Chapter 1

Microwave Active Filter Design

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Abstract

A simplified method for the project and design of microwave active filters is presented here. The presented design is based on the use of an active inductor that emulates an inductor behavior by implementing a passive variable phase- and amplitude-compensating network and amplifiers, forming a gyrator-C architecture. This method can be applied with success for the design of bandpass filters with very high performances in terms of integration and application from a few hundreds of MHz to tens of GHs with filter high dynamic range and frequency tuning capability.

Keywords: filters, active filter, active inductor, high dynamic range, gyrator

1. Introduction

Nowadays advancements in communications systems, in conjunction with the great increasing request for miniaturization and size reduction of wireless systems and devices, have led to an increase of performances and space reduction of modern communications systems and devices. Several RF building blocks have consequently been successfully implemented as integrated circuits (ICs) with different technologies. In this scenario, analog tunable filters working, in general, at RF and microwave frequencies still remain the most difficult part to be integrated in a single chip [1–18] due to requirements in input and output impedance matching, stability, and dynamic range. In fact, most commercial wireless receivers only use off-chip filters that are typically implemented with discrete components. This scenario leads to the occupation of a large area at the front-end. Even though the integrated version of filters has been presented in several papers, it must be noticed that in monolithic integrated microwave circuits (MMIC) most of the semiconductor area is occupied by passive elements, while active devices play a marginal role in area occupation. It is in fact generally known that passive inductors are largely responsible for the area occupation in IC. Classical spiral inductors, in fact, require large amounts of substrate area or air bridges and present a very limited bandwidth, a general high series resistance, and crosstalk problems. For this reason, it can really be important to replace spiral inductors with active circuits behaving like inductors, but characterized by a very reduced semiconductor area occupation. It is well known that an inductor can be simulated by a gyrator loaded by a capacitance; in the literature and in the commercial market, a lot of solutions have been proposed that use the gyrator principle to simulate an inductive behavior at microwave frequency range [15, 16, 18–32]. Analog filters using Active Inductors (AIs) are good candidates for high-frequency operation; therefore, considerable interest has been shown in their use in active filters. In this perspective, in particular, band-pass filters designed by means of active inductor, usually suffer from a very low dynamic range due to the combination of a high noise level introduced by the AI and the relatively low compression power level of the AI itself. For all these reasons, there are only a few commercial solutions available on the market based on active inductors. Moreover, complex arrangements with many transistors tend to degrade the stability, preventing their use in practical applications. All these considerations show how the availability of simple, stable AIs with relatively high-power handling is therefore a critical issue, for their use as a replacement for spiral inductors in integrated filters especially when high-order filters with tunable characteristics are required. In the following paragraph, we will see how filters using active inductors have good potential to operate at high frequency, with high dynamic range, maintaining a constant quality factor (Q) in tunable applications when coupled with varactors diode, so considerable amount of interest has been shown in their use. A very simple approach for the design of tunable, high-quality factor, stable active inductors by means of an active inductor is presented here.

2. The active inductor design

Generally, active filters suffer from a very low dynamic range (see **Figure 1**) [33–53] due to the combination of a high noise level introduced by the AI and the relatively low compression power level of the AI itself. An interesting way to design active filters, in order to overcome this problem, is through the use of AI. The first high-Q AIs were reported over two decades ago, and the use of these inductors for the implementation of inductor-capacitor (L-C) resonator-type active filters was presented both in several papers and conferences.

However, as already mentioned, high-frequency active bandpass filters designed by means of AI usually suffer from a low dynamic range due to the combination of high noise level introduced by the AI and of low compression point of the AI itself, which limits the maximum handled power. Several kinds of AI have been proposed in the past [1–52], especially grounded inductors that are usually based on gyrator scheme loaded with a capacitor (**Figure 2**). With reference to **Figure 2**, the input voltage through the inverting amplifier (that can be based on a common-source stage) drives the loaded capacitor, producing a 90° delayed voltage; this in turn drives the output current through a noninverting transconductance, typically a source follower.

