Acoustic Wave Transversal Filter for 5G N77 Band

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Abstract—This work presents the complete synthesis procedure and circuit transformations needed to go from the characteristic polynomials to the transversal network formed by electroacoustic resonators. This filter topology allows the possibility to achieve any filter response in spite of limited electroacoustic coupling. This is crucial for obtaining very wideband filters with arbitrary positions of transmission zeros. Additional circuit transformations reveal the flexibility of such topology to control other important parameters of the acoustic filter such as the impedance of the resonators and their resonant frequencies, which directly affect the manufacturing process.

The transversal topology for electroacoustic filters has been verified by the synthesis, design, fabrication, and measurement of a very wideband filter to meet the requirements of 5G N77 frequency band, centered at 3.75 GHz with 910 MHz of bandwidth.

Index Terms—Bulk acoustic wave (BAW), electroacoustic filters, surface acoustic wave (SAW), synthesis, 5G.

I. INTRODUCTION

The last decade had witnessed the birth and evolution to maturity of 4G technology of mobile communications. The success and evolution of this generation has been driven by the development of RF technologies which contribute heavily to the communication chain in fixed and portable devices. Among those RF devices, the filtering functionality by means of electroacoustic filters has been without any doubt one of the most significant contributors [1].

During this period, the advance in RF electroacoustic technology were developed in response to the increasingly demanding and stringent requirements of spectrum constrains, i.e., the use of new bands, adaption to carrier aggregation approach and the continuous need of further miniaturization of devices, offering high performance filters in super compact form factors allowing the integration of over 70 filters into a single portable device [2][3]. Nowadays there is still no technology that can beat the performance of these sub-6 GHz filters with similar degree of miniaturization [3].

The forthcoming new 5G scenario [4] calls for even more stringent requirements on filter performance, in terms of bandwidth, operation frequency, sharpness of the frequency roll-off at the band edges and compactness of design. Yet continuous reduction of insertion loss is desired. In particular, eMBB (enhanced mobile broadband) targets continuous wideband at operating frequencies above 3.5 GHz [3]. This new demand can be barely met with the current state of the art filters, mostly due to the intrinsic restrictions of the commonly used topologies in the design of electroacoustic filters, primarily, the well-known ladder topology. Such filter topology allows for very sharp narrowband filters. Nevertheless, the filter bandwidth is governed by the electroacoustic coupling of the resonators, which at the same time prescribes the position of the transmission zeros, therefore limiting the type of frequency responses to be achieved [5].

This work presents the development of a transversal filter topology based on electroacoustic resonators where the performance of the synthesized network is not limited by electroacoustic coupling of the material [6] [7]. In practice, however, the selected value of the electroacoustic coupling has effects on the performance of implemented filters. These effects will be discussed latter in the paper.

The paper starts by reviewing the concept, benefits and drawbacks of the transversal topology applied to Bulk Acoustic Wave (BAW) or Surface Acoustic Wave (SAW) filters. Section III fully details the synthesis procedure and demonstrates the ability of this topology to overcome the limitations of electroacoustic coupling. Section IV extends the approach through additional circuit transformations to offer control over the resonant frequencies and impedances of the resonators. Finally, section V, presents the design, fabrication and measurement of a transversal filter at N77 band (3.75 GHz) exhibiting a super wideband of 910 MHz. Section VI concludes on the results of this topology and outlines possible future steps. Additional circuit analysis details and intermediate formulation are provided in the Appendices.

II. ELECTROACOUSTIC RESONATOR AND CONCEPT OF TRANSVERSAL ACOUSTIC FILTER TOPOLOGY

Fig. 1 shows the lumped circuit model to emulate the narrow band frequency response of an electroacoustic
resonator, known as BVD (Butterworth Van Dyke) model. The BVD model is widely used for the synthesis and design of electroacoustic filters based on resonators in BAW or SAW configurations and for evaluating their narrow band responses. The LC branch – represented by $L_s$ and $C_p$ – accounts for the acoustic resonance due the propagation of the acoustic wave along the piezoelectric, whereas the $C_p$ accounts for the electrostatic capacitance, which represents the capacitance created between the electrodes of the resonator and defines its low frequency impedance [5].

![Fig. 1. BVD equivalent circuit model.](image)

Direct analysis of the BVD model reveals the existence of two resonances, at the frequencies with zero impedance and infinite impedance. Those can be respectively referred to as the series resonance, $f_s$, ($f_s = \frac{1}{2\pi\sqrt{L_sC_p}}$) and the parallel resonance $f_p$, ($f_p = \frac{1}{2\pi\sqrt{L_sc_p}}$). The frequency spacing between the two resonances is related to the so-called electroacoustic coupling $k^2_p$ using [5]:

$$k^2_p = \frac{\pi f_s}{2f_p}\tan\left(\frac{\pi f_s}{2f_p}\right) \equiv \pi^2 f_s\frac{f_p}{4f_p}\left(1 - \frac{f_s}{f_p}\right)$$

which in turn can be related to the circuit elements of the BVD model. Of interest for this work is the ratio between the electrostatic capacitance and acoustic capacitance given by:

$$k = \frac{C_p}{C_s} = \frac{1}{\left(1 - \frac{4}{\pi^2}k^2_p \left(1 + \frac{4}{\pi^2}k^2_p\right)^{-1}\right) - 1}$$

Note that $k$ is constant for a given electroacoustic coupling coefficient $k^2_p$. Details on the intermediate steps to obtain (2) are outlined in Appendix I.

Such frequency behavior with high and low impedance resonances allows to create filter responses using a ladder arrangement of the resonators, where the high impedance of the series resonators produces transmission zeros at the upper sideband of the filter, whereas the low impedance of the shunt resonators creates transmission zeros at the lower sideband of the filter. This creates very sharp frequency responses, however due to the limited value of the electroacoustic coupling this results as well in a limited bandwidth and a prescribed position of the transmission zeroes [5]. This also limits the type of responses to be achieved.

