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Inline Microstrip Bandpass Filter With Two Transmission Zeros and Increased Order Using Spurious Resonance of Frequency-Dependent Inverter

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***Abstract*—A design method for a class of fourth-order inline microstrip bandpass filter with two transmission zeros and 20% fractional bandwidth is presented. The filter consists of two quarter-wavelength transmission-line resonators coupled by a frequency-dependent inverter. The inverter is composed of two open-ended stubs that are connected by an interdigital capacitor and introduces two poles and two transmission zeros in the filter response. One of these poles is obtained from the spurious resonance of the capacitor, which leads to a very compact filter structure. An equivalent circuit model of the frequency-dependent inverter is provided along with a detailed coupling-matrix-based synthesis procedure to design the filter prototype. The design theory is validated with a constructed 2-GHz proof-of-concept prototype. Measured results are in close agreement with the synthesis and EM-simulated ones, hence verifying the devised design approach.**

I. Introduction

The still growing number of users of telecommunication services requires the development of advanced wireless systems with increased amount of transmitted data. Therefore, the demands imposed on the specifications of their microwave components become more stringent and pushes the designers to their very limits. To cope with this challenging task, more-efficient design methods for RF devices are required. In the case of coupled-resonators microwave filters, the synthesis techniques based on the coupling matrix offer reliable design approaches and high architecture flexibility when selecting a target implementation technology [1], [2]. Moreover, coupling-matrix-based synthesis methods allow the designer to easily include transmission zeros in the filter response, which is crucial for meeting the demands of high selectivity, potentially suppressing spurious bands, or patterning a specific group-delay profile for the filter [3], [4]. However, the introduction of transmission zeros in form of cross-coupled networks often leads to complicated filter topologies, making their fabrication more difficult and sensitive to tolerances. An alternative solution to avoid this issue is the realization of the transmission zeros by means of dispersive couplings. Such type of coupling is represented in the coupling-matrix model by an inverter with frequency-dependent impedance and can produce additional transmission zeros. Reported solutions using these classes of couplings have resulted in very-compact filters with multiple transmission zeros for inline circuit arrangements, which is very convenient in terms of easiness of physical implementation [5]. Thus, the utilization of frequency-dependent inverters opens up vast design possibilities. For example, in [6], a type of dispersive inverter that introduces two transmission zeros and one transmission pole was proposed. Such solution leads to further reduction of the filter size.

It is then widely accepted that dispersive structures have great potential to develop new and more-robust RF devices. Within this context, this work presents a design method for inline coupled-resonators microwave filters with frequency-dependent inverters. Specifically, the proposed filter is composed of two resonators that are coupled by the dispersive inverter. The inverter structure consists of two quarter-wavelength open-ended stubs that are connected through a capacitor. Normally, such inverter produces two transmission zeros and one pole above them. However, in this specific design, by properly choosing dimensions and number of sections of such interdigital capacitor, it can exhibit a spurious resonance very close to the main filter passband. Some related approaches were engineered for dual-mode resonator filters. For example, in [7], the additional resonance is introduced by removing a part of a dielectric in SIW cavity, while in [8] it is obtained by adding a metal post in the waveguide cavity. Such spurious resonance can be then exploited as an additional pole and included in the passband characteristic to effectively increase the filter order. In this manner, the resulting filter response exhibits four poles and two transmission zeros. The design method of such a filter is described in the following sections. Furthermore, for experimental-validation purposes, a microstrip prototype is manufactured and characterized.

II. Inverter model

The impedance matrix of the proposed dispersive inverter is needed during the coupling-matrix-based synthesis of the filter. Therefore, this section presents how to obtain such matrix.

The equivalent circuit presented in Fig. 1 is employed to model the behaviour of a physical circuit composed of two open-ended quarter-wavelength stubs interconnected by an interdigital capacitor. The referred stubs, which are responsible for transmission-zero generation, are represented by two series-type LC branches connected to ground. The interdigital capacitor is modelled using a capacitor between the two short-ended LC circuits and its spurious resonance is introduced by a series-type LC resonator followed by an impedance inverter. The impedance matrix Z_{inv} of the overall dispersive inverter

$$\boldsymbol{Z} = \begin{bmatrix} -jR_{in} + Z_{11in} & Z_{12in} & 0 & 0 \\ Z_{21in} & Z_{22in} + Z_{11inv} & Z_{12inv} & 0 \\ 0 & Z_{21inv} & Z_{22inv} + Z_{11out} & Z_{12out} \\ 0 & 0 & Z_{21out} & Z_{22out} - jR_{out} \end{bmatrix} \quad (1)$$

can be derived from $ABCD$ matrix formulation as described by the following equation:

$$\boldsymbol{A} = \boldsymbol{A_1} \cdot \boldsymbol{A_2} \cdot \boldsymbol{A_3} \quad (2)$$

where $\boldsymbol{A_1}$, $\boldsymbol{A_2}$, and $\boldsymbol{A_3}$ are the $ABCD$ matrices of the subcircuits of Fig. 1. Note that the inverter described by matrix $\boldsymbol{A_3}$ is a regular inverter with frequency-invariant behavior.