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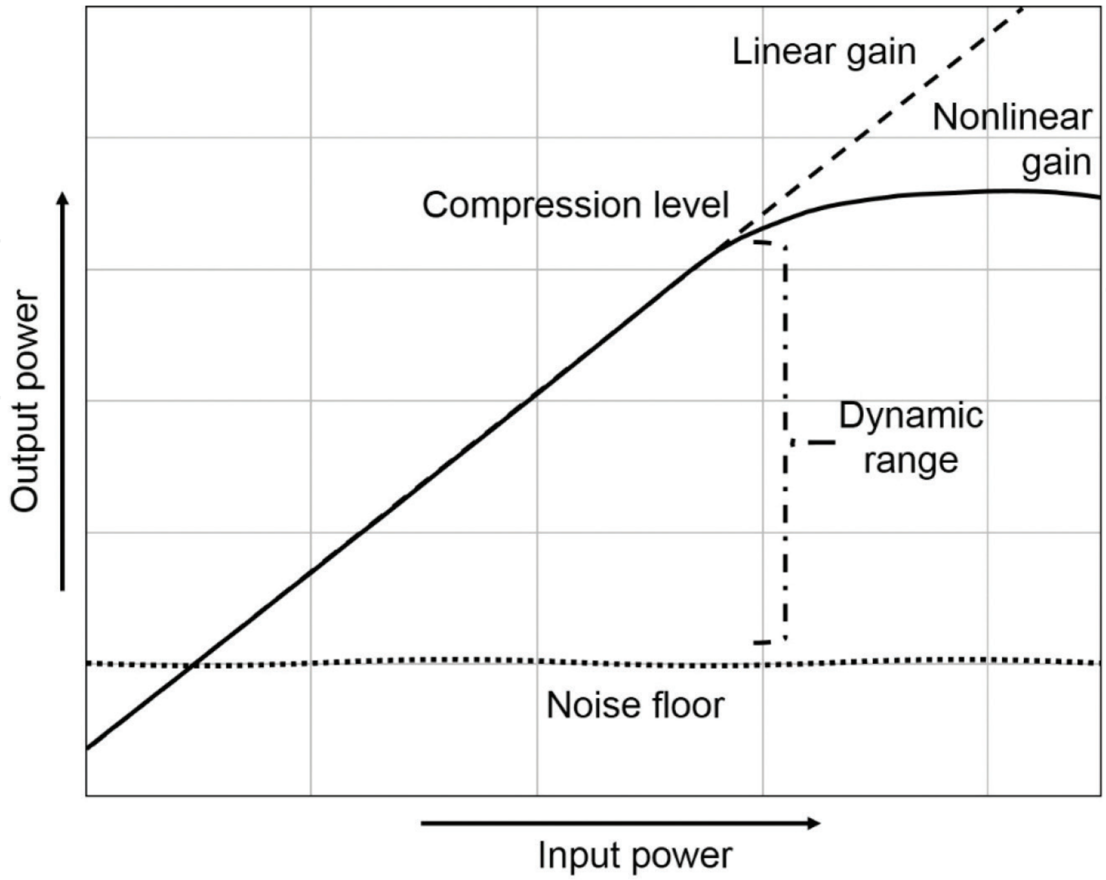

Figure 1. Dynamic range for an active filter.

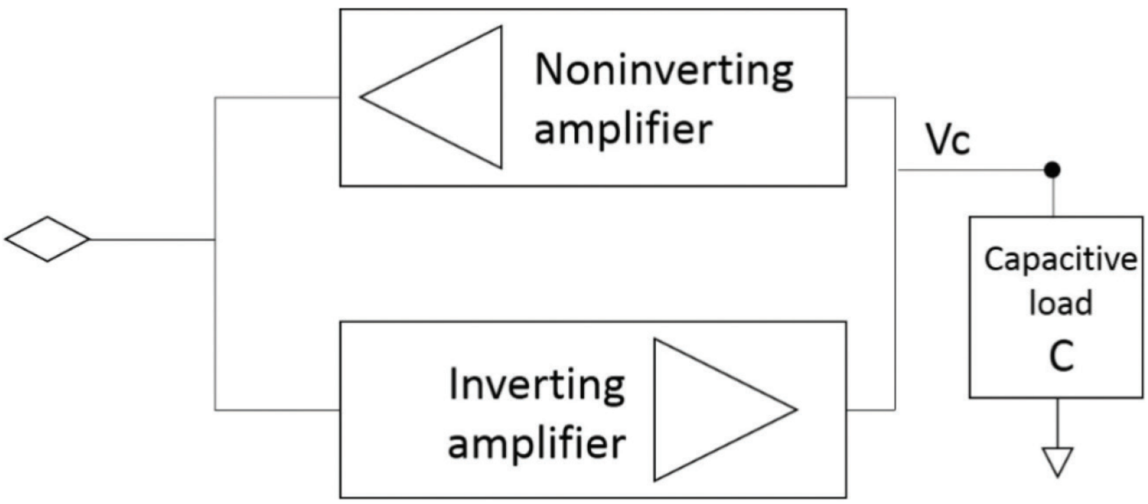

Figure 2. A gyrator-based active inductor scheme.

From **Figure 2** circuit, if we assume ideal amplifiers, the following equations can be derived [tcas]:

$$V_C = -\frac{g_{m1}}{j\omega C} \cdot V_{in} \tag{1}$$

$$I_{in} = -g_{m2} \cdot V_C = \frac{g_{m1} \cdot g_{m2}}{j\omega C} \cdot V_{in} = \frac{1}{j\omega L_{eq}} \cdot V_{in} \tag{2}$$

Finally, the equivalent inductance of the gyrator-based active inductor can be expressed as:

$$L_{eq} = \frac{C}{g_{m1} g_{m2}} \tag{3}$$

Eq. (3), of course, is valid only if the two amplifiers are ideal and zero phase delays and zero losses are present in the architecture. It is important to notice that if nonidealities in the amplifiers are considered, two unwanted effects appear: (1) a parasitic resistance (positive or negative) arises in series to the equivalent inductance; (2) the bandwidth of the inductive impedance becomes finite. **Figure 3a** shows that a phase relation between input less than 90° gives positive resistance, while a phase relation in excess of 90° causes a negative resistance: the AI can now be considered as in **Figure 3b**

