Design Procedure for Waveguide Combline Filters Based on Flat Metallic Strips

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***Abstract*— In this work, we discuss the complete design procedure for a new topology of waveguide combline filters based on flat metallic strip resonators. The main advantages of the new topology that we propose are the ability to implement transmission zeros (TZ), compactness and low-cost practical realization. Furthermore, the position of the TZs can be controlled using offset resonators, thus achieving more selective filter responses. The design of several bandpass filter examples with finite-frequency TZs is also discussed. The correct operation of the filters is verified using commercial software packages, thus validating both the topology that we propose and the design procedure.**

***Keywords*— bandpass filters, combline filters, transmission zeros.**

I. Introduction

Bandpass filters based on coaxial resonators in a rectangular metallic enclosure, as combline or interdigital filters, are extensively used in different communication systems [1]–[3]. Classically, combline resonators are realized with cylindrical [2] or rectangular metallic posts [4], whose dimensions are optimized to obtain the desired resonant frequency. In this work, we discuss a new topology for waveguide combline bandpass filters with an inline configuration, based on planar metallic strips of a finite thickness (see Fig. 1). Some important features of this type of filters are their compactness, and that they can be manufactured at a very low cost with standard milling or wire erosion techniques, starting from a metallic sheet of the right thickness. Several resonators can, then, be easily stacked, using alignment pins to maintain mechanical accuracy. In addition, choosing the layers of the resonator with an appropriate thickness, will also allow the use of tuning screws, if needed.

To continue, we recall that it is well known that finite-frequency TZs can be realized using this type of filters [2], [3]. In our specific case, the TZs will be located above the passband, due to the magnetic nature of the cross-couplings between the strips. In this context, we propose to control the frequency position of the TZs by simply introducing an offset in the position of the combline resonators, thus achieving more flexible frequency responses. The filters are designed using the classical approach, based on the knowledge of the input/output and the inter-resonator couplings [5], and uses also the well-known Aggressive Space

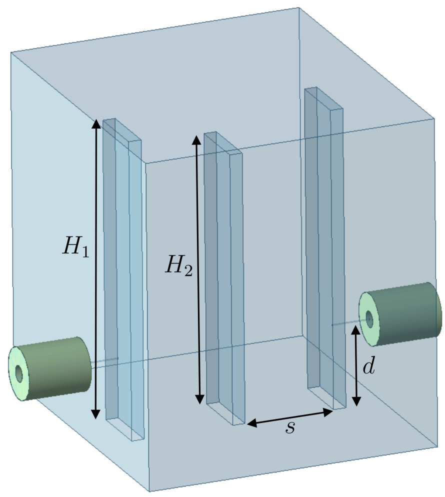

Fig. 1. Third-order symmetric combline filter with flat metallic strips.

Mapping (ASM) technique [6] with the aim of reducing the overall computation effort. In order to validate both the new topology that we propose and the design procedure, combline filters of orders 3 and 6 operating in the S-band have been designed. For validation purposes, the resulting structures have been successfully simulated using two different software packages (Ansys HFSS and CST Studio Suite) obtaining an excellent agreement.

II. Basic coaxial cavity and related equivalent circuit

The compact coaxial resonator that we propose is based on a rectangular cavity loaded with a flat metallic strip in the center. The metallic strip is connected to the metallic enclosure at the base (short circuit) and has an open circuit at the top, as shown in Fig. 2. The relevant dimensions of this structure are designed to obtain the fundamental resonance at the desired central frequency (in this case, $f_0 = 3$ GHz).

To this aim, the electromagnetic (EM) solver Ansys HFSS is used to analyze the one-port component obtained using, as the input port, the $y = 0$ plane in the resonator shown in Fig. 2. With this choice, the input port becomes a coaxial waveguide with rectangular inner and outer conductors. To continue, the dimensions of the component are optimized until the phase of

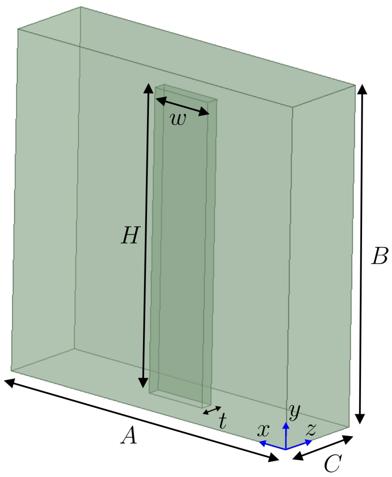

Fig. 2. Coaxial resonator loaded with a flat metallic strip.

