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Design of Multiple-band Microwave Filters Using Cascaded Filter Elements

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Abstract: In this paper, a new approach to design multiple-band microwave filters is presented. A transfer function of a cascade of a low pass prototype filter with one or more band reject filters is calculated. This transfer function is manipulated to enable its application to a single coupled-resonator filter. This method can be applied to design symmetric and asymmetric multiple-band filters.

1 Introduction

Recently, there has been an increasing demand on microwave filters that exhibit multiple band characteristics. A multi-band bandpass function can be achieved by optimization of the coupling matrix as in references [1]-[5]. However, optimization algorithms do not yield optimal solutions in all cases. This can be clearly observed when the complexity of the filters increases, such as in the case of an asymmetric response or higher order filters. The algorithm tends to converge to a local minimum instead.

Analytical methods to achieve multi-band filtering functions are usually limited to dual-band cases as in [6], [7]. Frequency transformation is used in [8] to achieve a triple-band filter. In [9], the authors have introduced new technique to design multiple band filters using lumped elements.

One way of designing a dual-band filter is to cascade a wide bandpass filter with a narrow stopband filter [10]. The main drawback of this method is the requirement for two different filter circuits. This results in a large size for the overall application, and added complexity as it requires simultaneous tuning for both of the circuits. However, the transfer function for the cascaded system can be manipulated so that it can be implemented in a single filter circuit as described here.

In this paper, a new methodology to design multi-band filters is introduced. The methodology can be applied in symmetric and asymmetric multi-band cases. The technique aims to find the filtering functions for multi-band filters by starting from cascade of a lowpass prototype filter with at least one bandreject filter. The filtering functions for the combination can be found and then manipulated to be applied to a single filter circuit. The design methodology for symmetric and asymmetric dual-band filters are introduced in the following sections. The design method can be easily generalized to incorporate filters with more passbands.

2 Symmetrical dual-band filters

In order to design a multi-band bandpass filter, a wideband bandpass filter can be cascaded with smaller band bandreject filter. Initially we will discuss a symmetric case where the stopband is in the middle of two passbands of equal bandwidth (see fig. 1). Using frequency transformation to the lowpass prototype frequency, the bandpass is transformed into a lowpass filter with cutoff of unity (Ω_A = 1). While the bandreject filter is transformed to highpass filter with cutoff frequency (Ω_B << 1).

The scattering parameters of for the lowpass filter are given by [11]

$$S_{11}^{A}(\Omega)=S_{22}^{A}(\Omega)=\frac{F_A(j\Omega)}{\varepsilon_{rA}E_A(j\Omega)},\ \text{and}\ \ S_{21}^{A}(\Omega)=S_{12}^{A}(\Omega)=\frac{P_A(j\Omega)}{\varepsilon_{A}E_A(j\Omega)}. \quad (1)$$

For the high, pass filter, the scattering polynomial (P_B, F_B, and E_B) can be calculated for a prototype highpass filter [4] and then transform by frequency transformation $\Omega \rightarrow \Omega\Omega_B$ to achieve

$$S_{11}^{B}=S_{22}^{B}=\frac{P_B(j\Omega)}{\varepsilon_{B}E_B(j\Omega)},\qquad \text{and}\qquad S_{21}^{B}=S_{12}^{B}=\frac{F_B(j\Omega)}{\varepsilon_{rB}E_B(j\Omega)}. \quad (2)$$

The scattering parameters for the cascade can be calculated using [12]

$$S_{21}=\frac{S_{21}^{A}S_{21}^{B}}{1-S_{22}^{A}S_{11}^{B}}, \quad (3)$$

$$S_{11}=\frac{S_{11}^{A}-\Delta_{s1}S_{11}^{B}}{1-S_{22}^{A}S_{11}^{B}}, \quad (4)$$