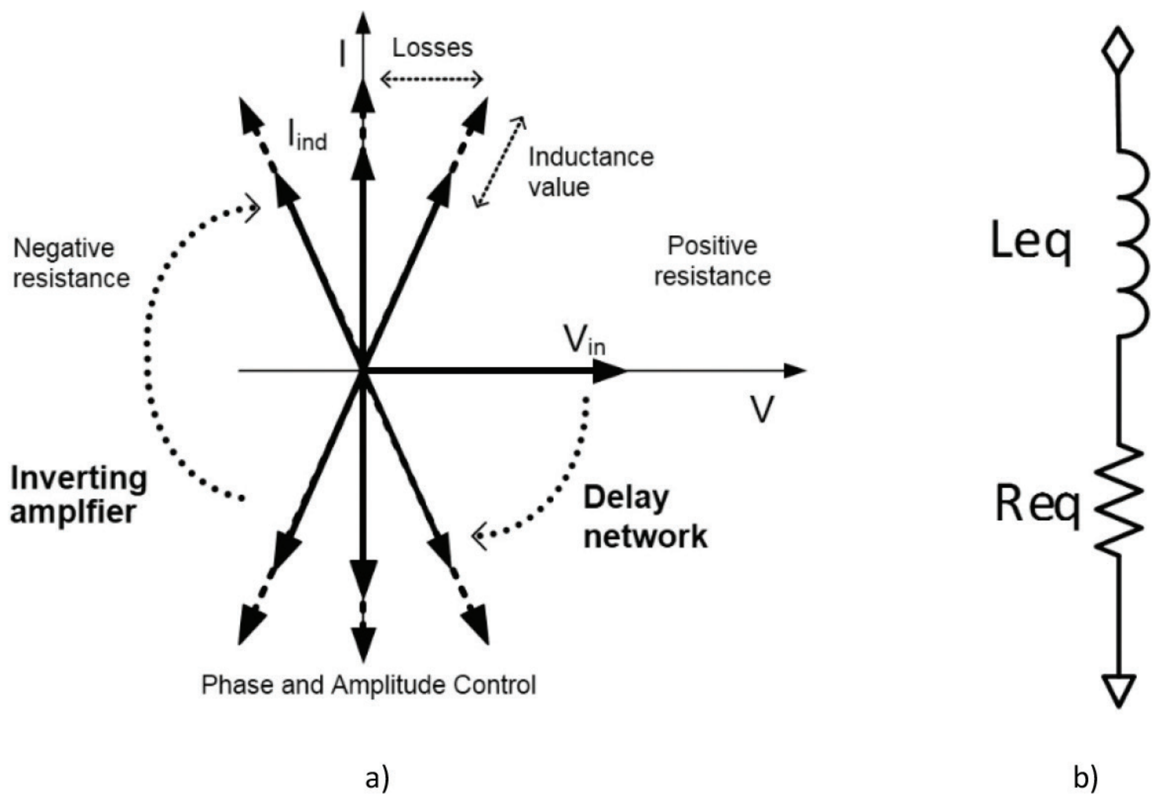

Figure 3. (a) AI voltage and current relations and (b) AI with series resistance.

If we were able to control both these phase relations, without acting on the bias condition of the amplifiers, the linearity of the active component of the gyrator will not be affected. This condition can be reached by the insertion of a suitable variable passive compensation network [35] (**Figure 4**).

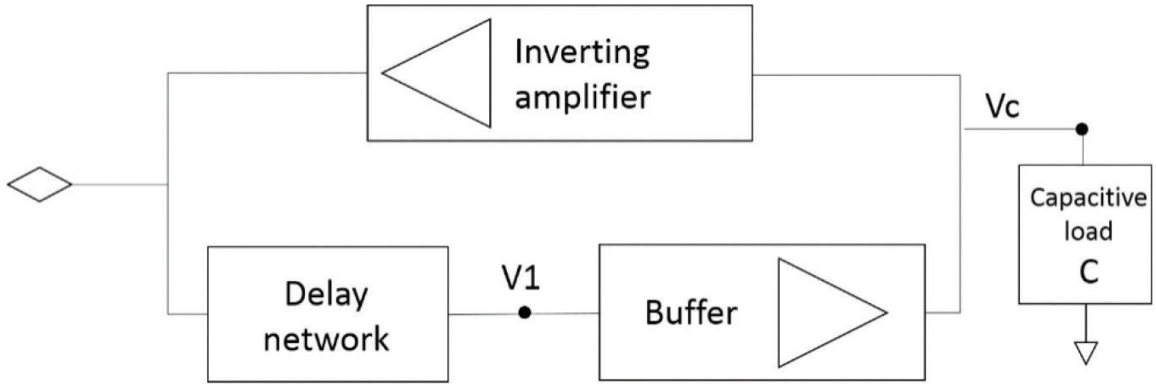

Figure 4. Block scheme of the active inductor with high dynamic range.

