Extended Configuration of Antiparallel Band Pass Filters with Two Independently Adjustable Transmission Zeros

Siniša P. Jovanović

Abstract – **This paper shows how a simple modification of the basic bandpass filter with an antiparallel configuration introduces a pair of transmission zeros close to the passband of the filter. Closed-form expressions for calculating all the components of the filter prototype depending on the frequency of the transmission zeros for different widths of the passband were derived as well. The methodology for filter realization with distributed parameter elements was also shown. The obtained extended configuration of the filter is very suitable for the synthesis of RF and microwave filters with favourable characteristics, as well as the practical realization in various planar structures and technologies.**

Keywords – **Band Pass Filter, Antiparallel Configuration, Adjustable Transmission Zeros, Printed Filter, Microstrip.**

I. Introduction

For the fast development of modern wireless and mobile telecommunications systems, the existence of an RF transceiver with a wide enough frequency spectrum, good sensitivity and adaptive characteristics which have small dimensions and are low cost is necessary. One of the essential components of such transceivers are Band Pass Filters (BPF) since they can efficiently select the desired signals from the working range and repress the unwanted signals from neighbouring or distant channels. For the realization of the needed selectivity without increasing the order of filters, several papers [1–4] have suggested BPFs whose transmission characteristics contain zeros close to the upper and lower border of the passband region. The existence of such zeros improves selectivity, that is increases the insertion loss in the upper and lower stopband of the BPF. During the synthesis, designing and realization of the BPF it is especially useful if the frequency position of the transmission zeroes can be set at desired frequency values in order to attenuate the dominant parasitic and unwanted signals [1,2].

The mutually similar characteristics of the printed filters presented in various source literature, and which their authors classified in different ways and analysed with different methods, suggest that the existence of transmission zeros close to the pass-band of the BPF can be a consequence of the specific topology of such filters [5-15]. This is especially the case with filters with capacitively coupled resonators published in [16-22] and reviewed in [23]. Those filters have an identical general topology, characterized by the fact that the two-port filter network is formed with an antiparallel connection of two identical asymmetrical subnetworks. In paper [24], the simplest configuration of the asymmetrical subnetwork whose antiparallel connection has the characteristics of a BPF is identified. For this Basic Antiparallel BPF (BABPF) a method for the synthesis of the ideal filters for different widths of the pass-band was presented. Such a BABPF had a limited practical applicability, but it was an excellent foundation for the development of an upgraded version which had significantly better characteristics and was more suitable for the realization of printed filters at RF and microwave frequencies, which is presented in this paper.

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II. A Basic BPF with an Antiparallel Configuration

Fig. 1 shows a circuit diagram of a BABPF whose characteristics are defined by the values of only three parameters: C - series capacitance; L - series inductance; and k - grounding to series capacitances ratio.

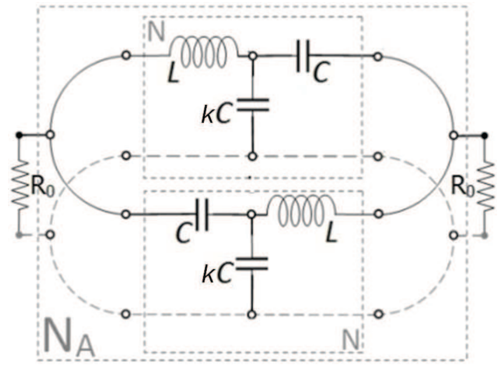


Fig. 1. Circuit diagram of a BABPF formed by an antiparallel connection of two identical asymmetrical subnetworks N

In papers [24, 25] there is a detailed presentation of how, with setting up the condition of the existence of a double transmission pole at the unity angular frequency, unambiguous real and positive values of the BABPF components can be calculated by solving a system of two nonlinear equations for unknown variables k and L, for every value of the series capacitance $C > 1$, which is selected as an independent variable. Alternatively, for $C > 2$ (which is sufficient for most practical applications), parameter k can be calculated, with a relative precision better than 1×10^{-5}, with the following empirical equation which is a unary argument function of C:

$$k \approx \sqrt{4C^2 - 1} - 2 - 0.20844\, C^{-3.047732} \qquad (1)$$

which allows obtaining the value of parameter L by applying:

$$L = \frac{1 + k(1 + C(C(k+2) + \sqrt{4C^2 - (k+2)^2}))}{C(C^2k^2 + (k+1)^2)} \qquad (2)$$

In that manner a family of prototypes of BABPF with the passband centered on the unity angular frequency, having various selectivity and various passband widths, can be obtained. In Table 1 there is an example of four sets of parameters obtained from (1) and (2) and the most important characteristics of corresponding BABPFs, while Fig. 2 shows the frequency characteristics of their S parameters.

