Active-Tunable Inductor Effected by a Novel Impedance Synthesis Circuit

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Abstract: This paper describes a novel impedance synthesis (IS) circuit: the impedance of an inductor is made tunable as a function of the frequency by means of a transconductance function $g_m(s)$ that is located in the feedback loop. In order to show the potential of this novel IS circuit, the effect of several basic transconductance functions onto the synthesized impedance is presented. The specific case of $g_m(s) \sim s^{-1}$ brings up the 'coil enhancement' feature of this IS circuit: the synthesized impedance is purely inductive with an actual boosted inductance level compared to the inductance of the physical inductor. Based upon this 'coil enhancement' property, the IS circuit is incorporated in new 'plain old telephone service' (POTS) low-pass filter which is fully compliant to the 'TS 101 952-1-1 V1.2.1' standard[1] of the European Telecommunications Standards Institute (ETSI).

Key–Words: inductor, transformers, passive filter, active filter, feedback

1 Introduction

Passive electronic filters have been used for decades and are still preferable over other filter topologies in the lower frequency range. These cost-effective passive filters offer several advantages such as guaranteed stability, zero power consumption and good large signal endurance. However, for very demanding applications (low passband ripple and high stopband attenuation with a small pass- to stopband window) higher order and complex passive filters are required which are notoriously area-consuming.

The opposite evolution, namely area-reduction, is found in the continuous miniaturisation of passive components. However, not all passive elements follow the same scaling rate: package sizes of resistors and capacitors have dropped drastically over the last two decades, whereas package sizes of inductors (and transformers) seem to have reached a lower limit and stagnate. This difference in size between the inductors and the other passive elements turns the inductors into the area-qualifying factor in passive filter design.

It is, hence, not surprising that several techniques exist for reducing or circumventing the use of inductors. For the very small inductance values, one may consider a monolithic integration of the inductor[2]. But for the majority of inductance values, the widespread method of actively generating the inductor behaviour is applicable. Popular circuits that are selected for such inductor emulation are based upon CMOS active inductors[3],[4], or gyrator implementations[5],[6] or rely on current conveyor combinations[7],[8]. However, filters that require a certain robustness against overvoltages (e.g. 1.5kV), interchanging all inductors in a filter with their emulation circuit is not possible. The circuit implementations that can handle such voltages will not be area-saving (and cost-efficient).

In this paper, a new interesting area-saving option is elaborated: associating extra functionality with the inductors. Instead of the linearly rising impedance versus frequency characteristic of inductors, one may benefit from a more complex frequency domain behaviour. This adaptable inductor approach can reduce the amount of filter stages at the cost of a few extra components. In the following, a circuit for generating such an adaptable inductor is discussed in detail and it's employment in a POTS low-pass filter is presented.

2 Concept

A single inductor is electrically represented by its inductance L. In the Laplace domain, the inductor's impedance is formulated as:

$$\frac{v_L}{i_L} = sL \tag{1}$$

where s embodies the complex frequency. The inductor's reactance $X_L = \omega \cdot L = 2\pi \cdot f \cdot L$ is proportional to the signal frequency f and the inductance L. Hence, once a specific value for L is chosen, the inductor's impedance for all frequencies is predetermined.

In order to obtain more flexibility in impedance control, the circuit from fig. 1 is developed. This so-called IS circuit comprises a pair of coupled inductors L_1-L_2 and a transconductance amplifier. The first inductor L_1 is placed directly in-between the input terminal node t_1 and the output terminal node t_2. The second inductor L_2 is wound onto the same core as the first inductor L_1 and forms the transformer combination L_1-L_2. One terminal of L_2 is connected to ground, whereas the other end of L_2 is attached to the input of a transconductance amplifier. The output of the transconductance amplifier is fed back to the second inductor L_2. The combination of the second inductor L_2 with the transconductance amplifier constitutes a feedback circuit.