It has already been demonstrated in reference [35] that

$$R_{eq} = \frac{1}{g_m A_{\mathrm{var}}}\cos(\varphi_{\mathrm{var}}) + \frac{\omega R_{par} C_{tot}}{g_m A_{\mathrm{var}}}\sin(\varphi_{\mathrm{var}})$$
$$L_{eq} = -\frac{1}{g_m A_{\mathrm{var}}}\sin(\varphi_{\mathrm{var}}) + \frac{\omega R_{par} C_{tot}}{g_m A_{\mathrm{var}}}\cos(\varphi_{\mathrm{var}}) \tag{4}$$

From Eq. (4), it can be noticed that the phase of the compensating network must be such that the two addends cancel in the expression of the equivalent resistance, in the band of interest. This leads to a complex scenario. In order to simplify the design, a single active block AI and a compensation network can be also adopted as shown in **Figure 5**. In **Figure 6**, an equivalent circuit of the single active block AI [38] is shown. In this case, a simplified model of the capacitance gyrator with a single active block is shown.

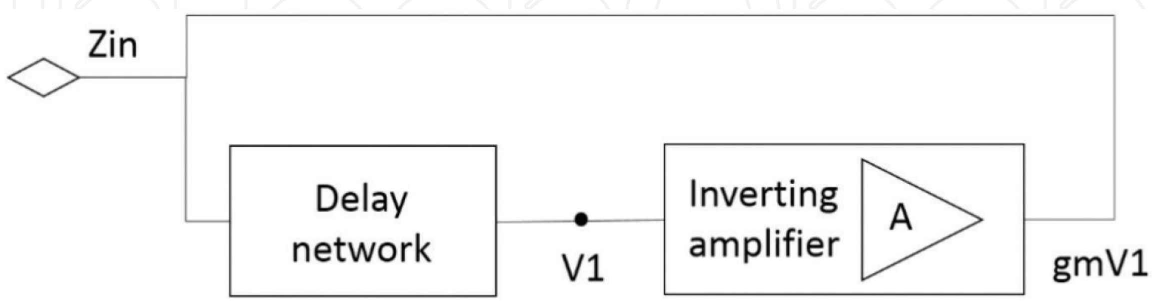

Figure 5. Block scheme of the proposed single active block active inductor.

An exact set of values of the resistor-capacitor (R-C) network gives a correct phase delay between input and output (voltage and current), minimizing the real part of the equivalent impedance. In the ideal case [38], the circuit input impedance is given by the following relation:

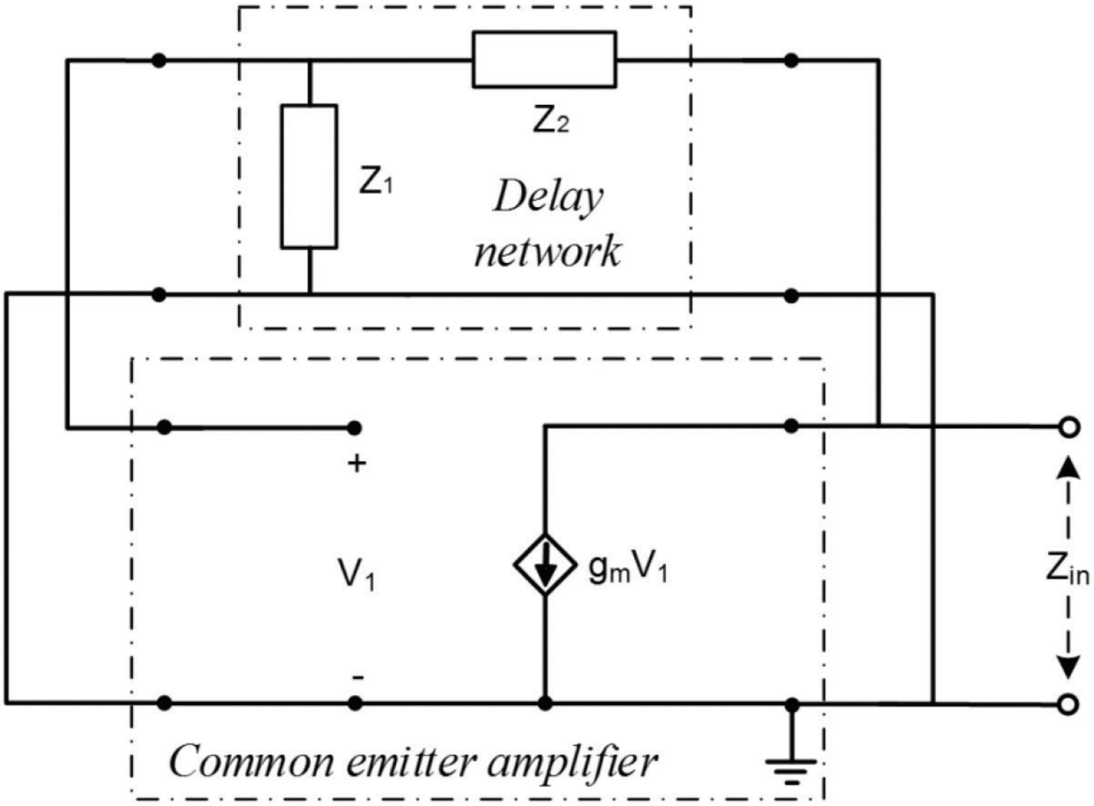

Figure 6. Simplified schematic model of the single-transistor AI.

