Compact microstrip band-pass filters based on semi-lumped resonators

J. Bonache, I. Gil, J. García-García and F. Martín

Abstract: A new design approach for compact microstrip band-pass filter based on semi-lumped resonators are proposed. The resonators, which are coupled through quarter wavelength meander lines acting as admittance inverters, are shunt connected to the line. They consist of parallel combination of a grounded (inductive) stub and a narrow metallic strip followed by a capacitive patch to ground. With this topology, the necessary degree of flexibility to design narrow and broad-band-pass filters with compact dimensions and good out-of-band performance was obtained. Another key advantage of the devices, as compared to previous lumped or semi-lumped element-based structures reported by the authors, is the absence of ground plane etching. To illustrate the potentiality of the proposed approach, a third-order (30% fractional bandwidth) and a ninth-order (35% fractional bandwidth) Chebyshev band-pass filters have been designed and fabricated. The measured frequency responses are very symmetric and exhibit low in-band losses as well as good out-of-band rejection up to approximately 3\(f_0\). Filter dimensions are as small as 0.40\(\lambda\) \(\times\) 0.12\(\lambda\) (third-order prototype) and 0.62\(\lambda\) \(\times\) 0.16\(\lambda\) (ninth-order prototype), \(\lambda\) being the guided wavelength at \(f_0\). With these dimensions and performance, and the possibility to synthesise microstrip filters with controllable bandwidth over a wide margin, it is believed that the reported approach can be of actual interest for the design of planar filters at microwave frequencies.

1 Introduction

High data-rate communication standards are renewing the interest in planar broad-band filters [1, 2]. In applications where size is a critical aspect, the use of miniaturisation techniques becomes crucial [3]. Some of these techniques are based on the use of lumped resonators instead of distributed approaches, which can be implemented by means of small size in-house soldered components such as parallel-plate chip capacitors and air-wound inductors. Although such an approach intrinsically provides small physical size and low cost solutions, it is not compatible with full-planar circuit technology, and such components have limited \(Q\)-factors and bandwidth at microwave frequencies. Alternatively, semi-lumped element microwave filters entirely fabricated using printed circuit or thin-film technologies have been proposed [4, 5] where capacitors have been implemented by the well-known interdigitated concept or by means of signal-to-ground metal patches, and inductors have been designed by means of wire loops and narrow strips (a good description of semi-lumped element filters and related references are given in [6]). Remarkable are also the series-resonators proposed by Mathaei et al. [7], which are very useful for shunt connections in microstrip technology. They consist of the combination of a narrow metallic strip, which behaves inductively, and a capacitive patch to ground. These elements have successfully been applied, for instance, to the design of miniature low pass filters with attenuation poles properly located (by adequately dimensioning the cited resonators) in order to obtain a sharp cut-off [7]. In this work, such series-resonators are combined with grounded stubs to form a shunt-connected planar resonant element with two characteristic frequencies, that is the frequency that nulls its admittance, \(f_0\), and the frequency that nulls its impedance (transmission zero frequency, \(f_z\)). By coupling these semi-lumped resonant elements through 90° (at \(f_0\)) transmission lines, it will be shown that microstrip band-pass filters with compact dimensions and good out-of-band performance (thanks to the presence of attenuation poles) can be implemented. The technique can be applied to the design of band-pass filters with fractional bandwidths up to 35% (as will be shown) and beyond. Other microstrip filters recently reported by the authors and based on semi-lumped resonators [8] or complementary split rings resonators (CSRRs) [9, 10], have provided comparable performance and dimensions, but the fabrication process is more complicated since two metal layers are needed.

The frequency responses measured on fabricated prototype devices (a third - and a ninth-order Chebyshev band-pass filters centered at 3 and 1 GHz, respectively) designed by means of the approach presented in this work point out good in-band insertion and return losses, as well as the possibility to enhance the stop band by properly adjusting the attenuation poles of the structure. For these reasons and the compact dimensions, the authors are confident on the practical application of the structures presented in this paper.