TABLE 1
ELEMENTS AND CHARACTERISTICS OF BABPF PROTOTYPES

	parameter values			main characteristics of BABPF				
	C	L	k	BW_{3dB}	ω_c	ω_{z1}	ω_M	$S_{21}(\omega_M)$
i	2	0.811	1.848	75.6%	1.0684	0.465	0.144	-5.6 dB
ii	4	0.321	5.934	36.2%	1.0046	0.335	0.032	-12 dB
iii	8	0.142	13.968	17.8%	1.0005	0.243	0.008	-18.dB
iv	16	0.067	29.984	8.8%	1.00006	0.174	0.002	-24 dB

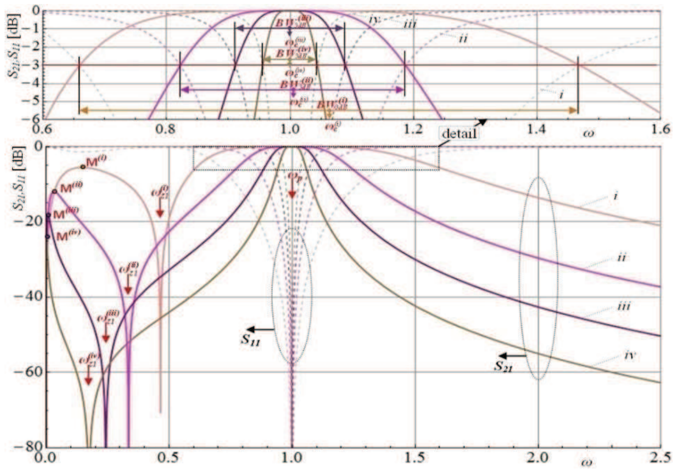


Fig. 2. Frequency characteristic of S_{21} and S_{11} parameters of four BABPF prototypes from Table 1

It can be seen in Fig. 2 that the network in Fig. 1, for values of parameters from Table 1, has the characteristics of a bandpass filter whose selectivity increases with an increase in capacitance values C (and with a decline in the values of L). A set of solutions (C, L and k) which fulfil the condition of the existence of a double pole at unity angular frequency will exist for arbitrarily large values of C, so that BABPF selectivity is constrained only by technological limitations. The isolation of the filter in the upper bandstop region increases monotonically with an increase in frequency, and also increases with an increase in values of capacitive prototype elements. In the lower bandstop region there are two transmission zeros for ω_{Z0}=0, as well as at ω_{Z1} which is given with the equation:

$$\omega_{Z1} = \frac{1}{\sqrt{LC(k+1)}} \qquad (3)$$

Between ω_{Z0} and ω_{Z1} zeros there is a point (ω_M) of local minimum of isolation $|S_{21}(\omega_M)|$ in the lower stopband. From Table 1 and in Fig. 2 it can be seen that the value of isolation in the point of local minimum $|S_{21}(\omega_M)|$ grows with the increase in value C, and at the same time, moves toward the lower frequencies, which is also the case with the frequency of transmission zero ω_{Z1} defined by equation (3).

The practical applicability of the BABPF is limited by the incapability for certain characteristics, such as BW_{3dB}, ω_{Z1}, ω_M, and $|S_{21}(\omega_M)|$, to be changed independently, so they must be chosen in „a package“, based on the characteristic that is the most crucial according to project specifications. Also, the fabrication of serial capacitance C which is not surrounded on both its ends with grounding capacitance but on only one is unsuitable for realization on many printed circuit substrates. The specified limitation of the BABPF required a modification, i.e. an extension, in the configuration of the basic subnetwork.

III. A BPF WITH AN EXTENDED ANTIPARALLEL CONFIGURATION

The described deficiencies of the BABPF can be overcome by modifying the configuration so that, instead of a pair of the simplest subnetworks N from Fig. 1, the filter is formed with a pair of subnetworks N_Y shown in Fig. 3a, or alternatively with a pair of subnetworks N_Δ equivalent to N_Y, shown in Fig. 3b.

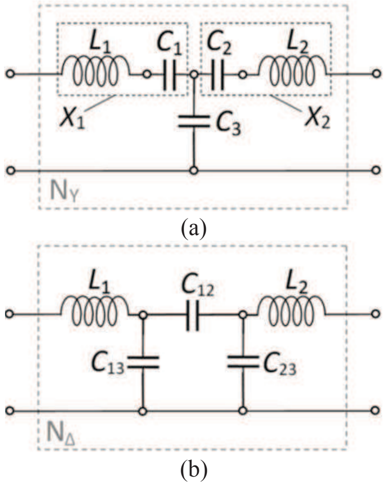


Fig. 3. Two mutually equivalent extended subnetworks: a) with capacitors in Y configuration; b) with capacitors in Δ configuration

The values of the subnetworks components from Fig. 3 are defined in a way shown in Table 2, by using the values of parameters C, L and k which form some of BABPF prototypes, with the introduction of two non-negative parameters m and n. The values of subnetwork N_Δ capacitances are obtained via the star-to-delta transformation of the capacitances from subnetwork N_Y, while the inductances in both cases are the same. For $m \geq 0$ and $n \geq 0$, all capacitances and inductances of subnetworks N_Y and N_Δ are positive, so that they can be realized with passive components.