In contrast with the ladder topology the transversal topology consists of an arrangement of resonators as outlined in Fig. 2, for the case of a fourth-order filter. In this topology each path is formed by a series resonator, whose impedance and resonant frequency are given by the synthesis method. As it will be seen in the following section the electroacoustic coupling may be arbitrary selected by the designer without detriment to the filter bandwidth and position and number of transmission zeros, therefore allowing the creation of any filter response [6] [7]. Note that this statement is only strictly valid from the synthesis point of view. In practice, however, the selected electroacoustic coupling, as we will see latter in the paper, has effects on the final implemented filter performance.

Fig. 2 also reveals the need of having a differential port at the input or at the output of the filter, and the existence of shunt inductances at the input and output ports. Their origin and effects will be mentioned and referred to along the paper.

![Fig. 2. Transversal topology based on acoustic resonators. Fourth-order filter.](image)

III. SYNTHESIS OF TRANSVERSAL ACOUSTIC FILTER

The synthesis procedure starts by defining the frequency response of the filter by means of the characteristic polynomials, $P(j\Omega)$, $F(j\Omega)$ and $E(j\Omega)$, corresponding to the transmission zeroes, reflection zeroes and poles of the transfer function, respectively, where $\Omega$ is the normalized frequency [9]. Additionally, constant scaling values, usually referred as $e_r$ and $e$, are used to fulfill with the energy conservation condition. From the characteristic polynomials it is possible to define the two port S-parameter matrix of the filter which in turns can be transformed to the two-port admittance matrix $Y_{2x2}(j\Omega)$ [9]. Partial polynomial expansion into $Y_{2x2}(j\Omega)$ gives rise to a synthesized transversal network of the filter response. Although widely reported [9], Fig. 3(a) outlines the topology of a transversal network based on conventional LC resonators, for a fourth-order filter. The black circles correspond to the resonators and the lines correspond to the couplings between source (S) and load (L) to the resonators.

Each path of the transversal network consists of a shunted resonator sandwiched between two admittance inverters, as indicated in Fig. 4(a). The values of $J_{SI}$, $J_{IL}$, $C_i$ and $B_i$ are obtained from the synthesis procedure [9]. The lowpass prototype of a shunt resonator consists of a capacitance ($C_i$) shunt connected to a frequency independent reactive element ($jB_i$). In an equally loaded input and output filter, the two inverters ($J_{SI}$ from the source to the resonator $i$, and $J_{IL}$ from the resonator $i$ to the load) of each path are equal in magnitude and might only differ by $180^\circ$ phase, i.e., $J_{SI} = +/- J_{IL}$. In fact,
for an N even order filter, in N/2 of the transversal paths, the inverters differ by 180° phase, and for an N odd order filter this occurs in (N-1)/2 of the paths [9].

Fig. 3. Outline of transversal topology (a) with conventional LC shunt resonator, (b) with additional direct coupling paths of opposite sign.

A. Transformation to transversal topology based on electroacoustic resonators

This subsection describes the circuit transformations applied to the transversal network of Fig. 3(a), to result in a transversal network as the one in Fig. 2. This starts by considering a single path, Fig. 4(a), and transforming the shunt resonator into a series resonator to obtain the equivalent circuit of Fig. 4(b) (or Fig. 4(c), depending on the phase between input and output couplings and the sign of the reactive element), where the values of $L_{si}$ and $X_{i}$ can be written, respectively, as:

$$I_{si} = \frac{C_{i}}{f_{si}^{2}} \quad \text{and} \quad X_{si} = \frac{B_{i}}{f_{si}^{2}} \quad (3)$$

Resulting in a lowpass series resonant frequency $\Omega_{k,i} = \frac{-X_{si}}{L_{si}}$.

Although the transformation applied is general, for simplicity let us consider the case of a second-order filter with symmetric frequency response. In this scenario the two branches resulting from the transversal network, would be as the ones detailed in Fig. 4(b) and 4(c), respectively. That is branches with equal impedance and mirrored normalized resonant frequency ($\Omega_{k}$ and $-\Omega_{k}$ respectively), due to the opposite sign of the reactive series value, $jX_{si}$ in Fig. 4(b) and $-jX_{si}$ in Fig. 4(c) [9].

Taking the branches forming the filter, we can add at each branch an admittance inverter $jY_{SLi}$ of the same value and opposite sign, as detailed in Fig. 5(a) and 5(b), respectively. This concept is also illustrated in Fig. 3(b). Then, since the inverters are connected in parallel they simply cancel out and therefore do not affect the frequency response. Note that in a filtering structure of order N, this would result in N additional branches and the cancellation should occur from the addition of all branches.

From Fig. 5(a) and using the $\pi$-network circuit model defining the admittance inverter [10], it is then straightforward to obtain the equivalent circuit of Fig. 5(b). It can be also demonstrated that the circuit of Fig. 5(b) results in the equivalent circuit of Fig. 5(d) (see Appendix II).

Fig. 4. Single path of a conventional transversal network. (a) non-normalized values, (b) LC branch when equally phase input and output coupling (c) LC branch with opposite phase input and output coupling. The term T corresponds to a transformer of -1.

The two branches defined by Fig. 5(c) and 5(d) reveal that now each branch contains an electroacoustic resonator, indicated by its lowpass prototype representation [11] [12], (squared in dashed lines). This also reveals the existence of two shunt admittances ($Y = -jY_{SLi}$) at the input and output of each resonator.

Note that in the procedure above no conditions have been set on the values of $jY_{SLi}$. These values can be arbitrary selected, provided that the summation of all new introduced admittance inverters is zero, (or for the case of a canonical network [9], equal to that initially defined by the synthesis).

