A Novel Dual-Mode Wideband Band Pass Filter

A. Ahmadi [1], S. V. Makki [2], A. Lalbakhsh [1], and S. Majidifar [3]

[1] Department of Electrical Engineering, Kermanshah Branch
Islamic Azad University, Kermanshah, Iran
aahmadi70@gmail.com, ali.lalbakhsh@yahoo.com

[2] Department of Electrical Engineering
Razi University, Bagh-E-Abrisham, Kermanshah, Iran
v.makki@razi.ac.ir

[3] Department of Electrical Engineering
Kermanshah University of Technology, Kermanshah, Iran
s.majidi@kut.ac.ir

Abstract **—** A novel Wideband (WB) Bandpass Filter (BPF) with improved passband performance using a Defected Ground Structure (DGS) is presented in this paper. The proposed BPF is composed of two novel basic WB resonators and six Dumbbell-Shaped (DS) DGSs. By cascading resonators we can achieve better skirt characteristics, and by using DS-DGSs under interdigital coupled I/O lines we can improve the return loss within the passband. The simulated and measured results are found in good agreement with each other showing a wide passband from 4.93 to 11.62 GHz, a wide upper stopband with around 25 dB attenuation up to 20 GHz, and sharp roll-offs around 0.03 in lower and upper edges.

Index Terms **—** Dual mode, microstrip bandpass filter, wide passband.

I. INTRODUCTION

Microstrip filters are in high demand in modern communication systems according to their importance in selecting the desired frequency range among numerous neighbor bands, and also suppressing the undesired harmonics [1-7]. In the front-end of the Ultra-Wideband (UWB) systems, a small circuit size, high selectivity, and wide stopband microwave Bandpass Filters (BPF) are essential to provide better frequency responses. Since the federal communications commission has allocated 7.5 GHz of spectrum for unlicensed use of UWB devices [8], many attempts have been made to produce various kinds of wideband BPFs [9-19]. These wideband filters can be divided into two basic types. The first type is based on the design of a single Stepped Impedance Resonator (SIR) and strong I/O coupling, which is named as Multiple Mode Resonator (MMR) [9-10]. The second UWB filter was initially presented as an optimum distributed highpass filter [11], but some sections were added to the highpass filter in order to create transmission nulls which were used to control upper passband edge [12-13]. In [14], a single modified square slot-line resonator and parallel microstrip feed lines have been used for UWB filter design. Here, the improved upper stopband has been obtained by exploiting the frequency dispersive coupling behavior between the feed lines and the resonator. A wideband bandpass filter has been designed in [15], using Composite Right/Left Handed (CRLH) transmission lines and floating slot in the ground plane in order to get high coupling in comparison to conventional edge coupled microstrip line. In [16], parallel coupled line filters have been constructed, employing tight-coupling structures; however, the reported FBW only achieved around 64% of the bandwidth. Wideband bandpass filters using multiple-mode

Submitted On: January 30, 2013
Accepted On: September 5, 2014

resonators to achieve a fractional bandwidth of 100% have been reported in [17]. However, these wideband filters suffer from some problems, such as larger size and slow rejection skirt. Although in [18-19], the authors could realize sharp roll-off and good stopband characteristic using a new fractal geometry, the proposed filter did not provide wide passband. In this paper, a novel dual-mode wide band bandpass filter using symmetric interdigital coupled Input/Output (I/O) lines with the patterns of DGS is proposed. The mechanism of the proposed dual-mode resonator filter is investigated in detail using even and odd-mode analysis. Finally, this proposed filter is verified by simulation and measurement.

