Broadband Microwave Filters Based on Open Split Ring Resonators (OSRRs) and Open Complementary Split Ring Resonators (OCSRRs): Improved Models and Design Optimization

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Abstract. The paper is focused on the design of broadband bandpass filters at microwave frequencies. The proposed filters are based on a combination of open split ring resonators (OSRRs) and open complementary split ring resonators (OCSRRs) loaded in a host transmission line. Since these resonators (OSRRs and OCSRRs) are electrically small, the resulting filters are compact. As compared to previous papers by the authors on this topic, the main aim and originality of the present paper is to demonstrate that by including a new series inductance in the circuit model of the OCSRR, it is possible to improve the predictions of these filter models and better fit the measured filter responses. Moreover, the parameter extraction method of the new circuit model and an automated filter design technique is introduced and demonstrated. The paper is complemented with the design and comparison of several prototypes.

Keywords
Bandpass filters, metamaterials, open split ring resonator, open complementary split ring resonator.

1. Introduction

The wideband bandpass filters of the present paper are based on the open versions of the split ring resonator (SRR) [1] and the complementary split ring resonator (CSRR) [2]. SRRs and CSRRs, among other particles [3], [4], have been used for the synthesis of negative effective permeability and permittivity media, respectively, and also for the synthesis of left handed metamaterials. In planar technology, these particles have been used for the implementation of composite right/left handed (CRLH) transmission lines, namely, lines exhibiting backward wave propagation at low frequencies, and forward wave propagation at high frequencies [5], [6], and also for the implementation of broadband filters based on them [7].

However, by using the open versions of the SRR and the CSRR combined, i.e., the open split ring resonator (OSRR) [8] and the open complementary split ring resonator (OCSRR) [9], it has been recently demonstrated the possibility of implementing CRLH lines in both microstrip and coplanar waveguide technology [10] without the presence of a transmission zero below the first (left handed) transmission band (such transmission zero is present in the CRLH lines implemented by using SRRs or CSRRs [5], [6]). It has been also pointed out that by sacrificing periodicity in OSRR- and OCSRR-based CRLH lines, it is possible to implement compact wideband bandpass filters subject to specifications, and several prototype devices covering different orders and bandwidths have been reported [11], [12].

In the present paper we report an improvement of the circuit model of these OSRR and OCSRR-based filters and also a method to automatically synthesize the required filter responses. This new circuit model is necessary to better predict the response of the filter in the upper region of the band-pass and also the upper transition band. The paper is organized as follows: in section 2, the filter topology and the improved circuit model is presented. In section 3, the parameter extraction method to infer the circuit values of the new proposed model is shown. The automated synthesis method is reported in section 4. In section 5, the design of several prototype devices is reported. Finally, the main conclusions are highlighted in section 6.

2. Filter Topology and Circuit Models

The proposed filters are based on alternating sections of host lines loaded with series connected OSRRs and shunt connected OCSRRs. Fig. 1 presents the topology and the formerly proposed lumped element equivalent circuit models of these filter sections, corresponding to CPW technology [10]. CPW is the technology used for the implementation of the OSRR and OCSRR based filters since
the circuit model of a microstrip line section loaded with an OSRR is more complex and this complicates the design, as reported in [10].