2 Filter topology and design methodology

The filters proposed in this work are fabricated by cascading cells with the typical layout shown in Fig. 1. This is a quarter wavelength microstrip line loaded with a shunt inductor, \(L_p\), implemented by means of a grounded stub,
and a series-resonator \( L_s - C_s \), achieved through the combination of a narrow strip and a capacitive patch. The equivalent circuit model of the structure is depicted in Fig. 1b. The shunt-connected admittance, formed by the parallel combination of \( L_p \) and the series resonator (from now on the shunt resonator), exhibits two singularities at the resonant, \( f_x \), and anti-resonant, \( f_o \), frequencies. These frequencies null the impedance and admittance, respectively, of the shunt resonator and are given by

\[
f_x = \frac{1}{2\pi\sqrt{L_sC_s}} \quad (1)
\]

\[
f_o = \frac{1}{2\pi\sqrt{(L_p + L_s)C_s}} \quad (2)
\]

The 90° transmission lines comprised between adjacent resonators, which are implemented by means of meander lines to shorten the device, act as admittance inverters with normalised admittance \( l = 1 \). Thus, the band-pass filters of this work can be described by the generalised band-pass filter network depicted in Fig. 2 [6]. According to this network model, the central frequency of the filter is given by the frequency that nulls resonator’s admittance, hence for our filter topology, it will be given by \( f_o \). The fractional bandwidth is controlled by the elements that form the shunt resonator. If these resonators were parallel LC tanks (as in Fig. 2), the element values, \( L \) and \( C \), would be given by frequency and element transformation from the low pass filter prototype, according to the well-known formulas [6]

\[
C = \left[ \frac{1}{\text{FWB} \cdot \omega_0 \cdot Z_0} \right] g_i \quad (3)
\]

\[
L = \frac{1}{\omega_0^2 C} \quad (4)
\]

where \( g_i \)'s are the elements of the low pass filter prototype, \( \text{FWB} \) is the fractional filter bandwidth and \( \omega_0 = 2\pi f_o \). For the filters described by the model given in Fig. 1b, an additional equation [(besides (1) and (2))] is necessary to univocally determine the three element values for each shunt resonator. This equation arises by comparing the susceptance slope of this resonator (at the central filter frequency) with that of the LC resonant tank that results from (3) and (4) at the corresponding filter stage. After some tedious calculation this leads to

\[
\frac{C(L_s + L_p)^2}{L_p^2} = C \quad (5)
\]

Thus, (1) through (5) allows us to infer the element values per filter stage from filter specifications (\( f_o \), \( \text{FBW} \)) for any given filter approximation, namely from the tabulated values of \( g_i \).

With regard to the transmission zero (attenuation pole) given by (1), it can be adjusted to reject the first spurious band of the filter [11], which is expected to be situated in the vicinity of \( 2f_o \) on account of the presence of parasitic half-wavelength resonators at this frequency. Alternatively, the attenuation pole could be used to improve the fall-off of the upper transition band [12].

Once the element values \( L_p, L_s \) and \( C_s \) have been determined for all the cells, shunt resonator’s topologies are determined through well-known formulas. Thus, the lengths of the inductive strip and capacitive patch forming the \( LC \) series resonator to ground are roughly given by [13]

\[
I_{L_s} = \frac{L_f}{Z_h} \lambda(f_o) \quad (6)
\]

\[
I_{C_s} = Z_h C f_o \lambda(f_o) \quad (7)
\]

where \( \lambda(f_o) \) is the guided wavelength at the central filter frequency and the widths, \( w_L \) and \( w_C \), of the inductive strip and capacitive patch are arbitrarily set to obtain a high/low characteristic impedance \( Z_h/Z_c \). More accurate expressions to determine these dimensions are given in ref. [7], but expressions (6) and (7) suffice since, as will be explained in the next section, final dimensions are inferred by means of an optimisation procedure. The length of the grounded stub is determined by an expression formally identical to (6), that is

\[
I_{L_p} = \frac{L_p f_o}{Z_h} \lambda(f_o) \quad (8)
\]

\( Z_r \) being the characteristic impedance of this stub. The maximum bandwidth achievable with this technique is determined by the values of \( Z_L \) and \( Z_C \) realisable with the resolution of the fabrication process.