$$\Delta_{s1}=S_{11}^{A}S_{22}^{A}-S_{12}^{A}S_{21}^{A}. \quad (5)$$

But as $S_{22}^{A}S_{11}^{B}\approx 0$, this yields

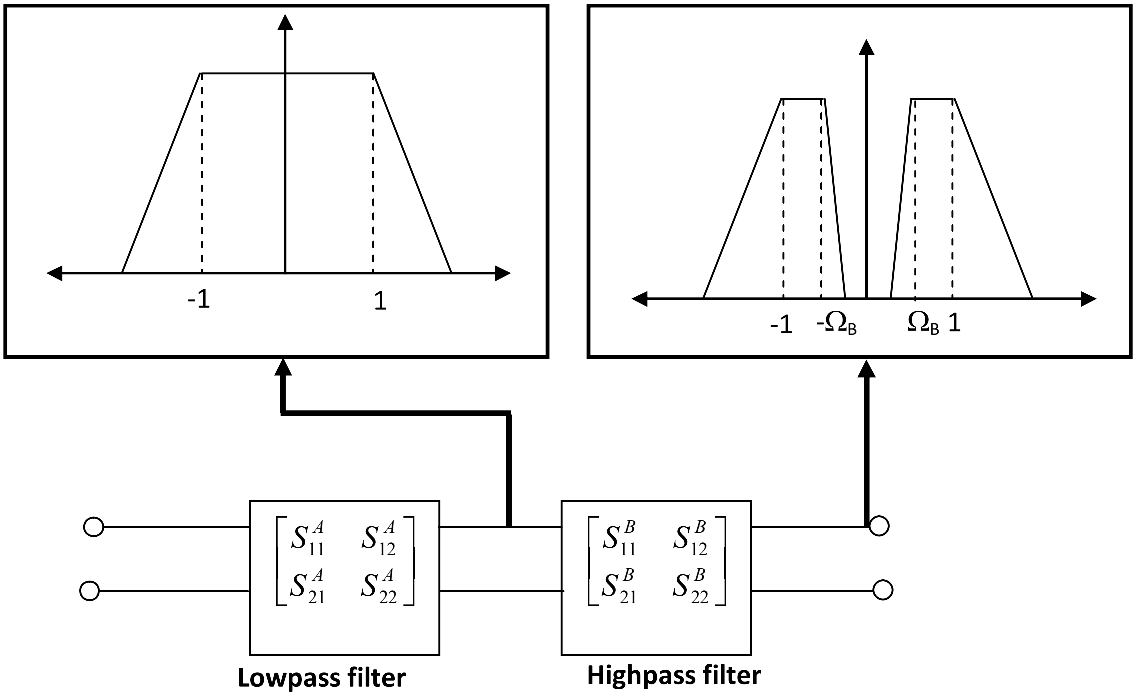


Fig. 1. The cascaded system at lowpass prototype frequencies

$$S_{21} \approx S_{21}^A S_{21}^B, \tag{6}$$

$$S_{11} \approx S_{11}^A - \Delta_{s1} S_{11}^B. \tag{7}$$

The polynomials for the insertion loss of the resultant network can be calculated simply using the polynomials of the cascaded networks. From (1), (2), and (6)

$$S_{21} = S_{12} = \left(\frac{P_A(j\Omega) F_B(j\Omega)}{\varepsilon_A \varepsilon_{rB} E_A(j\Omega) E_B(j\Omega)} \right) = \frac{P(j\Omega)}{\varepsilon E(j\Omega)}, \tag{8}$$

Where *P*, *F, E* and ε are the filters polynomials and the ripple factor, respectively. As a result,

$$P(j\Omega) = P_A(j\Omega) F_B(j\Omega), \tag{9}$$

$$E(j\Omega) = E_A(j\Omega) E_B(j\Omega) \text{ and,} \tag{10}$$

$$\varepsilon = \varepsilon_A \varepsilon_{rB}. \tag{11}$$

Substituting (1) and (2) in (7), it can be easily proved that the denominators of (6) and (7) are not equal. This means F for the equivalent network cannot be derived directly from the numerator of (7). However, the polynomials $E(j\Omega)$, $P(j\Omega)$ and $F(j\Omega)$ are interdependent and any two are sufficient to find the third one by Starting from the energy conservation formula

