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Implementing Quasi-Elliptic Microstrip Filters Using Terminating Half Sections

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***Abstract*—This letter presents a technique for the introduction of transmission zeros (TZs) using half section matched terminations for implementing high selectivity filters in planar technology. The proposed technique does not require any cross coupling between nonadjacent resonators or multiple signal paths, and it can be easily incorporated on any planar filter using lumped, quasi-lumped, and/or distributed elements depending on the desired response. Two different bandpass filter topologies are employed for applying the proposed technique. Both prototypes are designed, manufactured, and measured. The measured results show that the TZs can be independently located very close to the passband edges, without degrading the original filter response.**

***Index Terms*—Lumped elements, quasi-elliptic filter, transmission zeros (TZs).**

I. Introduction

FILTERS with elliptic or quasi-elliptic response are of great interest for obtaining higher selectivity with minimum filter order and more compact size. Thus, the introduction of transmission zeros (TZs) at the upper and lower stopband enables to improve rejection close to the passband, while keeping an acceptable level of out-of-band attenuation.

Several techniques for implementing elliptic and quasi-elliptic filters have been presented in the literature. While cross couplings between nonadjacent resonators are broadly used for the insertion of attenuation poles at finite frequencies [1], [2], other extended approaches are based on topologies having several signal paths connecting the filter source and load [3], [4]. Furthermore, extracted-pole synthesis techniques [5] can also be applied in complex conjugate symmetrical networks, and more recently alternate circuits based on a distributed implementation of the elliptic filter prototype have been proposed [6].

This letter presents a simple and general approach for the introduction of two independently adjustable TZs at both sides of the filter passband. The main advantage of the proposed technique is that it can be applied to any filter implementation, disregarding the filter order, topology, and resonator employed. The approach is based on the introduction of low-pass (LP) and high-pass (HP) terminating half-section cells at the filter input and output ports. These sections can be implemented using lumped, quasi-lumped, and/or distributed elements depending on the filter specifications.

This technique was already demonstrated for implementing bandpass filters with ultrawide stopband [7]. However, in that case, the TZs were not located very close to the passband edges, as required in a highly selective response. Therefore, two different filter prototypes with quasi-elliptic transfer functions have been designed, manufactured, and measured based on conventional bandpass filter structures without any intrinsic TZ. The proposed structures show that a pair of TZs can be located very close to each filter passband with negligible disturbance of the original filter response. The measured results present an excellent agreement with the simulations, and the proposed approach can be used for enhancing the selectivity of any filter topology and implementation in planar or integrated technology.

II. Terminating Half Sections

A. Ideal Lumped-Element Model

The circuit model of the proposed terminating network, based on a series reactance connected to a shunt resonator, was already developed in [7] for the LP section. The HP section can be easily derived following the same considerations, but changing the reactive element employed. The lumped-element model, design equations, and ideal response are summarized in Fig. 1.

Basically, these sections can introduce an adjustable TZ at the upper (i.e., LP section, series inductor L) or lower (i.e., HP section, series capacitor C) stopband of the filter, while providing a good matching at the center frequency f_0.

As can be easily recognized from the previous expressions, the nearness between f_0 and f_{TZ} has a great impact on the shunt-resonator element values. This effect is graphically represented in Fig. 2 by obtaining the resonator element values for $f_0 = 4$ GHz, $Z_F = 35\ \Omega$, and $Z_P = 50\ \Omega$ at different TZ locations.

For very selective filters, where the ratio $f_{\mathrm{TZ}}/f_0 \rightarrow 1$, the inductive part of the resonator (L_1 and L_2) becomes higher and the capacitive part (C_1 and C_2) becomes lower. In this case, the very high inductance values for L_1 and L_2 will prevent from using lumped inductors, which would require a high Q-factor value and a self-resonant frequency far enough from the filter passband. A distributed implementation in microstrip technology could also be considered by means of quasi-lumped elements or distributed resonators