TABLE 2
COMPONENTS DEFINITION FOR SUBNETWORKS N_Y AND N_Δ

N_Y:	L_1	C_1	C_3	C_2	L_2
	$(m+1)L$	$1/(mL)$	kC	$C/(n+1)$	n/C
N_Δ:	L_1	C_{13}	C_{12}	C_{23}	L_2
	$(m+1)L$	$(n+1)kC/K_\Delta$	C/K_Δ	mkC^2L/K_Δ	n/C

where: $K_\Delta = n+1+mLC(1+(n+1)k)$

For the components defined as indicated in Table 2, admittances X_1 and X_2 from Fig. 3a can be expressed as:

$$X_1 = \omega L_1 - \frac{1}{\omega C_1} = \omega(m+1)L - \frac{mL}{\omega} =$$

$$\omega L + mL(\omega - \frac{1}{\omega}) = X_L + mL(\omega - \frac{1}{\omega}) \tag{4}$$

$$X_2 = \omega L_2 - \frac{1}{\omega C_2} = \frac{n\omega}{C} - \frac{(n+1)}{\omega C} =$$

$$-\frac{1}{\omega C} + \frac{n}{C}(\omega - \frac{1}{\omega}) = X_C + \frac{n}{C}(\omega - \frac{1}{\omega}) \tag{5}$$

Equations (4) and (5) show that at a unity angular frequency (ω=1), as well as for m=0 (or n=0) serial admittances X_1, (or X_2) of the extended subnetworks N_Y are reduced to corresponding serial admittances X_L (or X_C) of basic subnetwork N. The definitions of subnetwork N_Y components shown in Table 2 are postulated in order to achieve such behavior.

By applying the same procedure through which (3) was derived in [24], equations for circular frequency transmission zeroes of the filter composed of an antiparallel connection of a pair of subnetworks N_Y are obtained:

$$\omega_{EZ1} = \sqrt{p_{mn} - \sqrt{p_{mn}^2 - q_{mn}}} \tag{6}$$

$$\omega_{EZ2} = \sqrt{p_{mn} + \sqrt{p_{mn}^2 - q_{mn}}} \tag{7}$$

where p_{mn} and q_{mn} are:

$$p_{mn} = \frac{1}{2nk} + \frac{1}{2(m+1)kCL} + \frac{2mn+m+n+1}{2n(m+1)}$$

$$q_{mn} = \frac{m}{n(m+1)k} + \frac{n+1}{n(m+1)kCL} + \frac{m(n+1)}{n(m+1)}$$

Compared to a BABPF, the BPF with an Extended Antiparallel configuration (EABPF) has one additional transmission zero, whereby the following equations apply:

$$\lim_{m\to 0 \wedge n\to 0} \omega_{EZ1} = \omega_{Z1}, \quad \lim_{m\to 0 \wedge n\to 0} \omega_{EZ2} \to \infty \tag{8}$$

$$\lim_{m\to \infty \wedge n\to \infty} \omega_{EZ1} \to 1, \quad \lim_{m\to \infty \wedge n\to \infty} \omega_{EZ2} \to 1 \tag{9}$$

where ω_{Z1} is defined in (3)

Equations (6-9) show that, compared to the BABPF, the EABPF has a much higher level of freedom in terms of the number and position of transmission zeroes. Instead of one zero in the lower stopband, whose position cannot be altered without a significant change in other BABPF characteristics (primarily the width of the passband), the EABPF has one zero in each stopband whose position can be altered within broad boundaries by varying parameter m, that is n, from 0 to +∞.