$$Z_{in} = \frac{1 + j\omega RC}{g_m + j\omega C} \tag{5}$$

Eq. (5) can be also written in a more simple way as real and imaginary parts as in the following:

$$\mathrm{Re}\{Z_{in}\} = \frac{g_m + \omega^2 RC^2}{g_m^2 + \omega^2 C^2} \tag{6}$$

$$\mathrm{Im}\{Z_{in}\} = \frac{\omega g_m RC - \omega C}{g_m^2 + \omega^2 C^2} \tag{7}$$

Now it can be seen that the input impedance of the network is not always inductive [38]. By analyzing the imaginary part, we notice that the imaginary part of the AI impedance has a zero in $\omega = 0$, and it is always positive only if:

$$g_m R > 1 \tag{8}$$

The last is a very important condition if we want to obtain an inductive behavior. By considering also the amplifier block parasitic, we obtain the real and imaginary part of the input impedance of the single transistor (common emitter stage as the transconductance amplifier) gyrator as shown in **Figures 7** and **8**.

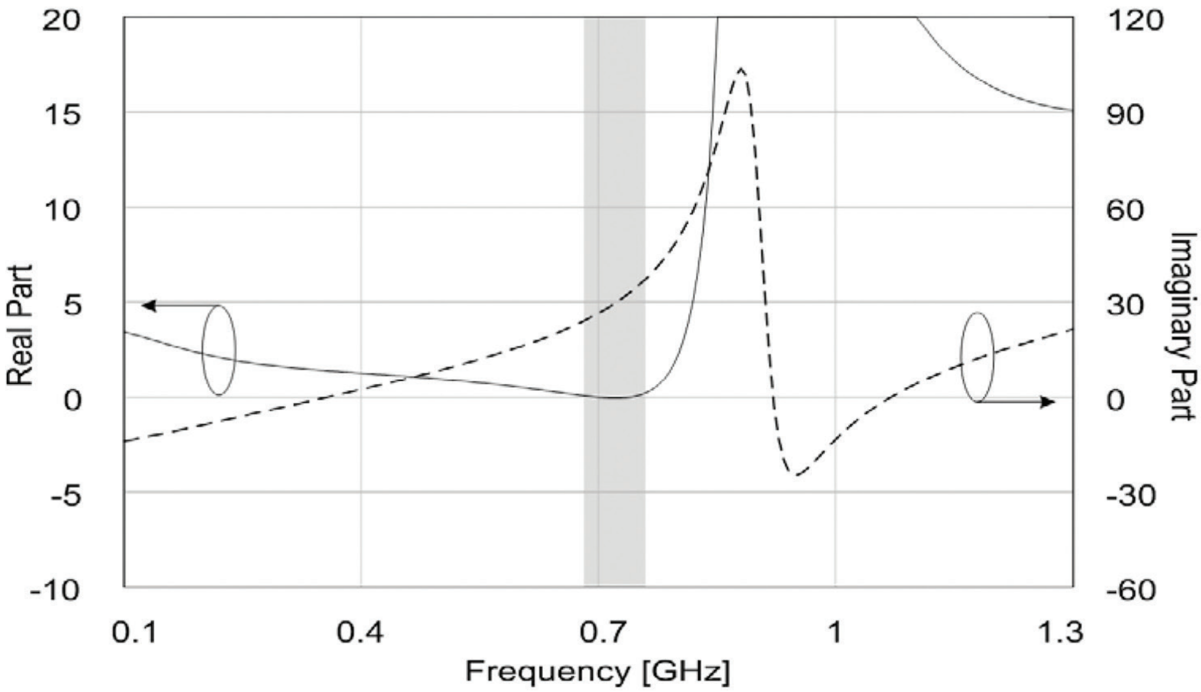

Figure 7. Real and imaginary part of the input impedance of the single transistor gyrator.

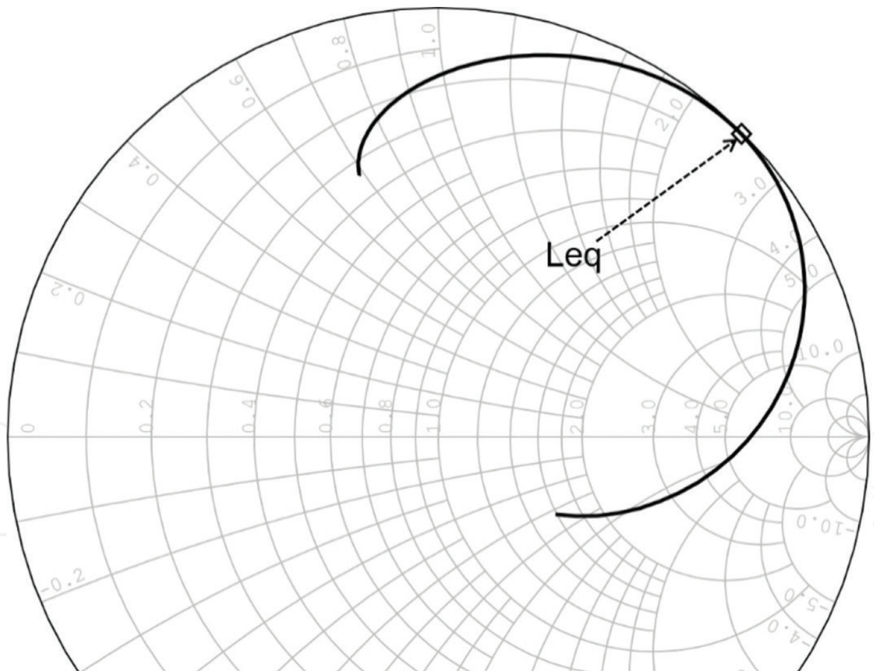

Figure 8. Complex input impedance of the AI.

