DUAL BAND FILTER DESIGN USING REAL FREQUENCY TECHNIQUE AND FREQUENCY TRANSFORMATION

Serkan YILDIZ[1], Ahmet AKSEN[2], Sıddık B. YARMAN[3]

[1]Department of Electronics and Communication Engineering, Istanbul Technical University, Turkey
[2]Department of Electrical-Electronics Engineering, Işık University, Turkey
[3]Department of Electrical-Electronics Engineering, Istanbul University, Turkey
seryildiz@itu.edu.tr, ahmet.aksen@isikun.edu.tr, yarman@istanbul.edu.tr

Abstract: *In this work, we present a frequency transformation based dual-band filter design approach using scattering based real frequency technique. In the design process, conventional low-pass to band-pass frequency transformation is integrated with the simplified real frequency technique to construct double pass-band filters. The approach is particularly advantageous for designing matching filters between different termination resistances. As an alternative to a direct low-pass to dual-band frequency transformation, the use of dual-band mapping on a normalized band-pass prototype is investigated for efficient control of the passbands. Application of the proposed approaches for dual-band filters is presented by comparative design examples.*
Keywords: *Dual-band filters, frequency transformation, real frequency technique, band-pass prototype.*

1. Introduction

Nowadays, frequency spectrum is allocated by many wireless communications protocols such as GSM, UMTS, WiMAX and communication devices are asked to perform concurrently in multiple protocols. Therefore, multiband systems providing many advantages has gained importance in this track. Multiband filtering is one of the major issues in communication subsystems. Design of multiband filters are well elaborated in literature [1]. One of the major approaches uses transmission zeros on a band-pass filter response and splits frequency bands into sub-bands [2]. In this case, transmission zero producing elements are generally synthesized as Brune sections which are difficult to realize in lumped or distributed domain [3]. Different multiband implementation techniques have been extensively investigated in the literature [4-6].

In most of the design methods, major aim is the transfer function based analytical characterization of multiband filters [7]. A major track in this approach is the use of frequency transformation on a prototype network. In literature, there are many studies that use this approach [8-13]. The usual application in this method is the employment of sequential low-pass (LP) to band-pass (BP) transformation on a low pass prototype filter transfer function to result in multiband characteristic. However from the network realization point of view, this doubles the element count of the prototype after each application of LP to BP transformation to create more pass bands [14].

In this study, the conventional LP to BP transformation approach is taken as the basis of dual-band design approach. Development of an integrated tool for real frequency based design of dual band matching filters is aimed along with proper frequency mapping from prototype designs. For efficient control of the passbands in a double band design, traditional frequency transformation approach is modified and the design process is integrated with the simplified real frequency technique (SRFT).

This paper is an extended version of our work presented in ELECO2016 [13]. Based on the SRFT design of double passband filters, the use of frequency transformation approach and application of frequency mapping on two reference prototype networks are presented. First method involves two sequential transformation step on a low pass prototype network. Second method includes construction of a transformerless BP prototype network and single transformation step on this prototype to result in dual band filter characteristic. In both of the approaches, dual band filter characteristic is aimed with transformer free realizations.

In Section 2, LP to dual band frequency transformation is introduced. In Section 3, real frequency technique based design of dual band filters is discussed and two design approaches are presented. In Section 4, comparison of the

Received on: 13.03.2017
Accepted on: 31.05.2017

proposed approaches are presented along with dual band filter design examples.

2. Low Pass to Dual Band Frequency Transformation

Dual band filter characteristic can be generated via sequential application of the well-known LP to BP transformation on a low pass prototype network as shown in Fig. 1.

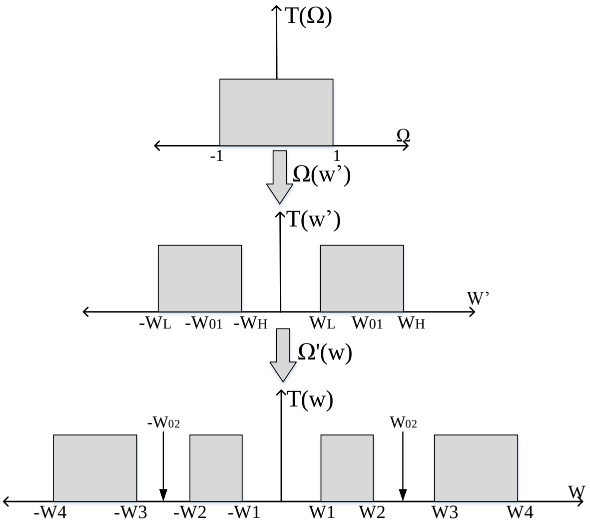

Figure 1. Low pass to dual-band transformation

A normalized LP prototype characteristic is first mapped to a BP response by applying the transformation,

$$\Omega(w') = \frac{w'^2 - w_{01}^2}{B_1 w'}, \quad B_1 = w_H - w_L, \quad w_{01} = \sqrt{w_L w_H} \tag{1}$$