II. ANALYSIS AND DESIGN

A. Structure and equivalent circuit of the proposed basic wide band resonator

Figure 1 (a) shows the schematic diagram of the proposed basic WB resonator. The dimensions of the basic resonator in Fig. 1 (a) are: $L1 = 7.43, L2 = 6.38, L3 = 4.3, W1 = 0.11, W2 = 0.32, W3 = 0.1, W4 = 0.1, W5 = 0.56$ (all in millimeter). The equivalent circuit model of the proposed resonator is shown in Fig. 1 (b), which consists of Ls=4.6 nH (Ls is the inductance of upper and lower inductive lines), Lt=13.1 nH (Lt is the inductance of central inductive line), Cc=0.07 pF (Cc is the coupling capacitance between these inductive lines), and Cb=0.05 pF (Cb is the capacitance between the edge parts of the resonators and the ground plane). The obtained values of those parameters are extracted using methods discussed in [11]. This structure is designed on RO4003 with a dielectric constant of 3.38, height of 0.508 mm, and loss tangent of 0.0021.

The simulated S-parameters of the basic resonator are shown in Fig. 1 (c). As it is shown in Fig. 1 (c), the simulated S-parameter of the designed resonator has ripple in the passband and also suffers from big return loss within the passband. For solving this problem, we proposed a Defected Microstrip Structure (DMS) to suppress unwanted ripples and decrease the return loss. Frequency responses and the layout of this structure are shown in Fig. 2. The dimensions of the layout in Fig. 2 are: $P1 = 6.8$, $P2 = 4.4$, $M1 = 0.4$, $M2 = 1$, $M3 = 0.4$, $M4 = 0.3$, $M5 = 0.63$ (all in millimeter), while the widths of lines are the same as mentioned.

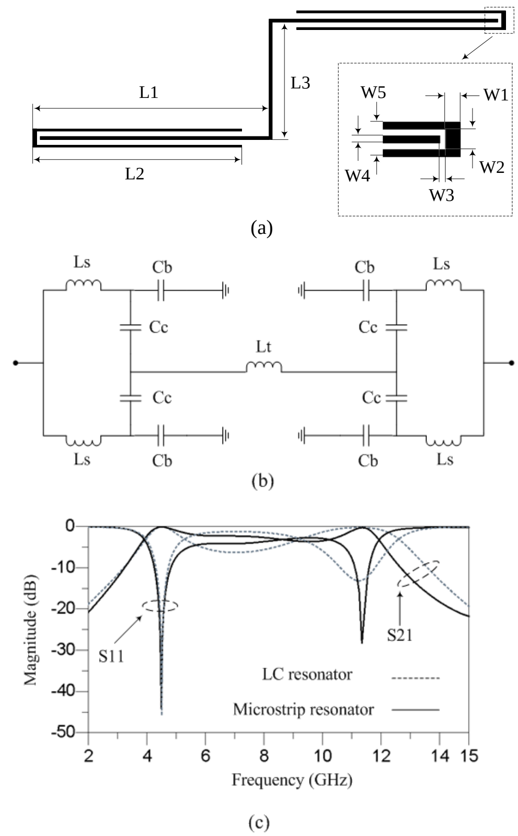


Fig. 1. Basic resonator: (a) layout, (b) LC model, and (c) frequency responses.

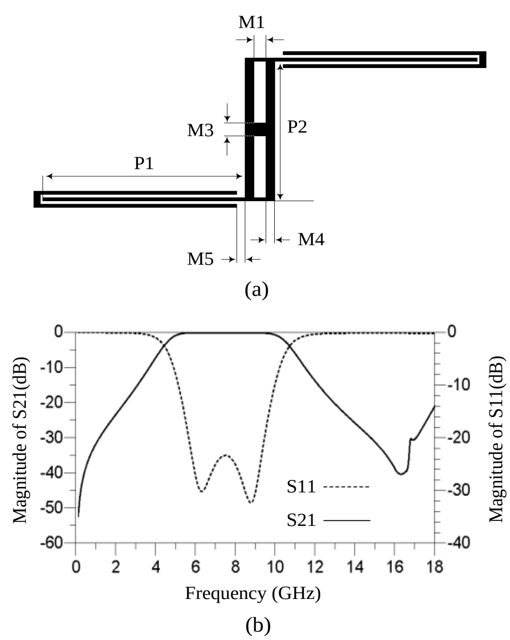
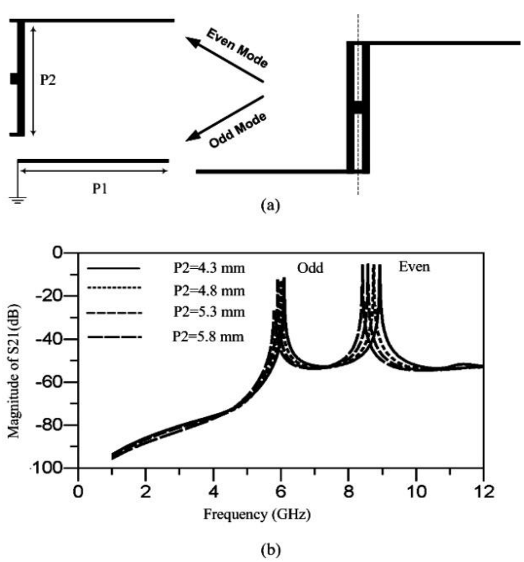


