Bandwidth-Enhancement g_m-C *Filter with Independent* ω_O *and* Q *Tuning Mechanisms in Both Topology and Control Loops*

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***Abstract*—This paper deals with the proposal of a new topology for a g_m-C continuous time filter which allows the adjustment and tuning of its characteristic parameters (ω_O and *Q*) in an independent way (without cross-tuning), thereby extending the *Q* range of the filter for a particular ω_O value. Additionally a comparison of three different *Q*-tuning algorithms is presented. It is shown that an LMS-based *Q*-control strategy allows to overcome the intrinsic dependence between the *Q* and ω_O tuning loops. The combination of both the proposed filter topology and the selected control loop algorithms results in an enhanced transient performance as well as an improvement in terms of cross-detuning.**

I. INTRODUCTION

The number of wireless communication standards is currently large and is expected to steadily grow. This fact, together with the interest in the reduction of area and cost of handheld portable terminals, results in the need of developing versatile programmable transceivers with multistandard capabilities. In this work the topological improvement of a programmable base-band continuous time filter (CTF), including the selection of its more suitable tuning subsystem is addressed, targeting bandwidth enhancement and independent ω_O and *Q* tuning mechanisms.

The filter considered in this work is, as a starting point, a second order g_m-*C* topology, which is commonly used in baseband noise rejection filters in RF transceivers for communication applications capable of multistandards such as GSM, Bluetooh, CDMA2000 and WiMAX [1]-[3]. From the four conventional alternatives to implement high-order filters [4], a synthesis based on biquads has been considered [5]. High order filters can be implemented as a cascade of 2nd-order cells (plus a 1st-order in case of an odd-order filter). In addition, cascade filters are easier to build and tune. Thus, they are the most extended approach in order to implement high-order filtering systems [1]. On the other hand, the election of the g_m-*C* technology is justified because of bandwidth, silicon area, noise and power consumption reasons as well as its ease of tunability. The bias current adjusts the OTA transconductance (g_m), allowing to modify both the central frequency, ω_O, and the quality factor, *Q*, of the 2nd-order filter cell.

In this paper, a two-fold contribution is proposed. The first one is at filter topology level by presenting a topology with independent tuning of its characteristic parameters (central frequency ω_O and quality factor *Q*); the second one is at quality-factor tuning loop level by identifying the most suitable tuning subsystem after comparing both transient and steady state performances for the three existing alternatives. These two contributions have as main objective to improve the overall transient performance and tunability range.

The paper is organized as follows: In Section II the selected starting point topology of the second-order CT filter is presented. In this section it is discussed the problem associated to the fact that, although theoretically the frequency and the quality factor can be tuned independently, in practice, there is a limitation in the values of *Q* since the value of the quality factor also depends on the signal that tunes the central frequency. Thus, in Section III, an improvement in the original filter is proposed, avoiding this limitation. In Section IV the main characteristics of the master-slave self-tuning method and the ω_O-control loop are discussed, together with the three implementation alternatives for the *Q*-control loop. Finally, a comparative between them is discussed in Section V, showing the overall dynamic improvement when combining the enhanced filter topology and the selected control loop.

II. 2ND ORDER G_M-C CT FILTER

The second-order g_m-*C* continuous-time filter considered in this work is shown in figure 1 [1]-[3]. This filter implements a band-pass function between the input v_{in} and the output v_{bp} and a low-pass function between v_{in} and the output v_{lp}. The transfer function between v_{in} and the output v_{bp} is given by expression (1):

$$H(s)=\frac{V_{BP}(s)}{V_{IN}(s)}=\frac{\frac{g_0}{C_1}s}{s^2+\frac{g_3}{C_1}s+\frac{g_2g_1}{C_2C_1}}=\frac{\omega_o s}{s^2+\frac{\omega_o}{Q}s+\omega_o^2} \quad (1)$$

