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Wideband coupled-line BPF with high-selectivity based on parallel transmission line signal interference technique for cellular base stations applications

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ABSTRACT

This paper presents a novel approach involving a modified wideband parallel coupled line bandpass filter (BPF). The proposed modification aims to achieve both enhanced skirt selectivity and simplified configuration, while ensuring the prevention from discontinuities between adjacent segments. The improvements in the performance of the filter's structure are achieved through the integration of a signal-interference filtering model using parallel transmission line with distinct impedance values and electrical lengths at the input/output of the filter. This integration gives rise to the generation of multiple transmission zeros (TZs), thereby bolstering attenuation within both in-band and out-of-band frequency ranges. For the purpose of concept validation, a 3^{rd} wideband band-pass filter with f_0=2.45 Ghz, accompanied by a fractional bandwidth of 50% was designed and simulated using microstrip RO6010 substrate. The outcomes of the simulation exhibit good performance, characterized by minimal insertion loss, wide bandwidth and the presence of seven TZs within the passband, resulting in high selectivity and sharp stopband rejection level.

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1. INTRODUCTION

Nowadays, microwave filters with low insertion loss and high selectivity operating in the radio-frequency (RF)/microwave communication systems play a crucial and irreplaceable role, particularly in high-frequency elements of RF transceiver frontends. Among these, RF/microwave bandpass filters (BPFs) emerge as fundamental components. Their significance lies in their ability to accurately capture desired RF signals, effectively defining the operational bandwidth of the RF system. Over the last few years, many different techniques for designing wideband BPFs have been explored, such as multi-mode stepped impedance resonators, defected-ground structure (DGS), aperture backed-line, fork-shaped resonators, multilayer technology and coupled structures including interdigital filters, combine filters, hairpin filters, and parallel coupled-line filters [1]–[19].

BPFs with parallel coupled lines are the most commonly used filters because of their straightforward design synthesis, low fabrication cost and easier integration with other components [1]. However, to realise a wide bandwidth (FBW>10%), the unachievable small gap size presents the main limitation in the filter implementation and the fabrication tolerance increasement. To overcome this shortcoming, parallel three coupled-lines BPF is suggested [8], [9]. The gap coupling between lines is significantly relaxed and the fractional bandwidth of more than 10% is realised. Nonetheless, the filter still suffers from very poor selectivity and low insertion loss. Aperture technique is employed to improve the coupling strength in parallel coupled-line, so as so realizing wideband BPF [10], [11]. Folded fork-shaped structure with loaded open-ended superconductor-insulator-superconductor (SIS) is adopted in [12] to design compact wideband BPF. Cross-coupled wideband filters featuring a stub-loaded resonator and an extremely miniaturized SIRs are presented [13], [14]. Explore an alternative approach that involves investigating various DGS-shaped configurations etched into the ground plane as a means to enhance both the bandwidth and sharpness of microstrip BPFs [16]–[18]. However, the filtering behaviors are degraded due to increased losses resulting from the decreasement of the quality factor. Finally, in addition to the above techniques, multilayer printed board technology has attracted significant attention with the aim of boosting the response of classical coupled line structure [19], [20]. The aforementioned techniques give effective solutions to strengthen the coupling gap and enhancing the bandwidth. Nonetheless, most of them suffer from radiation loss, complexity in the structures design and are typically struggling dissatisfaction of high selectivity with transmissions zeros (TZs) at both sides of the passband.

Recenty, signal interference technique has been reported to design several microwave circuits [21]–[27]. In the wideband applications [21]–[23], parallel transmission lines modal of distincts characteristic impedances and electrical lengths and connected at their two ends establishes the fundamental filter's structure. The transmission signal applied at the first end propagates through those parallel paths and interferes at the other end with distinct phases and magnitudes. Varying these phases and magnitudes allows for the creation of multiple finite TZs resulting in sharp rejection attribute.

The approaches investigated the impact of the coupling gap in fabrication and bandwidth enhancement. But they did not explicitly address its influence on considering simple synthesis design, bandwidth and good performances simultaneously. To overcome this limitation, a wideband parallel coupled line filter based on interference technique is proposed in this study. Wide-band performance is accomplished by using three parallel coupled-lines BPF connected at it two ends to a bi-path transversal signal interference section. The design guidelines are provided, clearly indicating that multiple TZs can be created and independently controlled by the bi-path section. For validation, a wideband BPF operating at 2.45 GHz with a fractional bandwidth of 50% is designed. Based on the simulated results, the proposed design validates its potential appropriateness for cellular base stations applications.