For designing the filter it is useful to derive expressions for the value of parameters m and n depending on the given values ω_{EZ1} and ω_{EZ2}. This can be obtained by solving (6) and (7) as a system of two equations for unknowns m and n, which gives:

$$m = \frac{1}{2}\left(\frac{\omega_{Z1}^2 - \omega_{EZ1}^2}{\omega_{EZ1}^2 - 1} + \frac{\omega_{Z1}^2 - \omega_{EZ2}^2}{\omega_{EZ2}^2 - 1} + \frac{(\omega_{Z1}^2 - 1)\sqrt{u+v}}{(\omega_{EZ1}^2 - 1)(\omega_{EZ2}^2 - 1)} \right) \tag{10}$$

$$n = \frac{2\left(k^{-1}+1\right)}{\left(\omega_{EZ1}^2 - 1\right) + \left(\omega_{EZ2}^2 - 1\right) + \sqrt{u+v}} \tag{11}$$

where: $u = (\omega_{EZ1}^2 - \omega_{EZ2}^2)^2$ and $v = 4\dfrac{(\omega_{EZ1}^2 - 1)(\omega_{EZ2}^2 - 1)}{k(\omega_{Z1}^{-2} - 1)}$

In most practical cases, it applies that u is much greater than v, which, applied to (10) and (11), leads to significant simplification:

$$m \approx \frac{\omega_{Z1}^2 - \omega_{EZ1}^2}{\omega_{EZ1}^2 - 1} \tag{12}$$

$$n \approx \frac{k^{-1}+1}{\omega_{EZ2}^2 - 1} \tag{13}$$

Equations (12) and (13) confirm the intuitive assumption that parameter m has a dominant influence on the zero in the lower stopband (ω_{EZ1}), while parameter n has a dominant influence on the zero in the upper stopband, which means that the position of zeroes can be independently altered within broad boundaries: $\omega_{Z1} \leq \omega_{EZ1} \leq 1$, that is $1 \leq \omega_{EZ2} < \infty$. The ability of a precise controlling of the narrowband frequency ranges with maximum attenuation, especially near the passband region, is a very desirable characteristic of a BPF which is especially employable in suppressing dominant parasitic signals such as the „leakage“ of local oscillator signals or their harmonics, as well as the higher harmonics of the useful RF signal.

Using (10) and (11) and the expressions from Table 2 it is possible to determine the values of all inductances and capacitances of subnetwork N_Y, that is N_Δ, which forms the EABPF with transmission zeroes at desired frequencies ω_{EZ1} and ω_{EZ2}. Such a EABPF can be based on any BABPF prototype defined by the value of independent parameter C as well as L and k parameter values determined by (1) and (2). Table 3 lists the component values for four examples of the EABPF, all based on the same BABPF prototype with $C = 4$, with four different combinations of transmission zeros ω_{EZ1} and ω_{EZ2}: a) 0.5, 2.0; b) 0.5, 1.5; c) 0.75, 1.5) and d) 0.75, 2.

TABLE 3

COMPONENTS, PARAMETERS AND FEATURES FOR FOUR EABPFS DERIVED FROM THE BABPF PROTOTYPE WITH $C = 4$

	ω_{EZ1}	m	L_1	C_1	C_3	C_2	L_2	n	ω_{EZ2}
a	0.5	0.178	0.378	17.47	23.737	2.875	0.098	0.391	2.0
b	0.5	0.174	0.376	17.92	23.737	2.059	0.236	0.942	1.5
c	0.75	1.017	0.647	3.064	23.737	2.062	0.235	0.940	1.5
d	0.75	1.023	0.649	3.047	23.737	2.876	0.098	0.391	2.0

Y-Δ

	BW_{3dB}	$S_{21}(\omega_{ML})$	ω_{ML}	C_{13}	C_{12}	C_{23}	$S_{21}(\omega_{MH})$	ω_{MH}	ω_c
a	24.8%	-15.2 dB	0.076	9.409	1.140	1.548	-44.1 dB	3.031	0.9991
b	19.3%	-17.3 dB	0.079	9.732	0.844	1.118	-36.8 dB	2.047	1.0017
c	13.4%	-20.1 dB	0.222	2.520	0.219	1.696	-39.3 dB	2.092	1.0033
d	15.7%	-18.8 dB	0.195	2.438	0.296	2.302	-47.7 dB	3.070	1.0031

Frequency characteristics of S parameters of four different EABPFs are shown in Fig. 4, and the distinctive values are also outlined in the shaded columns of Table 3. It can be noticed that the change in frequency of transmission zeros ω_{EZ1} and ω_{EZ2} simultaneously affects the width of the passband, selectivity, as well as the value of the insertion loss in the upper and lower stopband. The EABPF is significantly more suitable for shaping a filter characteristics, since this can be performed by changing three parameters (C, m, n), instead of just one (C) as was the case with the BABPF.

