Design methods for micro-electro-mechanical bandpass filters with transmission zeros

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ABSTRACT. A design method for high-order electro-mechanical filters allowing the synthesis of architectures of electro-mechanical filters from equivalent electrical prototypes is presented. Conditions for existence of the equivalent mechanical system for a given electrical network are derived. This method is applied to the problem of the design of band-pass electro-mechanical filters with transmission zeros in the stopband. Electro-statical coupling of micro-mechanical resonators that avoids the use of a mechanical coupling spring is introduced. The application to the simulation and design of a bandpass filter with finite transmission zeros implemented in a thick-layer epi-poly silicon micro-machining technology is shown.

Keywords: Electro-mechanical filter, transducer, IF, devices and components, structured design methodologies.

1. Introduction.

There is a strong interest for miniaturized highly selective filters in mobile telecom applications. Silicon micromachined electromechanical (EM) filters are one of the possible solutions and gained recently considerable attention. The filtering function is completely achieved by an oscillating mechanical structure. Sensing and actuating are achieved by means of capacitive transducers. The interest of such a system for filtering applications is that mechanical oscillating structures have a much higher quality factor than their electrical LC counterparts. After it was shown that filtering is possible with a single-oscillator system, numerous studies have been carried out in order to develop high-order filter architectures meeting specifications of real telecommunication systems. Up to now, all high-order implementations have been based on the simple “coupled-resonators” structure[1] implementing “all-pole” transfer functions. To design such filters, the mechanical oscillating structure is decomposed into elementary linear mechanical elements with lumped parameters, and their equivalent electrical models are built. After that the existing design method for coupled-resonators electrical passive filters is applied to calculate the element values.

In many cases, finite transmission zeros in the stopband help to reduce the filter order. The coupled-resonator method for electro-mechanical filter design doesn’t offer this possibility, so another architecture has to be found to implement such filters. However the existing design methods for MEM filters don’t allow the synthesis of the actual filter architecture, they just allow to analyze the given filter architecture and to calculate parameter values for given specifications [1,5].

Our work aims to develop a design method that allows the re-use of existing design methods for electrical passive filters and to synthesize mechanical filter architectures from their electrical prototypes. This method should be applicable to the design of a filter with transmission zeros.

Firstly we have studied the properties of electro-mechanical and mechanical-electrical transformations, the conditions of their existence and the architectural properties of transformed systems. Then we derived the rules for these transformations allowing us to make conclusions about their application to mechanical filter design. Making use of the obtained results, we have conceived an architecture of electro-mechanical filters with transmissions zeros. A new method of electrostatical coupling was used instead of the usual coupling by mechanical springs. This architecture was then implemented in a silicon epi-poly surface-micromachining technology.

2.Transformations between mechanical and electrical domains.

To use the electrical prototype networks for the synthesis of mechanical systems, the transformation from electrical into the mechanical domain has to be defined. We will briefly show how this definition can be done.

It is well known that for a simple second-order oscillating LCR electrical network the equivalent mechanical system is composed of a spring, mass and damper (fig. 1). This equivalence is based on the similarity of mathematical equations describing both systems, it can easily be extended for more complex systems.

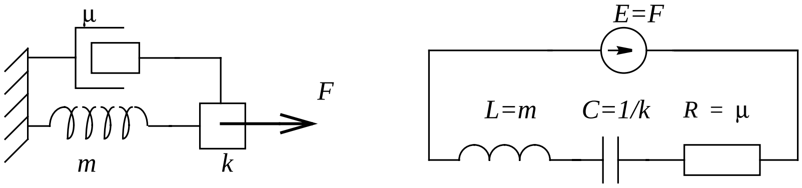


Figure 1. Elementary mechanical circuit and its electrical equivalent

2.1 Transformation from mechanical to electrical domain

To define the transformation from electrical to mechanical domain let us firstly define the inverse transformation. We note that all considered systems are submitted to sinusoidal excitations, so we are only interested in the small signal mode. Let us consider an arbitrary complex lumped-elements mechanical system consisting of masses, springs and dampers (fig. 2). To describe the behavior of such a system, Newton's second law equation has to be written for each mobile node. We call the mobile node a point to which springs and dampers are attached, a mobile node can contain mass or not. For the *i*-th node the equation in the small signal mode will have the following form (we assume that the *i*-th node is attached to some springs and dampers that have the other end attached to other nodes (*j*-th) or fixed):

$$F_i + k_i \cdot X_i + k_{ij} \cdot (X_i - X_j) + \mu_{ij} \cdot j \cdot \omega \cdot (X_i - X_j) + \ldots = (j \cdot \omega)^2 \cdot m_i \cdot X_i, \qquad (1)$$

where X_i and X_j denote the complex amplitudes of the displacement of the *i*-th and *j*-th node, m_i denotes the mass of the i-th node, k_{ij} and μ_{ij} denote the stiffness of the spring and the damping coefficients of the dampers attached to *i*-th and *j*-th nodes, k_i denotes the stiffness of the spring joined to the *i*-th nodes and fixed at the other end. (2)

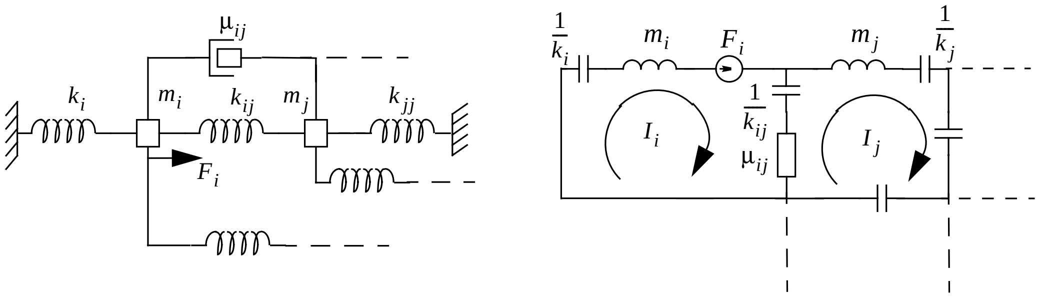


Figure 2. The equivalent electrical network for an arbitrary lumped mechanical system.