Essentially, the OSRR and the OCSRR are described by means of series and parallel resonators, respectively. However, some phase shift at resonance occurs, due to the presence of the host line, and the accurate models of the particles must include these parasitic effects (by means of the indicated phase shifting lines). In the OCSRR section, it is necessary to electrically connect the different ground plane regions (by means of backside strips and vias) to avoid the presence of the slot mode. Notice that the circuit models can be simplified, as Fig. 1 illustrates (i.e., $C_p' = 2(C + C_p)$ and $L_s' = 2L + L_s$). The representation of $S_{11}$ in a Smith Chart (Fig. 2) for both structures reveals the need to include the additional elements to account for the phase shift. The dimensions of the OCSRR are: external radius $r_{ext} = 1.6$ mm, width $c$ and distance between the slot rings $d$ are $c = d = 0.2$ mm. The values of the equivalent circuit elements are: $L = 0.65$ nH, $L_p' = L_p/2 = 0.97$ nH, $C_p' = 2(C_p + C) = 3.11$ pF. The dimensions of the OSRR are: $r_{ext} = 1.8$ mm, $c = 0.2$ mm, $d = 0.3$ mm. The values of the equivalent circuit elements are: $C = 0.2$ pF, $L_s' = L_s + 2L = 6.33$ nH, $C_p = 0.511$ pF. The previous element values have been obtained by means of the parameter extraction procedure detailed in [10].

The comparison of the insertion and return losses for both structures is depicted in Fig. 3. For the OSRR-loaded CPW section, the agreement between the electromagnetic and circuit simulation is excellent within the frequency range shown. For the OCSRR structure, it can be appreciated that some discrepancy appears at high frequencies. This is due to the connection between the central strip of the CPW transmission line and the inner metallic region of the OCSRR, which generates an inductive effect in series with the shunt resonator. Hence, in order to consider this particle for applications where a prediction far below and beyond the resonance frequency is required, such as wideband devices, this additional inductive effect should be taken into account. Fig. 4 shows the resulting wideband circuit model of an OCSRR-loaded CPW transmission line, where the approach of considering that the shunt capacitances $C$ could still be absorbed by $C_p'$ regardless of the presence of $L_{sh}$ has been considered. The comparison between the electromagnetic and circuit simulation of the wideband response of the structure of Fig. 1(a) is also depicted in Fig. 4, where the parameter extraction method that will be explained in section 3 has been considered to infer the circuit values. It can be appreciated the excellent agreement between both responses when the inductance $L_{sh}$ is included in the model. Indeed, this inductance is responsible for the presence of a transmission zero above the pass-band of the structure, and it is of interest to improve the filter selectivity at the upper transition band.

![Fig. 1. Topology and circuit model of the OCSRR (a) and OSRR (b) loaded CPW.](image1)

![Fig. 2. Representation of $S_{11}$ for the OCSRR (a) and OSRR (b) structures of Fig. 1, implemented on the Rogers RO3010 substrate with thickness $h = 0.254$ mm and dielectric constant $\varepsilon_r = 11.2$.](image2)

![Fig. 3. Insertion and return losses for the structures of Fig. 1. (a) OCSRR-loaded CPW section; (b) OSRR-loaded CPW section.](image3)
3. Parameter Extraction Method of the OCSRR Wideband Circuit Model

The parameters of the circuit model of Fig. 4 can be inferred from the electromagnetic simulation through close equations following a similar methodology as the one explained in [10]. The series inductance can be obtained from the intercept of the return losses with the unit resistance circle in the Smith chart, yielding

\[ L = \frac{\chi}{2\omega |_{\omega \rightarrow \infty}} \]  

where \( \chi \) is the reactance at the intercept point. Additionally, the \( LC \) tank resonates at this frequency, that is

\[ \omega^2 \right|_{\omega \rightarrow \infty} = \frac{1}{L'p C'_p} \]  

Another condition can be obtained from the reflection zero frequency \( \omega_0 \), where the characteristic impedance \( Z_0 \) is 50 \( \Omega \) and is given by

\[ Z_0(\omega) = \sqrt{Z_s(\omega) Z_p(\omega) + 2Z_p(\omega)} \]  

where \( Z_s \) and \( Z_p \) are the series and shunt impedance of the \( T \)-circuit model of Fig. 4(a), respectively. Finally, the last condition is obtained from the point at which the electrical length \( \beta \ell \) is 90\(^{\circ} \), given by

\[ \cos \beta \ell = 1 + \frac{Z_s(\omega)}{Z_p(\omega)} \]  