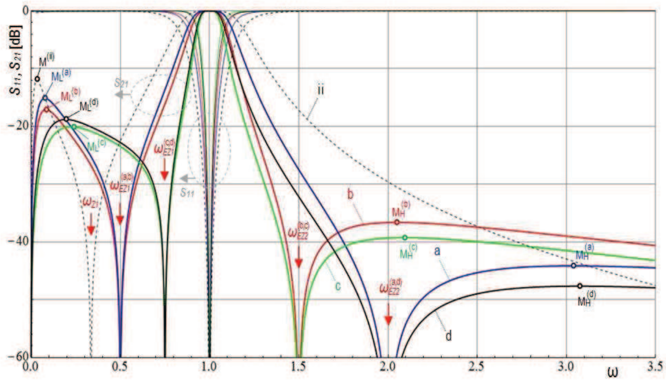


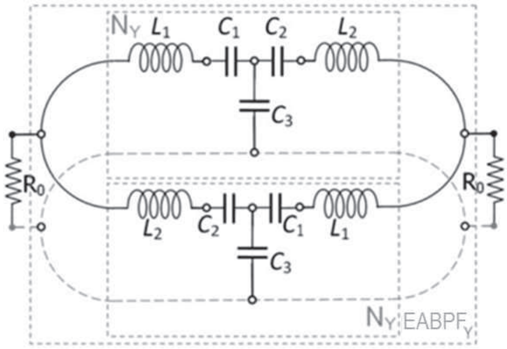
Fig. 4. Comparison of S parameter frequency characteristics of four EABPFs from Table 3 and a basic BABPF (*ii*) from Table 1

The examples in Table 3 also show that a large difference can exist in the capacitance values between different EABPF filters obtained from the same BABPF prototype, depending on the desired frequency of the transmission zeros. This way, for example, by comparing the Δ configurations, it can be seen that the sum of capacitances can vary from ΣC_{ij}=4.435 in example (c), to ΣC_{ij}=12.097 in example (a). This difference in capacitance values will also cause a difference in the physical size of the filters if they are realized in one of the planar technologies. This means that the comparison in the sizes of the planar filters, which is one of the frequently used criteria for the valorisation of the suggested solutions and configurations in literature, can be considered an objective norm only if all the characteristics of the observed filters are mutually comparable, and not only the standard characteristics such as center frequency, width of the passband, relative permittivity and thickness of the substrate.

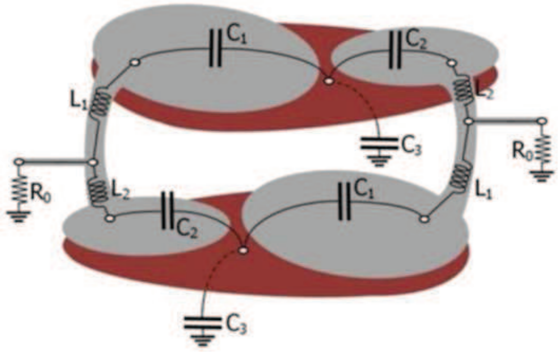
IV. The Application of the EABPF for the Realization of Planar Filters

EABPFs are suitable for realization in all known filter fabrication techniques due to a small number of different components which can easily be realized, either with different types of elements with concentrated parameters, either with printed transmission lines (TLs). Fig. 5a and Fig. 6a show EABPFs composed of N_Y and N_Δ subnetworks, respectively, and marked as $EABPF_Y$ and $EABPF_\Delta$. Comparatively, Fig. 5b and Fig. 6b show the corresponding general topologies of these filters for the realization in one of the planar techniques.

The common characteristic of the topology of these filters is that they have a central symmetry, which is a consequence of the antiparallel configuration of the filters. By comparing them it can be concluded that the $EABPF_\Delta$ is more suitable for practical realization in one of the planar technologies since it demands only two conductive layers, out of which one is the reference ground while the other layer contains all the other filter planar components, whereby the conductive layers are separated by one layer of a low-loss dielectric substrate.

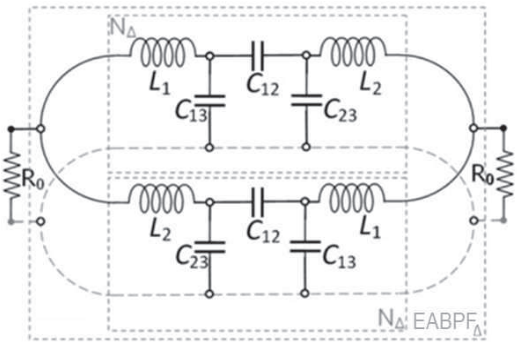


(a)

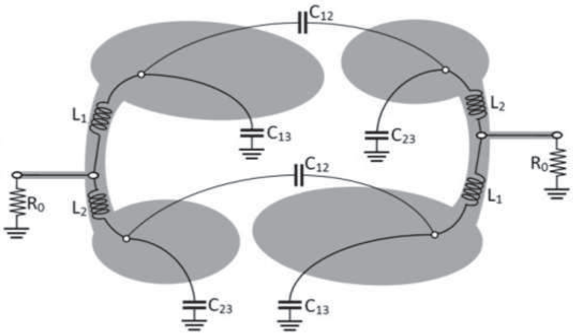


(b)