The electrical behaviour of the IS circuit of fig. 1 relies on the expressions of the coupled inductors L_1-L_2 (in the s-domain):

$$\begin{aligned} v_{L_1} &= (sL_{11}) \cdot i_{L_1} + (sM_{12}) \cdot i_{L_2} \\ v_{L_2} &= (sL_{22}) \cdot i_{L_2} + (sM_{12}) \cdot i_{L_1} \end{aligned} \tag{2}$$

where L_{11} and L_{22} are the self-inductances of the inductors L_1 and L_2 respectively and where M_{12} describes the mutual inductance between L_1 and L_2. The voltage v_{L_2} over L_2 is sensed by the transconductance amplifier and converted into a current i_{L_2} through some transconductance function $g_m(s)$. This configuration is mathematically expressed as:

$$i_{L_2} = g_m(s) \cdot v_{L_2} \tag{3}$$

Note that no current flows through the high-ohmic inputs of the transconductance amplifier. Combining the equations (2) and (3), further adopting a lossless coupling between L_1 and L_2 ($M_{12}=\sqrt{L_{11}\, L_{22}}$), leads to the following expression for the synthesized impedance Z_{syn} between the electrical terminals t_1-t_2:

$$Z_{syn} = \frac{v_{L_1}}{i_{L_1}} = sL_{11} \cdot \frac{1}{1 - g_m(s) \cdot sL_{22}} \tag{4}$$

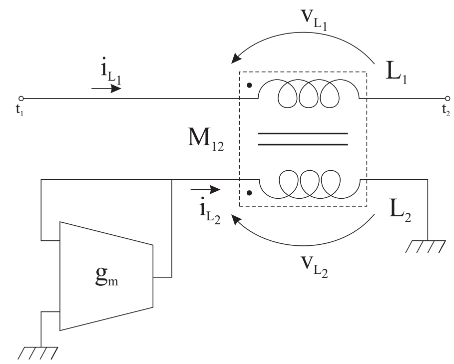

Figure 1: The impedance synthesis (IS) circuit: coupled inductors with v-sensing and current driving on secondary

This equation (4) reports that the synthesized impedance of the IS circuit of fig. 1 equals a product of two factors. The first factor matches the impedance of a single inductor L_1 without feedback circuit and is to be considered as a reference impedance (Z_{ref}). The second factor $(1 - g_m(s) \cdot sL_{22})^{-1}$ in (4) originates from the feedback circuit and hands out an instrument for altering the synthesized impedance between the output terminals t_1-t_2 from the reference impedance level.

The feedback inside the IS circuit draws extra attention to the stability analysis. Whenever some filter is designed with the IS circuit, the standard stability criteria are to be met onto the open loop gain (OLG):

$$OLG = g_m(s) \cdot \left(\frac{n_2}{n_1}\right)^2 \cdot (sL_{11} \,//\, Z_{ext}) \tag{5}$$

in which // symbolizes a parallel connection, n_1 and n_2 the number of windings on the transformer of the coils L_1 and L_2 respectively, and where Z_{ext} represents all the external impedances seen at the output terminals t_1-t_2.

3 Impedance synthesis

In contrast with the single inductor reference $Z_{ref} = sL_{11}$, the IS circuit of fig. 1 introduces multiple degrees of freedom for shaping the frequency behaviour, allowing higher order filters based on a single inductor. Depending on the transconductance function $g_m(s)$, the frequency behaviour of the synthesized impedance Z_{syn} is anything but the frequency behaviour of the reference. Table 1 presents several elementary examples of possible transconductance functions and the corresponding synthesized impedances. The transconductance function parameters k, z_1 and

ISSN: 1790-5117

ISBN: 978-960-474-076-5

p_1 are supposed to be real arbitrary constants. And keeping the stability of the complete filter in mind, the parameter p_1 must be furthermore restricted to exclusively positive real values.

An extra tuning option for the synthesized impedance arises when the transconductance function itself becomes steerable through external inputs: each set of inputs defines a different transconductance function. This extra measure of freedom increases the variety of realisable impedances of the IS circuit even more and can be utilized for fine-tuning purposes or for integration of two or more filters in one compact filter set-up.

4 Coil enhancement [9]

One of the major advantages of the IS circuit is the ability of realizing purely inductive impedances between the output terminals t_1-t_2 which have amplitude levels above the reference $|sL_{11}|$. The terms for an increased inductive Z_{syn} follow from:

$$\begin{cases} |1 - g_m(s) \cdot sL_{22}| < 1 \\ \measuredangle\,(1 - g_m(s) \cdot sL_{22}) = 0^\circ \end{cases} \tag{6}$$

The first condition $|1 - g_m(s) \cdot sL_{22}| < 1$ by itself determines transconductance functions that lead to impedances Z_{syn} with amplitudes larger than $|sL_{11}|$, but the realized Z_{syn} is not necessarily inductive any more. The situations at which the Z_{syn} is not only increased in amplitude but also purely inductive have to meet both requirements of (6).