Assuming that three of the four cell transconductance are considered equal ($g_0=g_1=g_2$), as well as the circuit capacitors ($C_1=C_2=C$), the following expressions are obtained for the central frequency, ω_O, and the quality factor, Q, of the filter:

$$\omega_0 = \frac{g_0}{C} \quad ; \quad Q = \frac{g_0}{g_3} \tag{2}$$

Notice that in these expressions it is explicit that adjusting the transconductance value g_0 modifies the central frequency of the filter. Modifying the filter quality factor requires to adjust the transconductance value g_3. The transconductance of an OTA can be modified varying its control bias current [6] according to the expression (3) where a is a constant with units V^{-1}:

$$g_m = aI_{control} \tag{3}$$

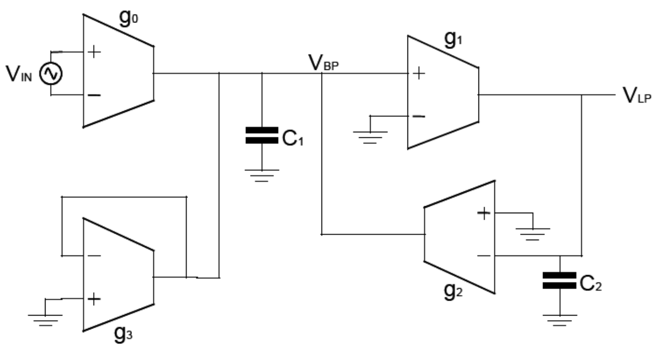

Fig. 1.- Second-order g_m-C filter cell.

Therefore, the transconductance g_0 and, in turn the central frequency ω_O, can be modified by a control current I_ω. On the other hand, the transconductance g_3 and, thus, Q, can be controlled through a control current I_Q. The relationships between the parameters ω_O and Q and the two control signals, assuming $a_\omega=a_Q=a$, are given by:

$$\omega_0 = \frac{aI_\omega}{C} \quad ; \quad Q = \frac{I_\omega}{I_Q} \tag{4}$$

As it is observed in expression (4), the quality factor depends not only upon the corresponding control current I_Q but also on the ω_O value of the filter (through the control current I_ω). This results in the fact that any error in ω_O-control current, I_ω^{error}, also impacts upon the Q control. In face of the previous argument, the maximum Q that is possible to achieve will be reduced when the tuned central frequency decreases as a result of the relationship between control variables just as it is shown in expression (5) and plotted in figure 2 (area a within red dashed line).

$$Q_{max} = \frac{I_\omega^{REF} + I_\omega^{error}}{I_Q^{min}} \tag{5}$$

On the other hand, the Q-control loop will present more problems for high quality factors since it requires small control current I_Q. Thus, the system is limited to a minimum value (I_Q^{min}) due to the non-idealities of the implementation.

III. Enhanced G_M-C Biquad Topology

In order to overcome the problem discussed in the previous section, an enhancement of the original filter topology in figure 1 is presented in figure 3. The target of this enhancement is to overcome the cross-dependence between the Q and ω_O tuning mechanisms, thereby expanding the possible filter operating area, shown in figure 2. The proposal is the filter presented in figure 3, obtained when the OTAs with transconductances g_A and g_B are added to the original structure. The transfer function between the output v_{bp} and v_{in} exhibits a band-pass characteristic, defined by the expression (6).

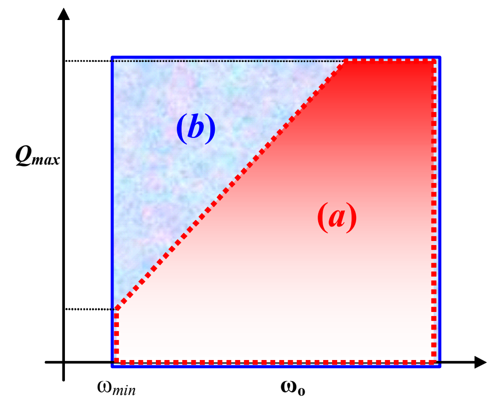

Fig. 2.- ***(a)*** Region (dashed line) containing the operation margin of the CT filter in figure 1. It can be seen that the maximum value of the quality factor depends upon the filter central frequency ω_O. ***(b)*** Region (solid line) containing the operation margin of the CT filter in figure 3, showing independence of the maximum Q value to the ω_O value.