Fig. 5. (a) and (b) outlined of the two paths, with additional inverters included in the circuits of Fig. 4(a) and 4(b), respectively. (c) and (d) outlined of the two paths, with $\pi$-network as inverters, with the low pass prototype of the BVD is squared.
which are negative (Fig. 5(c) and 5(d)), and in the bandpass domain represent shunt inductances.

C. From lowpass to bandpass prototype

To evaluate the practical values of the resulting electroacoustic filter we need to apply frequency and element transformation to go from the lowpass prototype to the bandpass prototype [10][9]. The LC series acoustic branch is defined by \( L_{s,i} \) and \( C_{s,i} \) as:

\[
L_{s,i} = \frac{L_{Si}}{FBW \cdot \omega_{s,i}} \quad C_{s,i} = \frac{1}{L_{s,i} \cdot \omega_{s,i}^2} \quad (8)
\]

The shunt capacitance should then be obtained to produce the parallel resonant frequency at \( f_{p,i} \). Note that \( f_{p,i} \) has been obtained from the bandpass transformation [10] of \( \Omega_{p,i}, (5) \).

To do that we may define the total admittance of the BVD model as:

\[
Y_{BVD} = \frac{1}{j \omega L_{s,i} + \frac{1}{j \omega C_{s,i}}} + j \omega C_{p,i} \quad (9)
\]

Evaluating the overall admittance at \( f_{p,i}, Y_{BVD} = 0 \), we can isolate \( C_{p,i} \) as:

\[
C_{p,i} = \frac{C_{s,i}}{\omega^2_{p,i} C_{s,i} L_{s,i} - 1} \quad (10)
\]

At this point, it is worth recalling that the frequency response from the lowpass prototype of the BVD model, Fig. 6, does not follow exactly the same frequency dependence that BVD of Fig. 1 [6][12][13]. This gives rise to slight differences between the synthesized lowpass and bandpass responses. This effect is more noticeable for wide band filters [13]. Possible limitation and solution to this discrepancy are discussed in section IV.D.

Once all the circuit parameters of the BVD model of the electroacoustic resonators are fixed, their impedances are also defined. Following the impedance definition [5][14] and using (8), (10), and (2) the impedances can be written as:

\[
Z_{s,i} = \frac{1}{\omega_{s,i} \cdot (C_{p,i} + C_{s,i})} = \frac{L_{s,i}}{FBW (k + 1)} \quad (11)
\]

This expression reveals that the resulting impedance of each resonator is defined by the synthesized lowpass value, \( L_{s,i} \) (so the characteristic polynomials), the prescribed electroacoustic coupling \( k_e^2 \), (which is related with \( k \)) and the fractional bandwidth of the filter.

To finalize this section, we need to consider the value of the input and output shunt inductances, of Fig. 2. As outlined in subsection above the shunt inductances result from the inclusion of the cross-couplings \( J_{SL,i} \), which give rise to negative shunt admittances at each transversal path, see Fig. 5(c) and 5(d). Note that these inductances are used to emulate
a negative capacitor coming from the \( \pi \)-network of the inverter. This creates additional differences between the lowpass synthesized resonator and the corresponding bandpass prototype. Again, only noticeable for wide band filters [13]. As in the case of discrepancies between the lowpass and the bandpass prototype of the BVD resonators, discussion on the limiting effect of the shunt inductances is provided in section IV.D.

The shunt inductance for each individual path, \( L_{\text{shunted},i} \), is obtained as:

\[
L_{\text{shunted},i} = \frac{1}{\omega_0^2 i C_p,i} \tag{12}
\]

Then the inductances of each individual path are all shunt connected to create the shunt inductance at the positive and negative ports of the differential output (see Fig.2), respectively, as:

\[
L_{\text{shunted},UP} = \frac{1}{\sum_{\text{positive branches}} \omega_0^2 i C_p,i}
\]

\[
L_{\text{shunted},DOWN} = \frac{1}{\sum_{\text{negative branches}} \omega_0^2 i C_p,i} \tag{13}
\]

The total \( L_{\text{shunted}} \) results from the shunt connection of \( L_{\text{shunted},UP} \) and \( L_{\text{shunted},DOWN} \).

The synthesis procedure described above has been used to illustrate the concept of electroacoustic transversal filters, where the bandwidth and position of the transmission zeros do not depend on the electroacoustic coupling \( k_e^2 \) [6]. This topology has been also successfully proven for the synthesis of multiband and multiplexer filtering function. In particular, it was demonstrated [15] that this topology allows for a direct connection of filters to create multiplexing responses, without the need of designing any additional matching network or design process, by designing and simulating a 9-plexer configuration where each filter contains 6 resonators, which results in 54 resonators for the whole multiplexer.

Practical implementation issues were also evaluated in [8], including the effects of the losses, sensitivity analysis due to deviation of the synthesized nominal values of the impedances and resonant frequencies and even the effects of a non-ideal BALUN. Reference [8] concluded that transversal topologies are very sensitive to deviations of design parameters that produce unbalancing effects between the transversal branches. Despite that, due to its flexibility, this topology also allows to partially recover the filter response by modifying the external components or scaling the impedance of the transversal branches, among others.

D. Example

This section applies the procedure above to synthesize a sixth-order filter with 100 MHz bandwidth centered at 2 GHz, with four transmission zeros, located at the normalized frequencies of: \( j1.8, j1.66, -j1.66, -j1.8 \). These values have been arbitrarily selected to provide an illustrative example. The conventional synthesis technique described in [9] is used to obtain \( I_{S1}, I_{L}, C_1 \) and \( B_1 \), for each transversal path, which are then used to extract \( L_{\text{shunted},1} \) and \( C_{p,1} \) by means of (3) and (8).