One case that is able to satisfy the 'enhanced coil' conditions is the pure integrator transconductance function $g_m(s) = k \cdot s^{-1}$ in which k is a real constant (see also table 1). Such a transconductance function is similar to placing a (positive or negative) inductive load in parallel with Z_{ref}. The synthesized impedance Z_{syn} is equivalent to Z_{ref} apart from a frequency independent scaling factor $(1 - kL_{22})^{-1}$. Four different k-zones can be identified:

$$\begin{array}{rcll} k < 0 & \overset{\forall s}{\Longrightarrow} & |Z_{syn}| < |Z_{ref}| & \measuredangle Z_{syn} = 90^\circ \\ 0 < k < \frac{1}{L_{22}} & \overset{\forall s}{\Longrightarrow} & |Z_{syn}| > |Z_{ref}| & \measuredangle Z_{syn} = 90^\circ \\ \frac{1}{L_{22}} < k < \frac{2}{L_{22}} & \overset{\forall s}{\Longrightarrow} & |Z_{syn}| > |Z_{ref}| & \measuredangle Z_{syn} = -90^\circ \\ k > \frac{2}{L_{22}} & \overset{\forall s}{\Longrightarrow} & |Z_{syn}| < |Z_{ref}| & \measuredangle Z_{syn} = -90^\circ \end{array}$$

Whereas generally spoken the product $g_m(s) \cdot sL_{22}$ is some arbitrary vector in the complex plane that shifts with changing frequencies, this product becomes purely real and constant for all frequencies. The specification $|1 - g_m(s) \cdot sL_{22}| < 1$ can therefore be translated into a limiting range for the parameter

g_m function	Z_{syn}	$Z_{syn}\big\|_{s\ll}$	$Z_{syn}\big\|_{s\gg}$
k	$\frac{sL_{11}}{1-k\,L_{22}\cdot s}$	sL_{11}	$-\frac{L_{11}}{k\,L_{22}}$
$k \cdot s$	$\frac{sL_{11}}{1-k\,L_{22}\cdot s^2}$	sL_{11}	$-\frac{L_{11}}{k\,L_{22}\cdot s}$
$k \cdot (s + z_1)$	$\frac{sL_{11}}{1-k\,L_{22}\cdot s\cdot(s+z_1)}$	sL_{11}	$-\frac{L_{11}}{k\,L_{22}\cdot s}$
$k \cdot s^2$	$\frac{sL_{11}}{1-k\,L_{22}\cdot s^3}$	sL_{11}	$-\frac{L_{11}}{k\,L_{22}\cdot s^2}$
$\frac{k}{s}$	$\frac{sL_{11}}{1-k\,L_{22}}$	$\frac{sL_{11}}{1-k\,L_{22}}$	$\frac{sL_{11}}{1-k\,L_{22}}$
$\frac{k}{s+p_1}$	$\frac{sL_{11}\cdot(s+p_1)}{p_1+s\cdot(1-k\,L_{22})}$	sL_{11}	$\frac{sL_{11}}{1-k\,L_{22}}$
$\frac{k}{s^2}$	$\frac{s^2\,L_{11}}{s-kL_{22}}$	$-\frac{s^2\,L_{11}}{k\,L_{22}}$	sL_{11}
$\frac{k\cdot(s+z_1)}{s}$	$\frac{sL_{11}}{1-k\,L_{22}\cdot(s+z_1)}$	$\frac{sL_{11}}{1-k\,z_1\,L_{22}}$	$-\frac{L_{11}}{k\,L_{22}}$
$\frac{k\cdot s}{(s+p_1)}$	$\frac{sL_{11}\cdot(s+p_1)}{p_1+s-k\,L_{22}\cdot s^2}$	sL_{11}	$-\frac{L_{11}}{k\,L_{22}}$

Table 1: Examples of the synthesized impedance Z_{syn} corresponding with some basic transconductance functions

k, namely $0 < k < 2 \cdot L_{22}^{-1}$. The phase requirement $\measuredangle\,(1 - g_m(s) \cdot sL_{22}) = 0^\circ$ also produces a bounding scope for k, namely $k < L_{22}^{-1}$. The intersection of both k ranges leads to $0 < k < L_{22}^{-1}$ and consequently defines the limiting k-range for obtaining 'coil enhancement'.

Another, more realistic, situation that results in 'coil enhancement' occurs when a multiple pole-zero transconductance function $g_m(s)$ behaves proportional to s^{-1} in a frequency range $[f_1,f_2]$. This is the case when each pole or zero in $g_m(s)$, represented by the factor $(s + c)$ in the transfer function, can be approximated as either s or c (with c is an arbitrary complex number) for frequencies in-between $[f_1,f_2]$. In other words, the roll over frequencies of the zeros and poles are located far away of the frequency range $[f_1,f_2]$. One example is presented below in order to demonstrate such an approximation operation on more complicated g_m function.