Let us rewrite this equation in a form suitable for the matrix representation, and expressing the node displacements via the node velocities (we assume that $V = j \cdot \omega \cdot X$):

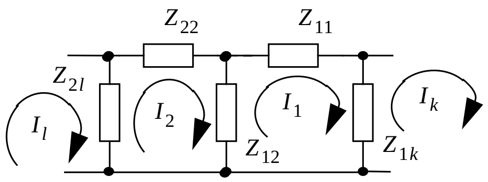
$$\left(j \cdot \omega \cdot m_i + \frac{k_i}{j \cdot \omega} + \frac{k_{ij}}{j \cdot \omega} + \mu_{ij} + \ldots\right) \cdot V_i - \left(\frac{k_{ij}}{j \cdot \omega} + \mu_{ij} + \ldots\right) \cdot V_j + \ldots = F_i \qquad (3)$$

Establishing similar equations for each mobile node of the system, we obtain the matrix equation of the system:

$$\begin{bmatrix} \frac{k_1}{j\cdot\omega} + j\cdot\omega\cdot m_1 + \frac{k_{1j}}{j\cdot\omega} + \mu_{1j} + \ldots & -\frac{k_{12}}{j\cdot\omega} - \mu_{12} & \ldots & -\frac{k_{1j}}{j\cdot\omega} - \mu_{1j} & \ldots & -\frac{k_{1N}}{j\cdot\omega} - \mu_{1N} \\ -\frac{k_{21}}{j\cdot\omega} - \mu_{21} & \frac{k_2}{j\cdot\omega} + j\cdot\omega\cdot m_2 + \frac{k_{2j}}{j\cdot\omega} + \mu_{2j} + \ldots & \ldots & -\frac{k_{2j}}{j\cdot\omega} - \mu_{2j} & \ldots & -\frac{k_{2N}}{j\cdot\omega} - \mu_{2N} \\ \ldots & \ldots & \ldots & \ldots & \ldots & \ldots \\ -\frac{k_{j1}}{j\cdot\omega} - \mu_{j1} & -\frac{k_{j2}}{j\cdot\omega} - \mu_{j2} & \ldots & \frac{k_j}{j\cdot\omega} + j\cdot\omega\cdot m_j + \frac{k_{jm}}{j\cdot\omega} + \mu_{jm} + \ldots & \ldots & -\frac{k_{jN}}{j\cdot\omega} - \mu_{jN} \\ \ldots & \ldots & \ldots & \ldots & \ldots & \ldots \\ -\frac{k_{N1}}{j\cdot\omega} - \mu_{N1} & -\frac{k_{N2}}{j\cdot\omega} - \mu_{N2} & \ldots & -\frac{k_{Nj}}{j\cdot\omega} - \mu_{Nj} & \ldots & \frac{k_N}{j\cdot\omega} + j\cdot\omega\cdot m_N + \frac{k_{Nj}}{j\cdot\omega} + \mu_{Nj} + \ldots \end{bmatrix} \times \begin{bmatrix} V_1 \\ V_2 \\ \ldots \\ V_j \\ \ldots \\ V_N \end{bmatrix} = \begin{bmatrix} F_1 \\ F_2 \\ \ldots \\ F_j \\ \ldots \\ F_N \end{bmatrix} \qquad (4)$$

From this matrix equation the equivalent electrical network can easily be found. For our study we assumed that the matrix describes a network by the mesh currents method (the node voltage method could also be used). The mesh current method consists in the following [3].

The electrical network is decomposed into a set of N independent non-redundant meshes that includes all branches of the network. It is assumed that to each mesh corresponds a current circulating only in the given mesh (fig. 3). The branch currents are created by the superposition of the mesh currents. The goal of the method is to find the mesh current values, from which all branch currents can be derived.



I_1, I_2, I_l, I_k are the mesh currents of meshes 1, 2, l, k;
Elements Z_{11}, Z_{22} belong to the proper branches of the meshes 1, 2;
Element Z_{12} belongs to the common branch of the meshes 1 and 2;

Figure 3. Demonstration of the current mesh method.

To achieve this, the impedence matrix and the voltage source vector are generated for the given network.

The impedance matrix is a $N \times N$ matrix, which includes two types of elements: (1) mesh impedances, that include sums of the impedances of all elements belonging to each current mesh, these elements are placed in the diagonal of the matrix and (2) common impedances of each couple of meshes that include sums of the impedances of elements belonging simultaneously to the two given meshes. The common impedances are taken with the positive sign if the directions of mesh currents in the common branch are the same, and with the negative sign, if their directions are opposite. The impedance matrix is symmetrical. In such a way two types of branches can be defined for any given mesh: proper branches, that are not shared with other meshes, and common branches, that are shared with some other meshes.

The N-elements source vector contains the sums of voltage sources acting in each mesh.

These two matrixes are used in the network equation $[Z] \times [I] = [E]$, where Z is the impedance matrix, I is the vector of the mesh currents, and E is the vector of the voltage sources of the each mesh. From this equation the vector of mesh currents I can be found.

It is obovious that matrixes Z and E contain all information about the topology of the network, and if these matrixes are known, the network can be derived from them.

Let us analyze the matrix equation (4) obtained for the mechanical system. It has the form of one written for an electrical network, where the correspondence between mechanical and electrtical values is defined in [4], [5], so the electrical network described by this equation can be found (fig. 2).