$$|S_{11}|^2 + |S_{21}|^2 = 1 \tag{12}$$

And substituting with

$$|F(j\Omega)|^2 = |E(j\Omega)|^2 - |P(j\Omega)|^2 / \varepsilon^2. \tag{13}$$

Filter Example

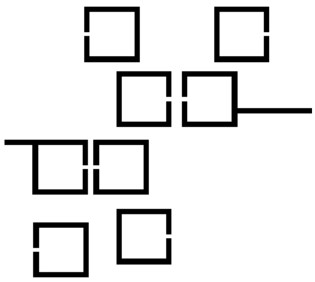


Fig. 2.The layout of the dual-band symmetrical band filter

The filter has the requirement

1) First passband: 4.9- 4.97GHz
2) Second passband 5.03- 5.1 GHz

The maximum reflection loss of the two passbands is to be -20 dB, while the maximum insertion loss in the middle stopband is to be -30 dB.

The specifications can be realized by cascading a four-pole bandpass filter with a bandwidth of 200 MHz and centre frequency of 5 GHz and a four-pole stopband filter with a bandwidth of 65 MHz and the same centre frequency. The lowpass prototype frequencies corresponding to the filter passbands are found to be [-j1.00 j0.33] Hz and [j0.33 j1.00] Hz.

To achieve a better roll-off factor for the stopband filters, four reflection-zeros have been inserted at [-*j*0.4466,-*j*1.1517, +*j*0.4466, +*j*1.1517] low pass prototypes frequencies. Using equations (9), (10), (11), and (13) the polynomials *E, F* and *P* are found and shown in Table 1.0

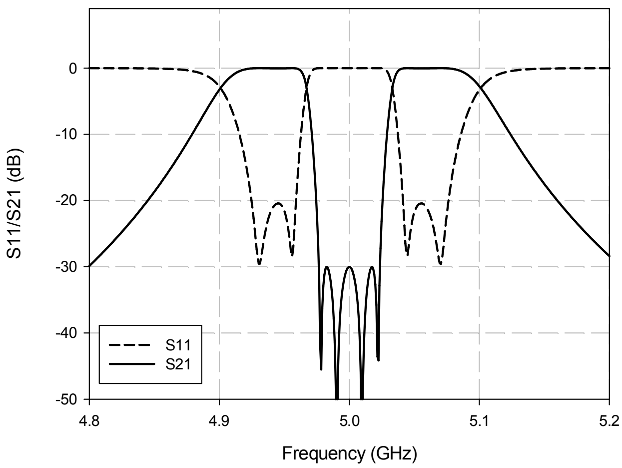
The achieved transmission/reflection losses are depicted in Fig. 2. After achieving the transfer functions (S_{11} and S_{12}), the general coupling matrix can be constructed using the synthesis method developed in [11].

TABLE 1: SINGULARITIES FOR AN EIGHT-POLE SYMMETRIC DUAL-BAND FILTER

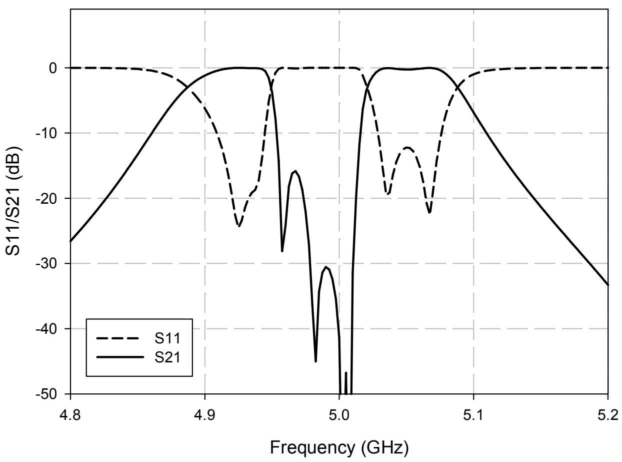
Transmission zeros	Reflection zeros	Transmission/reflection poles
+ 0.219917 *j*	0.026074 - 0.699284 *j*	-0.255564 - 0.924416 *j*
- 0.219917 *j*	0.286063 - 0.453411 *j*	-0.255564 + 0.924416 *j*
+ 0.097025 *j*	0.471245 - 0.186472 *j*	-0.616987 - 0.382906 *j*
- 0.097025 *j*	0.0167 - 0.439099 *j*	-0.616987 + 0.382906 *j*
	0.471245 + 0.186472 *j*	-0.066454 - 0.326679 *j*
	0.026074 + 0.699283 *j*	-0.066454 + 0.326679 *j*
	0.286063 + 0.453411 *j*	-0.291316 - 0.211385 *j*
	0.0167 + 0.439099 *j*	-0.291316 + 0.211385 *j*