This paper is structured in the following manner. Firstly, subsection 2.1 briefly summarizes the principle of signal interference filtering transversal section. In subsection 2.2, the technique outlined in subsection 2.1 is applied to construct a 3^{rd} order microstrip parallel coupled-line BPF with wide-bandwidth and enhanced performance. Section 3 focuses on examining the filter and conducting a comparison between the proposed microstrip BPF and the existing literature. Finally, the key conclusions drawn from this study are outlined in section 4.

2. METHOD

The geometrical structure of the proposed BPF is illustrated in Figure 1. It is made up of three parallel coupled-lines BPF connected at it two ends to a bi-path transversal signal interference section. The operational concept of this filter configuration involves acquiring a highly selective wideband BPF behavior by promoting constructive/destructive signal for the desired passband to create TZs. The transversal signal interference sections are folded in order to reduce the overall circuit size.

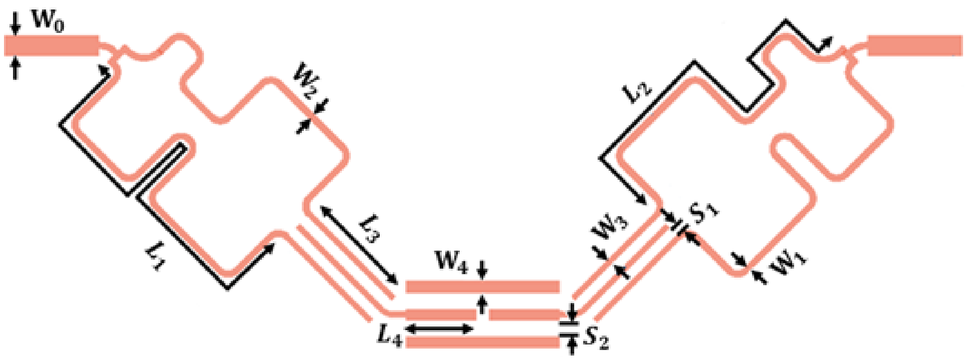


Figure 1. Design of the suggested wideband band-pass filter grounded on interference technique

2.1. Signal interference technique principle

The configuration of the used transversal section is depicted in Figure 2. As it can be observed, it consists of two different electrical transmission paths in parallel and connected to each other at their two input/output. A transmission line segment having an electrical length and characteristic impedance referred to as θ_1 and Z_1, respectively, is placed in path 1. In the second path, another transmission line is located with characteristic impedance Z_2 and electrical length θ_2. The operating interference principle consists of dividing the signal coming from one end into two components with the same frequency, and propagating at the other side with different amplitudes and phases depending on the characteristic parameters of the two-lines.

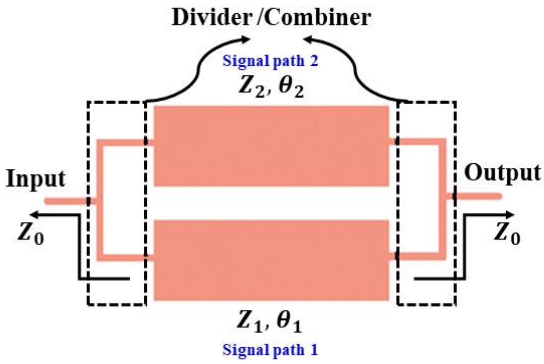


Figure 2. Configuration of the transversal transmission-line section

Building upon this principle allows to achieve multiple TZs, thereby contributing to the enhancement of performance. The location of the TZs is obtained by setting the scattering parameter $|S_{21}| = 0$. Thus, TZs will be established at the frequencies where the condition (1) is fulfilled:

$$\cot\left(\theta_{z1}\left(\frac{f_z}{f_0}\right)\right)\cot\left(\theta_{z2}\left(\frac{f_z}{f_0}\right)\right)+\frac{Z_2}{Z_1}=0 \quad (1)$$