The network described by equation (4) has the following properties: (1) since all common impedances (non-diagonal elements of the impedance matrix) are negative, any two mesh currents passing through a (common) branch have opposite directions regarding to each other. That is possible only for planar networks [3], consequently, the equivalent network is planar; (2) impedances corresponding to the inductors (to the masses in the mechanical system) belong only to diagonal elements of the matrix. Therefore, all inductors belong only to proper (non-shared) branches of meshes; (3) impedances corresponding to the two-end mechanical elements that join two mechanical nodes (capacitors and resistors in the network, springs and dampers in the mechanical system) are situated in the matrix in the crossing of two rows and two columns corresponding to these nodes. Therefore, these mechanical elements transform to electrical components of the network that belong to the common branches of the meshes corresponding to the mechanical nodes; (4) impedances corresponding to the two-end elements that are fixed at one end belong only to diagonal elements of the matrix, and therefore the electrical components corresponding to these mechanical elements belong to proper branches of the mesh; (5) from (4), (3) and (2) we can conclude that the mesh definition is made in such a way that all components of the network belong to not more than 2 meshes. Such a mesh definition is always possible for planar networks; (5) the force acting on a node corresponds to a voltage source put in the proper branch of the mesh corresponding to this node.

An example of the application of the transformation defined above to a complex mechanical system is shown in fig. 4.

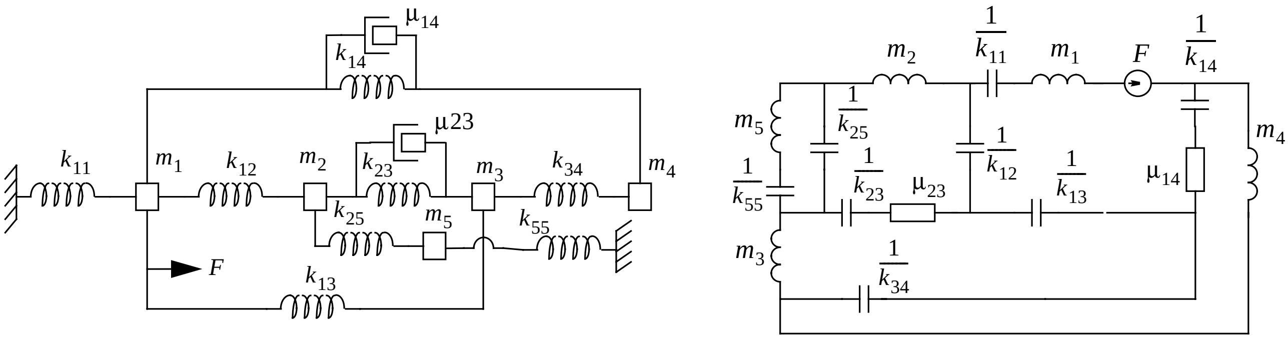


Figure 4. Example of a complex mechanical system and its equivalent electronic network.

2.2 Transformation from electrical to mechanical domain.

Since our reasoning has been made for an arbitrary mechanical system, we can conclude that any lumped mechanical system has an electrical equivalent network. It is easily seen that all the networks obtained from this transformation form a network sub-class, that is defined by the properties mentioned above. To make an inverse transformation and to find the equivalent mechanical system for a given electrical network, it is obvious that the latter has to belong to this sub-class. It means that not all electrical networks have an equivalence in the mechanical domain.

Let us resume this very important result and formulate the conditions of the existence of an equivalent mechanical system.

To have an equivalent lumped-elements mechanical system, the electrical network has to have the following properties:

- the network must be planar;
- the mesh definition must be possible, for which:
    inductors belong only to proper branches of the meshes;
    two-ended elements (resistors and capacitors) belong to not more than two meshes.

Networks that don’t satisfy these requirements have no equivalent mechanical system.

If an electrical network is convertible to the mechanical domain, i. e. it has all properties mentioned above, its transformation to the mechanical domain is obvious. One should proceed in the same way as we have done to find the equivalent electrical network: compare the matrix equations for both systems and synthesize the mechanical equivalent system. Let us resume the algorithm of such a transformation.

To find the equivalent mechanical system for the given electrical network that satisfies the existence conditions defined below, one should proceed as follow:

- Define meshes in the network in the way described in the conditions of existence. Each mesh corresponds to a mechanical mobile node;
- A mass should be placed in the node if there is an inductor in the corresponding mesh;
- A spring or a damper fixed at one end should be attached to the node if the mesh contains a capacitor or a resistor in its proper branch;
- A spring or a damper should be attached to two nodes if the common branch of two meshes contains a capacitor or a resistor;
- A force acts on the nodes that correspond to meshes containing voltage sources.

2.3 Electrical representation of losses in mechanical oscillating structures.

There is basicly two kinds of losses in mechanical structures: viscous losses due to the gaz damping and internal losses in the material due to the deformation of springs. The first are of extrinsic nature and can be reduced to a negligible level by putting the resonator in a deep vacuum. The second are of intrinsic nature and are determined by losses in the springs themselves and at the anchoring points. While increasing the vacuum level, the quality factor of a resonator increases, which reveals the decrease of total loss. But at some pressure level the quality factor saturates and doesn’t increase with the decrease of the pressure. In this case the intrinsic losses are dominating the extrinsic ones, and the latter can be neglected. Let us see how to represent losses on the equivalent electrical networks. In the lumped-parameter models of mechanical resonators their damping can be associated with a mobile node (were a mass can be present or not). It is represented by a damper connected to the mass and fixed at the other termination (fig. 5).

In the equivalent network this damper will be represented by a resistance in series with the inductor corresponding to the mass and will obviously belong to a proper branch of the mesh (since it is connected between a mobile node and a fixed

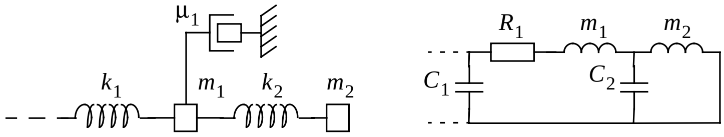


Figure 5. Air viscous losses modeling

point). Since real springs have a mass, in lumped mechanical representations of real mechanical systems there are always masses at the mobile nodes. Therefore in the electrical equivalent network there will always be an inductor in series with a resistor in all proper meshes. Consequently in the electrical domain we can represent the extrinsic losses due to the air damping as losses in inductors.

Let us now consider damping related to the intrinsic losses in springs and anchors. They are proportional (in a linear small-signal model) to the velocity of the spring deformation or to the difference between the velocities of the displacements of the ends of the spring. Consequently these losses can be considered in the lumped representation of mechanical resonating systems as dampers connected in parallel to a spring (fig. 6).