The electroacoustic coupling has also been arbitrarily set to 6.8\%, which results in a \( k=17 \) using (2). From (7), (9), and (10) we obtain \( C_{p,1} \). The resulting impedances (11) of the resonators are: 810\( \Omega \), 354\( \Omega \), 258\( \Omega \), 252\( \Omega \), 337\( \Omega \), and 766\( \Omega \), respectively, which as previously stated would give rise to impractical small resonators.

The results of the bandpass synthesized response are shown in red-solid line in Fig. 7. Blue-dashed line is used to indicate the characteristic polynomial response, showing very good agreement with the synthesized response, despite the different frequency dependence between the lowpass and bandpass prototype [13].

Fig. 7. Frequency response of the transversal filter synthesized as outlined in section III.D. Legend are used to indicate the correspondence of each curve.

IV. CIRCUIT TRANSFORMATION FOR FLEXIBLE APPROACHES

This section applies circuit transformations to the transversal network topology of Fig. 2 not only to prescribe the electroacoustic coupling from the very beginning but also to offer control over critical parameters of the electroacoustic resonators, namely, the impedance and the series resonant frequency \( f_s \).

A. Circuit transformations

To do that, we start by taking the BVD model of Fig. 1 (replicated in Fig. 8(a)) and transform it to its equivalent circuit depicted in Fig. 8(b) [16]. The values of the circuit elements of Fig. 8(b), \( C_2 \), \( L_2 \), and \( C_3 \), are defined from the initial values of the BVD, \( C_5 \), \( L_5 \) and \( C_p \), as outlined in Appendix IV.

The procedure continues by splitting the series capacitance \( C_3 \) in two series connected capacitances \( C_4 \) and \( C_5 \), as depicted in Fig. 8(c). The values of these new capacitances read as:

\[
C_4 = \frac{C_3}{\alpha} \quad C_5 = \frac{C_3}{1 - \alpha} \tag{14}
\]
where $\alpha$ is a new designing parameter and has to be positive and smaller than 1.

\[
C_6 = \frac{C_s + k}{\alpha + ak} \quad L_3 = L_s \left(\frac{1 + ak}{1 + k}\right)^2
\]

\[
C_7 = C_s \frac{k(1 + k)}{1 + ak}
\]

\[
C_5 = \frac{C_p + C_s}{1 - \alpha}
\]

The impedance (11) of the new electroacoustic resonator is:

\[
Z_{\text{new}} = \frac{1}{2\pi f_0(C_6 + C_7)} \approx \alpha Z_0
\]  

(16)

Which is scaled down by $\alpha$. In BAW resonators configuration, the area of the new resonator $A_{\text{new}}$ is increased to $A_{\text{new}} = A/\alpha$ [5], $A$ being the area of the initial resonator. In most cases this results in a more suitable resonator size.

While the new resonator maintains the parallel resonant frequency $f_p$, the new series resonant frequency is obtained as:

\[
f_s = \frac{1}{2\pi \sqrt{\frac{C_p L_s}{\alpha (1 + k\alpha)}}}
\]

(17)

This indeed changes the electroacoustic coupling coefficient, whose new $k$ value (2) ($k_{\text{new}}$) is given by:

\[
k_{\text{new}} = k\alpha
\]

(18)

This previous formulation introduces flexibility in the choice of resonators forming the filter at the expense of including a series capacitance, $C_{S,i}$, at each branch of the transversal network. An outline of the new topology is shown in Fig. 9, again for a fourth-order filter.

Fig. 8. (a) BVD equivalent circuit, (b) reciprocal BVD (c) Reciprocal equivalent circuit by splitting the $C_s$ component, (d) BVD, $L_3$, $C_6$, $C_7$, with a series capacitance $C_5$, (e) BVD with a series capacitance $C_6$.

Now the values of $C_2$, $L_2$, and $C_4$ are used to define a new electroacoustic resonator by means of its equivalent BVD model. The new circuit is outlined in Fig. 8(d) and (8(e)) [16]. The values of the new BVD resonator $L_3$, $C_6$, and $C_7$ and the external capacitance $C_5$ are defined in the set of equation below:

\[
C_6 = \frac{C_s + k}{\alpha + ak} \quad L_3 = L_s \left(\frac{1 + ak}{1 + k}\right)^2
\]

\[
C_7 = C_s \frac{k(1 + k)}{1 + ak}
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This previous formulation introduces flexibility in the choice of resonators forming the filter at the expense of including a series capacitance, $C_{S,i}$, at each branch of the transversal network. An outline of the new topology is shown in Fig. 9, again for a fourth-order filter.

Fig. 9. Transversal filter topology based on acoustic BAW resonators, after the transformation outlined in section IV.

**B. Flexible approaches**

The steps for the synthesis procedure in the flexible approach can be outlined as:

1- Obtain the synthesized $L_{s,i}$ and $C_{s,i}$ of each branch of the conventional transversal network [9].

2- Prescribed the desired electroacoustic coupling obtaining $k_{\text{new}}$ from (2).

3- Set the impedance of the resonators, $Z_{\text{new}}$, and extract $C_6$ by:

\[
C_6 = \frac{C_{\text{total}}}{1 + k_{\text{new}}}
\]

(19)

Where $C_{\text{total}}$ is the initial $L_{s,i}$ and $C_{s,i}$ of point 1.

4- Obtain $C_7$, from $C_7 = C_{\text{total}} - C_6$.

5- From (15) and (18), we obtain the following second order equation:

\[
\alpha^2 C_6 (1 + k_{\text{new}}) - \alpha C_s - k_{\text{new}} C_s = 0
\]

(20)

From which we can isolate the designing parameter $\alpha$.

6- Obtain the acoustic resonant frequency from (17).

7- The $\alpha$ and $k_{\text{new}}$ values are used to obtain the $k$ value from (18).

8- The $k$ value and $C_s$ are used to obtain $C_p$ by means of (2).

9- The $C_5$ values are obtained from (15).

10- The $C_p$ for each resonator are used to obtain the shunt inductance values (12) and (13).