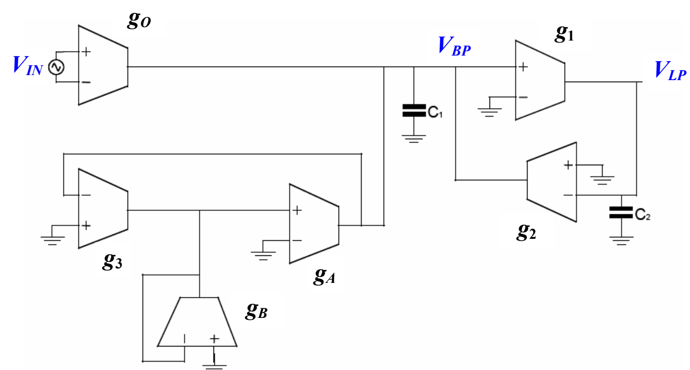

Fig. 3.- Proposed filter with crosstuning-free independent control between the central frequency ω_O and the quality factor Q.

$$H(s) = \frac{V_{BP}(s)}{V_{IN}(s)} = \frac{\frac{g_0}{C_1}s}{s^2 + \frac{g_3 g_A}{C_1 g_B}s + \frac{g_2 g_1}{C_2 C_1}} \tag{6}$$

Considering $g_0=g_1=g_2$ and capacitors $C_1=C_2=C$, the following expressions are obtained for ω_O and Q:

$$\omega_0 = \frac{g_0}{C} \quad ; \quad Q = \frac{g_0 g_B}{g_3 g_A} \tag{7}$$

Notice that if transconductances g_0 and g_A depend upon the same control current ($g_0=g_A$), ω_O and Q control adjustments will be independent. Thus, provided that:

$$g_0 = g_A = a \cdot I_\omega \;,\; g_B = a \cdot I_B \;\text{ and }\; g_3 = a \cdot I_q \tag{8}$$

the following expressions are obtained:

$$\omega_0 = \frac{a \cdot I_\omega}{C} \quad ; \quad Q = \frac{I_B}{I_q} \tag{9}$$

and the value of the maximum Q that can be obtained will not depend upon the value of the central frequency. In addition, possible errors in the ω_O-control loop will not influence the tuning of the Q-control loop. Accordingly, the region containing the operation margin of the CT filter in figure 3 is shown in figure 2 (area *b* within blue solid line). It can be seen, the maximum value of Q is independent of the ω_O value.

IV. PARAMETERS CONTROL LOOPS

A. *Filter Central Frequency Control Loop*

The need to implement a tuning system for CT filters to correct errors due to component tolerances is well known in the bibliography [7]. In this work, the tuning system is considered not only to correct drifts in the quality factor and the central frequency of the filter (fine adjustment or tuning) but also to program them according to changes in the operating band as required by the communication standard (coarse adjustment).

The most frequently used tuning system is, nowadays, the master-slave scheme [8]. In this tuning structure, a filter (the *master*) in the control loop is used. This filter is equal to a subcell (that is, the 2nd-order basic cell in a cascade structure) of the main filter that processes the signal of interest (*slave filter*). The master filter processes a reference signal (typically a sinusoidal tone). This reference signal together with the output signal of the master filter constitute the two inputs of the control loops, which are responsible for adjusting ω_O and Q of the master filter. The different cells of the slave filter are tuned with the same control signals obtained through these two control loops.

The control loop that tunes the central frequency is typically an analog loop. Its operation principle is based on the fact that if the reference input signal frequency (ω_{REF}) coincides with ω_O, the phase drift between the input and output signals is zero, just as it can be derived from expression (1). Thus, the frequency control loop obtains the phase difference between the filter input reference and the output signals. It adjusts the control current I_ω of the central frequency to minimize (ideally reducing to zero) this phase difference. The block diagram of this central frequency control loop is presented in figure 5 [8].