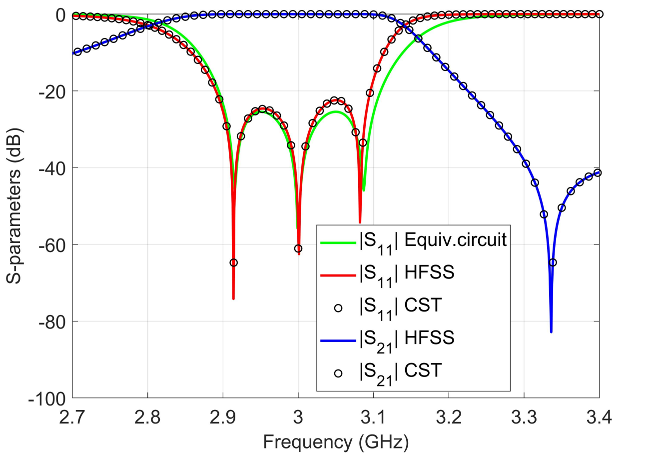

Fig. 3. Frequency response of the designed third-order combline filter with centered strips.

the S_{11} parameter is equal to zero at f_0. The final dimensions (in mm) of the resonator are (see Fig. 2): $A = B = 26$, $C = 7$, $H = 23.233$, $w = 5$ and $t = 1$. It is important to note that A, B, C, w and t are not optimized during the design process to be detailed next. In order to obtain a lumped-element equivalent circuit of the resonator (series combination of an inductance L and a capacitance C), the slope parameter is calculated [7] from the simulated result. The following values for the LC series network are then easily obtained: L = 2.747 nH and $C = 1.024$ pF.

III. Equivalent circuit of the filter

The objective of this section is to describe an equivalent circuit for the third-order bandpass filter shown in Fig. 1. It is important to note at this point that the equivalent circuit that we have developed is only valid for the in-band performance, and it will not exhibit TZs. However, the electrical response of the equivalent circuit will be used as reference (target) curve for recovering the in-band response of the final filters, including TZs, during the optimization stage of the design process.

The equivalent circuit of the filter is based on ideal impedance inverters (to represent both the input/output and the inter-resonator couplings) and LC series networks (to model the resonators). The values (in Ω) of the inverters needed for the third-order bandpass filter (whose electrical specifications are detailed in the next section) have been computed using the slope parameter of the resonator [7]. The values obtained are: $K_{01} = K_{34} = 36.169$ and $K_{12} = K_{23} = 4.211$.

IV. Design of the cascaded triplet section

In this section, the design of a symmetric third-order Chebyshev bandpass filter (f_0 = 3 GHz) is addressed. The electrical specifications are: 25 dB return loss, with a bandwidth of 200 MHz. In this first design, all metallic strips are located in a centered position (see Fig. 1). Using this arrangement, it is well-known that there is a magnetic-type (inductive) cross-coupling between metallic strips #1 and #3 [5]. However, the position of the TZ that will be obtained in the frequency response (located above the passband) cannot be controlled in this configuration, as long as the resonators are centered.

The design process is based on a classical approach [5], combined with the well-known ASM technique [6]. First of all, input and output resonators are designed to achieve the required external quality factor [5]. These resonators are excited using standard coaxial waveguide ports (r_{ext} = 2.05 mm, $r_{\text{in}} = 0.65$ mm, $\varepsilon_r = 2.2$). The input/output metallic strip resonators are in contact with the feeding probe of radius 0.065 mm. The feeding probe is placed at a distance d from the bottom of the filter, as shown in Fig. 1.