Fig. 2. Basic resonator with DMS: (a) layout, and (b) simulation results.

B. Dual mode characteristic

The resonator in Fig. 2 can be divided into two sections along the symmetric plane of the filter; an open-circuited resonator for even mode and a short-circuited for odd mode [11], as shown in Fig. 3 (a). The resonant characteristic of the resonator against the length of P2 is depicted in Fig. 3 (b). Weak coupling is applied in the simulation of decreasing the influence of the input and output lines of the resonator. As it is shown in Fig. 3 (b), resonant frequency of even modes of the resonator is mainly determined by P1 and P2. Keeping P1 stable and increasing the length of P2, the fundamental mode remains unchanged while the second resonant mode f_{even} of the resonator will move towards the fundamental mode. Thus, a dual-mode characteristic can be achieved by tuning the length of P2.

Figure 3 (c) shows the resonance frequencies of the two modes against the length of P1. The horizontal axis is the length of the left/right branch P1, and the vertical axis is the resonator frequency of the two modes. As is shown in Fig. 3 (c), when P1 varies from 3.9 mm to 9.9 mm, the resonance frequencies of the odd and even modes decrease almost linearly from 9.51 to 4.438 and 13.72 to 6.74 GHz, respectively. It is evident from Fig. 3 that the resonant frequency of the even mode is determined by both P1 and P2, while the resonant frequency of the odd mode is controlled by the length of P2. So changing the even-mode resonance frequencies will not affect the odd-mode resonance frequencies. Having this feature, the dual-mode filters with the flexible passband frequencies can be designed.



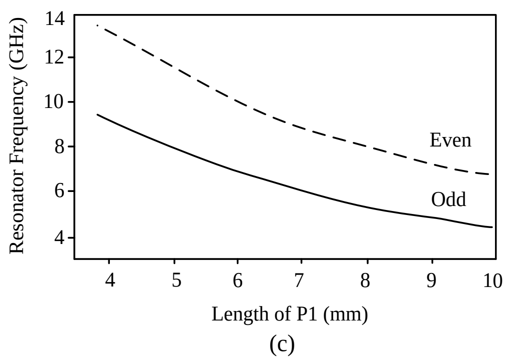


Fig. 3. (a) The configuration of the dual-mode microstrip resonator, (b) simulated resonances of the two modes against P2, and (c) simulated resonances of the two modes against P1.

C. Coupling coefficient

The coupling coefficient can be evaluated from the two dominant resonant frequencies for any two synchronously tuned coupled resonators If f_{p1} and f_{p2} are defined to be the lower-odd mode and higher-even mode of the two resonant frequencies respectively, the coupling coefficient can be obtained by [11]:

$$M = \frac{f_{p2}^2 - f_{p1}^2}{f_{p2}^2 + f_{p1}^2}. \quad (1)$$

The coupling coefficient of the proposed resonator in Fig. 2 (a) is only controlled by the width of the line in the middle of the interdigital coupled I/O lines-denoted by W4-the values of coupling coefficient M is given against the width of W4 by Fig. 4. As can be seen from this figure, by increasing the width of W4 and keeping unchanged the spaces between the coupled lines, the coupling coefficient increases.