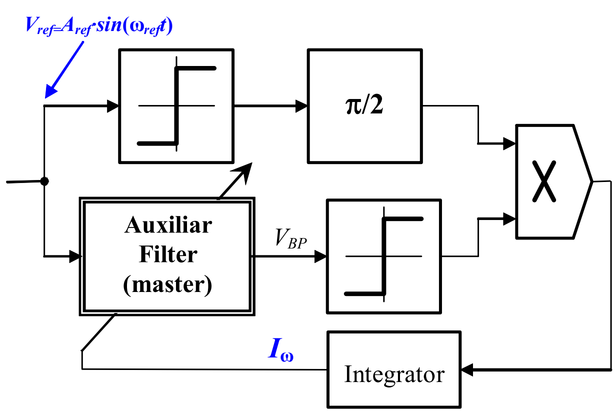

Fig. 5.- Block diagram of the central frequency control loop.

B. *Quality-Factor Control Loop*

The Q factor tuning loop can consist mainly of variations around three alternatives ***(a)*** selftuned envelope detector (SED, a modified version of the work shown in [9] which allows spreading its frequency range of application and, therefore, the programming range of the filter), ***(b)*** magnitude locked loop (MLL) [8], and ***(c)*** the least mean square (LMS) approach [1]. The first one (SED algorithm) extracts the amplitude of the filter output signal with an envelope detector and tries to minimize the difference between this amplitude and a reference constant value (set point). This reference value is equal to the input signal amplitude multiplied by the desired quality factor Q_{REF}. Both SED and MLL algorithms are based on the fact that, for an input tone equal to the filter central frequency (that is, $\omega_{REF}=\omega_O$), the output signal amplitude is Q times the input amplitude, just as it can be derived from expression (1).

The tuning algorithm MLL tries to minimize the difference between the square of the output signal amplitude and the square of the input signal amplitude amplified by the desired quality factor Q_{REF}. This algorithm obtains squares of the amplitudes to make good use of the fact that both input and output signals are sinusoidal tones. By squaring both signals, a constant term appears that is proportional to the square of the amplitude. This system and the SED method require that the ω_O of the master filter is correctly tuned, since the output signal amplitude depends upon the phase relationship between the input reference tone and output signals. Consequently, it is mandatory that in these two algorithms, the time constant (time response) of the quality-factor control loop should be designed much slower than the one for the ω_O-control loop in order to allow proper operation.

Finally, the LMS algorithm minimizes the quadratic error between the input signal amplified by a Q factor and the master filter output signal. That is, it minimizes the difference between the square of the output signal and the input-output crossed product. This tuning system has as a main advantage that it does not depend on the correct tune carried out by the central frequency control loop [1].

In figure 6 the block diagram for this tuning system is shown. As it can be observed, the quality-factor control current I_Q is obtained by the low-pass filtering of the difference between the square of the output signal and the product of the input and output signals, both being sinusoidal tones. This filtering, that will be implemented by a lossy integrator with high gain, provides the DC component of interest that is generated in such products. However, a ripple with a frequency that is twice the input signal frequency is also originated. Note that this ripple will constitute an error of the tuning signal.

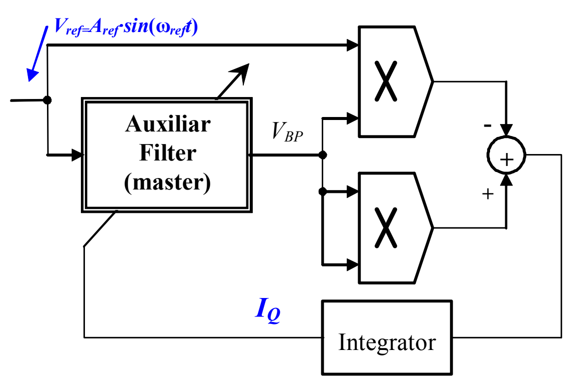

Fig. 6.- Block diagram of the Q control loop based on the LMS algorithm.