To continue, the inter-resonator couplings are calculated in terms of the separation s between the metallic strips (see Fig. 1), and an initial value for s is obtained taking into account the values of the required EM couplings. The complete filter is then assembled and optimized using the passband response of the equivalent circuit as the target curve. The simulations for all the previous steps are performed very efficiently (with a very low CPU effort) using the commercial software FEST3D with low accuracy settings (the coarse model). The accurate set of filter dimensions is finally obtained using ASM. In this case, Ansys HFSS is used to perform the accurate simulations (the fine model).

The final dimensions (in mm) of the filter are: $H_1 = 23.715$, $H_2 = 21.381$, $s = 5.829$ and $d = 5.763$. In order to validate both the new filter structure and the design procedure, the filter structure is next simulated using two different EM software packages, namely Ansys HFSS and CST Studio Suite. An excellent agreement between both sets of simulated data is obtained, as shown in Fig. 3, where the S_{11} parameter of the equivalent circuit has also been included in the figure for comparison purposes. As expected, we can observe a TZ located at the upper side of the passband ($f_{\text{TZ}} = 3.336$ GHz). Although the in-band response of the equivalent circuit has not been perfectly recovered (since the equivalent circuit does not account for the presence of the TZ), a bandwidth (centered at f_0) of 198 MHz, and a return loss higher than 22.6 dB have been obtained.

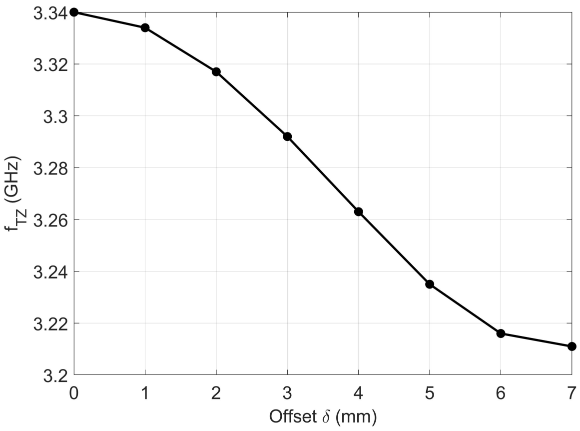

Fig. 4. Variation of the position of the TZ in terms of the offset δ. Data obtained for a separation $s = 5.829$ mm between the strips.

Table 1. Dimensions (all in mm) of the designed triplet sections.

	H_1	H_2	s	d
$\delta = 3$	23.630	21.499	5.417	5.485
$\delta = 4$	23.561	21.567	5.212	5.257
$\delta = 5$	23.460	21.611	5.240	4.946
$\delta = 6$	23.310	21.642	5.671	4.514

A. Control of the position of the TZ

As we have already mentioned, the position of the TZ cannot be easily controlled in a filter that uses only centered resonators. In this context, we demonstrate now that one possibility to change the frequency location of the TZ is to introduce an offset (along the x-dimension) in the position of the central resonator of the cascaded triplet section (see also the reference system shown in Fig. 2). If x_0 denotes the position of the center of the strip along the x-dimension, we can write $x_0 = A/2 + \delta$, where δ represents the offset applied to the strip position with respect to the center (i.e., $\delta = 0$ for a centered strip).

As we can see in Fig. 4, we can now easily change the frequency position f_{TZ} of the TZ by changing the value of the offset δ. In physical terms, the value of the offset changes the strength of the EM cross-coupling between resonators #1 and #3. Generally, as observed in Fig. 4, an increase in the value of δ results in a TZ that is closer to the passband.

Next, our objective is to bring the TZ closer to the passband, while maintaining an in-band performance as close as possible to the one obtained with centered strips. Fig. 5 shows the response (in terms of the S_{21} parameter) for several filters, which have been designed changing the value of the offset δ ($\delta = 3$, $\delta = 4$, $\delta = 5$, and $\delta = 6$ mm) in the central resonator. Again, two different EM solvers (Ansys HFSS and CST Studio Suite) have been used to perform the simulations. An excellent agreement between both results is evident in all cases. Finally, Table 1 collects the dimensions (in mm) of the filters we have designed, for the various values of δ that we have used.