The example starts from a third order transconductance function 7 with three real zeros and three real negative poles. This indicates that z_1, z_2, z_3 are real constants, that p_1, p_2 and p_3 are real positive con-

ISSN: 1790-5117

ISBN: 978-960-474-076-5

stants and that K is a real frequency independent factor. Boundary conditions that are further presupposed for frequencies in-between $[f_1,f_2]$ are $s \gg z_1$, $s \ll z_2$, $s \ll z_3$, $s \gg p_1$, $s \gg p_2$ and $s \ll p_3$.

$$\begin{aligned} g_m(s) =& K\frac{(s+z_1)\cdot(s+z_2)\cdot(s+z_3)}{(s+p_1)\cdot(s+p_2)\cdot(s+p_3)} \\ \cong& K\frac{s\cdot z_2\cdot z_3}{s\cdot s\cdot p_3} \quad \text{for } s \in [2\pi i\, f_1, 2\pi i\, f_2] \end{aligned} \tag{7}$$

If the net effect of the simplification for the frequency range $[f_1,f_2]$ yields a g_m function of the form $k \cdot s^{-1}$ and if $0 < k < L_{22}^{-1}$, then a 'coil enhancement' situation occurs. Concretised into the example (7), 'coil enhancement' arises if the constant factor $K \cdot z_2 \cdot z_3 \cdot p_3^{-1}$ is positive and less than L_{22}^{-1}.

5 Improvement

The major advantage of the 'coil enhancement' circuit of fig. 1 is already described in previous section: the value of the inductor L_1 can be adapted, by altering the gain of the feedback circuit. Another benefit that can be attributed to the 'coil enhancement' circuit is a good performance for common mode rejection: common mode signals from the feedback circuit will not affect the main circuit branch t_1-t_2 (and visa versa) due to the presence of the common mode blocking L_1-L_2 transformer. Next to that, the circuit is by nature protected against large overvoltages and overcurrents (e.g. nearby lightning strikes): only a fraction of the large overvoltages/overcurrents will arrive at L_1-L_2 transformer's secondary side as the feedback circuit becomes isolated from the transformer's primary side in saturated conditions. The small overvoltages that eventually appear at the secondary transformer side can be dealt with by a simple voltage-clamping protection.

One of the major issues that blocks the circuit of fig. 1 from being ready-to-use, is the influence of the series resistance R_{s2} of inductor L_2. This winding resistance R_{s2}, which is not shown on the circuit schematic, creates an additional voltage $i_{L_2}\cdot R_{s2}$ that also contributes to the voltage v_{L_2}. This gives an error on the generated current i_{L_2} at the output of the transconductance amplifier. The exact contribution of this series resistance R_{s2} to the synthesized impedance comes to the surface from the extended expression of Z_{syn}:

$$Z_{syn} = \frac{v_{L_1}}{i_{L_1}} = sL_{11}\cdot\frac{1}{1-\frac{g_m(s)\cdot sL_{22}}{1-g_m(s)\cdot R_{s2}}} \tag{8}$$

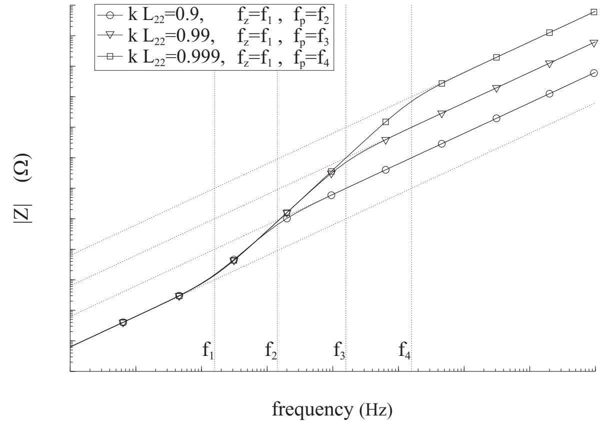

Figure 2: Influence of the series resistance R_{s2} on the synthesized impedance Z_{syn} for $k\,L_{22} = 0.9$, $k\,L_{22} = 0.99$ and $k\,L_{22} = 0.999$

Note that R_{s2} is not exactly known as it is a parasitic of the transformer L_1-L_2 and is hence difficult to compensate for.

