A 10.7-MHz Fully Balanced, High-Q, Low-Sensitivity Current-Tunable Gm-C Bandpass Filter

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Abstract

A 10.7-MHz fully balanced, high-Q, low-sensitivities current-tunable Gm-C bandpass filter is presented. The technique is relatively simple based on two fully balanced components, i.e. an adder and a low-Q-based bandpass filter. The Q factor is approximately equal to a typically high and constant value of a common-emitter current gain (β) and is independent of variables such as a center frequency. Sensitivities of either the Q factor or the center frequency are constant between -1 to 1 and are no longer strongly affected by the Q factor or variables. As a simple example at 10.7 MHz, the paper demonstrates the high-Q factor of 121. The center frequency is current tunable over three orders of magnitude.

Keywords: 10.7 MHz, fully balanced, high-Q, Gm-C bandpass filter, sensitivities.

1. Introduction

Bandpass filters are employed in many applications such as in a radio-frequency (RF) filter for image rejection or an intermediate-frequency (IF) filter for channel selection of a wireless receiver. Typically, FM radio receivers require an IF filter set at a center frequency (f_0) of 10.7 MHz based on off-chip devices such as discrete ceramic or surface acoustic wave (SAW) components [1,2]. As off-chip filters are bulky and consume more power to drive external devices, the need for possible on-chip filters for fully viable integrated receivers has increasingly been motivated. Recently, attempts at possible on-chip filters have particularly been demonstrated for 10.7-MHz IF filters based on, for example, switched capacitors (SC) [3-8], and Gm-C [9-13] techniques. Such techniques have, however, repeatedly suffered from low quality (Q) factors. In addition, most Q factors have generally been a function of variables such as a center frequency [14, 15]. For example, the quality factors of some existing Gm-C approaches [16, 17] have particularly been inversely proportional to the center frequency. Such variable quality factors have been difficult to tune as the variables may vary rapidly and drastically resulting in the need for additional or complicated Q-tunable circuits. In addition, sensitivities of neither the Q factor nor the center frequency at 10.7 MHz have been clearly reported, although sensitivities of the Q factor at other center frequencies have been undesirably a function of the Q factors [14, 15]. In this paper, a 10.7-MHz fully balanced, high-Q, wide-dynamic-range current-tunable Gm-C bandpass filter is presented. The technique is relatively simple based on two fully-balanced components, i.e. an adder and a low-Q-based bandpass filter. The Q factor is approximately equal to a typically high and constant value of a common-emitter current gain β. The high-Q factor is typically constant as it is no longer a function of variables such as a center frequency. This results in not only a great reduction in the need for additional Q-tunable circuits but also a much better sensitivity of the Q factor. The sensitivities of either the Q factor or the center frequency are constant values from -1 to 1 and are no longer strongly affected by the Q factor or variables.

As a simple example at 10.7 MHz, the paper demonstrates the high-Q factor of 121. The center frequency is current tunable over three orders of magnitude.

2. The Proposed 10.7-MHz High-Q Wide-Dynamic-Range Bandpass Filter

2.1 System Realization

Figure 1 shows the proposed system realization of a high-Q bandpass filter where the system is relatively simple based on two fully balanced components, i.e. an two-output adder A_D and a low-Q-based bandpass filter $A_{LQ}(s)$.

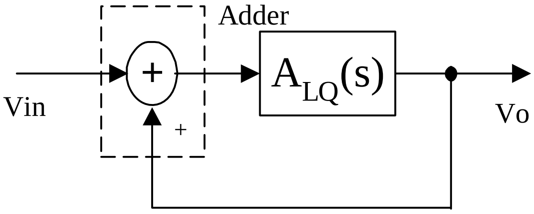

Figure 1. Proposed system realization of a high-Q bandpass filter

The transfer function of the low-Q-based bandpass filter $A_{LQ}(s)$ can be written as

$$A_{LQ}(s) = \frac{a_1\left(\frac{\omega_o}{Q_{LQ}}\right)s}{s^2 + \frac{\omega_o}{Q_{LQ}}s + \omega_o^2} \tag{1}$$