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$$L = \frac{Z_F\sqrt{\frac{Z_P}{Z_F}-1}}{\omega_0} \qquad C = \frac{1}{\omega_0 Z_F\sqrt{\frac{Z_P}{Z_F}-1}}$$

$$C_1 = \frac{\omega_{TZ}^2-\omega_0^2}{\omega_{TZ}^2\omega_0}\cdot\frac{\sqrt{\frac{Z_P}{Z_F}-1}}{Z_P} \qquad C_2 = \frac{\omega_0^2-\omega_{TZ}^2}{\omega_{TZ}^2\omega_0}\cdot\frac{Z_P-Z_F}{Z_P Z_F\sqrt{\frac{Z_P}{Z_F}-1}}$$

$$L_1 = \frac{1}{\omega_{TZ}^2 C_1} \qquad L_2 = \frac{1}{\omega_{TZ}^2 C_2}$$

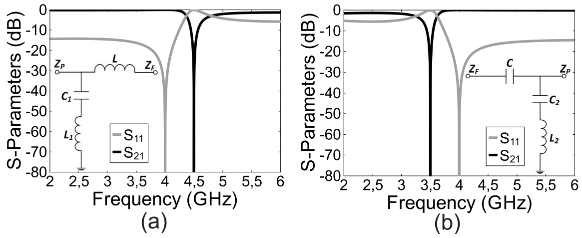

Fig. 1. Circuit schematics, equations, and responses of the ideal terminating half sections. (a) LP section and its S-parameters response for $f_{\mathrm{TZ}} = 4.5$ GHz, $f_0 = 4$ GHz, $Z_F = 35\ \Omega$, and $Z_P = 50\ \Omega$. $L = 0.9$ nH, $C_1 = 0.1$ pF, and $L_1 = 11$ nH. (b) HP section and its S-parameters response for $f_{\mathrm{TZ}} = 3.5$ GHz, $f_0 = 4$ GHz, $Z_F = 35\ \Omega$, and $Z_P = 50\ \Omega$. $C = 1.7$ pF, $C_2 = 0.16$ pF, and $L_2 = 13$ nH.

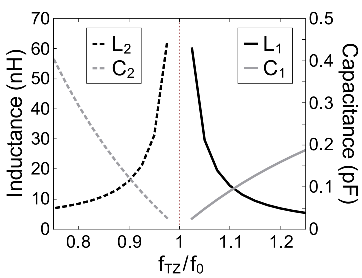

Fig. 2. Variation of C_1, L_1, C_2, and L_2 versus f_{TZ}/f_0 ratio for $f_0 = 4$ GHz, $Z_F = 35\ \Omega$, and $Z_P = 50\ \Omega$.

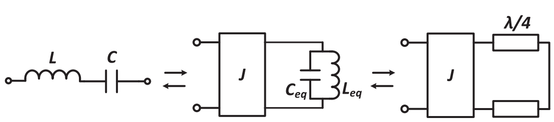

Fig. 3. Schematic synthesis and proposed equivalent circuit of the resonator of the terminating half sections for their practical implementation.

(e.g., open-ended $\lambda/4$ transmission line), nevertheless the required characteristic impedance would be totally unrealizable.

B. Practical Implementation With Inverters and Distributed Resonators

Thus, the equivalent scheme shown in Fig. 3 is proposed for implementing the series resonator by using an admittance inverter coupled to a parallel resonant circuit.

By equating the input impedance of both circuits, the next relationship can be obtained

$$J^2 = \frac{C_{\mathrm{eq}}}{L} = \frac{C}{L_{\mathrm{eq}}}. \tag{1}$$

As can be seen, a new degree of freedom is introduced by the admittance inverter. Given any resonator centered at f_{TZ}, a proper J-inverter can be found.

A parallel resonator can be implemented using a short-circuited transmission line of length $\lambda/4$ and characteristic impedance Z_0, where its equivalent capacitance and inductance are

$$C_{\mathrm{eq}} = \frac{\pi}{4\omega_0 Z_0} \quad L_{\mathrm{eq}} = \frac{4Z_0}{\pi\omega_0}. \tag{2}$$

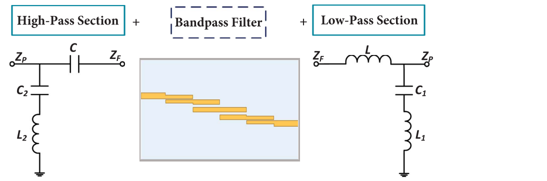

Fig. 4. Configuration of the filter with two TZs, one on each side of the passband.