A more comprehensive interpretation of the influence of the series resistance R_{s2} of inductor L_2 on Z_{syn} is obtained when constricting equation (8) to the 'coil enhancement' situation. The substitution of $g_m = k \cdot s^{-1}$ yields the following formula:

$$\begin{aligned} Z_{syn} &= \frac{v_{L_1}}{i_{L_1}} = sL_{11}\cdot\frac{1}{1-k\,L_{22}}\cdot\frac{s-2\pi\,f_Z}{s-2\pi\,f_P} \\ \text{with} \quad f_Z &= \frac{k\,R_{S2}}{2\pi} \\ f_P &= \frac{k\,R_{S2}}{2\pi\cdot(1-k\,L_{22})} \end{aligned} \tag{9}$$

Note that without taking the series resistance R_{s2} of inductor L_2 into consideration the amplification factor is calculated to be $(1 - k\,L_{22})^{-1}$ for all frequencies (cf. table 1). In case the series resistance R_{s2} of inductor L_2 is not disregarded, equation (9) states that this amplification factor $(1-k\,L_{22})^{-1}$ is obtained only for frequencies above f_P. Figure 2 shows the frequency plot according to equation (9) for three different $k\,L_{22}$-values: 0.9, 0.99 and 0.999.

At the expense of an extra winding onto the same transformer the detrimental influence of the series resistance R_{s2} can be eliminated: the improved IS circuit is shown in fig. 3. This improved IS circuit comprises a triple of coupled inductors L_1-L_2-L_3 and a transconductance amplifier. The first inductor L_1 is again placed directly in-between the input terminal node t_1 and the output terminal node t_2. The second inductor L_2 and third inductor L_3 are wound onto the same core as the first inductor L_1 and form the trans-

ISSN: 1790-5117

ISBN: 978-960-474-076-5

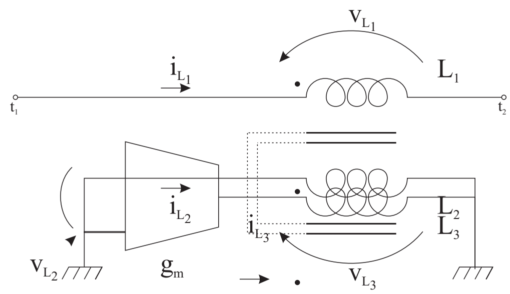

Figure 3: Improved IS circuit: coupled inductors with v-sensing on secondary and current driving on tertiary

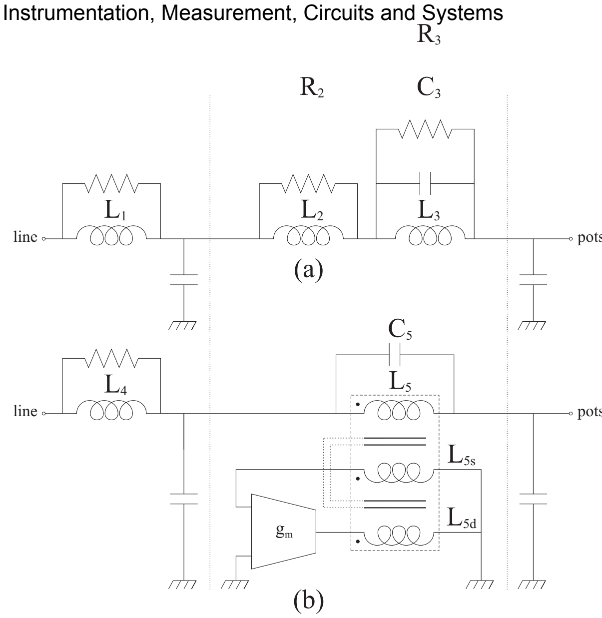

Figure 4: POTS low-pass filter circuit topologies: (a) the passive version, (b) the 'active' version

former combination L_1-L_2-L_3. One terminal of L_2 is connected to ground, whereas the other terminal of L_2 is attached to the input of a transconductance amplifier. Finally, one terminal of the third inductor L_3 is attached to the output of the transconductance amplifier, whereas the other terminal of L_3 is connected to ground. In the improved IS circuit, the feedback circuit is implemented by the combination of the second inductor L_2, the third inductor L_3 and the transconductance amplifier.