By applying this procedure for a range of suitable impedances one can obtain a set of design curves for each resonator of the electroacoustic transversal filter.

**C. Example**

The formulation above has been used to synthesize the same response of section III.D, by introducing control over the impedance and resonant frequency. Continuing with the prescribed value of 6.8% for the electroacoustic coupling. Fig. 10(a) and 10(b) outline the series resonant frequency (17) of each resonator as a function of its impedance (16), and the required series capacitance $C_{S,i}$ at each branch. Legends in the figures are used to identify each resonator.
By using the design curves in Fig. 10 one could independently select the impedance of each resonator to create the desired filter topology. As an example, here we select a uniform impedance of 50 Ω for all resonators. The response of such selection is depicted in Fig. 7 in black dotted-dashed lines. Note again that this slightly changes the frequency response of the filter.

D. Discussion on the limitations of the transversal network and selection of the solution

Before moving to the development and measurement of a transversal filter, this subsection discusses several implementation and design aspects rising from the synthesis and circuit transformation described above.

As stated in the introduction one of the major assets of the transversal network is that the filter response does not depend on the choice of electroacoustic coupling of the resonators made by the designer. This statement is general and true in the lowpass domain, however when moving to the bandpass domain there are discrepancies between the lowpass and bandpass responses which depend on the selected value of the electroacoustic coupling.

The first discrepancy occurs due to the different frequency behavior between the lowpass BVD model and the bandpass BVD model [12]. This generates differences mostly at frequencies far from the $f_p$ and $f_s$ of the resonators. Note that when implementing a wideband filter with low electroacoustic coupling these discrepancies occurs both in-band and out-of-band degrading the final synthesized bandpass response. Nevertheless, this effect can be completely solved by adding the admittance inverters ($j_{\text{SLI}}$ in section III.A and III.B) directly in the bandpass domain. Details on that solution are fully described in [13].

On the other hand, and as pointed out earlier, the added admittance inverters are created by means of a $\pi$-network where the shunt negative admittances are implemented with inductances, which is therefore a narrowband approach. The validity of this approach depends on the value of the inductances (12) (13), which depend on the FBW and electroacoustic coupling. When the product $FBW \cdot k$ is large - so the filter bandwidth is large in terms of the electroacoustic coupling - the resulting shunt inductances are small and then the in-band discrepancies between the lowpass and bandpass may become relevant [13]. Although this might have a limiting effect, optimization of the circuit parameters can be used to mitigate this effect, as done in the filter design in the following section.

The effect of having small shunt inductance at the input and output ports has also a significant effect on the losses, since these external inductances have small Q values (around 30).

We therefore conclude that in real implementation larger electroacoustic couplings help to obtain better wideband filters. Despite that the achievable bandwidth for a given electroacoustic coupling is larger than that of a conventional ladder configuration, and this can be achieved with a smaller number of resonators.

Continuing with practical aspects, the design curves of Fig. 10, reveal that for a given electroacoustic coupling there are many solutions to create a certain filter response corresponding to different choices of the impedance of each individual resonator. One might wonder therefore if a preferred solution exists.

In our opinion the preferred solution depends on manufacturing capabilities, operating frequency range, bandwidth, order of the filters, near band selectivity (position of transmission zeros) and losses [8].

For instance, when the impedances are not specified, the values obtained by synthesis are generally different which results in resonators with different areas and therefore different Q values. So, each path of the transversal network has different losses and therefore different contribution, which affects for instance the out-of-band position of transmission zeros, since those are created by cancellation [8]. This effect is more harmful when the synthesized impedances are widely spread which occurs when synthesizing sharp frequency responses, i.e., near band transmission zeros and/or high order filters.

![Fig. 10. Designing curves, (a) series resonant frequency for each resonator as a function of the impedance of the resonator, (b) external capacitance $C_s$ for each branch as function of the impedance of the resonator.](image-url)
resonator, so high impedances are given by small resonators, which are usually more affected by lateral mode effects, and nonlinearities [18] [19]. Note that this consideration is more important for filters operating at higher frequencies.

A foreseeable good solution is to select a reasonable resonator size with good performance and then setting all the resonators to equal or similar impedance which helps to keep the balance between all paths. However, this requires the inclusion of external series capacitances which might not be desirable. Additionally, the selected impedance also defines the value of the shunt inductances which have a limited Q, affecting therefore the overall losses of the filter.

The number and values of resonant frequencies is also an important aspect of the manufacturing process.

In our opinion there is no unique strategy for selecting the preferred solution and the selection of the final configuration depends on all the parameters listed above and on the configuration of the electroacoustic resonator, for instance if it is BAW or SAW.

Another important aspect of this topology is the need of a differential output (or input), which might be a disadvantage. When the differential port is realized using a magnetic BALUN, this might result in a poorer filter performance [8] and additional insertion losses.

V. DEVELOPMENT OF AN ACOUSTIC TRANSVERSAL FILTER PROTOTYPE

The specifications of the filter to be implemented partially correspond to the requirements of the N77 band of 5G NR (New Radio) located in Range 1 and included in the Sub-6 GHz bands [17]. Some electrical specifications are listed in Table I.

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Range (MHz)</th>
<th>Value (dB)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Insertion Loss</td>
<td>3300-4210</td>
<td>2</td>
</tr>
<tr>
<td>Return loss</td>
<td>3300-4210</td>
<td>15</td>
</tr>
<tr>
<td>Attenuation</td>
<td>2496-2570</td>
<td>30</td>
</tr>
<tr>
<td>Attenuation</td>
<td>2300-2482</td>
<td>25</td>
</tr>
<tr>
<td>Attenuation</td>
<td>4400-5000</td>
<td>10</td>
</tr>
<tr>
<td>Attenuation</td>
<td>5150-3925</td>
<td>25</td>
</tr>
<tr>
<td>Attenuation</td>
<td>6600-6740</td>
<td>40</td>
</tr>
</tbody>
</table>

The main specifications are the bandwidth, return losses, and insertion losses. Note that this makes a filter bandwidth of 910 MHz, and central frequency of 3.75 GHz, resulting in a fractional bandwidth of 24%. These specifications are very challenging and could be difficult to meet with a conventional and commonly used ladder filter topology, which is able to achieve wide bandwidths only with the aid of external inductors and employing higher order filters. The transversal topology designed in this work corresponds to a fourth-order filter.