Therefore, from equations (2)-(4), all the parameters of the shunt branch can be obtained as

\[ L'_p = \frac{\omega^2 |_{\omega \rightarrow \infty} - \omega^2 |_{\omega \rightarrow 0^+}}{2L_0^2 |_{\omega \rightarrow \infty} \omega^2 |_{\omega \rightarrow 0^+} \omega^2 |_{\omega \rightarrow \infty} - \omega^2 |_{\omega \rightarrow 0^+} \omega^2 |_{\omega \rightarrow \infty} \omega^2 |_{\omega \rightarrow 0^+}} \]  

\[ C'_p = \frac{1}{\omega^2 |_{\omega \rightarrow \infty} L'_p} \]  

\[ L_0 = \frac{L'p - L}{1 - \frac{\omega^2 |_{\omega \rightarrow 0^+}}{\omega^2 |_{\omega \rightarrow \infty}}} \]  

being thus all the values univocally determined.

4. Filter Design: An Automatic Optimization Routine

The first step for the design of filters based on OSRRs and OCSRRs is to determine the element values of the canonical circuit model, which is the model that results by cascading the simplified circuits of Fig. 1 without the presence of the parasitics, namely \( L \) and \( C \). Normally, the element values of the canonical bandpass filter are inferred from response and frequency transformations from the low pass filter prototype corresponding to a given standard response (Butterworth, Chebyshev, etc.), but this requirement is not actually necessary (indeed, in the examples reported later we have considered Chebyshev responses because of the higher frequency selectivity for a given order).

The second step consists to infer the layout of each section so that the elements of the resonator (obtained from the extraction procedure detailed in [10] and in section 3 for the models of Fig. 1 and Fig. 4, respectively) are those of the canonical circuit model. From the parameter extraction method, we obtain the values of the elements modeling the parasitics. Then we use the accurate model of the different sections (with the parasitic values inferred) and we tailor the resonator values until the specifications are satisfied at the circuit level (bandwidth and selectivity).

Finally, we modify the layout of the resonator until the response coincides with that of the circuit simulation. Following this procedure, the final filter layout can be obtained. Of course, due to the presence of the parasitics, a purely standard response cannot be obtained, but the effects of the parasitics are small and the proposed filter topology is roughly described by the canonical model [10].

The technique used to design this kind of filters has been proven to be simple for small order filters and/or small parasitic elements, since the element tuning to infer
the optimum response considering parasitics is realized at the circuit level. Fig. 5 reports an illustrative example corresponding to a 3rd order filter with central frequency \( f_0 = 2.9 \) GHz, 0.02 dB ripple and 35% fractional bandwidth. This filter has been fabricated on the Rogers RO3010 substrate with thickness \( h = 0.254 \) mm and measured dielectric constant \( \varepsilon_r = 11.2 \), where it has been found that by using thinner substrates the stop-band response is improved (i.e., the first spurious band due to the distributed resonances of the resonators is shifted to higher frequencies).

Fig. 5. Layout (a) and top (b) - bottom (c) photograph of the CPW wideband band-pass filter based on a combination of series connected OSRRs in the external stages and a pair of shunt connected OCSRRs in the central stage. Dimensions are: \( l = 9 \) mm, \( W = 5 \) mm and \( G = 0.55 \) mm. For the OCSRR: \( r_{ext} = 1.2 \) mm, \( c = 0.2 \) mm and \( d = 0.6 \) mm. For the OSRR: \( r_{ext} = 1.6 \) mm and \( c = d = 0.2 \) mm.