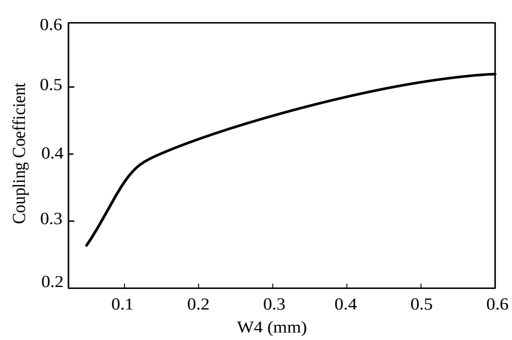


Fig. 4. Coupling coefficient of the proposed resonator versus the width of the central line.

III. FILTER DESIGN

A. Two cascaded resonators

Based on the above formulation, a bandpass filter characterizing sharp roll-off and wide bandwidth has been designed. The wide passband is realized by applying DMS, whereas the sharp roll-off is achieved by cascading two cells as illustrated in Fig. 5.

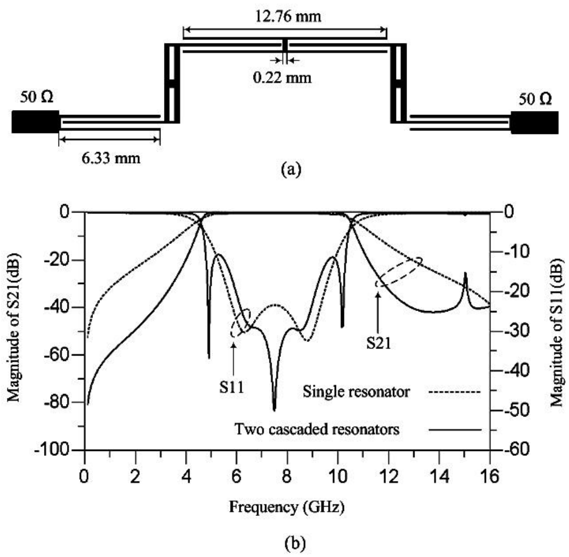


Fig. 5. Two cascaded resonators: (a) schematic, and (b) simulation.

It can be seen from Fig. 5 (b) that as a result of cascading two cells the lower and upper transition bands decrease from 2.11 and 2.3 GHz to 0.7 and 0.87 GHz, respectively, while the passband remains unchanged.

B. Applying dumbbell-shaped pattern defected ground structure

Unlike the previous researches which employed DGS to improve the stopband characteristics [20-21], here we have designed three Dumbbell-Shaped (DS) pattern Defected Ground Structure (DS-DGS) to decrease return loss within the passband.

To explain the DS-DGS effects, the LC model of this structure is extracted in Fig. 6. In this figure, the DS-DGS that is modeled by a parallel LC resonator [2], crates a transition pole. If the frequency of this pole matches with the transition zero frequency (in the passband of the proposed WB bandpass filter), then the passband performance will improve.

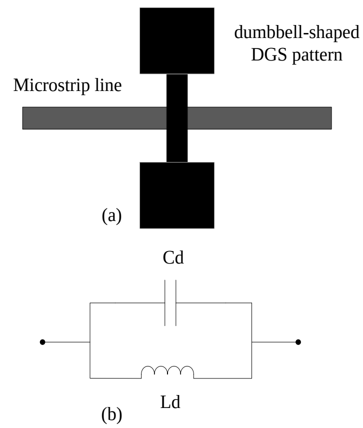


Fig. 6. DS-DGS: (a) layout, and (b) LC model.