where a_1 is a pass band gain, i.e. $a_1 = A_{LQ}(s)$ at $s = j\omega_o$ and Q_{LQ} is a relatively low-Q factor of $A_{LQ}(s)$. A closed-loop gain $A_{HQ}(s) = v_o/v_{in}$ is given by

$$A_{HQ}(s) = \frac{A_{LQ}(s)}{1 - A_{LQ}(s)} \tag{2}$$

Substituting $A_{LQ}(s)$ in (2) with (1) yields

$$A_{HQ}(s) = \frac{a_1\left(\frac{\omega_o}{Q_{LQ}}\right)s}{s^2 + \frac{\omega_o}{Q_{HQ}}s + \omega_o^2} \tag{3}$$

where the quality factor Q_{HQ} is given by

$$Q_{HQ} = \frac{Q_{LQ}}{1 - a_1} \tag{4}$$

It can be seen from (4) that Q_{HQ} may ideally approach infinite if the denominator $(1 - a_1)$ approaches zero. In practice, the denominator of (4) may be made relatively small, i.e. a_1 is in the proximity of 1, resulting in a relatively high quality factor Q_{HQ}.

2.2 Circuit Realization

Figure 2 shows the proposed circuit realization for Fig 1 through an example of a fully balanced high-Q current-tunable Gm-C bandpass filter (A_{HQ}). The circuit consists of two fully balanced components, i.e. a two-output adder (A_D) and a low-Q-based bandpass filter (A_{LQ}) whilst A_D = 1 (i.e. a direct connection), using matched transistors T1 to T8. Firstly, the low-Q-based bandpass filter A_{LQ1} is a modified version of an existing low-Q bandpass filter [18] and consists of a differential pair (T1, T2), two capacitors C_1 and $4C_1$, two current sinks 2I and four loading diode-connected transistors T3 to T6. A small-signal input voltage of A_{LQ1} is v_{AB} between the bases of T1 and T2 (or nodes A and B). A small-signal output voltage of A_{LQ1} is v_{DE} between the emitters of T7 and T8 (or nodes D and E).

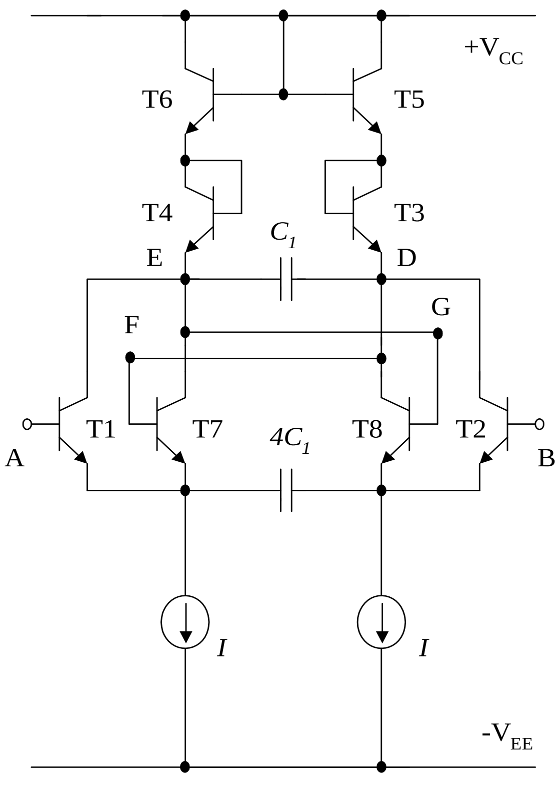

Figure 2. A 10.7-MHz fully balanced, high-Q current-tunable Gm-C bandpass filter

Secondly, the low-Q-based bandpass filter A_{LQ2} is a modified version of an existing low-Q bandpass filter [18] and consists of a differential pair (T7, T8), two capacitors C_1 and $4C_1$, two current sinks 2I and four loading diode-connected transistors T3 to T6. A small-signal input voltage of A_{LQ} is v_{DE} between the bases of T7 and T8 (or nodes D and E) and is obtained from the output v_o of A_D. A small-signal input voltage of A_{LQ2} is v_{FG} between the bases of T7 and T8 (or nodes F and G). A small-signal output voltage of A_{LQ2} is v_{DE} between the emitters of T7 and T8 (or nodes D and E). Finally, the transfer function of the low-Q-based bandpass filter is A_{LQ}=A_{LQ1}=A_{LQ2} and the transfer function of the high-Q bandpass filter is $A_{HQ} = v_O / v_{in}$ where $v_{in} = v_{AB}$ and $v_O = v_{DE}$. It can be seen from Fig. 1 that the circuit is fully balanced.