On the electrical level such a spring and its damper are represented by a capacitor and resistor in series. These two elements belong to the proper branch of two meshes, that correspond to the mobile nodes to which the spring is connected. In this way intrinsic losses of the spring are simulated by a resistor put in series with the capacitor modeling the spring.

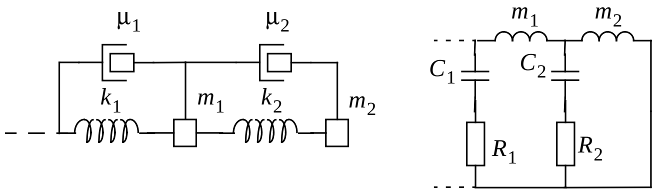


Figure 6. Internal losses modeling

While considering an oscillating system in deep vacuum, where the extrinsic damping is negligible, the losses are associated on the electrical level with the capacitances and are represented by resistors in series with capacitors. This is a considerable difference with respect to electrical filter networks, where losses are usually represented by resistances in parallel with capacitances and in series with inductors.

3. Application to the electro-mechanical filter design.

The main interest of the electro-mechanical transformation is the possibility of the re-use of traditional design methods of electrical passive filters by applying them to the mechanical filter design. The desired filter design flow is showed in the fig. 7.

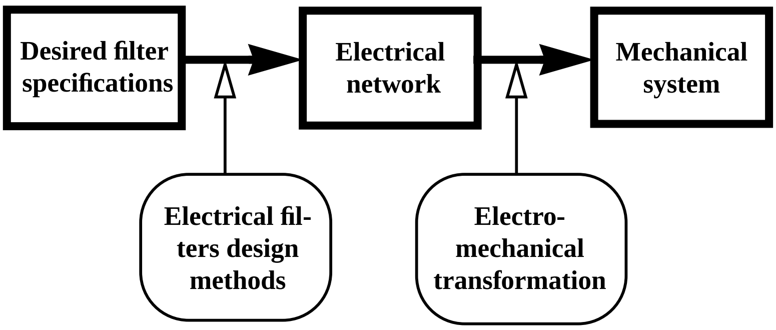


Figure 7. Desired design flow for electro-mechanical filters.

Using this approach all the filter analysis and synthesis can be made in the electrical domain, for which theory and design tools are well-developed. Also electronic designers are much more familiar with electrical passive networks rather than with complex mechanical systems, and the possibility to design mechanical filters by means of electrical networks is very attractive.

The main problem is that existing design methods of electrical passive filters result often in electrical networks that don't have the properties of one transformable into the mechanical domain described above. Synthesized networks may not have an equivalent mechanical system, and consequently the design isn't valid. Therefore the used electrical filter design methods must take into consideration the transformation existence criteria and result in networks satisfying the rules defined in the preceding section. Up to now no existing design method takes into account this issue. In the next sections we show how to design electro-mechanical filters from their electrical equivalent networks.

3.1 Ladder-network filters

We have chosen filters with ladder-network architecture as a basis of our study. Thanks to the inherently low sensitivity of the ladder structure, this class of circuits is well-studied in the literature, is widely used in passive filter design, and design methods are well-developed. We have explored the ladder filter architecture in order to determine if resulting networks are transformable to the mechanical domain, and how to use them to design electro-mechanical filters.

We have considered a general ladder passive network that is presented in fig. 8 a)

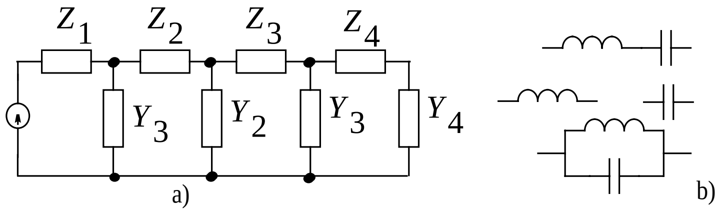


Figure 8. Passive ladder network. a) network architecture; b) Z and Y branches substitutes.

The elements Z and Y can be replaced by any of the branches shown in the fig. 8 b) or by their serial or parallel combination. To analyze the suitability of such ladder network for the conversion in the mechanical domain, let us define the meshes in the following way:

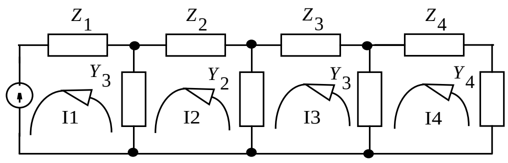


Figure 9. The choice of meshes.

Since the mechanical filter architecture should be derived from this network by the electro-mechanical transformation, the latter must belong to the transformable network sub-class defined above. The admittances Y belong to common branches of the meshes, so they can't contain inductors, but only capacitances and resistors. Impedances Z belong to proper branches, so they can contain serial RCL networks. They can also contain LC parallel circuits, under the condition that inductors aren't connected in the common branches. In the fig. 10 a) an example of a possible network architecture is shown, with appropriate meshes definitions. As we can see, no inductor is shared by two mesh currents, so the equivalent mechanical system exists, and is shown in the fig. 10 b).

The ladder shown in this example can be easily expanded or modified by adding series branches in a similar way. All corresponding parallel branches contain only capacitors.

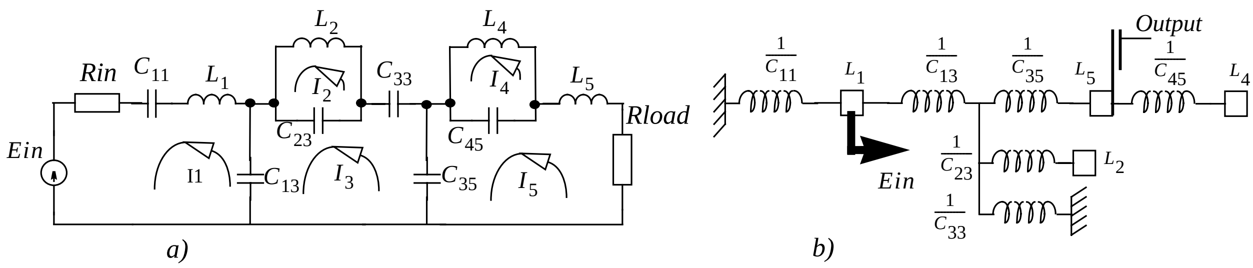


Figure 10. a) Example of a ladder network convertible to the mechanical domain; b) its equivalent mechanical system.