In another approach, to explain how DS-DGS affects the return loss, one half of the basic resonator (in the presence of the DS-DGS) in Fig. 1 (a) and its equivalent circuit are studied. Introducing DS-DGS causes a reduction in electric field between edge parts of microstrip lines and ground plane, as well as between microstrip lines themselves; the former leads to a decrease in capacitance of Cb and the latter causes a reduction in the capacitance of Cc. As far as magnetic field is concerned, the proposed DS-DGS increases magnetic field around the perimeter of defected area, which results in a small increase in the inductance of Ls and Lp (Lp is the inductance of the central strip line in the half-resonator). Figure 7 shows the effects of these changes on the frequency responses of the half-resonator. As it is shown in Fig. 7, decreasing the Cb decreases return loss drastically, while decreasing Cc increases the return loss very slightly-around one dB-and increasing Ls and Lp have no effects on the return loss. The resultant of these effects causes only a reduction in return loss, while the resonance frequency keeps unchanged because the variation of Cb has no effect on resonance frequency, and shifting effects on resonance frequency caused by variation of Cc and inductors neutralize each other. It should be noticed that the dominant effect of DS-DGS decreases the electric field between ground plane and microstrip lines, which causes a huge decrease in return loss. So Cb has a dominant impact on the return loss.

Based on above explanations, three applied DS-DGS improved the return loss to around -23 dB. As it is shown in Fig. 8, the dumbbell-shaped patterns have etching dimensions of: $S1 = 0.5, S2 = 0.16, S3 = 0.39, S4 = 2.09$ (all in millimeter). The final dimensions of studied filter are tabulated in Table 1.

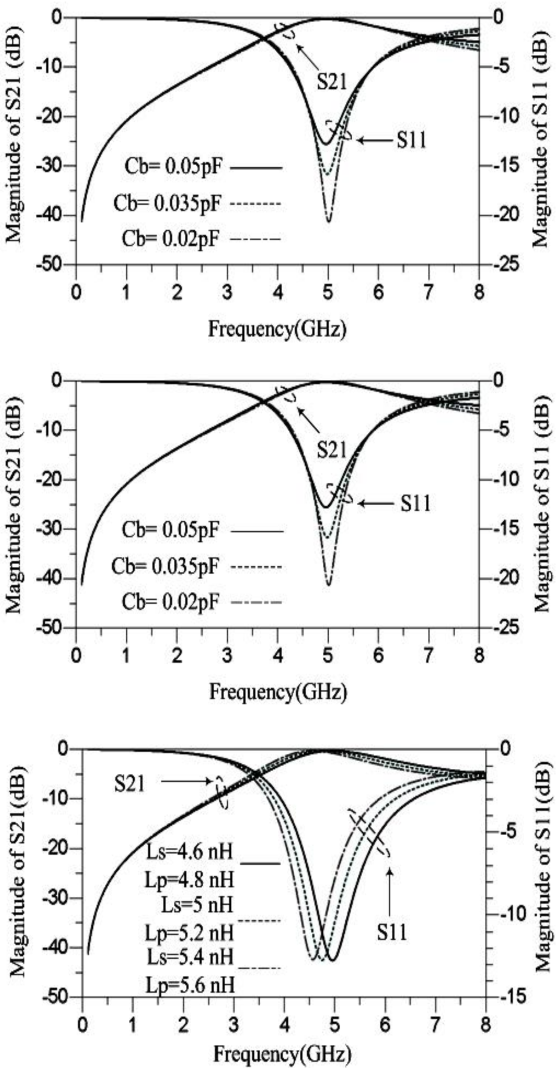


Fig. 7. Simulated results of the half-resonator model.

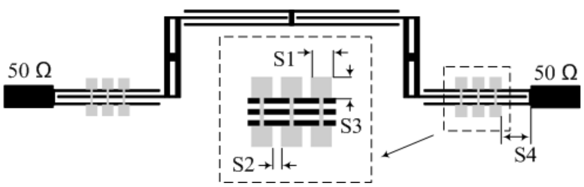


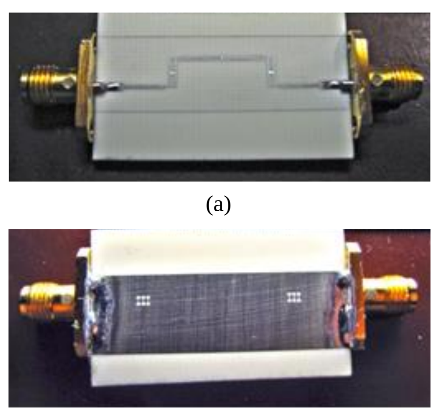
Fig. 8. Layout of the final filter.