V. COMPARISON OF THE Q TUNING SYSTEMS

In order to discriminate which of the three systems presents more advantages, this paper compares the dynamics of the convergence of the control variables, that is, the speed and the error [10]. As mentioned before, it is necessary that the quality-factor control loop that implements the SED and MLL algorithm should be designed slower than the central frequency tuning loop. For this same reason, the LMS results in the control algorithm being faster when the two control loops are taken into account; that is, they are operating simultaneously (figure 7).

In case there is an error in the central frequency tuning, the SED and MLL methods could tune to a quality factor Q higher than the desired Q_{REF}. Indeed, these two systems consider that the gain of the filter is Q_{REF} *independently of the value of the tuned central frequency*. It is apparent that this consideration is only true when the filter is tuned (ω_O=ω_{REF}).

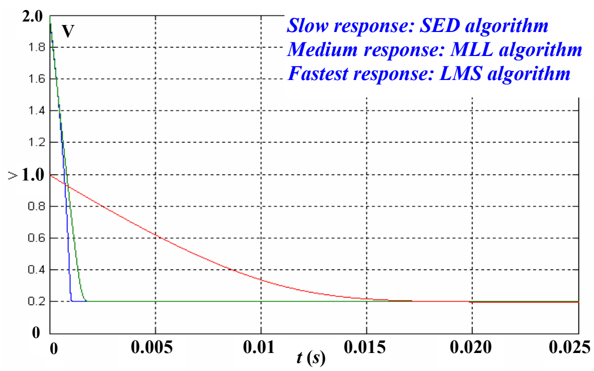

Fig. 7.- Transient response of the three considered Q control algorithms with Q_{REF}=5 and ω_{REF}=1 *MHz* in filter shown in figure 3.

On the other hand, the LMS algorithm achieves that the Q of the filter coincides with the Q_{REF}, although the central frequency control loop has not tuned ω_O yet (that is, independently of whether or not the reference input signal frequency ω_{REF} and the instantaneous ω_O coincide). Finally, in figure 8 the magnitudes of the frequency responses are shown for the three different Q-tuning algorithms considered when a central frequency tuning error exists. In short, the obtained results prove that the LMS method is the algorithm that presents better benefits. This algorithm contributes also to enhanced flexibility to the tuning system since it allows to adjust both the speed and the accuracy of the system depending on the final particular application.

VI. CONCLUSIONS

This paper has proposed a 2nd-order g_m-C CT filter with independent adjustment of their characteristic parameters (ω_O y Q). This independence leads to extend the Q range of the filter for a particular ω_O value.

On the other hand, in order to achieve the most suitable Q tuning, three different methods are compared. This comparative allows choosing the LMS-based Q-control algorithm as the one with best performance. This control strategy allows a cross-detuning between the Q and ω_O-control loops. The paper has shown that the LMS-based Q-control algorithm has the best dynamic behavior and allows to tune Q even if ω_O has not been tuned yet.

The performance characterization allows to validate that the combination of the proposed g_m-C CT filter together with the LMS-based Q-control algorithm results in an improvement and enhancement in the transient performance and in terms of cross-detuning.

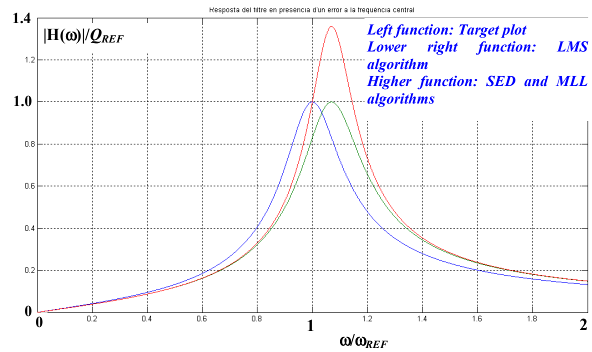

Fig. 8.- Frequency responses of the filter with the three Q-tuning algorithms in the case of an error in ω_O. Notice that only the LMS algorithm achieves to tune the accurate Q.

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