Fig. 6(a) shows the electromagnetic and circuit simulation of the filter without losses, considering for comparison purposes both the simple and wideband OCSRR model (i.e., considering the case with and without the additional parameter \( L_{sh} \) in reference of Fig. 4a). As can be appreciated, since the fractional bandwidth of the filter is moderate, the simple OCSRR model can still predict the band, although it begins to fail at higher frequencies. Fig. 6(b) also shows the wideband frequency response of the electromagnetic simulation considering dielectric, conductor and radiation losses, the circuit and ideal Chebyshev models, as well as the measurement of the fabricated device. The group delay and insertion loss of the electromagnetic simulation considering parasitic, conductor and radiation losses, the circuit and ideal Chebyshev models, as well as the measurement of the fabricated device. The group delay and insertion loss of the electromagnetic simulation and measurement are also depicted in Fig. 6(c). There is good agreement between all the curves in the region of interest, with the first spurious band at roughly 2.3 times the central filter frequency. Measured filter characteristics are good with insertion losses lower than 1 dB between 2.3 GHz and 3.5 GHz, and measured return losses better than 18 dB between 2.36 GHz and 3.41 GHz.

Furthermore, the circuit simulation considering the parasitic element \( L_{sh} \) predicts both the band-pass and the stop-band rejection, where due to this additional shunt inductance behavior, these types of filters will have the advantage of a more selective upper transition band compared to the Chebyshev responses, as well as an additional transmission zero above the band (present at higher frequencies than those depicted for the filter shown).

Nonetheless, if either the order or the parasitic element values dramatically increase, the manual tuning of
goals are achieved, but with a slight discrepancy in return
implemented in a commercial simulator such as
response (i.e., where the input impedance is 50
positions at the
automatic optimization routine that forces the frequency
which are given by the
the optimum circuit elements considering a set of fixed parasitics values can be
obtained for any filter order or any fixed parasitic value
through an automatic iterative process. However, the
element values as well as their possible swept ranges
should be chosen to be close to the optimal final solution in
order to achieve converge with the response that satisfies the
required specifications (i.e., ripple and bandwidth). To
this purpose, the circuit elements inferred by the parameter
extraction method can be considered as starting point.
These circuit elements should be the same or close to the
canonical circuit model elements, given that the magni-
tudes of the parasitics do not dramatically alter the final
optimal values as compared to the initial canonical circuit
elements. Hence, the requirements can be met.

As an example, this routine is applied to the already
designed filter, where the final circuit elements inferred by
the parameter extraction method are considered as a start-
ing point and optimized to fulfill the initial filter require-
ments considering constant parasitics (since these are the
only non-controllable parameters). Fig. 7 shows the
frequency response comparison between the circuit model
and the Chebyshev response. As can be observed, if the simple
OCSRR model is considered, a nearly perfect fit between
both curves is obtained with this method, with a slight
deviation only on the stop-band rejection at high frequen-
cies. On the other hand, if the same method is considered
with the wideband model of the OCSRR (i.e., considering
the additional inductance \(L_a\)) it can be seen how the same
goals are achieved, but with a slight discrepancy in return
loss, as well as a higher upper selectivity due to this factor.

5. Further Illustrative Examples

Once the design procedure has been established, the
limitation of this approach is studied by designing different
types of bandwidths and filter orders. In order to correctly
compare the above presented filter with the forthcoming
different ripple filters, the -3 dB fractional bandwidth pa-
rameter \(FBW_{-3dB}\) can be considered, related to the fractional
bandwidth of a Chebyshev filter \(FBW\) as [13]

\[
\frac{FBW_{-3dB}}{FBW} = \cosh \left( \frac{1}{N} \cdot \text{acosh} \left( \frac{1}{\sqrt{10^{\text{ripple}} - 1}} \right) \right)
\]

