), as shown in Fig. 4b. Therefore, it can be improved by using high-\( Q \) components for the parallel resonators in the circuit. It will also help to get a better in-band return loss (Sdd11 in Fig. 6b) for each differential band.

Finally, Table 1 illustrates the comparison between the engineered topology and the differential BSF designs already published. The second and third rows present the DM stopband frequency/frequencies and check whether the design can be extrapolated to realise multiple stopbands simultaneously. It indicates that the proposed design is one of the only two designs which can produce the arbitrary number of DM stopbands simultaneously. The fourth row shows that it exhibits better stopband rejection level than all other designs except [1]. As seen from the fifth row, this engineered filter is one of those designs with the minimum DM return loss. The sixth row indicates that it is one of the three designs which maintains a flat 0 dB CM passband response to deliver the optimum CMRR (seventh row), which is same as DM stopband rejection level. The last row depicts the most significant attribute of this work. The proposed filter in this work occupies the smallest area among all the designs in the table.

5 Conclusion

A multi-band balanced BSF architecture with all lumped elements is presented for the first time in this paper for UHF band applications. Each symmetrical bisection of this four-port ultra-compact structure consists of \( K \) series cascaded \( N \)-band cells to realise the \( K \)-th order \( N \)-band BSF response in DM operation. The effect of CM noise is inherently removed due to the symmetrical property of the structure. To illustrate the proposed topology, a second-order balanced dual-band BSF with centre frequencies 1.15 and 1.36 GHz is designed and fabricated. The FBW for each band is found to be about 6% and both bands maintain a CMRR value of about 25 dB. The simulation results of the optimised layout and measured results from the fabricated prototype are compared to show a good match with each other. Compared to the other differential BSF designs, the proposed filter has the advantages of compact size, the flexibility of realising the arbitrary number of stopbands in a great extent. As expected, the centre frequencies of the DM stopbands are found to be 1.15 and 1.36 GHz, respectively. The stopband rejection for both bands in DM operation (Sdd21) is close to 25 dB whereas Scc21 plot is a flat 0 dB passband response. The FBW of the lower band is measured to be about 6.3% whereas FBW of the upper band is 6.1%. On the other hand, the passband portion of the DM response shows an insertion loss of <1 dB. This insertion loss mainly depends upon the resonator \( Q \), as shown in Fig. 4b. Therefore, it can be improved by using high-\( Q \) components for the parallel resonators in the circuit. It will also help to get a better in-band return loss (Sdd11 in Fig. 6b) for each differential band.

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**Table 1** Comparison of various balanced BSFs

| Feature                                    | [6][a] | [15] | [16] | [17][a] | [18] |
|--------------------------------------------|--------|------|------|---------|------|
| DM centre frequencies, GHz                 | 2.4    | 2    | 2.45 | 0.83    | 1.22/1.68 |
| simultaneous realisation of multiple DM stopbands | No      | No   | No   | Yes     | Yes  |
| DM stopband rejection, S(dB)21, dB         | 50     | 18.89 | 15   | 20      | 26/21 |
| DM in-band return loss, S(dB)11, dB        | 2      | 2.5  | 2    | 5       | 2    |
| CM insertion loss, Scc21 (dB) within 3 dB bandwidth | 2.5    | 50.45 | 0    | 15      | 0    |
| [CMRR], dB                                 | 47.5   | 31.56 | 15   | 5       | 26/21 |
| effective size ($\lambda g^2$)             | 0.06   | 0.34 | 0.03 | 0.04    | 0.31 |

[a] $\lambda g$ denotes guided wavelength at centre frequency of all DM stopbands. [6, 17] are tunable designs and all the parameters are analysed at the centre frequency of the tunable range.

DM stopbands, low in-band return loss and satisfactory CMRR for each band. Further improvement in DM stopband response and CMRR value can be achieved by using high-Q inductors and capacitors in the circuit.

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