Fig. 5. (a) Circuit diagram of $EABPF_Y$ and (b) the corresponding general planar topology (for beter visibility, the reference ground plane and the dielectric layers are not shown)



(a)



(b)

Fig. 6. (a) Circuit diagram of $EABPF_\Delta$ and (b) the corresponding general planar topology (for beter visibility, the reference ground plane and the dielectric layer are not shown)

On the other hand, $EABPF_Y$ requires three conductive layers separated by two dielectric layers. Furthermore, as can be seen in Table 3, the total value of capacitances is, in the case of $EABPF_Y$, always significantly greater than in the case of the equivalent $EABPF_\Delta$, so that the planar realization of $EABPF_Y$ will always be larger than $EABPF_\Delta$ and therefore less suitable in almost all practical realizations. The only exception could be an application at higher microwave frequencies at which the required capacitance values of $EABPF_\Delta$ are too small for the available precision of the realization, in which case the application of $EABPF_Y$ could be a more suitable choice.

The multilayered substrate with grounded upper and lower conductive layers, which form a well-defined reference ground, would be the most suitable medium for the fabrication of both variants of the EABPF. Such groundwork would enable the realization of all possible values of inductances and capacitances needed for achieving various frequency characteristics of this type of bandpass filter.

V. The Application of the EABPF for Realization in the Microstrip Technique

Simple planar structures, such as microstrip, are suitable for the fabrication of $EABPF_\Delta$ versions of the filter which require relatively small values of series capacitance C_{12}, which can be realized by lateral coupling of the filter resonators. The other filter components can be realized by TLs which, for the realization of grounding capacitances (C_{13}, C_{23}) need to have characteristic impedances that are as low as possible, whereas the series inductances (L_1, L_2) should be realized with TLs having characteristic impedances as high as possible.

The electrical lengths (θ_L, θ_C) of inductive and capacitative TLs can be calculated using the following expressions for the approximation of inductance and capacitance with ideal TLs:

$$\theta_L = \frac{R_0 L_n}{Z_L} \quad (12)$$

$$\theta_C = \frac{Z_C C_n}{R_0} \quad (13)$$

where: L_n and C_n – normalized values of the prototype filter components; Z_L and Z_C – inductive (high) and capacitative (low) characteristic impedances; R_0 –filter's termination value

The corresponding microstrip TLs' lengths (l_L, l_C), for the given or favored corresponding TLs' widths (w_L , w_C), can be calculated by applying (12) and (13) in approximate expressions for microstip TLs [26], whereby the following expressions are obtained:

$$l_L = \frac{c_0 R_0 L_n}{120\pi \; f_c \ln\left(\frac{8h}{w_L} + \frac{w_L}{4h}\right)} \qquad \text{for } w_L \le h \quad (14)$$

$$l_C = \frac{60 c_0 C_n}{\varepsilon_e f_c R_0 \left(\frac{w_C}{h} + 1.393 + 0.667\ln(\frac{w_C}{h} + 1.444)\right)} \qquad \text{for } w_C \ge h \quad (15)$$

where: c_0 – the speed of light; f_c – the center frequency; w_L, w_C and l_L, l_C - widths and lengths of the microstrip TLs; h – the thickness of the dielectric; ε_e – the effective dielectric constant of the microstrip TL approximately determined by relative permittivity (ε_r) as: $\varepsilon_e=[\varepsilon_r+1+(\varepsilon_r-1)(1+12\, w_C/h)^{-1/2}]/2$.

By applying expressions (14) and (15) on the values of the $EABPF_\Delta$ prototype components, for example those given in Table 3, version *c*, the most important filter dimensions for the final layout of the microstrip filter can be obtained.