Then, the resulting BP characteristic is normalized by the upper cut off w_H and LP to BP transformation is reapplied as

$$\Omega'(w) = \frac{w'}{w_H} = \frac{w^2 - w_{02}^2}{B_2 w}, \quad w_{02}^2 = w_1 w_4, \quad B_2 = w_4 - w_1 \tag{2}$$

Instead of this sequential process, a direct compact formula for LP to DB transformation can be obtained by combining (1) and (2) as follows:

$$\Omega(w) = \frac{w^4 - (2w_{02}^2 + w_{01}^2 B_2^2)w^2 + w_{02}^4}{B_1 B_2 (w^3 - w w_{02}^2)} = \frac{1}{B_2 B_1}\frac{w^2 - w_{02}^2}{w} - (B_2 w_{01})^2 \frac{w}{w^2 - w_{02}^2} = \frac{1}{B_2 - B_S}\left(\frac{w^2 - w_{02}^2}{w}\right) - \frac{B_2 B_S}{B_2 - B_S}\left(\frac{w}{w^2 - w_{02}^2}\right). \tag{3}$$

where $B_S = w_3 - w_2$. The frequency mapping diagram of the direct LP to DB transformation given (3) is shown in Fig. 2.

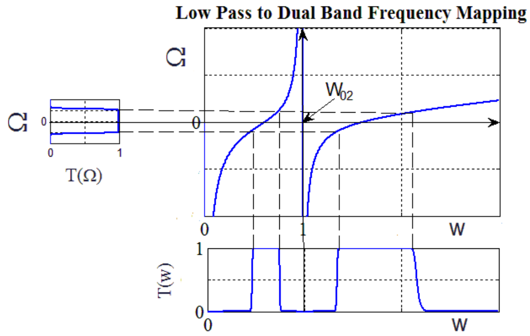

Figure 2. Low-pass to dual-band frequency mapping

As the mapping diagram implies, the transformation results in two passbands of unequal bandwidths which are geometrically symmetric around a stopband center w_{02}. The pass-band ripple characteristic of LP response is preserved through the transformation process. LP to DB mapping results in the following parameter relations;

$$w_{02} = \sqrt{w_1 w_4} = \sqrt{w_2 w_3}, \quad w_1 < w_2 < w_{02} < w_3 < w_4, \quad w_{01} = \sqrt{B_s / B_2}, \quad B_1 = 1 - B_s / B_2 \tag{4}$$

In the LP to DB transformation, three band edge frequencies of the DB response can be chosen independently, while the fourth one is set by (4), under geometric symmetry constraints of the mapping. In this case, selecting any three of the band edge frequencies, the mapping provides a direct definition of the transformation parameters B_2, B_S, w_{02}.

For example, if the normalized cut-off requirements are defined as $w_1 = 1, w_2 = 1.5, w_3 = 2$ then $w_4 = 3$ and the associated transformation parameters are obtained as $B_2 = 2, B_S = 0.5, w_{02} = 1.732$. With these parameters, if LP to DB transformation is applied on a Chebyshev LP prototype, the resulting transfer function magnitude is as shown in Fig. 3.

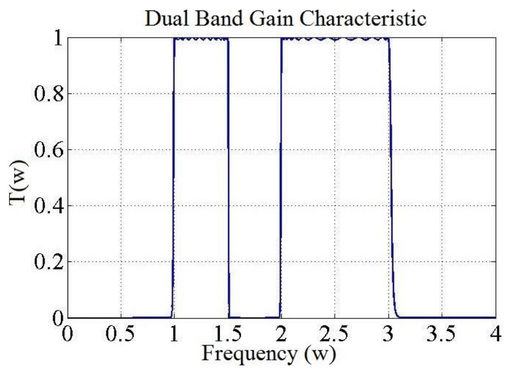

Figure 3. Gain response produced with LP to DB mapping

3. Scattering Based Real Frequency Technique for Dual-Band Filter Design

Simplified Real Frequency Technique (SRFT) is one of the well-known network design tools particularly suitable for matching problems [15]. SRFT is based on scattering parameter representation of a lossless two-port network. Consider the filter problem shown in Fig. 4.