3. Tunable microwave active filter example application

Thanks to the aforementioned approach, it is now possible to design an AI using, for example, medium-power bipolar transistor. The topology of the compensating network can be designed by means of an optimizer numerical software and the exact values of the network elements can be found by optimization, in order to take into account microstrip discontinuities and lines [38]. The ideal AI application is, of course, in monolithic integrated circuits, but as an example, a preliminary discrete design can be made on microstrip with discrete components. First, the AI design has to be performed, and consequently, a first-order (or greater) passband filter can be optimized (see **Figure 9**). As mentioned, the internal AI transconductance amplifier can be implemented by a bipolar transistor (e.g., the BFP420 bipolar junction transistor by Infineon in common-emitter configuration). **Figure 10** shows a simulated standalone AI real and imaginary part, while in **Figure 11**, the Smith chart complex input impedance is shown. In this case, the topology of the compensating network, and its components value, has been chosen in order to have an inductive behavior with minimum nonnegative series resistance around 2500 MHz. The standalone AI Q is shown in **Figure 12**. The implemented filter response in terms of S-parameters is shown in **Figure 13** showing the design success of microwave active filters by means of AI.

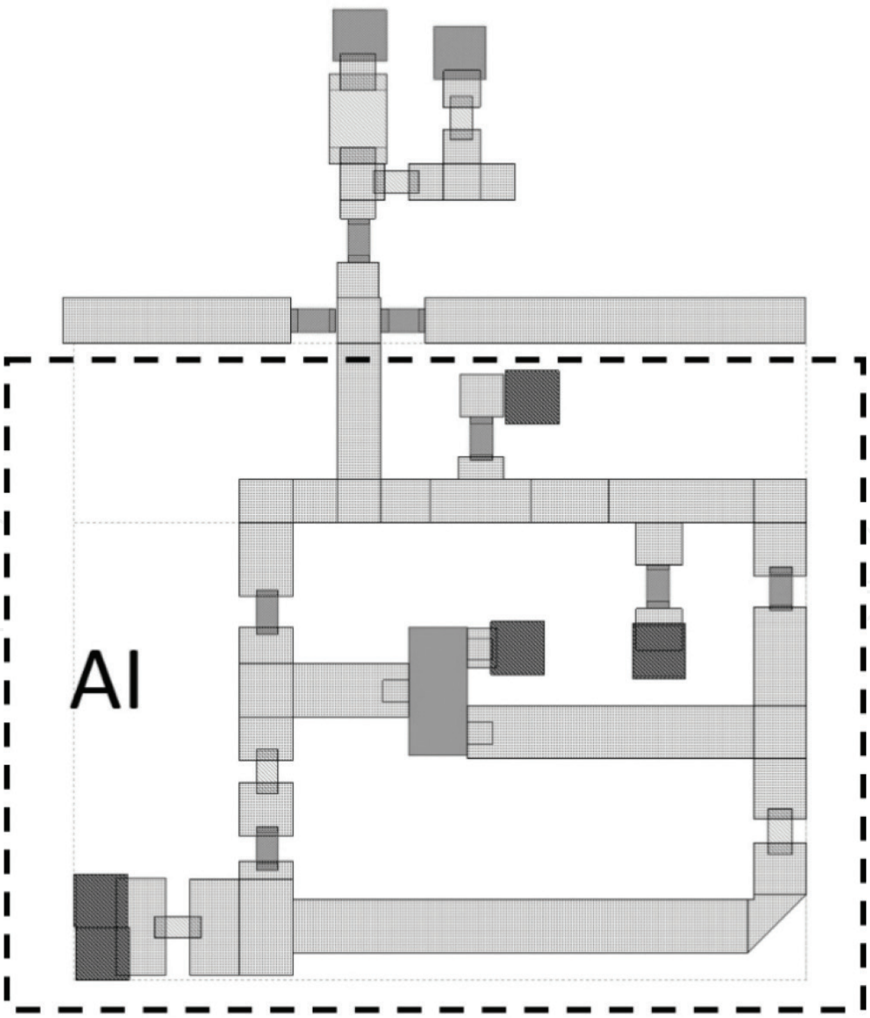

Figure 9. AI-based LC filter.