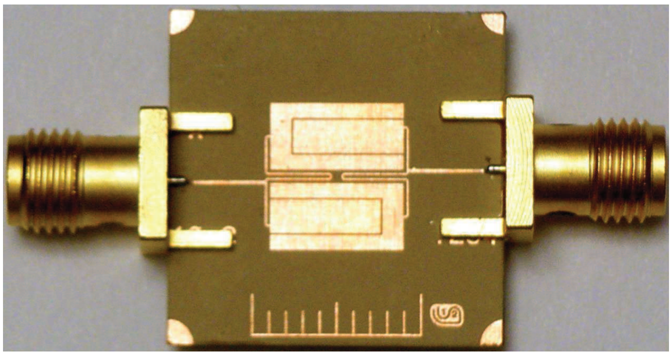


Fig. 7. Photo of the realized microstrip $EABPF_\Delta$

Figure 7 shows a photo of an assembled microstrip filter with SMA panel connectors. The filter is realized by the standard lithographic process with the line and gap width tolerances of ±5 µm. The filter's overall dimensions (excluding I/O 50Ω lines) are 8.52×8.74 mm ($0.076\lambda_g \times 0.078\lambda_g$) for f_c = 1 GHz and the RO3010 substrate with nominal ε_r = 10.2, h = 0.254 mm and $\tan\delta$ = 0.0023. The majority of the filter's dimensions are obtained from $EABPF_\Delta$ prototype values by closed-form expressions, with a limited use of electromagnetic analysis to determine an appropriate gap width for the correct value of C_{12} capacitance, as well to adjust the length of meandering inductive microstrip lines for the correct electrical length. The design process, as well as the obtained results, is comprehensively described in [25].

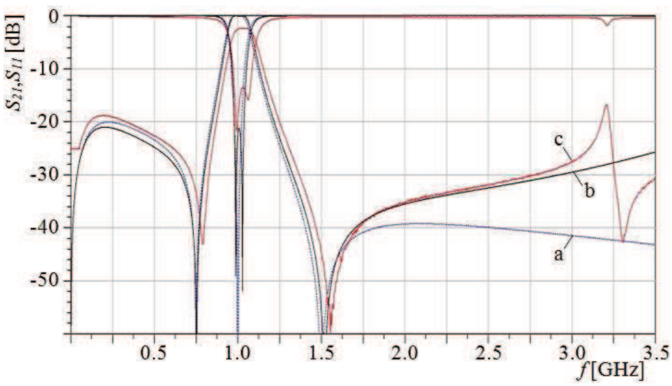


Fig. 8. Filter's S-parameters comparison between: a) scaled $EABPF_\Delta$ prototype; b) ideal TL model; c) measured results

As shown in Fig. 8, the measured electrical characteristics of the realized microstrip filter are, even without additional optimization, very similar to the designed ones (trace a), as well as to those obtained from an ideal TL model (trace b).

VI. Conclusion

This paper gives a detailed analysis of the extended version of a bandpass filter with an antiparallel configuration. The basic version of this type of filter is modified by introducing additional inductance and capacitance. This way, in the transmission characteristics of the filter, a pair of transmission zeros close to the passband of the filter are obtained, and their frequency position can be explicitly determined and adjusted to a wide extent. Equations are derived in the paper with which, in the form of closed-form expressions, a connection between the specified, i.e. suitable frequencies of the transmission zeros and parameters which define the values of all the filter components is established. Two types of an extended version of BPF are identified, which mutually differ based on the "Y", that is "Δ" configuration of the connection between the capacitances of the filter. Among them the Y configuration is more convenient for the analysis of filter characteristics, while the Δ configuration is more suitable for practical implementations. The proposed filter configuration is significantly more suitable for practical application for fabrication in a large number of planar technologies. In the paper, a closed-form equation for the calculation of the physical dimensions of the microstrip TLs based on the values of inductive and capacitive elements of the normalized filter prototype is derived.

Acknowledgement

This paper was partially supported by the Ministry of Education, Science and Technological Development of the Republic of Serbia under grant TR-32024.

References

[1] L. Zhu, V. Devabhaktuni, C. Wang and M. Yu, "A bandpass filter with adjustable bandwidth and predictable transmission zeros," *Int J RF and Microwave Comp Aid Eng*, vol. 20, no. 2, pp. 148–157, Mar. 2010.

[2] J. Long, C. Li, W. Cui, H. Jiangtao and R. Lixin, "A Tunable Microstrip Bandpass Filter With Two Independently Adjustable Transmission Zeros," *Microwave and Wireless Comp. Letters, IEEE*, vol. 21, no. 2, pp. 74-76, Feb. 2011.

[3] C. Jianxin, Y. Mengxia, X. Jun and X. Quan, "Compact microstrip bandpass filter with two transmission zeros," *Electronics Letters*, vol. 40, no. 5, pp. 311-313, Mar. 2004.

[4] Y.-Z. Zhu, Y.-J. Xie, "Novel microstrip bandpass filters with transmission zeros," *Progress In Electromagnetics Research*, vol. 77, pp. 29-41, 2007.

[5] V. Osipenkov, S. G. Vesnin, "Microwave filters of parallel cascade structure," *IEEE Trans. Microwave Theory Tech.*, vol. 42, no. 7, pp. 1360-1367, Jul 1994.

[6] C.-M. Tsai, S.-Y. Lee and C.-C. Tsai, "Performance of a planar filter using a 0° feed structure," *IEEE Trans. Microw. Theory Tech.*, vol. 50, no. 10, pp. 2362- 2367, Oct 2002.