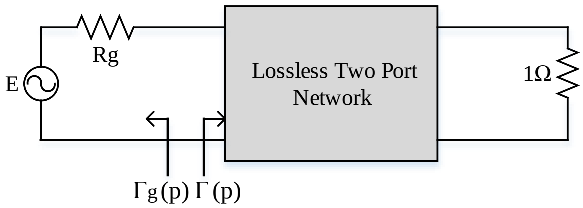

Figure 4. Lossless filter two port

Front-end junction reflection functions can be expressed as

$$\Gamma(p) = \frac{h(p)}{g(p)} = \frac{h_n p^n + h_{n-1} p^{n-1} + \ldots + h_1 p + h_0}{g_n p^n + g_{n-1} p^{n-1} + \ldots + g_1 p + g_0}$$
$$\Gamma_g = \frac{R_g - 1}{R_g + 1} \qquad (5)$$

Scattering parameters describing the two-port are defined in Belevitch canonic form as [16]

$$S_{11}(p) = \frac{h(p)}{g(p)}, \qquad S_{12}(p) = \pm \frac{f(-p)}{g(p)}$$
$$S_{21}(p) = \frac{f(p)}{g(p)}, \qquad S_{22}(p) = \mp \frac{h(-p)}{g(p)} \qquad (6)$$

where p is the complex frequency variable, $h(p)$ is an arbitrary polynomial and $g(p)$ is a strictly Hurwitz polynomial. The polynomial $f(p)$ is monic and represents the transmission zeros of the network [13]. For a reciprocal network including transmission zeros at DC, infinity and finite frequencies, the $f(p)$ polynomial can be written as

$$f(p) = a_0 p^k \prod_{i=1}^{n_z} (p^2 + w_i^2) \qquad (7)$$

where k represents the number of zeros at DC, a_0 is arbitrary constant and n_z the number of finite transmission zeros. These polynomials satisfy the losslessness condition

$$g(p)g(-p) = h(p)h(-p) + f(p)f(-p) \qquad (8)$$

which imposes following degree conditions;

$$\deg h \leq n, \qquad \deg f \leq n, \qquad n = \deg g \qquad (9)$$

In the above polynomial based characterization, the network topology is related with the $f(p)$ polynomial. If there are is transmission zeros at DC, these correspond to shunt inductors or series capacitors in network. If there are transmission zeros at infinity, they correspond shunt capacitor of series inductors in the structure. Real transmission zeros on the other hand correspond resonator sections.

If the complexity of the network and the form of $f(p)$ polynomial are selected, then $h(p)$ polynomial can be arbitrarily initiated, and from the losslessness condition (8), $g(p)$ polynomial can be extracted. Once the polynomials $f(p)$, $g(p)$ and $h(p)$ are specified, the transducer power gain (TPG) of the filter can be computed at real frequencies as

$$T(w) = \frac{(1-\Gamma_g^2)\left|f(jw)\right|^2}{\left|g(jw) - \Gamma_g h(jw)\right|^2} \qquad (10)$$

In the scattering based real frequency approach, choosing the coefficients of $h(p)$ polynomial as the unknowns of the problem, TPG is optimized using a nonlinear search routine to satisfy any prescribed passband requirements.

The real frequency technique outlined above is based on the bounded real scattering characterization of the two-port, and has well behaved numeric. The procedure provides the optimum network element values for a selected network complexity over the desired frequency band of operation.

For a dual pass-band filter requirement, the conventional frequency transformation approach outlined in Section 2 can directly be integrated into the above real frequency procedure. For this purpose, dual band TPG characteristic can be generated by applying the frequency transformation on a LP or BP type prototype, to end up with the construction of a dual band filter network.

3.1. Dual band mapping from a LP prototype

In this approach, a transformer free LP ladder prototype is constructed via SRFT where, the polynomial $f(p)$ is selected as a constant, $f(p) = 1$, and the constant term of the $h(p)$ polynomial is set as $h_0 = 0$. The remaining $h(p)$ coefficients can be initialized to deliver a maximally flat or equal ripple LP characteristic. For this setup, the application of the direct LP to DB transformation defined in (3), leads to the optimization of the filter gain function given by (10) over the resulting double band characteristic shown in Fig. 3 in real frequency domain. For this purpose we apply the DB frequency transformation of (3) in TPG form of (10), by substitution $\Omega \Rightarrow w$. In this case, for a double band gain optimization, the objective function to be minimized can be chosen as;

$$\delta = \sum_{i=1}^{N_1} (T(w_i) - T_0(w_i))^2 + \sum_{j=1}^{N_2} (T(w_j) - T_0(w_j))^2 \qquad (11)$$

Here N_1 and N_2 are the number of sampling frequencies in the first pass-band and the second pass-band respectively. $T(w)$ is the calculated gain, and $T_0(w)$ is the target gain characteristic of the dual band response.