Two sets of solutions can be derived. The first set is determined by considering the following conditions: $\theta_{Z1} = (2i-1)\pi/2$ and $\theta_{Z2} = (2j+1)\pi/2$. As shown in Figure 3(a), symmetrical TZs about the center frequency f_0 are obtained when $i = j = 1,2,3 \ldots$. By choosing $i \neq j$, the TZs appeared only at the lower or upper stopband with respect to f_0 as illustrated in Figure 3(b). For the second set, the symmetrical TZs locations are obtained if the conditions $\theta_{Z1} = i\pi/2$, $\theta_{Z2} = (j+2)\pi/2$ and $i = j = 2,4,6\ldots$ are satisfied as depicted in Figure 4.

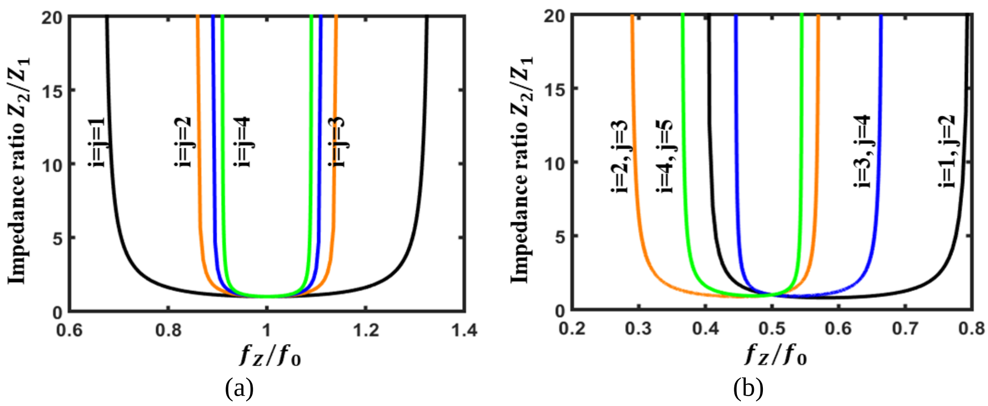


Figure 3. Variation of TZs location versus different values of impedance ratio for set 1; (a) i=j and (b) i≠j

According to the prior statement, it is clear that the TZs locations and bandwidth can be controlled by acting on both electrical-length and impedance parameters. Wide bandwidths are acquired as lower electrical lengths are chosen. Moreover, the higher impedance ratio is, the greater zero separation and stopband rejection level are.

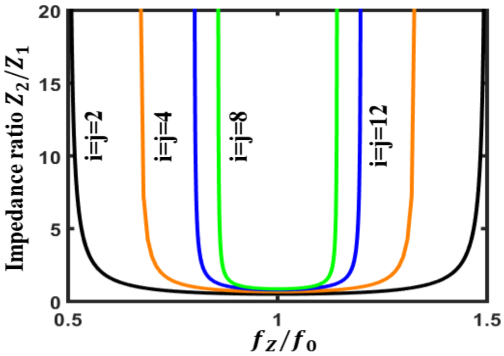


Figure 4. Variation of TZs under different values of impedance ratio for set 2

2.2. Bandpass filter design procedure

The design procedure of each quarter wavelength coupled stage is the same as in [8]. Each three-line section of six-port network can be transformed to two-port using currents and voltages termination conditions and then approximated by an admittance inverter in the middle of two quarter-wave transmission lines. The admittance inverter J values can be expressed as (2) to (4):

$$J_{01} = \sqrt{\frac{\omega b_1}{Z_0 g_0 g_1}} = \frac{1}{Z_0}\sqrt{\frac{\omega\theta}{g_0 g_1}} \tag{2}$$

$$J_{m,m+1} = \omega\sqrt{\frac{b_m b_{m+1}}{g_m g_{m+1}}} = \frac{\omega\theta}{\sqrt{g_m g_{m+1}}}\ ,\ m = 1,2,\ldots i-1 \tag{3}$$

$$J_{i,i+1} = \sqrt{\frac{\omega b_i}{Z_0 g_i g_{i+1}}} = \frac{1}{Z_0}\sqrt{\frac{\omega\theta}{g_i g_{i+1}}} \tag{4}$$

where, b_m's are the susceptance slope parameters, ω is the relative bandwidth and g_m's are the element values of the chosen lowpass prototype filter response. Z_0 present the terminating line characteristic impedance and is chosen as 50 Ω. Once J inverter is obtained, even/odd-mode characteristic impedances of each section can be deduced as (5) and (6):

$$Z_{even} \approx \frac{Z_0}{2}(J^2 {Z_0}^2 + JZ_0 + 1) \tag{5}$$

$$Z_{odd} \approx \frac{Z_0}{2}(J^2 {Z_0}^2 - JZ_0 + 1) \tag{6}$$

Overall, the design process of the proposed coupled-line BPF shown in Figure 1 can be succinctly outlined as follows:

- Calculate the admittance inverter J and even-odd mode impedances (Z_{even}, Z_{odd}) according to the design (2) to (6). Therefore, the initial line widths W_j and gaps S_j are determined with the given center frequency and substrate characteristics. The line length of each section is calculated using (7) [4]:

$$L = \frac{c}{4f_0\sqrt{\varepsilon_{eff}}} \tag{7}$$

where ε_{eff} is the effective dielectric constant, and it can be expressed as (8) and (9):

$$\varepsilon_{eff} = \frac{\varepsilon_r+1}{2} + \frac{\varepsilon_r-1}{2}\left(\frac{1}{\sqrt{1+\frac{12h}{t}}}\right) + 0.04\left(1-\frac{w}{h}\right)^2, for\ \frac{w}{h} < 1 \tag{8}$$

$$\varepsilon_{eff} = \frac{\varepsilon_r+1}{2} + \frac{\varepsilon_r-1}{2}\left(\frac{1}{\sqrt{1+\frac{12h}{t}}}\right),\ for\ \frac{w}{h} \geq 1 \tag{9}$$

- Determine the electrical and physical values Z_1, Z_2, θ_1, and θ_2 of the transversal filtering transmission line section for the chosen set of solution according to the desired propreties.
- As per the condition in (1) and the two sets of solutions, refer to Figures 3 and 4 to locate the TZs and their corresponding impedance ratio Z_2/Z_1.
- Parameter optimization is required to ensure that the filter response meet the desired performance.

To demonstrate the feasibility of the design procedure described above, a third order microstrip coupled three-line filter with Chebyshev response (0.5 dB ripple) is designed and optimized on Rogers 6010 substrate having a dielectric permittivity of $\varepsilon_r = 10.2$, a thickness of $h = 1.27$ mm and a loss tangent of 0.0023. The center frequency is selected to be $f_0 = 2.45$ GHz. The electrical parameters of the transversal section Z_1, Z_2, θ_1, and θ_2 were set to 106 Ω, 90 Ω, 180°, and 360°, respectively. As a result, the impedance ratio is $Z_2/Z_1 = 0.89$. By mapping this value on Figure 4, the corresponding band-in-band TZs f_{z1} and f_{z2} are expected to be obtained at 1.745 GHz and 3.14 GHz, respectively. The finally optimized physical values are labeled in Table 1.

Table 1. Physical dimensions of the designed filter

Symbol	Value (mm)
L_1	21.4
L_2	19.7
L_3	12.69
L_4	11.8
S_1	0.304
S_2	0.41
W_1	0.181
W_2	0.11
W_3	0.19
W_4	0.304

3. RESULTS AND DISCUSSION

The electromagnetic (EM) simulated frequency response is accomplished using both PathWave advanced design system (ADS) and HFSS as shown in Figure 5. A good agreement and excellent performances are exhibited, with the exception of certain frequency shift due to the coupling effects of the coupled-lines and the numerical approach used in each simulator. The center frequency is located at 2.45 GHz with a fractional bandwidth of 50% which lead to cover L-band and S-band applications. The in-band insertion loss and return loss are better than 0.6 dB and 15 dB, respectively, for the whole band. As expected from the transversal section theory, two TZs are obtained close to the passband at 1.76 dB and 3.12 dB. Besides, five other TZs are created at 0.518 dB, 1.705 dB, 3.22 dB, 4.31 dB, and 4.42 dB due to the optimal impedance selection. An optimal selection of the impedances Z_1 and Z_2 leads to split the reffered TZs into multiple TZs located around the passband. The side-band selectivity is satisfactory with better than 30 dB rejection level, and the upper stopband attenuation is about 25 dB. Table 2 shows the detailed comparison between the parallel coupled-line filter proposed in this work and others state-of-the-art reported ones. Distinctly, the filter has the merits of low insertion loss, wide bandwidth, high selectivity, sharp rejection stopband and simple design process.