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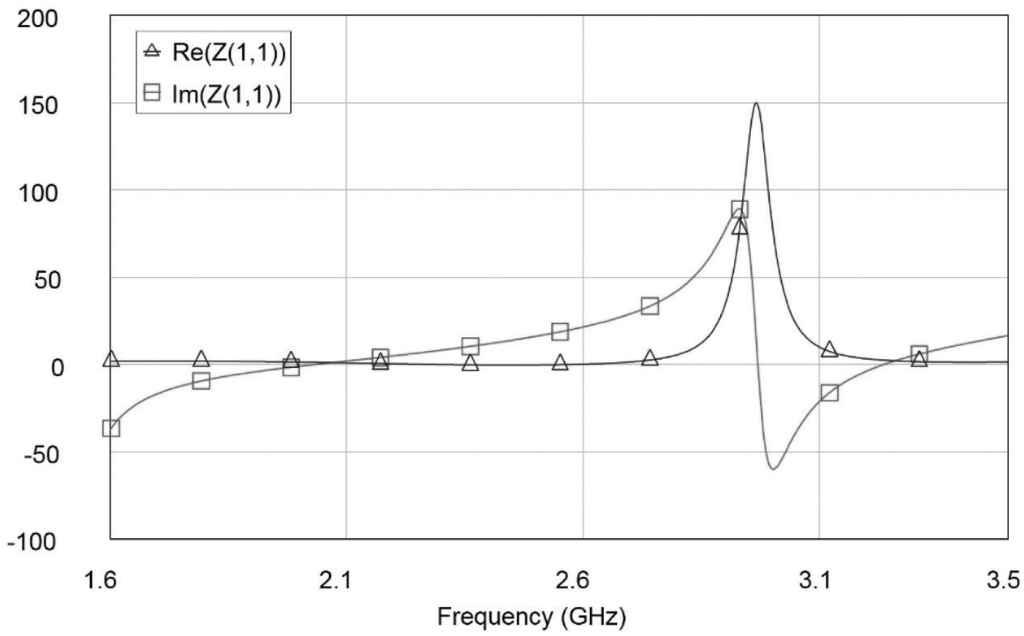

Figure 10. AI real and imaginary part.

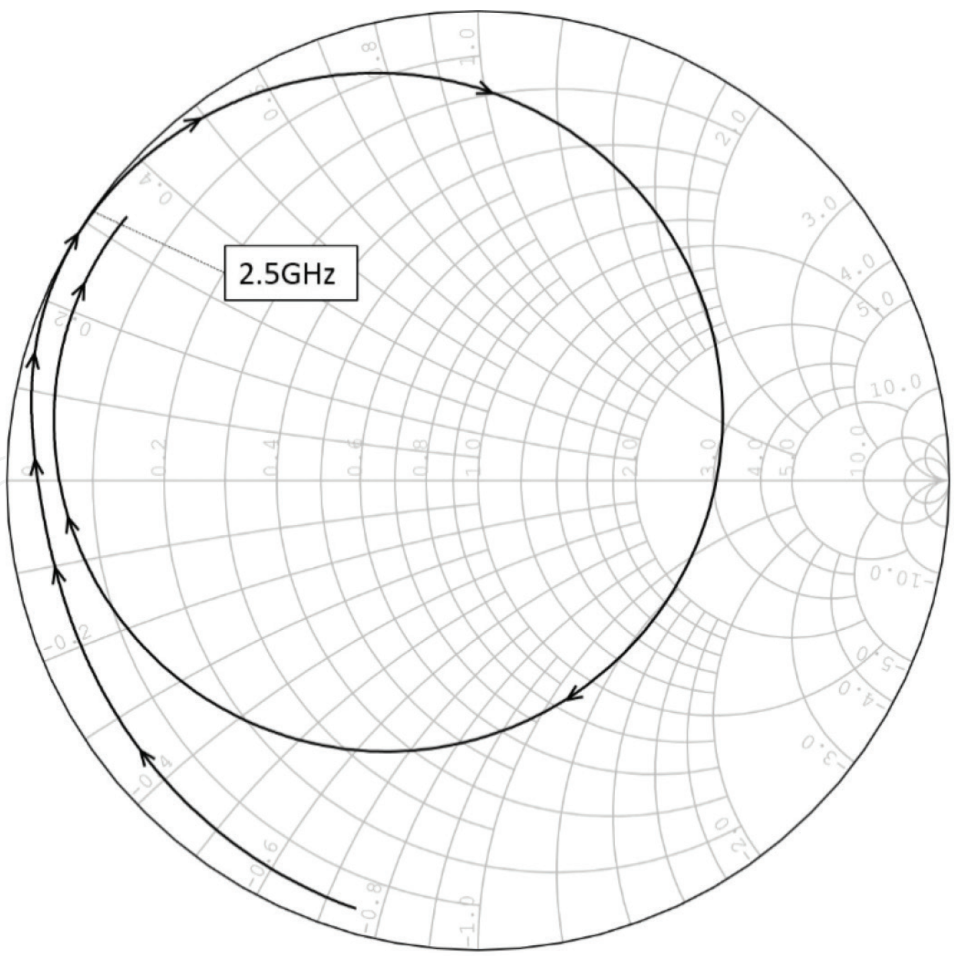

Figure 11. Complex input impedance of the active inductor.