It should be noticed in this process that, LP to DB transformation form is directly utilized instead of the conventional sequential LP-BP transformations. In this projection, all the pass band specifications generated in equations (1) to (4) are directly incorporated in the selective pass band optimization scheme. The result of the optimization is the best LP prototype, which under the DB transformation leads to the desired transformer free dual band filter network. The synthesis of network elements can be obtained from the optimized scattering polynomials, using the conventional ladder realization routines [16], and then by replacement of each L and C elements with the appropriate resonance sections.

3.2. Dual band mapping from a BP prototype

In this approach, instead of the direct LP to DB transformation form of (3), first a normalized BP type prototype is constructed. Then, the conventional first order LP to BP transformation is applied to form the double band mapping from the normalized BP prototype characteristic. Although this approach is a subset of the previous case, its usage along with SRFT provides flexibility in the control of pass bands and roll off characteristic [17]. The computational steps for the construction of a transformer free BP prototype are outlined in the following.

It is well known that, a transformer free BP network can be constructed by proper cascading of LP and HP type networks as shown in Fig. 5. A transformer free LP network can be constructed by selecting $f_l(p)=1$, and the constant term of the $h_l(p)$ polynomial is set as $h_0=0$. For an HP networks, since there exist transmission zeros at DC, the polynomial $f(p)$ is selected as $f_h(p)=p^k$. For a transformer free characterization, coefficient of the highest order term of $h_h(p)$ should be set as $h_n=0$. The cascade connection of the LP and HP sections defined as above, will result in a transformer free BP prototype network.

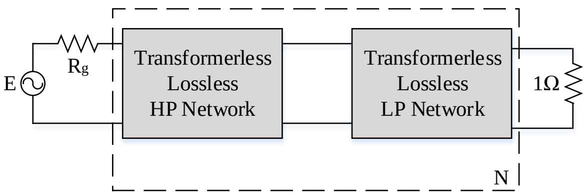

Figure 5. Construction of BP prototype network

The transfer scattering matrix of lossless cascaded network [N] can be written as [17];

$$S_T(p)=\frac{1}{f(p)}\begin{vmatrix}\mu g(-p) & h(p)\\ \mu h(-p) & g(p)\end{vmatrix},\quad \mu=\frac{f(-p)}{f(p)}=\pm 1 \tag{12}$$

If the transfer matrices of LP and HP sections are also written as

$$\begin{aligned}S_{Tl}(p)&=\frac{1}{f_l(p)}\begin{vmatrix}g_l(-p) & h_l(p)\\ h_l(-p) & g_l(p)\end{vmatrix},\\ S_{Th}(p)&=\frac{1}{f_h(p)}\begin{vmatrix}\mp g_h(-p) & h_h(p)\\ \pm h_h(-p) & g_h(p)\end{vmatrix}\end{aligned} \tag{13}$$

then, the transfer matrix of cascaded BP section can be written in terms of $S_{Tl}(p)$ and $S_{Th}(p)$ as;

$$S_T(p)=S_{Tl}(p)S_{Th}(p) \tag{14}$$

The scattering polynomials of cascaded structure can be written in terms of LP and HP polynomials as;

$$\begin{aligned}g(p)&=g_l(p)g_h(p)+\mu h_l(-p)h_h(p)\\ h(p)&=h_l(p)g_h(p)+\mu g_l(-p)g_h(p)\\ f(p)&=f_l(p)f_h(p),\ \mu=\mu_l\mu_h\end{aligned} \tag{15}$$

Thus, characterization of the cascaded structure is obtained in terms of the HP and LP subsection parameters. For this case, the initialized unknowns of the problem are set as $h_h(p)$ and $h_l(p)$ polynomials of the HP and LP prototypes respectively. Then, using the relations (8) and (15), the polynomials $g(p)$, $h(p)$ and $f(p)$ of BP structure are constructed. At this point, we employ LP to BP frequency transformation of (2) on the normalized BP prototype. That is upper band-edge of the band-pass is set to 1, and the lower cut-off is determined by (4). In this way, the BP prototype is considered as a LP, and by LP to BP transformation as depicted in the last schematic of Fig. 1, a DB frequency response $T(w)$ is generated. $T(w)$ is then optimized using (11).

The dual band mapping diagram from the normalized BP prototype is as shown in Fig. 6.

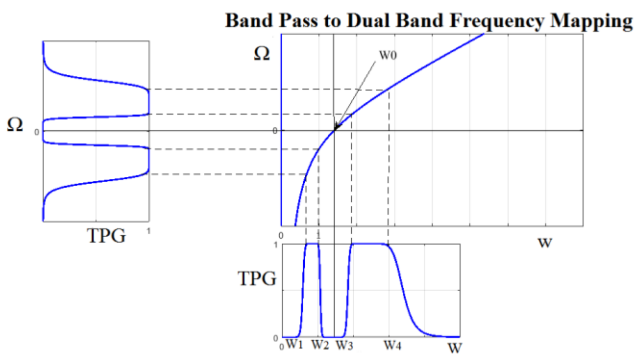

Figure 6. Band-pass to dual-band frequency mapping

In this approach, the unknowns of the problem are $h_h(p)$ and $h_l(p)$ polynomials of the HP and LP prototypes respectively. The output of the dual band optimization will then be the LP and HP subsections, whose ladder realizations can directly be utilized to yield resonance sections of the DB filter, using the conventional LP to BP element mappings [12].