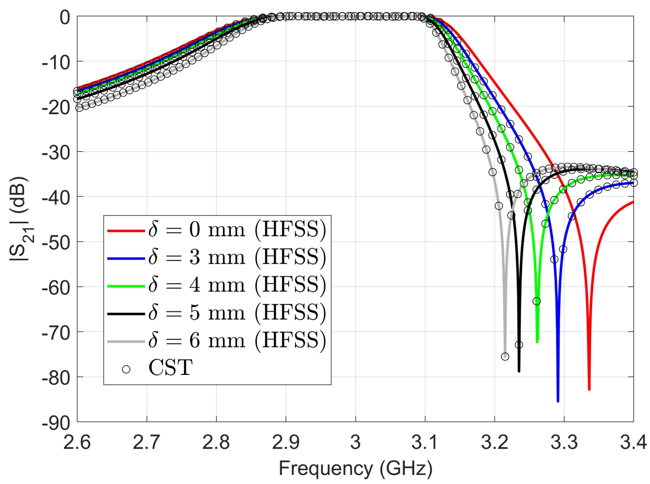

Fig. 5. S_{21} parameter of the designed triplet sections with off-centered strips.

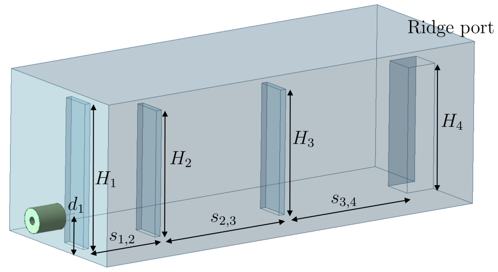

Fig. 6. First triplet section obtained from the sixth-order filter.

V. Design of Combline Filters with Two TZs

In this section, we discuss the use of two triplet sections including off-centered strips, to obtain two TZs in the frequency response. Both TZs will be located above the passband in a prescribed position. To this aim, we start with the design of a sixth-order symmetric combline filter with centered strips ($f_0 = 3$ GHz, $BW = 200$ MHz, 25 dB return loss), following the design guidelines described in the previous sections. The electrical response of this filter is shown in Fig. 9 (see blue curves labeled as "Centered strips").

After the initial design, the sixth-order filter is split into two triplet sections. The first triplet section is composed of the first three resonators of the filter, with input and output ports implemented in coaxial and ridge waveguide, respectively. The second triplet is composed of resonators #4, #5 and #6 of the initial filter. For the second triplet, input and output ports are implemented in ridge and coaxial waveguide, respectively. As an example, Fig. 6 shows the structure of the first triplet section, whose dimensions (in mm) are: $H_1 = 23.298$, $H_2 = 20.250$, $H_3 = H_4 = 20.1145$, $s_{1,2} = 11.127$, $s_{2,3} = 21.114$, $s_{3,4} = 23.324$, and $d_1 = 5.213$ (note that these are also the dimensions of the symmetric sixth-order filter designed with centered strips).

The in-band responses of the two triplet sections are identical due to the symmetry of the initial sixth-order filter. They will be used as a target curve in the next steps of the design process. To continue, Fig. 7 shows the variation of the

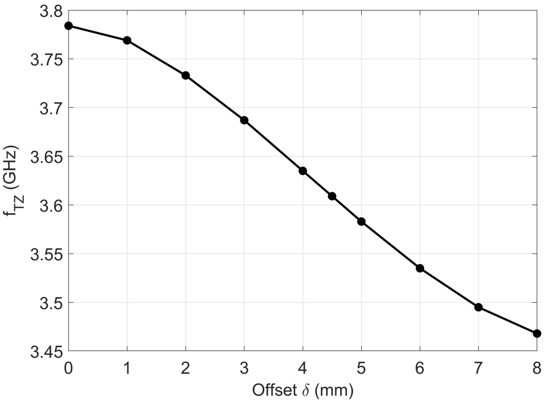

Fig. 7. Variation of the position of the triplet sections TZ.