A. Synthesized filter network

We apply the synthesis method reported in section III, using a uniform electroacoustic coupling of 13%, which is defined by the material and resonator configuration used in the design and implementation. The value of the electroacoustic coupling has been obtained from previous measurements of resonators, designed to operate at the same frequency band. The corresponding $k$ value (2) for such electroacoustic coupling is 8.3. The resulting topology is as outlined in Fig. 2, whose values are listed in Table II. Note that due to the symmetry of the resonators forming the filter $L_{\text{shunted, } \uparrow}$ and $L_{\text{shunted, } \downarrow}$ are equal to $2L_{\text{shunted}}$.

<table>
<thead>
<tr>
<th>Resonator</th>
<th>Branch</th>
<th>$C_r$ (pF)</th>
<th>$f_v$ (MHz)</th>
<th>$k_r^2$ (%)</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>+</td>
<td>0.29</td>
<td>3033.6</td>
<td>13</td>
</tr>
<tr>
<td>2</td>
<td>-</td>
<td>0.684</td>
<td>3267.9</td>
<td>13</td>
</tr>
<tr>
<td>3</td>
<td>+</td>
<td>0.684</td>
<td>3758.9</td>
<td>13</td>
</tr>
<tr>
<td>4</td>
<td>-</td>
<td>0.29</td>
<td>4244.6</td>
<td>13</td>
</tr>
</tbody>
</table>

$I_{\text{shunted}}=1.28$ nH

It is also very important to note that the filter topology does not include any external capacitance. Recall that external capacitances were included to reduce resonator impedances, requiring therefore larger resonators, being more suitable from a manufacturing point of view. This is not needed in this case due to the large FBW used for the design, and the larger electroacoustic coupling (11). In this case feasible impedances can be achieved by simply doubling the resonators and connecting them in series, as indicated in Fig. 11. Now each resonator has double the static capacitance, and therefore double the area.

Fig. 11. Network topology of the synthesized filter for the final prototype, doubling resonators.

As mentioned earlier the shunt inductances come from synthesized negative shunt capacitances. In a narrowband filter this has a minor effect on the in-band performance, but for such a wideband filter this affects the in-band matching [13]. Circuit optimization of the initial values of the synthesized network has been performed to meet the in-band specifications.

The optimized values of the network are used to proceed with the design of the filter layout, that is the arrangement of the resonators, areas and layers. Description of the design and simulated results are detailed in the following subsection.
B. Filter Design

The actual shapes, dimensions and arrangement of the resonators are shown in Fig. 12, along with the values listed in Table III. Note that the capacitances of Table III are approximately double of the ones in Table II, this is because the resonators are doubled and serially connected.

<table>
<thead>
<tr>
<th>Table III</th>
<th>PARAMETERS OF THE TRANSVERSAL TOPOLOGY OF FIG. 12</th>
</tr>
</thead>
<tbody>
<tr>
<td>Resonator</td>
<td>C_p (pF)</td>
</tr>
<tr>
<td>1</td>
<td>0.58</td>
</tr>
<tr>
<td>2</td>
<td>1.398</td>
</tr>
<tr>
<td>3</td>
<td>1.63</td>
</tr>
<tr>
<td>4</td>
<td>0.85</td>
</tr>
</tbody>
</table>

As foreseen in the analysis performed in [8], the transversal topology is a very sensitive network configuration. This is also true in the case of parasitic effects like the self-inductance existing between the different paths forming the network. Which need to be the same within ~1 pH. The design process also reveals that mutual inductance between paths needs to be balanced.

These restrictions call for a very symmetric design, as observed in the proposed layout of Fig. 12. Note moreover that the 4 GNDs at the corners of the die are dummy GNDs added for design rule reasons. Dimensions between those GNDs patches are approximately 400 μm by 550 μm.

The layout reveals that each path is created by two identical electroacoustic resonators, which are series connected to create an equivalent resonator of half the area. As outlined above, the final layout is very symmetric. The shape of the resonators is not a conventional rectangle to reduce the effect of lateral modes [5]. Additionally, although not seen in the layout, each resonator is designed with a border ring (BO) to further reduce such effect [5].

Simulations of the final design are shown in Fig. 13 and are indicated in blue thin-solid line. The simulated results include acoustic and metallic losses as well as the parasitic effects. Fig. 13(a) shows the wideband response of the transmission coefficient (S21), Fig. 13(b) and 13(c) outline the in-band details of the transmission and reflection coefficients, respectively. Grey rectangles and grey dashed lines are used to indicate the electrical specifications of Table I. The
transmission coefficient is obtained from the differential output (Out- and Out+).

![Diagram](image.png)

**Fig. 14.** Layout mounted in the laminate.

### C. Fabricated and measured filter

The layout of the filter above has been fabricated and mounted on a laminate to enable proper measurements. The laminate uses long signal traces to reduce the undesired coupling between probes when measuring. Those laminate traces will be de-embedded to extract the 3-port filter response to be then connected to a BALUN in a circuit simulator. Then the filter response is constructed. Fig. 14 shows the die mounted on the laminate.

The measured results are shown in the Fig. 13, in dashed-magenta line, along with the optimized model, in blue-thin-solid line and the results of the hybrid simulation in thick-solid black line. The hybrid simulation is obtained by taking measurements of individual resonators (see Fig. 15) and connecting them to create the topology of Fig. 11.