[7] J. Zhang, J.-Z. Gu, B. Cui and X.-W. Sun, "Compact and harmonic suppression open loop resonator bandpass filter with tri-section SIR," *Progress In Electromagnetics Research*, vol. 69, pp. 93-100, 2007.

[8] G. Dai, M. Xia, "Novel miniaturized bandpass filters using spiral-shaped resonators and window feed structures," *Progress In Electromagnetics Research*, vol. 100, pp. 235-243, 2010.

[9] S.-Y. Lee, C.-M. Tsai, "New cross-coupled filter design using improved hairpin resonators," *IEEE Trans. Microw. Theory Tech.*, vol. 48, no. 12, pp. 2482-2490, Dec. 2000.

[10] A. Namsang, P. Akkaraekthalin, "A Microstrip bandpass filter using asymmetrical stepped-impedance resonators," *Microwave Conference*, APMC 2007, Asia-Pacific, pp. 1-4, 11-14 Dec. 2007.

[11] J.-T. Kuo, E. Shih, "Microstrip stepped impedance resonator bandpass filter with an extended optimal rejection bandwidth," *IEEE Trans. Microwave Theory Tech.*, 2003, vol. 51, no. 15, pp. 1154–1159.

[12] W.-N. Chen, W.-K. Chia, C.-F. Yang and C.-L. Shih, "Improving the Coupling Characteristics of Bandpass Filters by Using Multilayer Structure and Defect Ground Units," *Microwave Conference*, APMC 2007, Asia-Pacific, pp. 1-3, 11-14 Dec. 2007.

[13] Q.-X. Chu, H. Wang, "Planar Quasi-Elliptic Filters with Inline EM Coupled Open-Loop Resonators," *IEEE MTT-S International Microwave Workshop Series on Art of Miniaturizing RF and Microwave Passive Components*, pp. 47-50, Dec. 2008.

[14] W. Krongkitsiri, C. Mahatthanajatuphat, P. Akkaraekthalin, "Wideband bandpass filters using parallel-coupled sirs with wide spurious supprression," *Progress In Electromagnetics Research C*, vol. 27, pp. 69-83, 2012.

[15] K. C. Lee, H. T. Su and M. K. Haldar, "Compact microstrip bandpass filters using triple-mode resonator," *Progress In Electromagnetics Research Letters*, vol. 35, pp. 89-98, 2012.

[16] S. Jovanovic, A. Nesic, "Microstrip bandpass filter with new type of capacitive coupled resonators," *Electronics Letters*, vol. 41, no. 1, pp. 19-21, January 2005.

[17] S. Jovanovic, A. Nesic, "Printed Band-Pass Filter at L Band," 14th IST Mobile & Wireless Communications Summit, Dresden, 19-23 June 2005.

[18] S. Jovanovic, A. Nesic, "A New Microstrip Bandpass Filter for UHF Range," *7th TELSIKS 2005*, Niš, Serbia, Proceedings of Papers, vol. 1, pp. 167-169, 28-30 Sep. 2005.

[19] S. Jovanovic, A. Nesic, "New Filter Type Suitable for Miniature Printed Bandpass Filters at RF & Microwave Frequencies," *European Microwave Conf., Paris*, pp. 25-28, 3-7 Oct. 2005.

[20] S. Jovanovic, A. Nesic, "Capacitive Coupled Microstrip Band-Pass Filter with Asymmetrically Capacitive Loaded Resonators," *Mediterranean Microwave Symposium MMS 2007*, Budapest, Hungary, pp. 113-116, 14-16 May, 2007.

[21] S. Prabhu, J. S. Mandeep and S. Jovanovic, "Microstrip Bandpass Filter at S Band Using Capacitive Coupled Resonator," *Progress In Electromagnetics Research, PIER*, vol. 76, pp. 223–228, 2007.

[22] J. S. Mandeep, S. Prabhu and S. Jovanovic, "Design of Capacitive Coupled Resonator Microstrip Filter," *Microwave and Optical Technology Letters*, vol. 50, no. 2, pp. 460 462, Feb. 2008.

[23] S. Jovanovic, "An Overview of Microwave Bandpass Filters with Capacitive Coupled Resonator," *Microwave Review*, vol. 16, no. 2, Dec. 2010.

[24] S. Jovanovic, "Analiza elementarnog filtra propusnika opsega sa antiparalelnom konfiguracijom," LVI Conference of ETRAN, Proceedings of Papers, MT 2.7, Zlatibor, Serbia, June 2012.

[25] S. Jovanovic, B. Milovanovic and M. Gmitrovic, "Theory and realization of simple bandpass filters with antiparallel configuration," *Progress In Electromagnetics Research*, vol. 136, pp. 101-122, 2013.

[26] D. M. Pozar, *Microwave Engineering*, 4th edition, John Wiley & Sons, New York.