3.2 Filter architectures

The application that has motivated this study is the filter design for IF stages in wireless devices. For this application transmission zeros in the filter stopband are often required. A typical characteristic of a bandpass IF filter is shown in fig. 11. The poles are distributed in the passband, and transmission zeros are placed symmetrically aside from the center frequency in the stopband. To achieve such transmission characteristics, the existing design methods generate ladder networks that contain inductors in parallel branches, consequently these methods are not suitable for the design of mechanical filters. We propose an approach that makes the design of EM filter with such characteristics possible.

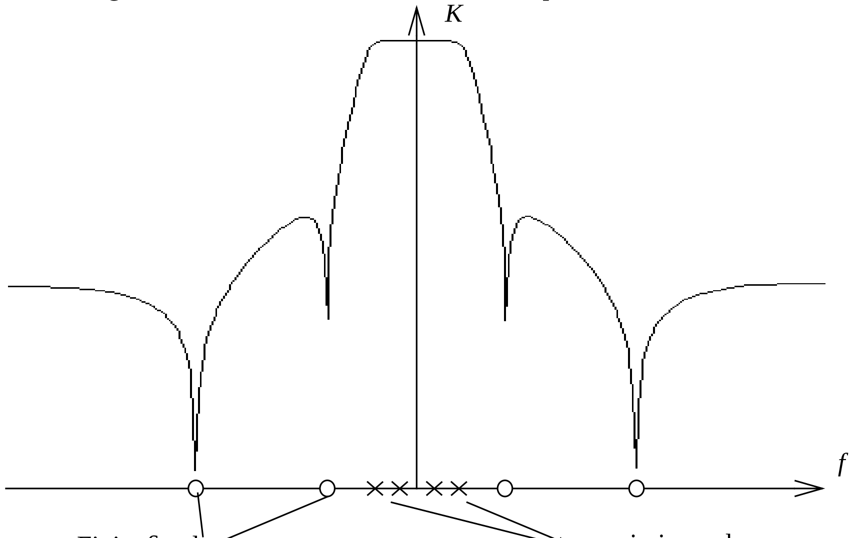


Figure 11. A typical IF filter characteristic with finite transmission zeros.

The existing coupled-resonator method achieves filters with similar characteristics as in the fig. 11, but without transmission zeros. The schematic of such a filter is presented in fig. 12. To achieve a proper design of coupled-resonator filters, all meshes must have the same serial-resonance frequency (mesh impedance zero frequency) equal to the center frequency of the filter. In this case, thanks to the coupling between meshes, poles are distributed on the $j\omega$ axis around the center frequency. The stronger the coupling, the farther the poles are away from the center frequency, so large-band characteristics can be achieved.

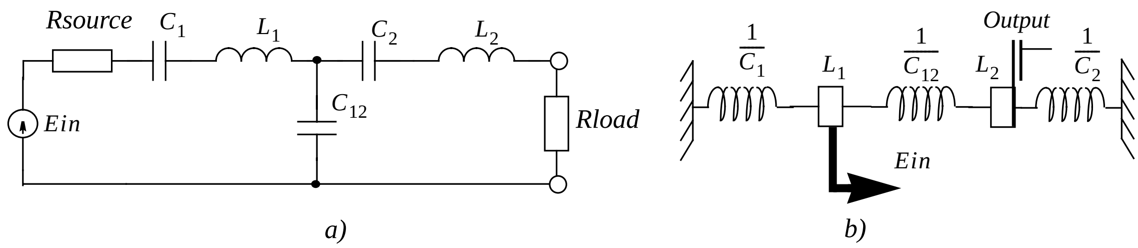


Figure 12. a) Coupled-resonator filter (solid line), filter with transmission zeros (dot-line); b) Mechanical schematic of the filter with the transmission zeros equivalent to the circuit showed in a).

To achieve a filter with transmission zeros, we have taken this filter as a basis. Effectively, we don't need to modify the pole distribution in the passband, but only introduce transmission zeros in the stopband. For this purpose we propose to replace the serial branches of the ladder network of fig. 12 by the networks shown in the fig. 13 a).

New serial LCL and LCC branches have both parallel and series resonances. Since the goal is only to add zeros without modifying poles distribution, the added elements must not modify neither the mesh series resonance frequency values, nor the coupling coefficients (that are responsible for the pole distribution near the center frequency). Therefore after adding the elements L_{20} and C_{10} (and so fixing transmission zeros), the series resonance frequency of meshes in the network of the