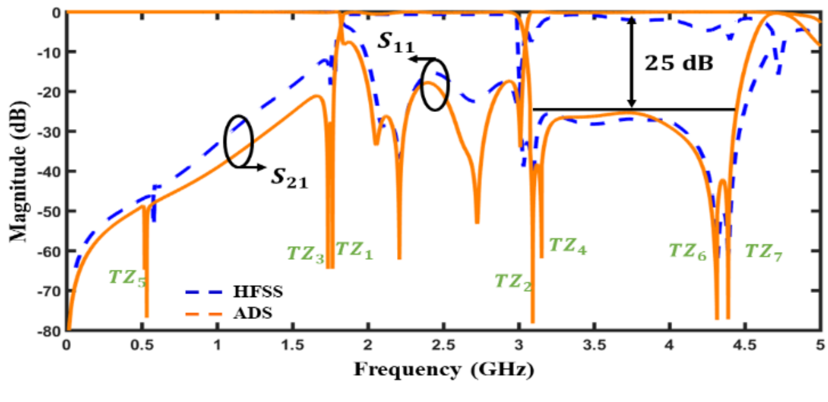
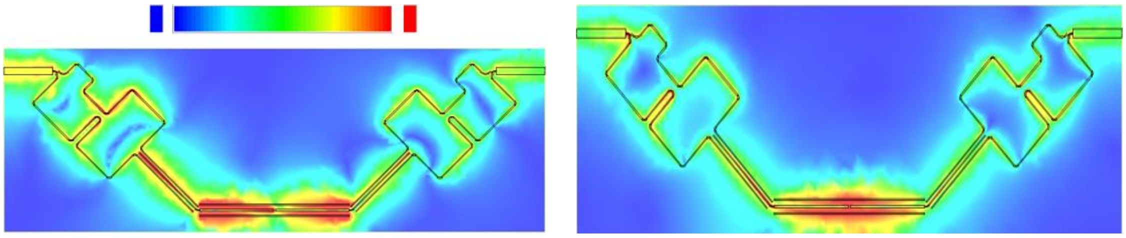


Figure 5. Simulated results of proposed parallel coupled-line BPF

Table 2. Comparison between the proposed wideband coupled-line BPF and those reported in literary works

References	CF (GHz)	FBW (%)	IL (dB)	RL (dB)	TZs	Stopband rejection level
[12]	2.2	45	1.4	10	4	20
[13]	1	21	2.05	NA	1	30
[14]	10.81	44.5	0.6	20	3	20
[18]	2.42	19.4	1	24.5	3	20
Proposed BPF	2.45	50	<0.6	>15	7	>25

Figures 6(a) and (b) depict the electric and the magnetic field distribution at the center frequency. It can be noticed that all the colors on the scale are spread out across the surface of the coupled lines. Red color represents the stronger coupling effect, whereas the blue color indicates a weaker one.



(a) (b)

Figure 6. Field distribution of the proposed filter at the center frequency; (a) electric field and (b) magnetic field

4. CONCLUSION

Transversal signal interference section has been employed in this work to design wideband BPF with enhanced proprieties. It made up of two distinct transmission-line segments in parallel and connected to coupled three-line structure, thereby creating two paths for signal propagation. Design method and rules are discussed to reveal the physical parameters and desired characteristics of the filter in accordance with its specifications. The proposed filter realizes about 50% of fractional bandwidth while creating seven TZs that improve the selectivity and stopband rejection level, which demonstrates its suitability for cellular base stations applications. As a limitation, the filter exhibits a large area of $0.5\,\lambda_0 \times 0.2\,\lambda_0$. To mitigate this drawback, a potential avenue for future could involve the incorporating of cascade stepped impedance signal interference sections into the design.

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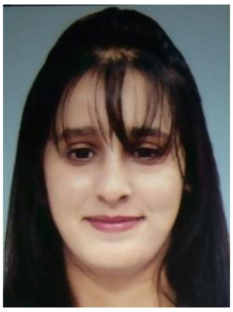

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