$\varepsilon = 2.0683$.

The required four transmission zeros can be realized with two cascaded quadruplet filters (see Fig. 2). The general coupling matrix is synthesized to achieve the required



(a)



(b)

Fig. 3: Reflection and insertion losses for the symmetrical dual band filter (a) from coupling matrix (b) simulation results

topology following method used in [13]. Table 2 gives a solution for the coupling matrix. Input and output external quality factors take values of 1.4249 and 0.6559, respectively.

Fig. 2 shows a layout of the 8-pole canonical filter. Hairpin resonators have been chosen to achieve the 3 % bandwidth at 5 GHz. The resonator dimensions at 5 GHz have been found to be 3.5 mm × 3.5 mm, the gap width is 0.3 mm and resonator line-width is 0.3 mm. The dielectric height and the dielectric constant are assumed to 0.5 mm and 9.65, respectively.

The filter design was carried out using the coupled-resonators design methodology [14].

The filter has been simulated and optimized using Sonnet software [15] with a resolution of 50 μm × 50 μm. The simulation results are depicted in Fig. 3(b).

TABLE 2: COUPLING MATRIX FOR AN EIGHT-POLE SYMMETRIC DUAL-BAND FILTER

	1	2	3	4	5	6	7	8
1	0	0.02452	0	-1.23047	0	0	0	0
2	0.02452	0	0.22393	0	0	0	0	0
3	0	0.22393	0	0.40565	0	0	0	0
4	-1.23047	0	0.40565	0	0.487029	0	0	0
5	0	0	0	0.487029	0	0.5523	0	-0.43155
6	0	0	0	0	0.5523	0	0.01292	0
7	0	0	0	0	0	0.01292	0	0.55957
8	0	0	0	0	-0.43155	0	0.55957	0

3 Asymmetrical dual-band filters

For an asymmetrical dual-band filter, the centre frequency of stopband and passband filters are not the same, as seen in fig. 4. At normalized lowpass frequencies, the stopband filter should have centre frequency of $\Omega_y \neq 0$. In this case, the same technique of design as used in the dual-band symmetrical filter can be followed. A frequency shift can be implemented to move the highpass filter to a centre frequency of Ω_y. The polynomials of the highpass filter can be shifted along the frequency axis by using the transformation $\Omega' = \Omega - \Omega_y$.

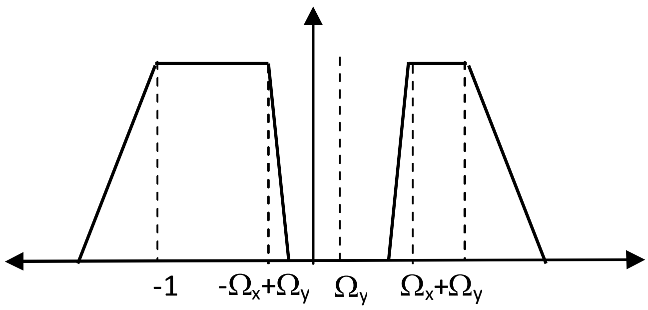


Fig . 4 : General transmission-loss response for an asymmetric dual-band filter

One example of an asymmetrical filter is given in the following section. The specification is based on the US mobile communications band for operator B [16].