4. Dual Band Filter Design Examples

In this section, two dual band filter design example is presented to illustrate the application of proposed approaches. Both examples are designed with the same normalized design specifications to compare their band control efficiencies. Desired dual-band filter is specified with three normalized pass-band edge frequencies: $w_1 = 0.7$, $w_2 = 1$ and $w_3 = 2$, which yield $w_4 = 2.84$.

Example 1: *Dual Band Filter Design using LP to DB Transformation*

For this example a sixth order LP prototype complexity is chosen. Following the procedure mentioned in Section 3.1, a sixth degree $h(p)$ polynomial is introduced to SRFT routine. After the LP to DB frequency transformation, the final synthesized DB filter will have 24 elements.

The direct DB transformation parameters are calculated according to the selected normalized filter specifications by using (1) to (4) as; $w_{02} = 1.41$, $B_2 = 2.14$, $B_s = 0.998$, $w_{01} = 0.6795$, $B_1 = 0.5383$. With these parameters, direct DB transformation in (3) is applied to the TPG of prototype LP as mentioned in Section 3.1. The optimized LP prototype network is as shown in Fig. 7.

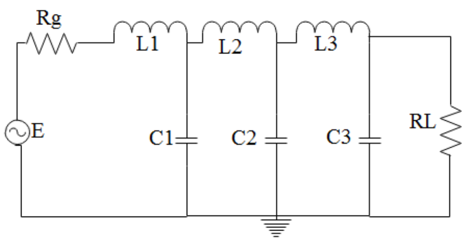

Figure 7. Synthesized normalized LP prototype network (Rg=1Ω, L1=0.34H, C1=0.85F, L2=1.09H, C2=1.04F, L3=0.77H, C3=0.28F, RL=1Ω)

Using the LP to BP element transformations given in Fig. 8 with (16), on each elements of Fig. 7, the dual band filter network elements can be obtained in two sequential steps as shown Fig. 9. The gain plot of the dual band filter of Fig. 9 is shown in Fig. 10.

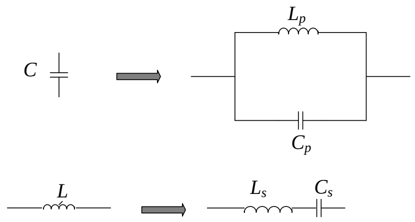

Figure 8. LP to BP element transformations

$$L_s = L/B, \quad C_s = B/w_0^2 L$$
$$C_p = C/B, \quad L_p = B/w_0^2 C \tag{16}$$

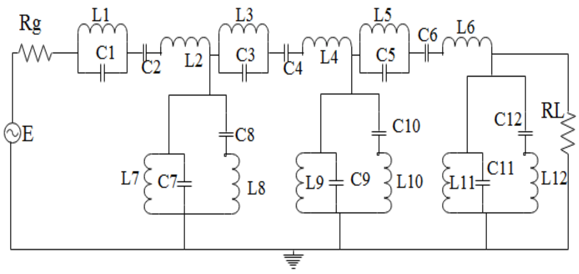

Figure 9. Synthesized normalized dual-band filter (Rg=1Ω, L1=0.3139H, C1=0.041F, L2=0.3H, C2=0.043F, L3=1.01H, C3=0.013F, L4=0.95H, C4=0.014F, L5=0.711H, C5=0.018F, L6=0.67H, C6=0.02F, L7=0.0173H, C7=0.738F, L8=0.0162H, C8=0.785F, L9=0.0141H, C9=0.903F, L10=0.0133H, C10=0.9602F, L11=0.0524H, C11=0.243F, L12=0.05H, C12=0.259F, RL=1Ω)

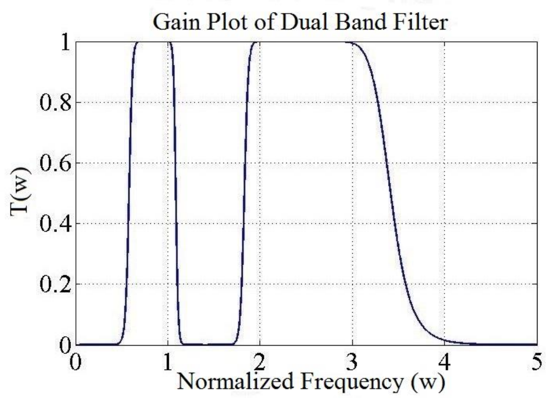

Figure 10. Filter TPG response of Example 1

Example 2: Dual-band filter design using BP to DB transformation

As outlined in Section 3.2, the proposed technique requires the characterization of BP prototype in terms of cascaded LP and HP subsections. In this example, both of the LP and HP sections are chosen as sixth order networks and the termination resistances are assumed unity.