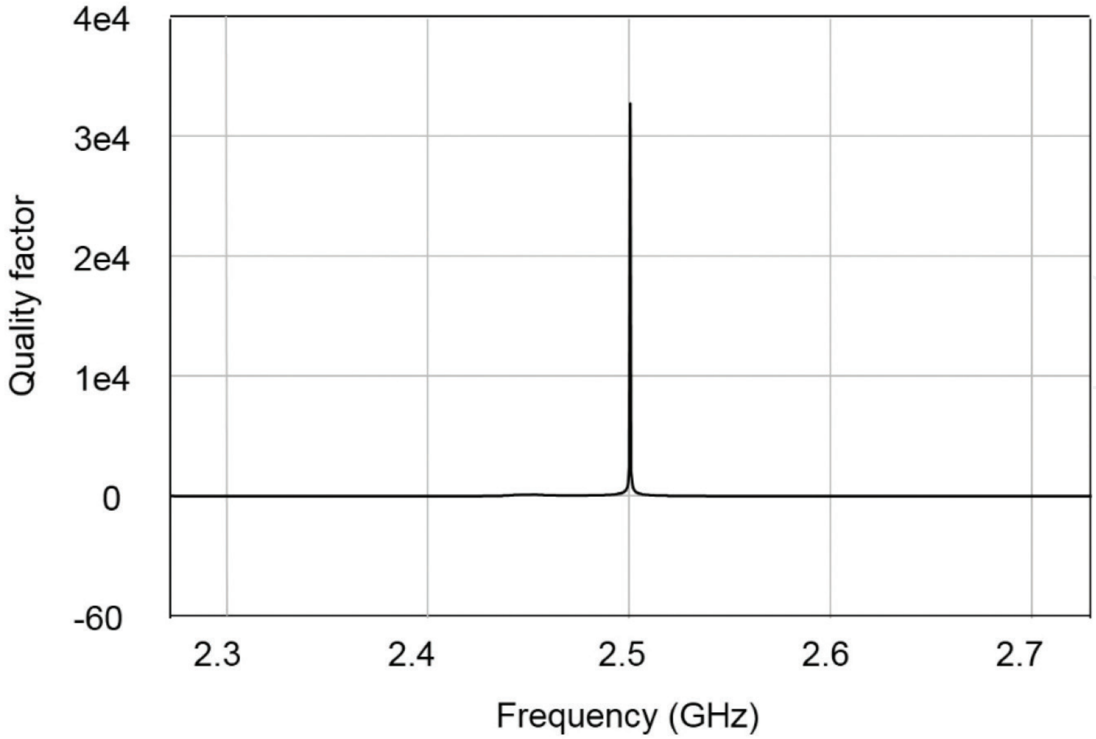

Figure 12. Quality factor of the active inductor.

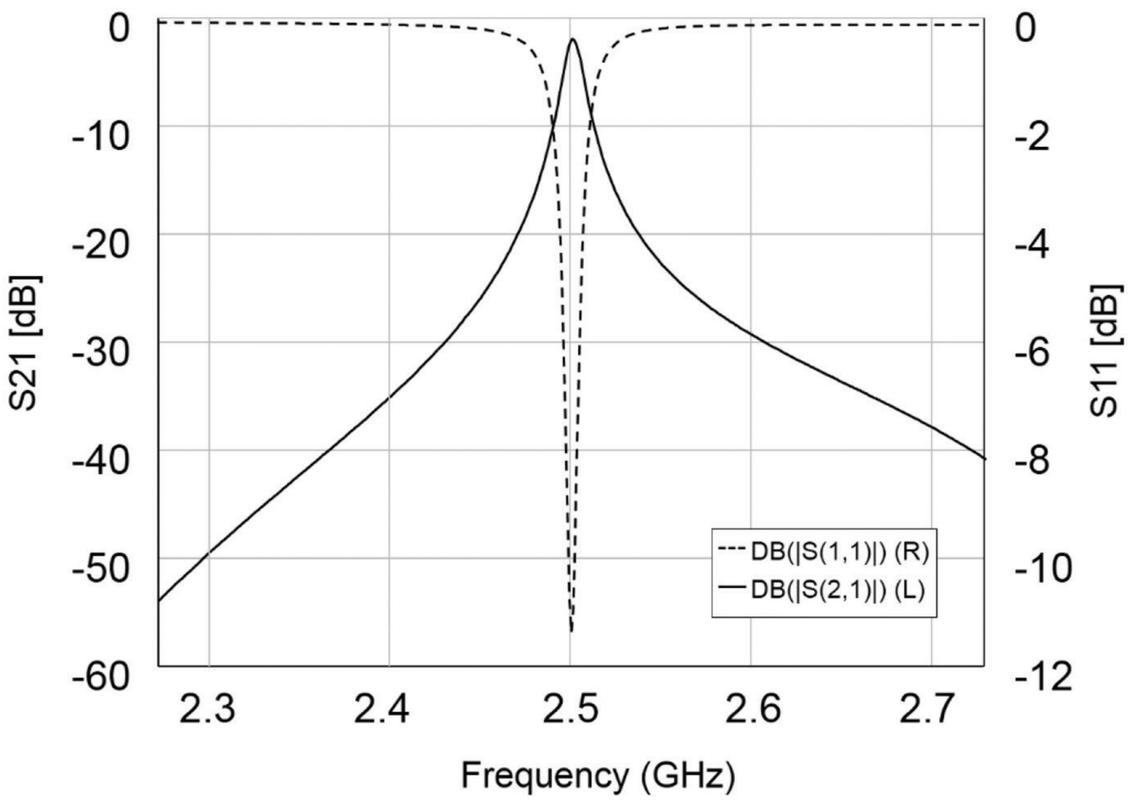

Figure 13. S-parameters of the designed active filter.

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