Filter example

A 12th order filter is used to realize two asymmetrical passbands extending across the lowpass prototype frequency ranges of [-*j*1.00 +*j*0.43] and [+*j*0.64 +*j*1.0]. The maximum reflection loss within the passbands is to be -22 dB and the maximum insertion loss within the stopband is to be -50 dB. The filter can be implemented by cascading an 8-pole bandpass filter with a 4-pole bandstop filter. The stopband filter has been designed to have cutoff frequency of Ω_x = 0.105 Hz and the centre of the asymmetric stopband is at 0.537 Hz. To achieve better rejection performance, reflection zeros at [-1.46*j* -0.122*j* 0.543j 1.86*j*] Hz have been introduced.

Using equations (9), (10), (11), and (13), the polynomials *E, F* and *P* for the resultant filter are given in Table 3.

The topology of the filter is to be four cascaded triplets as shown in Fig. 5. The filter polynomials in Table 3 are used to build the full coupling matrix which is then synthesized to the desired topology and given in Table 4. The input and output external quality factors are given by 0.9326 and 1.2434, respectively. The filter simulated response is depicted in Fig. 6.

4 Conclusion

A methodology for synthesizing multiple-band passband filters has been presented. The multiple-band performance is realized by cascading a combination of bandpass and stopband filters. The transfer function of the system is manipulated to be implemented in a single cross-coupled filter. All the resultant passbands have the same maximum reflection-loss which corresponds to the passband filter. Each stopband has the maximum insertion loss of the corresponding stopband filter. The bandwidth of the passbands can be achieved accurately by controlling the stopbands of each stopband filter. The algorithm provides sufficient results when the stopbands are smaller than the passbands. For wider stopbands, either a higher order can be assigned for the stopband filters or reflection zeros can be used to produce a steeper stopband roll-off.

For a quadruplet-passband filter or higher, the requested filter order increases significantly which makes the filter awkward to be realized practically. Examples of symmetric and asymmetric dual-band filters are given with full coupling matrices. A complete example has been given where the filter have been designed up to the simulation level. Although these filters have not been constructed, previous experience with building multiple-passband filters with HTS material [3] has shown the simulation agrees well with experimental measurements.

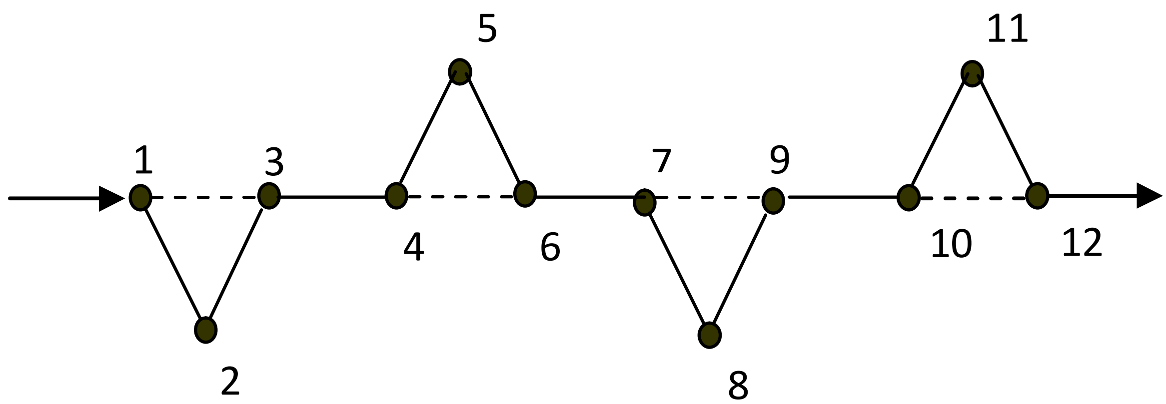


Fig. 5. Topology used to realize the asymmetrical dual passband response. The three cross couplings enable realisation of 4 transmission zeros in the filter's stopband.