### 3 Illustrative examples

To demonstrate the possibilities of the proposed approach, two illustrative designs are presented in this section. The first design corresponds to a third-order Chebyshev (0.16 dB ripple) band-pass filter centered at \( f_o = 3 \text{ GHz} \) with 30% fractional bandwidth. For the considered in-band ripple and order, the element values of the low pass filter prototype are identical, and hence do the three filter stages. To determine the element values of the equivalent circuit model, (1)–(5) have been solved by forcing the transmission zero frequency to be \( f_z = 2f_o \). The results are given in Table 1. From these values and expressions

![Fig. 1 Fabrication of band-pass filters](image)

*a Topology of the basic filter cell  
*b Equivalent circuit model*

![Fig. 2 Generalised band-pass filter network with shunt resonators and admittance inverters](image)

**Table 1: Element values for the equivalent circuit model of the third-order filter**

<table>
<thead>
<tr>
<th>G</th>
<th>( L_p ) (nH)</th>
<th>( C_s ) (pF)</th>
<th>( L_s ) (nH)</th>
</tr>
</thead>
<tbody>
<tr>
<td>1.15</td>
<td>0.77</td>
<td>2.74</td>
<td>0.26</td>
</tr>
</tbody>
</table>
(6)–(8), the dimensions of the semi-lumped shunt resonator have been obtained. However, these dimensions have been optimised in order to fit the final filter response to that of the equivalent circuit model. To this end, we have used the optimise tool included in the Agilent Momentum electromagnetic solver, and we have swept dimensions around the nominal values until a satisfactory adjustment between the electrical response of a single cell and that obtained from electromagnetic simulation has been found. The final filter layout and dimensions are depicted in Fig. 3.

The device has been fabricated on a Rogers RO3010 substrate with dielectric constant $\varepsilon_r = 10.2$, loss tangent $\tan \delta = 0.0023$ and thickness $h = 0.63$ mm (a standard photo/mask-etching technique has been used). The simulated (using Agilent Momentum) and measured (by means of the Agilent 8720ET vector network analyser) frequency responses of the filter are depicted in Fig. 4 (conductor vector network analyser frequency). As can be seen, thanks to the presence of the transmission zero, located at the appropriate position, the spurious band otherwise centered in the vicinity of $2/f_o$ has been substantially eliminated. The frequency response is quite symmetric with measured in-band losses below $IL < 1$ dB, and return losses greater than $RL > 11$ dB. Thanks to the 90° meander lines, the final filter dimensions have been compacted, these being as small as $0.40 \times 0.12 \lambda$, where $\lambda$ is the guided wavelength at $f_o$. The slight discrepancy between simulation and experiment in the stop band is thought to be because of the ideal vias considered in the simulation. These vias have been placed at the end of the shunt stubs and have been implemented by means of vertical metallic walls (zero thickness) with the same width as the stubs. However, as will be shown in the next prototype device example, if vias are implemented by considering vertical cylinders, which is a more realistic approach, then not only a good agreement is obtained in the pass band region, but also in the whole frequency range considered. The measurement also shows some level of losses around 8 GHz. This effect is because of the behaviour of the connectors at this frequency, and it cannot be attributed to radiation in the structure, since losses do not appear in the simulated results (which have been performed by taking into account radiation losses).

The second design is a ninth-order Chebyshev (0.5 dB ripple) band-pass filter centred at $f_o = 1$ GHz with 35% fractional bandwidth. The main challenge in this design was to achieve a rejection level greater than 80 dB at half the central filter frequency, with in-band losses and device dimensions as small as possible. To achieve such rejection level, nine filter stages have been needed. From the tabulated $g_i$'s of the low pass filter prototype, the elements of the equivalent circuit model as well as resonator’s topology have been inferred from the previous equations for each filter stage (see Table 2). As for the previous design, layout optimisation has been applied cell-by-cell to match to the electrical responses. The final device layout is depicted in Fig. 5, where final dimensions are in this case $0.62 \times 0.16 \lambda$ (identical substrate as in the previous design has been considered). As desired, insertion losses are better than $IL > 80$ dB (indeed we have measured the noise floor) at $f = f_o/2$, while measured in-band losses are $IL < 3$ dB in the interval between 0.83 and 1.11 GHz (see Fig. 6). We also want to mention that return losses better than $RL > 10$ dB have been measured within the range delimited by 0.82 and 1.14 GHz. Except for the fact that ohmic and dielectric losses have not been taken into account in the simulations, good agreement between the measured and the simulated frequency response has been obtained (cylindrical vias of 300 $\mu$m diameter have been considered in the simulation and physically implemented in the fabricated prototype). The absence of losses in the simulation results shows that the structure is not affected by radiation effects (that have been taken into account in the simulations).