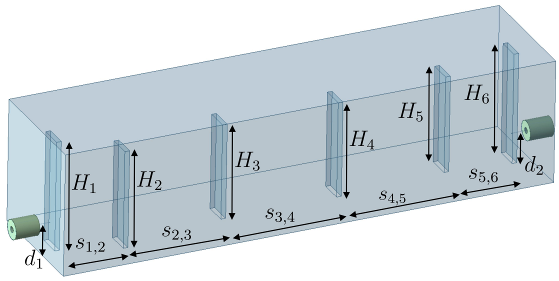

Fig. 8. Layout of the sixth-order combline filter with offset resonators.

position of the TZ generated by the triplet, as a function of the offset δ of the central resonator.

Next, our objective is to re-design the sixth-order combline filter in order to place the TZ provided by each triplet section in a given position. To this aim, an offset $\delta_1 = 4.5$ mm is used in the central strip of the first triplet to generate a TZ at $f_{\mathrm{TZ}_1} = 3.61$ GHz (see Fig. 7), while an offset $\delta_2 = -2$ mm is used for the central strip of the second triplet to obtain a TZ at $f_{\mathrm{TZ}_2} = 3.73$ GHz. It is important to note that the strips are off-centered in opposite directions to minimize other potential cross-couplings in the structure. After setting the initial dimensions, both triplet sections are optimized separately. Two target curves are used: (1) the in-band response provided by the triplet section with centered strips, and (2) the out-of-band response with the TZ in the desired position. Finally, the whole filter is assembled and a final optimization of all dimensions is performed (see the filter layout in Fig. 8).

The electrical response of the filter is shown in Fig. 9, using green (Ansys HFSS data) and red lines (CST data). A very good agreement between both data sets is evident. The response of the sixth-order combline filter with centered strips has been also included in the figure (using blue lines) for comparison purposes. It is important to note that the in-band response of the filter with centered strips has been recovered almost perfectly. Furthermore, two TZs can be observed in the out-of-band response of the filter, at $f_{\mathrm{TZ}_1} = 3.55$ GHz and $f_{\mathrm{TZ}_2} = 3.64$ GHz, which are very close to the design values, providing an extremely high rejection at the upper stopband of up to 150 dB. The dimensions (in mm) of the designed filter are (see also Fig. 8): $H_1 = 23.303$, $H_2 = 20.721$, $H_3 = 20.189$, $H_4 = 20.195$, $H_5 = 20.456$, $H_6 = 23.415$, $s_{1,2} = 10.122$,

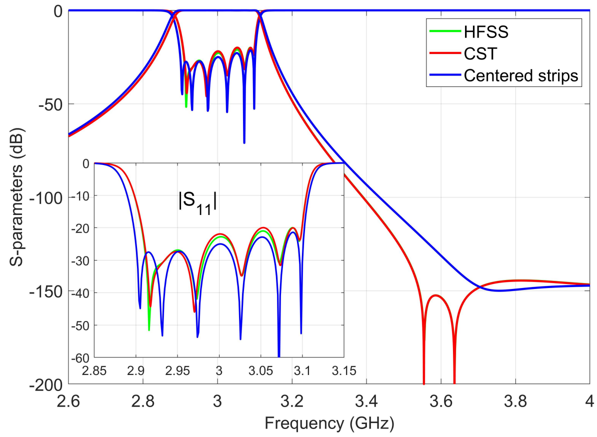

Fig. 9. S-parameters of the designed sixth-order filters with centered (blue lines) and off-centered strips (green and red lines).

$s_{2,3} = 20.010$, $s_{3,4} = 23.458$, $s_{4,5} = 20.930$, $s_{5,6} = 11.156$, $\delta_1 = 4.036$, $\delta_2 = -1.582$, $d_1 = 5.010$ and $d_2 = 5.142$.

VI. Conclusion

In this work, a new topology for combline bandpass filters based on flat metallic strip resonators has been discussed. The structures that we propose are compact and can realize TZs at prescribed frequencies above the passband. As we clearly show in the paper, the position of the TZs can be controlled by changing the offset of some of the resonators. The design of several bandpass filters has been discussed in detail. Furthermore, as a validation, the filters have been simulated using two different commercial software tools. The results obtained are in excellent agreement, thus validating both the proposed topology and the design procedure.

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