For LP and HP subsections, coefficients of $h_l(p)$ and $h_h(p)$ polynomial are initialized in ad hoc manner. Then, the polynomials $h_l(p)$ and $h_h(p)$ are determined by optimization, to produce a dual pass band TPG characteristic under the frequency transformation (2). Here the normalized transformation parameters are calculated as $w_0 = 1.41, B = 2.14$ according to the assigned cut off frequencies and using (2).

As a result of optimization, $h_l(p)$, $g_l(p)$ and $h_h(p)$, $g_h(p)$ polynomials are obtained as in Table 1.

Table 1. Polynomial coefficients of optimized BP prototype (Coefficients are ordered from highest to lowest degree)

$h_l(p)$	0.106	0.093	0.4472	0.5205	0.48	0.84	0
$g_l(p)$	0.106	0.416	1.2218	2.3037	2.96	2.58	1
$h_h(p)$	0	0.067	0.1992	0.2322	0.62	0.18	0.432
$g_h(p)$	1	3.170	5.0195	5.0794	3.50	1.59	0.432

The synthesized transformerless BP prototype filter is shown in Fig. 11.

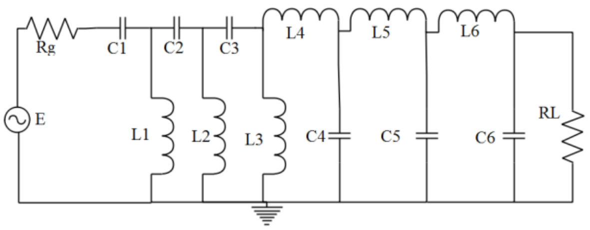

Figure 11. Synthesized normalized BP prototype (Rg=1Ω, L1=0.816H, C1=1.63F, L2=0.72H, C2=0.68F, L3=2.05H, C3=0.86F, L4=0.66H, C4=0.65F, L5=1.72H, C5=0.67F, L6=1.03H, C6=0.42F, RL=1Ω)

Using the LP to BP element transformations given in Fig. 8 with (16) on each elements of Fig. 11, the dual band filter network can be constructed as in Fig. 12. The associated gain response of the dual-band filter of Fig. 12 is as shown in Fig. 13.

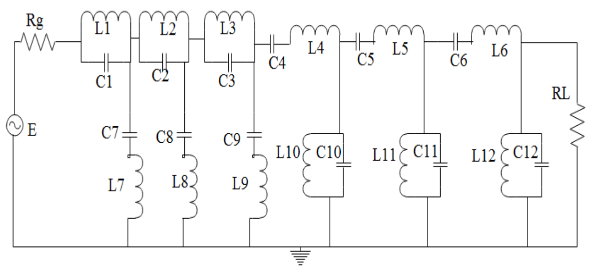

Figure 12. Synthesized normalized dual-band filter (Rg=1Ω, L1=0.0167H, C1=0.7617F, L2=0.0401H, C2=0.318F, L3=0.0317H, C3=0.402F, L4=0.3084H, C4=0.0413F, L5=0.8037H, C5= 0.016F, L6=0.4813H, C6=0.0265F, L7=0.3813H, C7= 0.0334F, L8=0.3364H, C8=0.038F, L9=0.958H, C9=0.0133F, L10= 0.042H, C10=0.304F, L11= 0.0407H, C11=0.3131F, L12=0.065H, C12=0.1963F, RL=1Ω)

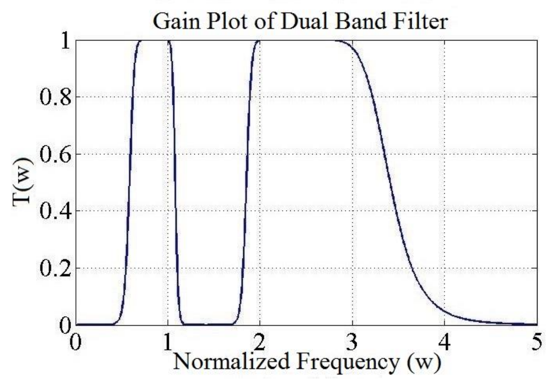

Figure 13. Filter TPG response of Example 2

The synthesized dual-band filters are given in normalized form, which can be properly denormalized by frequency and impedance scaling based design specifications.