Table 1: Final dimension of the proposed filter (unit: mm)

W1	W2	W3	W4	W5	P1	P2	M1
0.1	0.31	0.11	0.10	0.54	6.34	3.95	0.59
M2	M3	M4	M5	S1	S2	S3	S4
0.90	0.40	0.15	0.61	0.5	0.16	0.39	2.09

IV. SIMULATION AND EXPERIMENTAL RESULTS

The proposed WB filter is fabricated on the RO4003 substrate with a dielectric constant of 3.38, a thickness of 0.508 mm and loss tangent of 0.0021. The size of the fabricated filter is $27.15\times10.8\ \mathrm{mm}^2$, which corresponds to $1.12\ \lambda_g\times 0.45\ \lambda_g$, where λ_g is the guided wavelength at the center frequency. Measured results of the filter are characterized in an Agilent network analyzer N5230A. Figures 9 and 10 show the photograph and the simulated and measured results of the fabricated filter respectively. The measured results of the filter have 3 dB Fractional Bandwidth (FBW) of 88% at 7.57 GHz; the maximum insertion loss within the whole wide passband is 1.9 dB and the minimum return loss is around 15 dB. There is no undesired passband on the left side of the wide passband. On the right side of the passband, the 24 dB rejection band is extended from 12.27 to 20 GHz. The return loss in the stopband region is very small indicating negligibly small radiation loss. The lower transition band from 4.931 to 4.39 and the upper transition band from 11.62 to 12.27 GHz with -3 and -20 dB are 0.54 and 0.65GHz respectively, showing that the filter has excellent skirt performance.



(a)

(b)

Fig. 9. Photograph of the fabricated filter: (a) top view, and (b) bottom view.

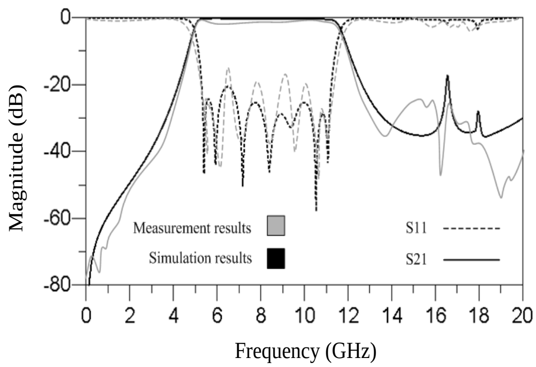


Fig. 10. Simulated and measured results.

V. DEVELOPMENT OF THE PROPOSED DESIGN

The proposed method is to scale all of the basic resonator dimensions with a constant factor. This method is suitable to make significant variations to the central frequency of the bandpass filter. Figure 11 shows the central frequency and fractional bandwidth of basic resonator as a function of scaling factor.

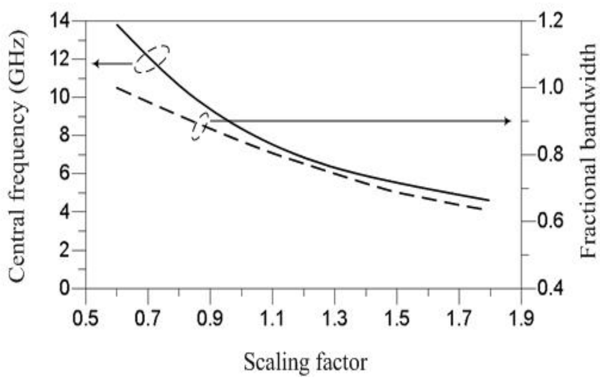


Fig. 11. The central frequency and fractional bandwidth of the basic resonator as a function of the scaling factor.