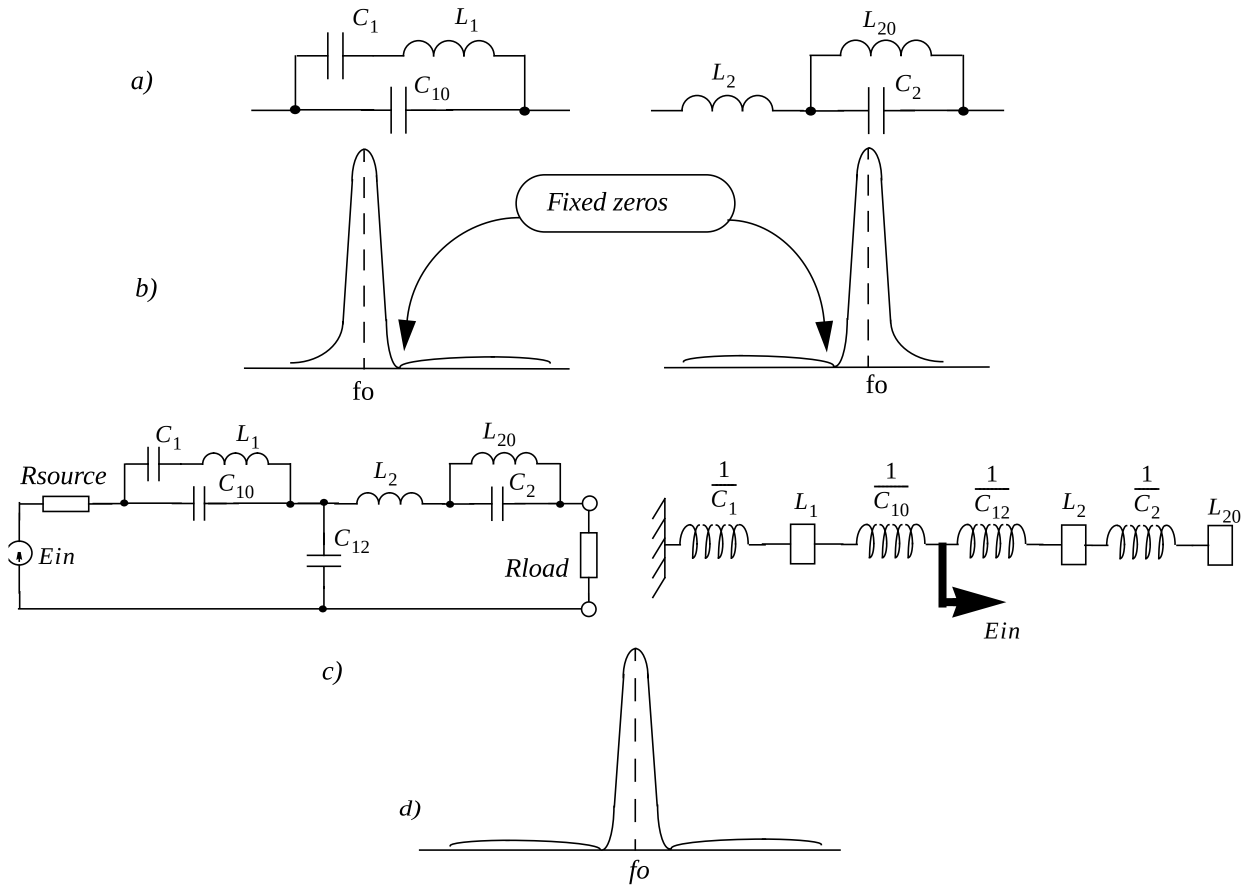


Figure 13. Architecture of a filter with transmissions zeros: a) Fixed transmission zeros implementing networks; b) Impedances of these networks; c) The electrical prototype of the modified coupled-resonator filter, that include the networks of a) in the serial branches and its equivalent mechanical system; d) the sketch of the transmission characteristic of the obtained filter

fig. 13 c) will not change. In this case pole positions in the passband also will not change, and finite transmission zero positions will be defined by the values of LCL and LCC branch elements.

The design flow of the overall filter is as follows. Firstly an all-pole filter with the given passband specification is synthesized. Then the finite zero positions are found from the stopband characteristics (with the help of existing analytical or numerical methods). Then the values of elements L_{20} and C_{10} are found and values of elements C_1, C_2, L_1, L_2 are modified in order to maintain the series resonance frequency and coupling coefficients unchanged.

3.3 Design and realization issues.

In order to test the exposed approach we have designed a filter with two resonators, in which transmission-zeros generating elements are included. The original electrical prototype of the filter is shown in fig. 13 c). To realize this prototype we need to design mechanical structures equivalent to resonating serial branches and to couple them (capacitor C_{12}). For clarity, resistances associated with the capacitors are not shown. These resistances will determine the quality factors of poles and zeros. Up to now, the coupling function was implemented by a mechanical spring [1,2]. This method has several disadvantages: increase of mechanical losses in the anchors with frequency, inability to tune the coupling coefficient, at high frequencies the masses of springs must be taking into account. We propose the electrostatical coupling of two resonators instead (fig. 14). The principle is the following: two resonators with input-output capacitive transducers are electrically connected together in a manner that the sensing electrode of one is connected to the actuating electrode of the other (node *A*). The node A is DC biased in the manner to be floating at the signal frequencies. This can be achieved by connecting the node A to the a DC source (or mass) via a high-value inductor or resistor (since there is no DC current through the node A, the last is also well suitable). While the resonator 1 is excited by the input source, the motional current is injected into the node A, and therefore a motional voltage is created, that excited the resonator 1. In the same way, oscillations of resonator 2 excite resonator 1,

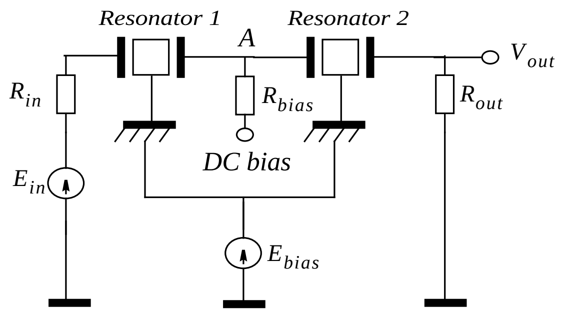


Figure 14. The principle of an electro-mechanical filter with electrostatical coupling.

and the two resonators are coupled. It can be shown that the electrostatic coupler acts on the mechanical resonators in the same way as a coupling spring.

The stiffness of this spring (i. e. coupling coefficient) is controlled by the DC bias applied to the net A. In this way filter tuning can be achieved.

We use this technique to design our fixed zero-transmission passband filter in the epi-poly silicon MEMS technology. The layout of the filter is shown on the fig. 17 and its electro-mechanical schematics are depicted on the fig. 15. Comparing the equivalent mechanical and electro-mechanical schematics, we can see that the coupling spring $\frac{1}{C_{12}}$ function is realized by an electrostatical coupler. Since in the equivalent mechanical structure (fig. 12, b) the elements L_2, $\frac{1}{C_2}$, L_{20} are maintained only by the spring $\frac{1}{C_{12}}$, and this spring is replaced by the electrostatical coupler in the electro-mechanical filter (fig. 15), these elements have to interact with other elements only by electrostatic coupling. In the same time, these element have to be fixed on the substrate. To achieve these two contradictory requirements, we introduce in the system a soft sustaining spring, that weakly affects the forces distribution while oscillating, but maintains the elements in a fixed position.