It should be noted that, BP to DB approach gives the flexibility of choosing the number of HP and LP elements for prototype network. This provides a direct control of the pass bands and roll-off characteristic in dual band filter. LP to DB method does not however provide such a flexibility. Once the element complexity is chosen in LP prototype, then the resultant DB characteristic implicitly results in unequal pass bands around the center of stop band. The result of a test run is given in Fig. 14 as an example, to display the band control and roll of variation for increased number of LP elements in BP prototype. From the practical implementation point of view, realization of BP to DB based design is relatively easier, especially if distributed resonance section equivalencies are considered [18-19].

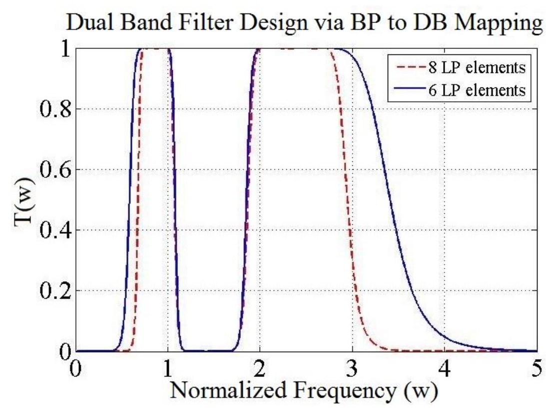

Figure 14. Filter response generated by BP to DB approach with different number of LP elements in BP prototype

5. Conclusion

In this study, a frequency transformation based dual-band matching filter design approach is presented. The scattering based real frequency technique is integrated with dual-band frequency transformation to design double band matching filters. The conventional frequency transformation method is applied on low-pass and band-pass type prototype networks to result with dual band gain characteristic. Explicit design steps for transformer free BP prototype network generation are introduced. Mapping diagrams are investigated to extract parametric relations of variables in transformation functions. Dual band filter design examples are provided with LP to DB and BP to DB approaches. It has been shown that, BP to DB design approach leads to advantageous topologies, while providing roll-off control flexibility to the designer.

6. References

[1] V. Crnojevic´-Bengin, "Advances in Multi-Band Microstrip Filters", Cambridge University Press, Cambridge, United Kingdom, 2015.
[2] A. Johannes, G. Malherbe, A. Swiatko, "Modified Chebyshev band-stop filter with transmission zeros at real frequencies", Microwave and Optical Technology Letters, vol. 53, no. 1, pp. 177–180, January 2011.
[3] H. Fujimoto, K. Murakami, S. Kitazawa, "Equivalent circuits and transmission zeros of the coupled square-loop resonator", IEICE Electronics Express, vol. 4, no. 18, pp. 575-581, 2007.
[4] M. Weng, S. Wu, S. Jhong, Y. Chang, and M. Lee, "A novel compact dual-mode filter using cross-slotted patch resonator for dual-band applications," Proceedings of International Microwave Symposium, Honolulu, Hawaii, pp. 921–924, 2007.
[5] X. Zhang, Q. Xue, and B. Hu, "Planar tri-band bandpass filter with compact size, "IEEE Microwave and Wireless Components Letters, vol. 20, no. 5, pp. 262–264, May 2010.
[6] H. Wu, Y. Chen, and Y. Chen, "Multi-layered dual-band bandpass filter using stub-loaded stepped-impedance and uniform-impedance resonators," IEEE Microwave and Wireless Components Letters, vol. 22, no. 3, pp. 114–116, March 2012.
[7] R. J. Cameron, R. Mansour, and C. M. Kudsia, "Microwave Filters for Communication Systems: Fundamentals, Design and Applications," Wiley-Interscience, New Jersey, 2007.
[8] A. García-Lampérez, and M. Salazar-Palma, "Single-Band to Multiband Frequency Transformation for Multiband Filters," IEEE Transactions on Microwave Theory And Techniques, vol. 59, no. 12, pp. 3048-3058, Dec. 2011.
[9] J. Lee, and K. Sarabandi, "A Synthesis Method for Dual-Passband Microwave Filters," IEEE Trans. Microw. Theory Techn, vol. 55, no. 6, pp. 1163-1170, June 2007.
[10] T. G. Brand, P. Meyer, R. H. Geschke, "Designing multiband coupled resonator filters using reactance transformations," Int. J. of RF and Microwave Computer-Aided Engineering, vol. 25, no 1, pp. 81–92, Jan. 2015.
[11] S. Yıldız, A. Aksen, S. B. Yarman, "Multiband Matching Network Design via Transformation based Real Frequency Approach", ISFEE, Romania, 29 June -1 July 2016.
[12] S. Yıldız, A. Aksen, S. B. Yarman, "Dual Band Matching Network Design via Real Frequency Technique by Mapping from a Band Pass Prototype," Mediterranean Microwave Symposium, UAE, 14 Nov. - 16 Nov. 2016.
[13] S. Yıldız, A. Aksen, S. B. Yarman, " SRFT ile Frekans Transformasyonunu Kullanarak Band Geçiren Filtre Prototipi Üzerinden Çokluband Filtre Tasarımı," International Conference on Electrical and Electronics Engineering, Bursa, Turkey, 01 -03 Dec. 2016.
[14] N. Nallam and S. Chatterjee, "Design of concurrent multi-band matching networks," International Symposium on Circuits and Systems, Rio de Janeiro, Brazil, 2011, pp. 201-204, May 15-19 2011.
[15] S. B. Yarman, "Design of Ultra Wideband Power Transfer Networks", 1st ed., John Wiley & Sons Ltd., UK, 2010.
[16] B. S. Yarman, H. J. Carlin, "A Simplified Real Frequency Technique Applied to Broadband Multistage Microwave Amplifiers", IEEE Transactions on Microwave Theory and Techniques, vol. 30, no. 12, pp. 2216-2222, 1982.
[17] B. S. Yarman, (2013, June 25). *High Precision Synthesis of a Richards Immittance via Parametric Approach*, [Software], Available: www.siddikyarman.com.
[18] B. S. Yarman, A. Aksen, et al. "Computer Aided Darlington Synthesis of an All Purpose Immittance Function," IU-JEEE vol. 16 no. 1, pp. 2027-2037, 2016.
[19] N. Yıldırım, "Filpro Manual: A textbook-like manual for the filter design software Filpro", Middle East Technical University, Ankara, Turkey, 1997.