where \(N\) is the order of the filter and the ripple is given in
decibels. Therefore, if a Chebyshev response with 3 dB
ripple is considered, the hyperbolic arc cosine is zero and
the ratio of fractional bandwidth reduces to unity as
expected, since both fractional bandwidths are the same
in this case. In a similar manner, the above designed filter
with 35% \(FBW\) presents a \(FBW_{-3dB} = 60\%\). Additionally,
two third-order filters with \(f_0 = 2\) GHz, ripple of 0.15 dB
and fractional bandwidth of 70% (\(FBW_{-3dB} = 93\%\) and
90% (\(FBW_{-3dB} = 119\%\)), respectively, have been designed.
The layouts and photographs of each filter are shown in
Fig. 8. In Fig. 9(a)-(b) the lossless frequency response
comparison of these filters is shown, including the elec-
romagnetic simulation and the circuit simulation, consid-
ering the simple and wideband OCSRR model (i.e., with
and without considering \(L_a\)). In Fig. 9(c)-(d) the wideband
frequency response of these filters including measurement
is depicted.

Fig. 7. Frequency response comparison between the ideal
Chebyshev and circuit simulation obtained by auto-
matic matching point optimization considering both the
simple and wideband model of the OCSRR. The
circuit values are: For the case of considering the simple
OCSRR model (in reference to Fig. 1a): \(C = 0.19\) pF, \(L = 0.4\) nH, \(C_r = 0.55\) pF, \(L_{r,c} = 3.17\) pF and \(L_{r,p} = 0.86\) nH. For the case of considering the wideband OCSRR model (in reference to Fig. 4): \(C = 0.19\) pF, \(L = 0.4\) nH, \(L_a = 0.19\) nH, \(C_r = 0.52\) pF,
\(L_{r,c} = 5.76\) nH, \(C_{r,c} = 3.24\) pF and \(L_{r,p} = 0.8\) nH.

Fig. 8. Topology (a), (c) and photograph (b), (d) of the de-
signed third-order filter with 70% and 90% fractional
bandwidth, respectively. The considered substrate is
the Rogers RO3010 with thickness \(h = 0.254\) mm and
measured dielectric constant \(\varepsilon_r = 11.2\). Dimensions are:
(a)-(b): \(l = 13.38\) mm, \(W = 5\) mm, \(G = 0.547\) mm, \(a =
0.16\) mm, \(b = 6.27\) mm, \(c = 0.73\) mm, \(f = 3\) mm. For the
OCSRR: \(r_{rel} = 1.8\) mm, \(c = 0.3\) mm, \(d = 0.16\) mm. For the
OSRR: \(r_{rel} = 1.73\) mm, \(c = d = 0.16\) mm. (c) - (d): \(l =
12.32\) mm, \(W = 5.6\) mm, \(G = 0.574\) mm, \(e = 1.34\) mm.
For the OCSRR: \(r_{rel} = 2.1\) mm, \(c = 0.5\) mm, \(d = 0.16\) mm.
For the OSRR: \(r_{rel} = 1.9\) mm, \(c = 0.3\) mm, \(d = 0.16\) mm.
As can be seen, the higher the bandwidth, the worse the model prediction of the upper transition band. Nonetheless, the wideband model of the OCSRR has been proven to be useful for filters with fractional bandwidths as high as 90%. For higher bandwidth requirements, the model will not accurately predict the upper transition band due to the limited range of operation of the lumped element equivalent circuit model. However, since the limitation to design wider band filters would come only from the model rather than from the layout, it would still be possible to consider this approach for wider band filters, where enhanced nominal bandwidths in the design could be considered to compensate the model inaccuracies. Good agreement between the measured frequency responses and the simulations is obtained. Wide stop-bands are achieved in both cases (the 70% fractional bandwidth filter is free of spurious bands in the depicted range, i.e., four times the central filter frequency). Filter size is as small as 0.13λg × 0.15λg and 0.13λg × 0.17λg for the 70% and 90% fractional bandwidth filter, respectively, where λg is the guided wavelength at the central filter frequency.

Higher order structures have also been studied [12], [14]. An order-5 Chebyshev filter with f0 = 2 GHz, ripple of 0.05 dB and fractional bandwidth of 50% (FBW-3dB = 59%) and an order-7 filter with f0 = 1.87 GHz, ripple of 0.25 dB and fractional bandwidth of 53% (FBW-3dB =...
55%), have been designed. The layouts and photographs of these filters are shown in Fig. 10 and the frequency response in Fig. 11.