So, in this case, the admittance inverter constant can be computed as

$$J = \sqrt{\frac{\pi}{4Z_0\omega_0 L}} = \sqrt{\frac{\pi\,\omega_0 C}{4Z_0}}. \tag{3}$$

Therefore, the use of an admittance inverter enables us to reach feasible element values for the equivalent resonator modeled by a parallel LC circuit. Thereby, the cells can be easily implemented at microwave frequencies using both lumped and distributed elements.

III. Design and Implementation

A. Parallel-Coupled Microstrip Bandpass Filter With Pseudoelliptic Response

To demonstrate the technique for designing high selective filters using the proposed terminations, the same filter proposed in [8] is considered. However, now the TZs are added symmetrically and closer to the passband, one on each side. Hence, both terminating sections (i.e., HP and LP) are needed. The configuration adopted is shown in Fig. 4.

The filter structure keeps unmodified, and the cascaded connection with the two terminating sections is simply made. The frequencies of the TZs are set symmetrically at 430 MHz away from f_0. Applying the equations of Fig. 1 for $Z_P = 50\ \Omega$, $Z_F = 35\ \Omega$, $f_0 = 4$ GHz, $f_{\mathrm{TZ1}} = 3.57$ GHz, and $f_{\mathrm{TZ2}} = 4.43$ GHz, the synthesis values of Table I are calculated. The technological realization is based on a hybrid solution of lumped and distributed elements. Thus, the series elements of the TZs networks (L and C) are implemented by lumped elements, whereas the elements of the parallel branch (i.e., resonators L_1, C_1, L_2, and C_2) are implemented using the proposed equivalent network based on inverters and distributed resonators of length $\lambda/4$. In this manner, the needed parameters for its implementation appear in Table I. As the required coupling values, J_{TZ1} and J_{TZ2}, are reasonably high, lumped elements (C_{TZ1} and C_{TZ2}) have also been used to achieve them.

The simulation results of the pseudoelliptic filter are compared with the response of the bandpass filter without TZs in Fig. 5(a). These results allow us to verify that the introduction of TZs using the method described in this letter significantly improves the out-of-band performance in terms of rejection and selectivity, without disturbing the in-band response of the original filter without TZs. Thus, insertion losses at midband frequency are about 3 dB, while the matching worsens slightly within the passband with a minimum value of 17.5 dB. Lastly, the TZs are correctly located achieving a rejection of 54.1 dB at 3.57 GHz and 52.3 dB at

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TABLE I
ELEMENTS OF THE "HP SECTION" AND "LP SECTION"

Parameter	Synthesis Value	Implementation
C	1.7 pF	1.2 pF 0402 AVX–Accu-P
C_2 L_2	0.13 pF 15 nH	$\lambda/4$ short-circuited line $[W = 1.1$ mm $(Z_0 = 50\ \Omega)$, $l = 10.25$ mm $(f_{TZ1} = 3.57$ GHz$)]$ $J_{TZ1} = 0.007 \rightarrow C_{TZ1} = 0.25$ pF
L	0.9 nH	1.3 nH Murata LQP15MN1N3W02
C_1 L_1	96.2 fF 13 nH	$\lambda/4$ short-circuited line $[W = 1.1$ mm $(Z_0 = 50\ \Omega)$, $l = 8.74$ mm $(f_{TZ2} = 4.43$ GHz$)]$ $J_{TZ2} = 0.006 \rightarrow C_{TZ2} = 0.2$ pF