Serkan Yıldız received his B.Sc. degree in Electrical and Electronics Engineering form Istanbul University, M.Sc. degree in Telecommunication Engineering from Istanbul Technical University, Turkey, in 2009 and 2011, respectively and he is currently working towards the Ph.D. degree at Electronics Engineering of Isik University. Since 2011, he has been a teaching assistant in Istanbul Technical University since 2013.

Ahmet Aksen received B.Sc. and M.Sc. degrees in Electrical and Electronics Engineering from the Middle East Technical University, Ankara, Turkey in 1981 and 1985 respectively and the Ph.D. degree from the Ruhr University, Bochum, Germany in 1994. He had been with the "Lehrstuhl für Nachrichtentechnik" of Ruhr University Bochum between 1989 and 1994. He was with Electrical Engineering Department of Istanbul University in 1995-1997 and since 1997, he is with Electrical and Electronics Engineering of Isık University, Istanbul. Currently, he is the chairman of Department of Electrical and Electronics Engineering of Işık University. His research interests are in the area of multivariable network theory and in the design of microwave filters, broadband matching networks and amplifiers.

Binboga Siddik Yarman (M'76–SM'94–F'04) received his B.Sc. in Electrical Engineering from Technical niversity of Istanbul (Feb. 1974), M.Sc. degree from Stevens Institute of Technology, NJ, USA (1978), Ph.D. degree from Cornell University, Ithaca, NY, USA (1982). He had been Member of Technical Staff at David Sarnoff Research Center where he was in charge of designing various satellite transponders for various commercial and military agencies in the US. He returned to Turkey in 1984 and served as Assistant, Associate and full Professor at Anatolia University Eskisehir, Middle East Technical University-Ankara, Technical University of Istanbul, and Istanbul University. He had been the chairperson of Department of Electronics Engineering, Defense Technologies and Director of School of Technical Sciences of Istanbul University over the years 1993-1996. He was the Founding President of Isik University. He had been a visiting professor at Ruhr University Bochum, Germany (1987-1994) and Tokyo Institute of Technology, Japan (2006-2008). Currently, he is the chairman of Department of Electrical-Electronics Engineering and the Scientific Research Projects Coordinator of Istanbul University. He also serves as the chairman of Board of Trustees of Isik University. He holds four US patents assigned to US Air Force. He is the author of the books titled "Design of Ultra Wideband Antenna Matching Networks" by Springer 2008 and "Design of Ultra Wideband Power Networks" by Wiley 2010. He received the Young Turkish Scientist Award in 1986, the Technology Award in 1987 of National Research and Technology Counsel of Turkey. He received the Research Fellowship award of Alexander Von Humboldt Foundation, Bonn, Germany, in 1987. He became the Member of New York Academy of Science in1994. He was named as the "Man of the year in Science and Technology" in 1998 of Cambridge Biography Center, UK and elevated to IEEE Fellow for his contribution to "Computer Aided design of Broadband Amplifiers".