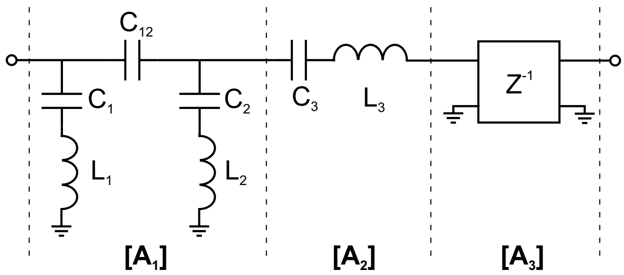

Fig. 1: Circuit model of the dispersive inverter.

III. Filter Design

The first step of the filter design method is finding the characteristics polynomials based on the provided specifications for the filter. This is done by using the technique described in [1]. Once the polynomials are known, the corresponding coupling matrix can be then derived. This is performed through the zero-pole optimization method reported in [9]. In this case, the coupling matrix has a form of 1, where R_{in} and R_{out} are the input and output impedances and $\boldsymbol{Z_{inv}}$ is the impedance matrix of the inverter. The input and output couplings are in a form of 90° coupled-line sections. In addition, the 90° line section acts as a quarter-wavelength resonator. $\boldsymbol{Z_{in}}$ and $\boldsymbol{Z_{out}}$ are the impedance matrices of the input and output coupled lines as follows:

$$Z_{11in/out} = Z_{22in/out} = \frac{-1}{2}(Ze_{in/out} + Zo_{in/out}) \cos\theta_{in/out} \quad (3)$$

$$Z_{12in/out} = Z_{21in/out} = \frac{1}{2}(Ze_{in/out} - Zo_{in/out}) \csc\theta_{in/out} \quad (4)$$

where $Ze_{in/out}$ and $Zo_{in/out}$ are the impedances of the even and odd modes of coupled-line section and $\theta_{in/out}$ is the electrical length of the lines. It should be noticed that entries of the coupling matrix are directly related to the circuit elements, so that finding the circuit network of the filter is straightforward. The initial circuit is composed of ideal elements, which are only an approximation to real components. Therefore, the final design should involve a more-sophisticated model of the filter structure to attain accurate results.

IV. Experimental Results

To validate the correctness of the proposed method, a fourth-order equiripple-type bandpass filter with two transmission zeros is selected to be designed. The filter is centered at $f_0 = 2.05$ GHz and has a bandwidth equal to 410 MHz with 20-dB minimum return-loss level. Transmission zeros are placed below the passband at 1.509 GHz and 1.772 GHz.

The characteristic polynomials are calculated using the approach described above. The roots of the polynomials are listed in Table I.

TABLE I: Roots of the characteristic polynomials of the filter in the s plane

poles	reflection zeros	transmission zeros
$-0.5499 + 1.2914j$	$0.8793j$	$-3.1121j$
$-0.9512 + 0.0901j$	$-0.9614j$	$-1.4625j$
$-0.1299 - 1.1067j$	$-0.6133j$	
$-0.5311 - 0.8351j$	$0.1350j$	

Next, the coupling matrix is found by using the procedure presented in the previous section. The obtained response is very close to the target response and only a slight optimization with the circuit simulator is needed to achieve the required in-band return-loss level. The responses of the filter calculated from the polynomials, the coupling matrix, and the initial circuit after the optimization are compared in Fig. 3. Note that the responses from the coupling matrix and initial circuit before optimization are identical. The filter structure resulting from the coupling-matrix formulation is shown in Fig. 2.

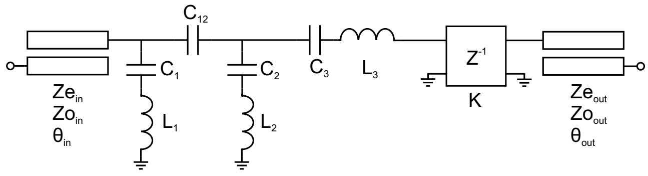

Fig. 2: Structure of the filter based on the coupling matrix. Elements values: $Ze_{in} = 76.68\ \Omega$, $Zo_{in} = 130.42\ \Omega$, $\theta_{in} = 85.69°$ at f_0, $L_1 = 2.04$ nH, $L_2 = 5.05$ nH, $L_3 = 5.29$ nH, $C_1 = 5.44$ pF, $C_{12} = 5.19$ pF, $C_2 = 1.6$ nF, $C_3 = 1.12$ pF, $K = 13.46\ \Omega$, $Ze_{in} = 77.92\ \Omega$, $Zo_{in} = 128.33\ \Omega$, and $\theta_{out} = 88.7526°$ at f_0.