From this analysis, it can be deduced that the parameter $L_a$ has to be also considered for high order filters although the bandwidth of the filter is relatively small compared to the cases analyzed before. This is because the small error of the simple OCSRR model is multiplied for each OCSRR stage, resulting in higher inaccuracies both below and above the band as the order increases (whereas for low order filters the simple OCSRR only results in inaccuracies above the band).

This phenomenon can be clearly appreciated in Fig. 11(b), where the seventh-order filter presents a very good agreement between the circuit and electromagnetic response around the band-pass but tends to fail at the edges if the parameter $L_a$ is not considered. On the other hand, it is demonstrated that the design of high order filters is also possible with this approach if the wideband model of the OCSRR is considered, obtaining similar results in measurement (although higher losses are observed for the seventh-order filter due to the non-consideration of conductor losses and fabrication tolerances). In addition, a high selectivity is obtained, with a rejection level better than -30 dB for at least 2.5$f_0$ and 3$f_0$, with filter dimensions as small as $0.24\lambda_g \times 0.17\lambda_g$ and $0.37\lambda_g \times 0.17\lambda_g$ for the fifth- and seventh-order filter, respectively.

![Figure 11](image_url)

**Fig. 11.** Frequency response without losses (a), (b) and wideband frequency response (c), (d) of the designed fifth- and seventh-order filter, respectively. The element values for the circuit simulation without considering $L_a$ are (in reference to Fig. 1): (a), (c): For the external OSRR: $C = 0.274 \text{ pF}$, $C_s = 0.763 \text{ pF}$, $L_s = 8.501 \text{ nH}$; For the central OCSRR: $C = 0.274 \text{ pF}$, $C_s = 0.763 \text{ pF}$, $L_s = 8.501 \text{ nH}$; For the OCSRRs: $L = 0.474 \text{ nH}$, $C_s = 0.274 \text{ pF}$, $C_s = 0.763 \text{ pF}$, $L_s = 8.501 \text{ nH}$; For the OCSRRs of stages 1 and 7: $L = 1.124 \text{ nH}$, $C_s = 0.274 \text{ pF}$, $C_s = 0.763 \text{ pF}$, $L_s = 8.501 \text{ nH}$; For the OCSRRs of stages 2, 4 and 6: $L = 11 \text{ nH}$, $C_s = 0.588 \text{ pF}$, $C_s = 0.26 \text{ pF}$; For the modified values of the OCSRR considering the wideband model with the additional parasitic element $L_a$ are (in reference to Fig. 4): (a), (c): $L = 0.385 \text{ nH}$, $C_s = 0.44 \text{ pF}$, $L_s = 1.259 \text{ nH}$, $L_a = 0.35 \text{ nH}$; (b), (d): For the OCSRRs of stages 1 and 7: $L = 1.56 \text{ nH}$, $C_s = 0.436 \text{ pF}$, $C_s = 0.763 \text{ pF}$, $L_s = 8.501 \text{ nH}$; For the OCSRRs of stages 3 and 5: $L = 0.48 \text{ nH}$, $L_a = 0.31 \text{ nH}$; For the OCSRRs of stages 2, 4 and 6: $L = 6.28 \text{ pF}$, $L = 0.301 \text{ nH}$, $L_a = 0.33 \text{ nH}$.

### 6. Conclusion

High-order CPW broadband bandpass filters have been implemented by combining OSRRs and OCSRRs. The main relevant aspects of this paper have been the introduction of an inductance $L_s$ in the OCSRR circuit model to better describe the wideband response of the particle, as well as the proposal of an optimization routine for filter design. With these aspects, it has been found that the models provide a good description of the proposed filters, and the design of filters subjected to specifications can be achieved.

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