The parallel-plate transducers are used for actuating, sensing and electrical coupling. All transducers have the same dimensions, shown in the table 1.

The element values of the actual electromechanical prototype of the filter shown in the fig. 15 are presented in the table 1. These values are the same for both the mechanical and the equivalent electrical elements, if they are expressed in MKS units.

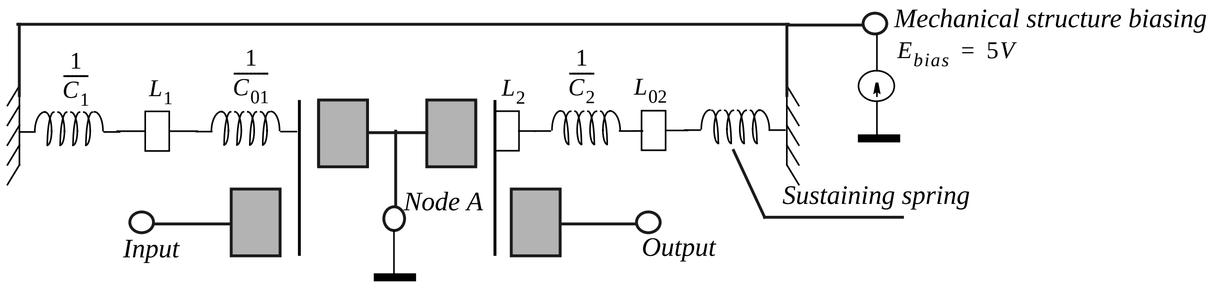


Figure 15. Electro-mechanical schematics of the designed filter.

The designed filter was simulated in the CADENCE environment. An AHDL model of the electrostatic coupler has been used to achieve mixed-domain system simulations. Input and output impedance of the filter actually depends on the bias voltage of the electromechanical transducers. Typical values for the input/output impedances are on the order of 100 kOhmd to 10 MOhms. In fig. 16 the simulated transmission characteristics of the filter are shown. This simulation has been done with open-circuit output and 50 Ohms resistance at the input. As we can see, two transmission zeros are generated outside of the passband. A theoretical work should be done in order to develop a proper and universal synthesis method for this type of filters, allowing the optimal approximation of given specifications. This is a subject of ongoing researches.

Table 1: Elements parameter values.

L_1, *H, kg*	$11.3 \cdot 10^{-12}$	C_1, *F*, $\frac{m}{N}$	$1.029 \cdot 10^{-3}$	Transducers gap d_0, *m*	$0.2 \cdot 10^{-6}$
L_2, H, kg	$86.31 \cdot 10^{-12}$	C_{10}, *F*, $\frac{m}{N}$	$7.345 \cdot 10^{-3}$	Transducer surface S, m^2	$225 \cdot 10^{-12}$
L_{20}, *H, kg*	$13 \cdot 10^{12}$	C_2, *F*, $\frac{m}{N}$	$1.029 \cdot 10^{-3}$	Sustaining spring stiffness, $\frac{N}{m}$	20

The filter has been realized in the thick-layer epipoly technology, with a stractural layer thiskkness 15 um. The oscillations are achieved in the lateral mode. The advantage of the used technology is the big area of the capacitive transducers, that allows to achieve relatively high transducer efficiency. The springs are implemented as flexural-mode beams. The actual layout of the filter is shown in fig. 17.

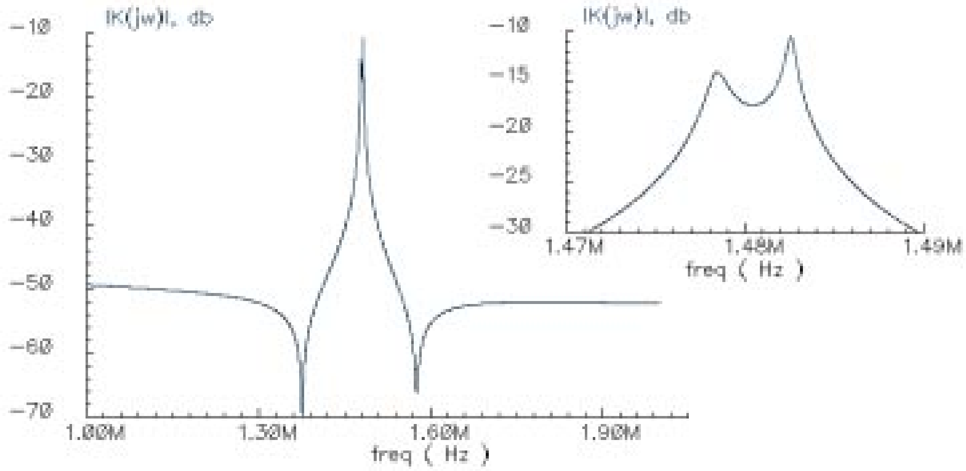


Figure 16. Simulated transmission characteristics of the designed filter.

In practice, a parasitic transmission pole is generates at a low frequency with respect to the useful passband. This phenomenon can be explained by analyzing the equivalent electrical circuit of the filter: on the low frequencies inductors L_{02} and L_{22} become series connected with the high-value coupling capacitor C_{12}, and a low-frequency resonance appears. To reduce this undesirable phenomenon we propose to add some selective stages to the filter, with a strong attenuation on the parasitic resonance frequency. Another alternative is to design a compensation filter stage that introduce a transmission zero on the this frequency. This issue is a subject of ongoing research.

3.4 Problems related to the realizations.

Let us consider some practical realization issues that are common to all MEMS filtering devices, but that are especially important in the case of complex-architecture filters.