The filter is implemented in microstrip technology on a substrate with dielectric constant $\epsilon_r = 3.48$, thickness of 0.762 mm, and loss tangent equal to 0.0018. Hence, the next step is finding the dimensions of the inverter modelled by the open-ended microstrip stubs and the interdigital capacitor. This is done by the optimization process—the characteristic of the microstrip model should be as close as possible to the response of the model from Fig. 1. When the inverter dimensions are found, the whole filter can be analyzed in the full-wave simulator. The final design was optimized in ADS Momentum. The dimensions of the physical structures are provided in Fig. 4. The characteristics of the final design and the fabricated filter prototype are presented in Fig. 5.

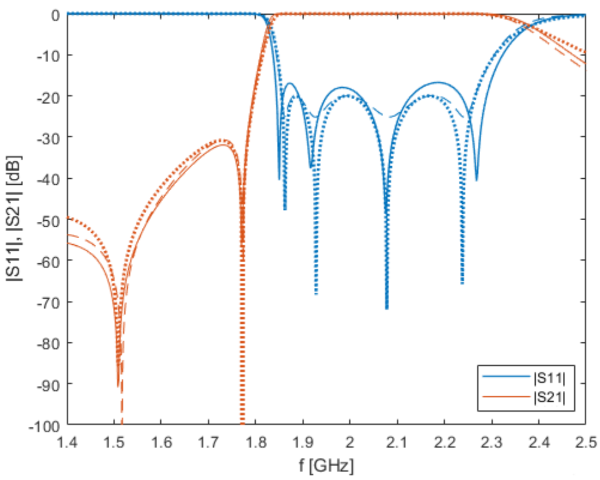

Fig. 3: Power transmission ($|S_{21}|$) and reflection ($|S_{11}|$) responses of the designed filter (dotted line: response calculated from polynomials; solid line: response based on coupling matrix; dashed line: response of optimized circuit with ideal elements).

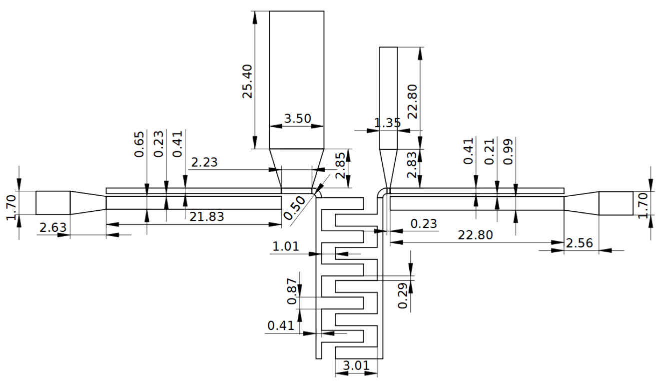

Fig. 4: Dimensions in millimeters of the fabricated filter.

Although there are some minor discrepancies between simulations and measurements, the main expected characteristics of the filter are visible, including the two transmission zeros produced below the passband. The minimum return-loss level is $RL_{min} = 12.15$ dB and the bandwidth referred to this value is 447 MHz with a center frequency of 2.03 GHz. The measured transmission zeros are located at 1.736 GHz and 1.505 GHz. The minimum in-band insertion-loss level is 1.02 dB and the estimated Q-factor is equal to $Q_{est} = 86$. The first spurious band occurs at 3.64 GHz. All the measured filter

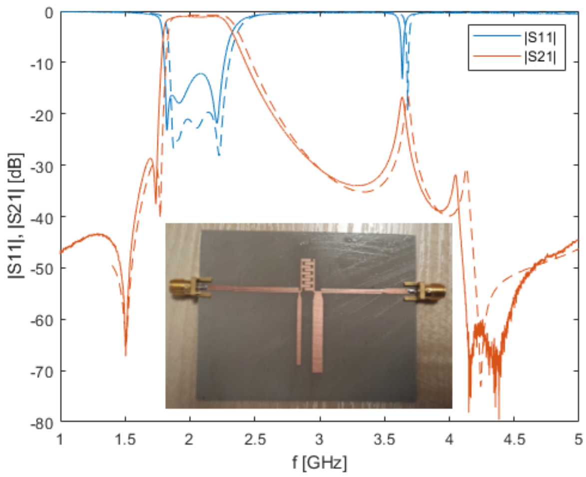

Fig. 5: Power transmission ($|S_{21}|$) and reflection ($|S_{11}|$) responses of the manufactured filter prototype (dashed line: full-wave simulation; solid line: measurement).

parameters, except the return loss, are very close to the values predicted by the simulation. Further investigation showed that the deviations observed can be caused by fabrication defects, especially in the interdigital structure. Indeed, the simulations indicate that the response is very sensitive to changes of the line-width parameters of the interdigital capacitor.

V. Conclusion

A design technique to synthesize inline quasi-elliptic-type microstrip filters based on a type of frequency-variable coupling inverter is presented. The proposed inverter is based on two open-ended quarter-wavelength stubs connected through an interdigital capacitor, and produces two additional poles in the passband and two transmission zeros in the stopband range of the filter. A detailed design procedure has been provided, along with experimental demonstration of an example of fourth-order microstrip filter prototype.

VI. Acknowledgement

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