The extended IS circuit of fig. 3 fulfills the same functionality as the basic version of fig. 1 but the voltage sense and current drive action are now separated:

$$i_{L_3} = g_m(s) \cdot v_{L_2} \tag{10}$$

Note that $i_{L_2} \cong 0$ due to the high-ohmic input of the g_m-amplifier and, hence, the effect of the series resistance R_{s2} is ruled out. The synthesized impedance Z_{syn} realized by the improved IS circuit is expressed as:

$$\begin{aligned} Z_{syn} = \frac{v_{L_1}}{i_{L_1}} &= sL_{11} \cdot \frac{1}{1 - g_m(s) \cdot s\sqrt{L_{22} \cdot L_{33}}} \\ &= sL_{11} \cdot \frac{1}{1 - \frac{n_3}{n_2} \cdot g_m(s) \cdot sL_{22}} \\ &= sL_{11} \cdot \frac{1}{1 - \frac{n_3 \cdot n_2}{n_1^2} \cdot g_m(s) \cdot sL_{11}} \end{aligned} \tag{11}$$

in which n_1, n_2 and n_3 are the number of windings of inductors L_1, L_2 and L_3 on the transformer core respectively that comply to $L_{11} = (n_1/n_2)^2 \cdot L_{22}$, $L_{11} = (n_1/n_3)^2 \cdot L_{33}$ and $L_{22} = (n_2/n_3)^2 \cdot L_{33}$.

6 Application

The single ended circuit topology of a typical, fully passive, POTS low-pass filter is shown in fig. 4(a) and is connected to the telephone exchange by means of the 'pots' terminal and is attached to the line (=twisted pair) through the 'line' node. The filter consists of three stages and three inductors: L_1, L_2 and L_3. Figure 4(b), however, presents the circuit topology of a new 'active' POTS low-pass filter that is based upon the improved IS circuit of fig. 3. This 'active' filter version consists of only two stages and one inductor L_4 and one transformer L_5-L_{5s}-L_{5d}.

The dashed lines on the fig. 4 mark out which impedances in both filters correspond: the series connection of $L_2//R_2$ with $L_3//C_3//R_3$ in the passive POTS filter is implemented by the $L_5 - L_{5s} - L_{5d} - g_m(s)//C_5$-combination in the 'active' POTS filter. And as L_3=L_5, the feedback loop of the IS circuit has to realise (in a certain frequency window) an inductive value of $L_2 + L_3$ through 'coil enhancement'.

Figure 5 shows the top and bottom side of a PCB with four ETSI 'TS 101 952-1-1 V1.2.1'-compliant 'active' POTS low-pass filters. The top side of the panel is reserved for the passive elements of the forward loop (inductors/transformers, resistors and capacitors). The bottom side of the board, on the other hand, is filled with the feedback loop elements that generate the required $g_m(s)$ and some voltage-clamping protections. This passive-to-active switch-

ISSN: 1790-5117

ISBN: 978-960-474-076-5

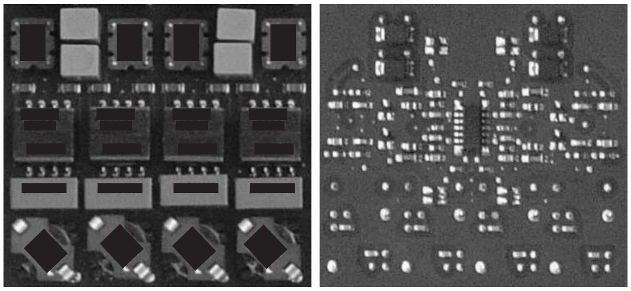

Figure 5: Top (left) and bottom (right) view of 4 active POTS low-pass filters

over in filter topology caused an area profit of 40%.

7 Conclusions

In this paper, a novel IS circuit is elaborated that creates a dirigible synthesized impedance through a transconductance function $g_m(s)$ in the feedback loop. An overview of several transconductance functions and their resulting synthesized impedances is presented. One very interesting characteristic of this IS circuit is its ability for realizing 'coil enhancement': the original inductor is boosted. The conditions for reaching 'coil enhancement' are derived and discussed.

The practical use of the IS circuit is blocked by the detrimental impact of the series resistance of the second inductor. A work-around for this issue is given through the use of a third coupled inductor.

The IS circuit, in 'coil enhancement' mode, is incorporated in a newly developed 'active' POTS low-pass filter. This 'active' POTS filter originates from a fully passive version in which two of the inductors are merged together into one 'active' inductor. The 'active' POTS filters are compliant to the ETSI 'TS 101 952-1-1 V1.2.1' standard. The passive-to-active conversion leads to a significant 40% area reduction.

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ISSN: 1790-5117

ISBN: 978-960-474-076-5