4.43 GHz. This implies an improvement of more than 20 dB compared with the attenuation values of the filter without TZs.

B. Highly Selective Combline Microstrip Bandpass Filter

Another example with the TZs even closer to the passband and using a different filter topology is addressed. Thus, the design of a two-pole Chebyshev bandpass filter with 0.1-dB passband ripple and 5% FBW at a design frequency $f_0 = 4$ GHz employing a microstrip combline structure is now contemplated. The input/output filter impedance is set as $Z_F = 45\ \Omega$ just to end with a typical port impedance of $Z_P = 50\ \Omega$, once the terminations are cascaded. The combline resonators are of 85-Ω impedance and $\lambda/4$ long for no capacitive loading, and the TZs are located at 250 MHz away from f_0. Therefore, the elements values of the sections are $L = 0.6$ nH, $C_1 = 30$ fF, and $L_1 = 46$ nH for the LP Section, and $C = 2.7$ pF, $C_2 = 36$ fF, and $L_2 = 49$ nH for the HP section.

As in the previous example, the technological realization is based on a hybrid solution of lumped and distributed elements, and the only difference is the implementation of the TZs' coupling values (i.e., J_{TZ1} and J_{TZ2}), which are made through interdigital capacitors instead of SMD components. The simulation results of the combline pseudoelliptic filter are compared in Fig. 5(b) with the response of the combline filter without TZs. The presence of the TZs on the bandpass filter response can be seen at 3.75 GHz with a rejection of 16.76 dB and at 4.25 GHz with 29.45 dB of attenuation, while keeping the in-band response.

IV. EXPERIMENTAL RESULTS

The first filter has been fabricated in a 0.51-mm-thick Rogers RO4003C substrate ($\epsilon_r = 3.55$ and tan$\delta = 2.7 \cdot 10^{-3}$) with a finishing of 35-μm copper and 5 μm of Ni–Au. The second one has employed the same characteristics except for the substrate thickness, which is 1.524 mm. The manufactured prototypes are shown in Fig. 5. The photographs show the different parts of the filters according to the scheme outlined in Fig. 4. Their final dimensions are 9.74×3 cm^2 for the parallel coupled structure and 3.7×3.7 cm^2 for the combline. The introduced terminating networks require some additional space compared to the filter without TZs, although some modifications on the half-section layout (e.g., folding the series-shunt resonator) could be carried out in order to reduce this extra size. The measurements together with the simulations are plotted in Fig. 5(a) and (b). The experimental characterization fits almost perfectly with the simulations results. In the first filter [see Fig. 5(a)], the insertion losses at f_0 reach 3.3 dB, and the return losses are better than 18.4 dB over the entire bandwidth. The TZs present a minimal frequency shift, being located at 3.61 and 4.48 GHz, while achieving an attenuation of 45.2 and 52.2 dB, respectively. For its part, the combline example presents 2.2 dB of insertion losses at f_0 with return losses better than 16 dB on its passband. The TZs accomplish a very selective response in comparison with the response without them, appearing at 3.76 and 4.29 GHz obtaining 12.5 and 25.4 dB of rejection, respectively.

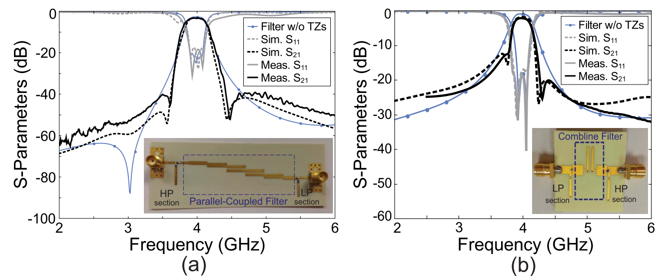

Fig. 5. Measured and simulated results of the fabricated prototypes in comparison with the bandpass filter responses without TZs. (a) Parallel-coupled structure. (b) Combline structure.

V. CONCLUSION

A novel technique for the introduction of TZs in filter responses has been presented. The technique is general, simple, and easy to apply, enabling us to introduce a pair of TZs very close to the filter passband. The usefulness of this technique was verified with a pair of pseudoelliptic filter examples, obtaining a significant improvement of rejection at prescribed frequencies.

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