Design process of a WB bandpass filter with a different central frequency is performed by the following steps:

1. Selection of the appropriate scaling factor, according to the desired central frequency using Fig. 11.
2. Cascading two basic resonators to improve sharp roll-off, as illustrated in Fig. 5.
3. Applying dumbbell-shaped pattern defected ground structure in order to passband performance improvement.

VI. CONCLUSION

In this paper, a novel dual-mode wideband BPF based on the odd and even mode resonance frequency analysis are presented. The passband of the filter is completely adjustable and provided a sharp roll-off in the upper and lower edges. In addition, a new approach to DGS was introduced in order to improve the return loss in the passband.

REFERENCES

[1] L. Yang, Y. Hongchun, W. Yawei, and X. Shaoqiu, "Ultra-wideband bandpass filter based on parallel-coupled microstrip lines and defected ground structure," *Appl. Comp. Electro. Soc. (ACES) Journal*, vol. 28, no. 1, pp. 21-25, January 2013.

[2] N. M. Garmjani and N. Komjani, "Improved microstrip folded tri-section stepped impedance resonator bandpass filter using defected ground structure," *Appl. Comp. Electro. Soc. (ACES) Journal*, vol. 25, no. 11, pp. 975-983, November 2010.

[3] M. Hayati, A. Khajavi, and H. Abdi, "A miniaturized microstrip dual-band bandpass filter using folded UIR for multimode WLANs," *Appl. Comp. Electro. Soc. (ACES) Journal*, vol. 28, no. 1, pp. 35-40, January 2013.

[4] S. Lotfi, S. V. Makki, and S. Majidifar, "Design of microstrip low-pass filter with ultra-wide stopband and sharp rejection," *International Journal of Engineering & Technology Sciences (IJETS)*, vol. 2, no. 1, pp. 47-57, 2014.

[5] S. V. Makki, A. Ahmadi, S. Majidifar, H. Sariri, and Z. Rahmani, "Sharp response microstrip LPF using folded stepped impedance open stubs," *Radioengineering Journal*, vol. 22, no. 1, pp. 328-332, 2013.

[6] H. Sariri, Z. Rahmani, A. Lalbakhsh, and S. Majidifar, "Compact LPF using T-shaped resonator," *Frquenz Journal*, vol. 67, no. (1-2), pp. 17-20, 2013.

[7] M. Hayati, S. Majidifar, and O. S. Fathabadi, "Compact microstrip lowpass filter with wide stopband and high attenuation," *Microwave Journal*, April 2012.

[8] "Revision of part 15 of the commission's rules regarding ultra-wideband transmission systems," *First Note and Order Federal Communications Commission*, ET-docket 2002, pp. 98-153, 2002.

[9] R. Li and L. Zhu, "Compact UWB bandpass filter using stub-loaded multiple-mode resonator," *IEEE Microw. Wireless Compon. Lett.*, vol. 17, no. 1, pp. 40-42, January 2007.

[10] S. Sun, L. Zhu, and H. H. Tan, "A compact wideband bandpass filter using transversal resonator and asymmetrical interdigital coupled lines," *IEEE*