TABLE 3: SINGULARITIES FOR A 12-POLE ASYMMETRIC DUAL-BAND FILTER

Transmission zeros	**Reflection zeros**	**Transmission/reflection poles**
0.577521 *j*	-0.00057 - 0.890656 *j*	-0.073354 - 0.963979 *j*
0.554205 *j*	-0.001972 - 0.755069 *j*	-0.208895 - 0.817222 *j*
0.496572 *j*	-0.00447 - 0.504559 *j*	-0.312634 - 0.54605 *j*
0.520336 *j*	-0.01067 - 0.177464 *j*	-0.368777 - 0.191747 *j*
	-0.007457 + 0.891665 *j*	-0.073354 + 0.963979 *j*
	-0.144701 + 0.640154 *j*	-0.208895 + 0.817222 *j*
	-0.178182 + 0.531638 *j*	-0.368777 + 0.191747 *j*
	-0.032795 + 0.752282 *j*	-0.312634 + 0.54605 *j*
	-0.030362 + 0.702766 *j*	-0.03037 + 0.63743 *j*
	-0.147415 + 0.413878 *j*	-0.099852 + 0.591864 *j*
	-0.031837 + 0.369814 *j*	-0.102619 + 0.48532 *j*
	-0.031092 + 0.174185 *j*	-0.033138 + 0.43402 *j*

$\varepsilon = 22.055$

TABLE 4: COUPLING MATRIX FOR A 12-POLE ASYMMETRIC DUAL-BAMD FILTER

	1	2	3	4	5	6	7	8	9	10	11	12
1	0.10867	0.0171	-0.7375	0	0	0	0	0	0	0	0	0
2	0.0171	-0.4995	0.2526	0	0	0	0	0	0	0	0	0
3	-0.7375	0.2526	0.0088	0.5047	0	0	0	0	0	0	0	0
4	0	0	0.5047	0.107	0.12025	-0.4797	0	0	0	0	0	0
5	0	0	0	0.12025	-0.60058	0.198	0	0	0	0	0	0
6	0	0	0	-0.4797	0.198	0.0688	0.468	0	0	0	0	0
7	0	0	0	0	0	0.468	0.0248	0.1694	-0.5363	0	0	0
8	0	0	0	0	0	0	0.1694	-0.5519	0.1026	0	0	0
9	0	0	0	0	0	0	-0.5363	0.1026	-0.0078	0.5982	0	0
10	0	0	0	0	0	0	0	0	0.5982	-0.0603	0.1531	-1.0318
11	0	0	0	0	0	0	0	0	0	0.1531	-0.5839	0.05375
12	0	0	0	0	0	0	0	0	0	-1.0318	0.0537	-0.1541

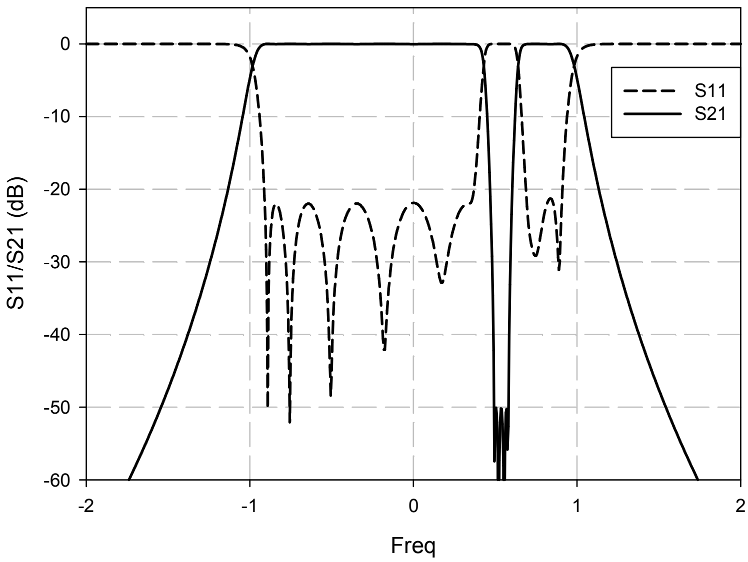


Fig. 6.: Simulated transmission and reflection losses for a 12-pole dual-band asymmetric filter from the coupling matrix.

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