![Fig. 3 Topology (layout) of the fabricated third-order Chebyshev band-pass filter and relevant dimensions. The meander lines are 0.58 mm wide, corresponding to a 50-characteristic impedance.](image-url)

![Fig. 4 Simulated (thin line) and measured (bold line) insertion and return losses for the filter shown in Fig. 3.](image-url)

<p>| Table 2: Elements values for the equivalent circuit model of the ninth-order filter and geometrical parameters |</p>
<table>
<thead>
<tr>
<th>Cell</th>
<th>$g$ (pF)</th>
<th>$L_p$ (nH)</th>
<th>$C_r$ (pF)</th>
<th>$L_r$ (nH)</th>
<th>$l_{ap}$–$w_{ap}$ (mm)</th>
<th>$l_{cr}$–$w_{cr}$ (mm)</th>
<th>$L_{cr}$–$W_{cr}$ (mm)</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>1.750</td>
<td>3.128</td>
<td>5.295</td>
<td>1.655</td>
<td>5.6–0.5</td>
<td>3.5–0.5</td>
<td>4.5–5.1</td>
</tr>
<tr>
<td>2</td>
<td>1.269</td>
<td>4.315</td>
<td>3.839</td>
<td>2.283</td>
<td>7.3–0.5</td>
<td>6.0–0.5</td>
<td>3.6–3.6</td>
</tr>
<tr>
<td>3</td>
<td>2.667</td>
<td>2.052</td>
<td>8.070</td>
<td>1.086</td>
<td>4.4–0.5</td>
<td>2.1–0.5</td>
<td>4.8–7.1</td>
</tr>
<tr>
<td>4</td>
<td>1.367</td>
<td>4.004</td>
<td>4.136</td>
<td>2.118</td>
<td>6.8–0.5</td>
<td>5.4–0.5</td>
<td>3.6–4.1</td>
</tr>
<tr>
<td>5</td>
<td>2.723</td>
<td>2.010</td>
<td>8.240</td>
<td>1.063</td>
<td>4.4–0.5</td>
<td>2.1–0.5</td>
<td>4.8–7.1</td>
</tr>
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<td>6</td>
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<tr>
<td>9</td>
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<td>5.295</td>
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<td>5.6–0.5</td>
<td>3.5–0.5</td>
<td>4.5–5.1</td>
</tr>
</tbody>
</table>
A rejection level better than 70 dB has been measured above the pass band of interest up to 2.5 GHz. This wide rejection has been achieved by adjusting the position of the transmission zeros properly, and this has required several iterations since the unwanted spurious band has not been found to exactly appear at $2f_0$ in this case. Indeed, the parasitic half-wavelength resonators are coupled through shunt impedances that introduce some phase shift. Therefore, the procedure has been to initially set the transmission zeros at $2f_0$, obtain the layout according to the previously described scheme, simulate the frequency response and adjust the position of the transmission zeros to match the parasitic spurious band. In Fig. 7, the measured frequency response obtained in a similar band-pass filter (with larger bandwidth), but without adjustment of the transmission zeros, is depicted (layout not shown). As can be seen, the resulting spurious band is very close to the upper edge of the band, this being unacceptable in certain applications.

With these dimensions and performance, it is believed that the proposed approach can be of interest for microwave filter engineers. The approach seems to be suitable for the design of planar filters with wide bandwidths.

### 4 Conclusion

A new design approach for the synthesis of compact microstrip band-pass filters with wide bandwidths and good out-of-band performance has been proposed. This is based on the use of shunt-connected semi-lumped resonators consisting of parallel combination of grounded stubs and series LC resonators, coupled by means of quarter wavelength meander lines. The key factor to achieve the wide stop bands reported is the presence of attenuation poles, which can be tailored in order to suppress the undesired spurious bands, inherently present in the structures. Two prototype device band-pass filters have been designed and fabricated to demonstrate the possibilities of the approach presented in this work. The measured frequency responses are indicative of good filter performance while device dimensions are small on account of the semi-lumped elements considered. Specifically, we have measured in-band losses below $IL < 1$ dB for the third-order prototype centered at 3 GHz (dimensions $15.6 \times 4.8$ mm) and below $IL < 3$ dB for the ninth-order prototype centered at 1 GHz (dimensions $72.8 \times 18.4$ mm). In the latter case, the measured fractional filter bandwidth is as wide as 35%. Thus, the presented approach suits well to the design of wide band-pass filters with good out-of-band performance.

### 5 Acknowledgments

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### 6 References