Fig. 13(c) shows the reflected coefficient S11, showing all very good in-band agreement. Fig. 13(b) shows the in-band insertion losses for the three cases, also showing very good agreement, but the effect of the BO that appears in the measured results. This effect produces an additional ripple at the in-band lower frequency range, typical from filters with resonators with BO [5].

Fig. 13(a) depicts the wideband rejection response for the three cases. While the optimized circuit model response (thin-solid blue line) does not show any peak out-of-band, the hybrid response (thick-solid black line) and the measured response (dashed magenta line) shows undesired peaks at the lower side band and at the upper side band of the rejection, which correspond to undesired spurious resonances appearing in the response of resonators of the individual paths. These out-of-band peaks are pointed by black arrows.

To complement this explanation, Fig. 15 shows the measured responses of the four individual resonators fabricated in the same wafer, where the existence of these resonances is also pointed with black arrows. Fig. 15 also reveals that the out-of-band impedances of the resonators are equal in pairs which guarantees the filter out-of-band rejection.

Although not measured in this work the second harmonic is expected to be negligible. This statement comes from the fact that each branch is created by two identical resonators connected in an (anti-)series manner [18]. Note that the anti-series connection does not affect the linear performance of the filter [18] but provides cancellation of the second harmonic produced by equal area resonators within each transversal filter branch. Having resonators of double the area also foresees a reduction of other manifestations of the nonlinear effects, such as intermodulation products. Further insight into second harmonic generation in transversal configurations is reported in [19].

![Wideband response](image.png)

**Fig. 15.** Wideband response of the four resonators. Blue-thick-solid line (resonator 1), black-thick-dashed line (resonator 2), black-thin-dashed line (resonator 3) and blue-thin-solid line (resonator 4). The black arrows indicate the peaks corresponding to spurious resonances appearing in the stack of materials.

### VI. Conclusion

This paper has developed the mathematical formulation and synthesis of transversal filters based on electroacoustic resonators, where the designer can select the electroacoustic coupling coefficient from the very beginning of the synthesis process. An additional circuit transformation procedure has been created to offer flexibility in network synthesis, whereby control of the impedance and resonant frequency of the resonators is possible.

The discrepancies between the lowpass and passband synthesized responses outlined the limitation on the filter performance due to the selected electroacoustic coupling, which have been extensively discussed in section IV.D.

The feasibility of this network has been proven by the synthesis of a 24% fractional bandwidth filter, which was subsequently fabricated and measured. The simulated and measured results show the feasibility of the transversal filters based on acoustic wave resonators and the possibility to overcome the sensitivity of the network.
A significant drawback of this configuration is the degradation of filter out-of-band rejection due to spurious resonances exhibited by constituent resonators. In a ladder configuration, these undesired resonating modes are usually filtered by the cascaded ladder stages and do not significantly affect the final response. Techniques to mitigate the spurious modes effects could be used to enhance the filter performance.

This opens the possibility of creating novel configurations based on transversal arrangement, such as combination with ladder topology to target advanced frequency responses. Note as well that this concept could be applied to other configurations of acoustic resonators such as coupled acoustic resonators (CRF) [20], in which case the BALUN could be removed from the final configuration.

Further research on this new configuration is also required to study important topics, such as power handling and non-linear effects.

APPENDIX I: EXTRACTION OF THE DESIGNING PARAMETER k

The k value in (2) defines the ratio between the shunt capacitance $C_p$ and the series capacitance $C_s$ of the BVD equivalent circuit of Fig. 1. This parameter is used along the design process as outlined in section III and IV, and it is related with the electroacoustic coupling $k_2^2$ as indicated in (2). Recalling the expression of the series resonance frequency, $f_s$, ($f_s = \frac{1}{2\pi\sqrt{C_pL_s}}$) and parallel resonance frequency $f_p$, $f_p = \frac{1}{2\pi\sqrt{C_p+C_s}}$ as a function of its circuit parameters, $C_p$, $C_s$, and $L_s$, it is straightforward to obtain the ratio between $C_p$ and $C_s$ as:

$$\frac{C_p}{C_s} = \left(\frac{f_p}{f_s}\right)^2 - 1 \quad (21)$$

Also, by taking the expression of the electroacoustic coupling (1), $k_2^2 \cong \frac{n^2}{4}\left(1 - \frac{f_s}{f_p}\right)^2$, we can extract:

$$\frac{f_s}{f_p} \cong 0.5 + \sqrt{0.25 - \frac{4}{n^2}k_2^2} \quad (22)$$

Which by applying a Taylor approximation the relationship yields:

$$\frac{f_s}{f_p} \cong 1 - \frac{4}{n^2}k_2^2\left(1 + \frac{4}{n^2}k_2^2\right) \quad (23)$$

Now by substituting (23) into (21) we obtain the expression in (2).

Note that although this is an approximation the value of $k$ as a function of $k_2^2$, from (2) and the one resulting without any approximation differ less than 0.3% within the range of electroacoustic couplings from 1% to 20%.