The main problem we anticipate for the realization of the described filter is the mismatch of zero and pole positions due to fabrication tolerances. Theses errors become especially crucial for highly selective filters. The needed accuracy of poles and zeros positioning grows with the quality factor of the filter: poles should be placed inside the passband, so the distance between them is at least Q times smaller that the center frequency. The problem is similar with transmission zeros. In the work [6] a filter with 3 coupled resonators has been realized for frequencies of several hundred kilohertz. In order to compensate for the fabrication tolerances of single resonators, additional tuning electrodes have been used for each resonator. This approach is not possible for filters at higher frequencies, with more complex resonance modes and with more elements influencing the pole distribution. A stronger control of manufacturing process is required to minimise the fabrication tolerances and so the distortion of the filter characteristics.

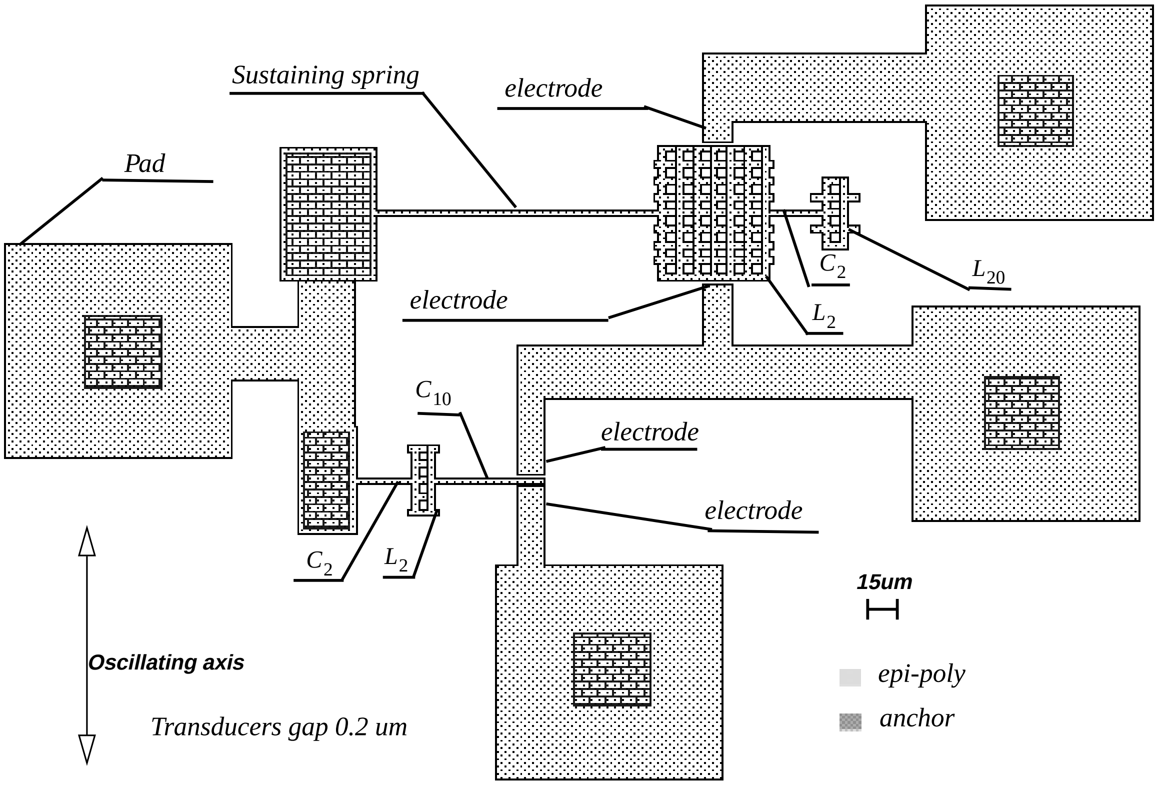


Figure 17. Layout of the designed electro-mechanical filter.

The second problem we have to emphasize is proper to all capacitive transducer. This is the parasitic capacitances intrinsicly present between signal electrodes themselves and between signal electrodes and the ground (in the most cases the resonator itself is DC biased and connected to the ground for AC signals, so a capacitance exists between electrodes and the mobile element). The biggest problem of the electrostatic coupler is the capacitance parallel to ground (fig. 18). This capacitance includes not only the electrode-to-resonator capacitance, but also the capacitances of the connection wires, pads etc. Since the impedance of the point A is high, a part of the the motional current goes into the capacitance Co. This phenomenon reduces the coupling efficiency. A possible solution to this problem is the reduction of connection capacitances by careful design. The ultimate solution is the co-integration of electronic and MEMS devices, eliminating the need for the pads, thus supressing the associated capacitances completely.

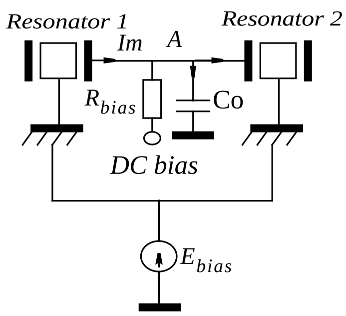


Figure 18. Parasitic capacitance Co of the electrostatic coupler

4.Conclusions.

In this paper we have shown the results of our work concerning the design of electro-mechanical filters for mobile telecommunication applications. This work has introduced an approach that allows to design architectures of electro-mechanical filters from electrical prototype networks. For this purpose the transformation from electrical to mechanical domain has been described. The most important result is that equivalent mechanical systems exist only for a sub-class of electrical equivalent networks defined in this study. Therefore an appropriate approach to electro-mechanical filter design is required.

These results have been applied to the design of an electro-mechanical filter with finite-transmission zeros. This design uses electrostatical coupling, that is a new method we have proposed for use in electro-mechanical filters. The electrostatical coupling approach allows to implement complex electro-mechanical systems by replacing some mechanical springs by electrically tunable electrostatic interaction.

Making use of these results, a bandpass filter with transmission zeros in the stopband has been designed in the epipoly silicon technology. The numerical parameters and the layout of this design are presented in this article.

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