Microw. Wireless Compon. Lett., vol. 18, no. 3, 2008.
[11] J. S. Hong and M. J. Lancaster, "Microstrip filters for RF/microwave applications," *New York: Wiley*, 200173-175, March 2008.
[12] H. Shaman and J. S. Hong, "Ultra-wideband bandpass filter with embedded band notch strctures," *IEEE Microw. Wireless Compon. Lett.*, vol. 17, no. 3, pp. 193-195, March 2007.
[13] H. Shaman and J. S. Hong, "Input and output cross-coupled wideband bandpass filter," *IEEE Trans. Microw. Theory Tech.*, vol. 55, no. 12, pp. 2562-2568, December 2007.
[14] P. Monda and M. K. Mandal, "Compact ultra-wideband bandpass filter with improved upper stopband," *IEEE Microw. Wireless Compon. Lett.*, vol. 17, no. 9, pp. 643-645, 2007.
[15] P. Mondal, A. Roy, and S. K. Parui, "Wide-band bandpass filter using CRLH transmission line and floating slot approach," *2nd Int. Conf. Computer, Communication, Control and Information Technology*, vol. 4, pp. 466-471, 2012.
[16] C. C. Chen, J. T. Kuo, M. Jiang, and A. Chin, "Study of parallel coupled-line microstrip filter in broadband," *Microw. Opt. Tech. Lett.*, vol. 48, no. 2, 373-375, February 2006.
[17] R. Li and L. Zhu, "Compact UWB bandpass filter using stub-loaded multiple-mode resonator," *IEEE Microw. Wireless Compon. Lett.*, vol. 17, no. 1, 40-42, January 2007.
[18] A. A. Lotfi-Neyestanak and A. Lalbakhsh, "Improved microstrip hairpin-line bandpass filters for spurious response suppression," *IEE Electron. Lett.*, vol. 48, no. 14, July 2012.
[19] A. Lalbakhsh, A. A. Lotfi-Neyestanak, and M. N. Moghaddasi, "Microstrip hairpin bandpass filter using modified minkowski fractal shape for suppression of second harmonic," *IEICE Trans. Electron.*, vol. E95-C, no. 3, pp. 378-381, 2012.
[20] J. Park, J. P. Kim, and S. Nam, "Design of a novel harmonic-suppressed microstrip low-pass filter," *IEEE Microw. Wireless Compon. Lett.*, vol. 17, pp. 424-426, 2007.
[21] M. H. AlSharkawy, D. Abd El-Aziz, and E. G. Mahmoud, "A miniaturized lowpass/bandpass filter using double arrow head defected ground structure with centered etched ellipse," *Progress In Electromagnetics Research Letters*, vol. 24, pp. 99-107, 2011.

Arash AhmadiI received his B.Sc. and M.Sc. degrees in Electronics Engineering from Sharif University of Technology and Tarbiat Modares University, Tehran, Iran, in 1993 and 1997 respectively. He joined soon after to the Electrical Engineering Department of Razi University, Kermanshah, Iran as a faculty member. He received his Ph.D. degree in Electronics from the University of Southampton, United Kingdom in 2008. From 2008 to 2010, he was a Fellow Researcher with the University of Southampton. He is currently an Assistant Professor in the Electrical Engineering Department, Razi University. His current research interests include implementation of signal processing systems, electronic circuit design and optimization and bio-inspired computing.

Seyed Vahab Al-Din Makki was born in Kermanshah. He received his Ph.D. degree in Electrical Engineering-Waves from Khaje Nasir Toosi University in 2008. His current research interests include modern digital radio propagation systems, microwave devices and radio transmitters. He is with the Electrical Engineering Department of Razi University in Kermanshah and Islamic Azad University, Kermanshah Branch, since 2008.

Ali Lalbakhsh was born in Tehran, Iran in 1985. He received his B. Sc. and M. Sc. degrees in Communication Eng. in 2008 and 2011 from Islamic Azad University, Borujerd Branch, Borujerd, Iran and Islamic Azad University, Science and Research Branch, Tehran, Iran. Since 2010 he has been teaching in the Department of Electrical Engineering at Islamic Azad University, Kermanshah Branch, Kermanshah, Iran. His research interests include microstrip filter, microstrip antenna, RF circuit design and evolutionary algorithm. Lalbakhsh is a member of IEEE and the Institute of Electronics, Information and Communication Engineers (IEICE). He is also a Reviewer to The Applied Computational Electromagnetics Society (ACES).

Sohrab Majidifar received his B.Sc. and M.Sc. degrees in Electrical Engineering from Razi University of Kermanshah in 2009 and 2011 respectively. He joined Kermanshah University of Technology in 2011 as a Lecturer. His research interests include microwave passive circuits, chipless RFID and RFIC.