APPENDIX II: ADMITTANCE MATRICES

The analysis in this section demonstrates the equivalence between circuits of Fig. 5(b) and 5(d). The following set of equations (23)-(27), outline the ABCD and Y matrix of basic building blocks used in such transformations, those are a series impedance $Z_{\text{series}}$, and admittance inverter J and a transformer (-1) T.

$$ABCD_{\text{series}} = \begin{bmatrix} 1 & Z \\ 0 & 1 \end{bmatrix} \quad (24)$$

$$ABCD_J = \begin{bmatrix} 0 & 1 \\ -jJ & 0 \end{bmatrix} \quad (25)$$

$$ABCD_{T=-1} = -\begin{bmatrix} 1 & 0 \\ 0 & 1 \end{bmatrix} \quad (26)$$

$$Y_J = \frac{1}{jJ} \begin{bmatrix} 1 & 1 \\ 0 & 1 \end{bmatrix} \quad (27)$$

$$Y_{\text{series}} = \frac{1}{Z} \begin{bmatrix} -1 & -1 \\ -1 & 1 \end{bmatrix} \quad (28)$$

The overall ABCD matrix of Fig. 5(d) through the following calculations:

$$Y_{\text{TOTAL}} = Y_{\text{series}} + Y_J = \begin{bmatrix} \frac{1}{Z} & \frac{1}{jJ} - \frac{1}{Z} \\ \frac{1}{jJ} + \frac{1}{Z} & \frac{1}{Z} \end{bmatrix} \quad (29)$$

$$ABCD_{\text{TOTAL}} = \frac{1}{jJ - \frac{1}{Z}} \begin{bmatrix} -\frac{1}{Z} & -1 \\ \frac{1}{jJ} - \frac{1}{Z} & \frac{1}{Z} \end{bmatrix} \quad \frac{1}{Z} \quad (30)$$

$$ABCD_{\text{TOTAL}} \cdot ABCD_{T=-1} = \begin{bmatrix} -1 & \frac{1}{jJ} - \frac{1}{Z} \\ \frac{1}{jJ} - \frac{1}{Z} & \frac{1}{Z} \end{bmatrix} \quad (31)$$

Analogously the ABCD matrix of the circuit of Fig. 5b

$$ABCD_{\text{series}} \cdot ABCD_{T=-1} = -\begin{bmatrix} 1 & Z \\ 0 & 1 \end{bmatrix} \quad (32)$$

$$ABCD_J \cdot ABCD_{T=-1} = -\begin{bmatrix} 0 & 1 \\ -jJ & 0 \end{bmatrix} \quad (33)$$

Transforming the above into Y matrices,

$$Y_{T=-1} = \frac{-1}{jJ} \begin{bmatrix} 1 & 1 \\ 1 & 0 \end{bmatrix} \quad (34)$$
\[
Y_{\text{series}T=\text{a-1}} = \frac{1}{Z} \begin{bmatrix} 1 & 1 \\ 1 & 1 \end{bmatrix}
\]

\[
Y_{\text{TOTAL}T=\text{a-1}} = Y_{\text{series}T=\text{a-1}} + Y_{J=\text{a-1}}
= \begin{bmatrix} \frac{1}{Z} & \frac{1}{Z} \\ \frac{1}{Z} & \frac{1}{Z} \end{bmatrix}
\]

\[
ABCD_{\text{TOTAL}T=\text{a-1}} = \frac{1}{1 + \frac{1}{Z}} \begin{bmatrix} \frac{1}{Z} & \frac{1}{Z} \\ \frac{1}{Z} & \frac{1}{Z} \end{bmatrix} - \frac{1}{Z}
\]

The previous and simple circuit analysis demonstrates that the network of Fig. 5(b) is identical to the Fig. 5(d).

APPENDIX III: EQUIVALENT CIRCUIT MODEL

From the conventional lowpass to bandpass frequency transformation, we can also write down:

\[
\Omega_s = \frac{1}{\text{FBW}} \left( \frac{f_s}{f_0} - \frac{f_0}{f_s} \right) \quad \Omega_p = \frac{1}{\text{FBW}} \left( \frac{f_p}{f_0} - \frac{f_0}{f_p} \right)
\]  

From (38):

\[
\Omega_s - \Omega_p = \frac{1}{\text{FBW}} \left( \frac{f_s}{f_0} - \frac{f_0}{f_s} - \left( \frac{f_p}{f_0} - \frac{f_0}{f_p} \right) \right)
\]  

which can be re-written as:

\[
\Omega_s - \Omega_p = \frac{1}{\text{FBW}} \left( \frac{f_s-f_p}{f_0} - \frac{f_0}{f_s} \frac{f_p-f_0}{f_p} \right) = \frac{2}{\text{FBW}} \left( \frac{f_s-f_p}{f_0} \right)
\]  

Where we consider the following $f_0 = \sqrt{f_s f_p}$, then

\[
\frac{f_0 f_p - f_0 f_s}{f_s f_p} = \frac{f_s-f_p}{f_0}.
\]

Now from the expression of the coupling coefficient $k_e^2$ (5),

\[
\frac{f_s}{f_p} \cong 1 - \frac{4}{\pi^2} k_e^2 \left( 1 + \frac{4}{\pi^2} k_e^2 \right).
\]

we can obtain the following relation:

\[
\frac{f_p-f_s}{f_p} \cong \frac{4}{\pi^2} k_e^2 \left( 1 + \frac{4}{\pi^2} k_e^2 \right)
\]  

which in turn can be considered in (6) to result in (7)

\[
J_{SL1} = \frac{\text{FBW}}{2L_s} \left( \frac{4}{\pi^2} k_e^2 \left( 1 + \frac{4}{\pi^2} k_e^2 \right) \right)
\]

APPENDIX IV: EQUIVALENT CIRCUIT MODEL

The BVD circuit model of Fig. 8(a), has an equivalent circuit model [16] as shown in Fig. 8(b). This is a circuit that provides the same frequency response by means of a different network. Such a transformation is very well established and allows to define the new circuit parameters, $C_2, L_2$ and $C_3$. The values of the series and shunt resonances now can be obtained from the circuit parameters defining the equivalent circuit. The set of equations are detailed in (43), below.

\[
C_2 = C_p + \frac{C_p^2}{C_s} \quad f_p = \frac{1}{2\pi \sqrt{L_2 C_2}}
\]

\[
C_3 = C_p + C_e \quad f_s = \frac{1}{2\pi \sqrt{L_2 (C_2 + C_3)}}
\]

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R. Perea-Robles et al., “Synthesis and Design of Filters and Multiplexers with Acoustic Transversal Topology” Accepted for publication at International Journal of Microwave and Wireless Technologies

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