2.3 A Current Tunable Gm-C Bandpass Filter

Parameters r_{e1}, r_{e2}, …, r_{e7} and r_{e8} are the small-signal emitter resistance of transistors T1, T2, …, T7 and T8, respectively, where ($r_{e1} = r_{e2}= r_{e7}= r_{e8}$) = $V_T/(I/2)$, ($r_{e3} = r_{e4}=r_{e5} = r_{e6}$) $\cong V_T/(\alpha I_2) \cong 2r_{e1}/\alpha$ for $\alpha = \beta/(\beta+1)$ and β is the common-emitter current gain of a BJT. The usual thermal voltage V_T is approximately 25 mV associated with an pn junction at room temperature. Firstly, the two-output adder A_D is considered. The output v_{DE} of A_D is obtained through superposition, i.e. $v_{DE} = v_{O1}+ v_{O2}$. The voltage v_{O1} is the output v_{DE} of A_D when the 1st-input v_{AB} of A_{LQ1} is activated, i.e. $v_{AB} = v_{in}$, but the 2nd-input v_{FG} of A_{LQ2} is temporary deactivated or separately connected to an ac ground, i.e.$v_{FG} = 0$. In contrast, the voltage v_{O2} is the output v_{DE} of A_D when the 2nd-input v_{DE} of A_{LQ2} is activated, i.e. $v_{DE} = v_O$, but the 1st-input v_{AB} of A_{LQ1} is temporary deactivated or connected to an ac ground, i.e. $v_{AB} = v_{in} = 0$. On the one hand, v_{O1} can be found at $v_{FG} = 0$. Therefore, v_{in} of A_{LQ1} enables a small-signal emitter current i_{e1} passing through the emitters of T1 and T2. The resulting small-signal collector current of T1 and T2 is i_{c1}. Most of i_{c1} passes through a loading impedance $Z_1 = 4r_{e2}$ // $(1/sC_1)$ formed by T3, T4, T5, T6 and C_1. As $v_{O1} \cong i_{c1}Z_1$, therefore $v_{O1}/v_{in} \cong A_{LQ1}$. On the other hand, v_{O2} can be found at $v_{AB} = 0$. Therefore, v_{FG} of A_{LQ1} enables a small-signal emitter current i_{e1} passing through the emitters of T1 and T2. The resulting small-signal collector current of T1 and T2 is i_{c1}. Most of i_{c1} passes through a loading impedance $Z_1 = 4r_{e2}$ // $(1/sC_1)$ formed by T3, T4, T5, T6 and C_1. As $v_{O2} \cong i_{c1}Z_1$, therefore $v_{O2}/v_{in} \cong A_{LQ2} = A_{LQ1}= A_{LQ}$. Consequently, $v_{DE} = v_{O1} + v_{O2} \cong A_{LQ}\, v_{AB} + A_{LQ}\, v_{FG}$. As $v_{DE} = v_{FG} = v_O$ and $v_{AB} = v_{in}$ therefore

$$v_o = A_{LQ} v_{in} + A_{LQ} v_o \qquad (5)$$

Secondly, the low-Q-based bandpass filter A_{LQ1} is considered. The input v_{AB} of A_{LQ1} enables a small-signal emitter current $i_{e1} = v_{AB}\,(4sC_1)/(1+s\tau_1)$ passing through the emitters of T1 and T2, where $\tau_1 = 8r_{e1}C_1$. The resulting small-signal collector current of T5 and T6 is $i_{c1} = \alpha i_{e1}$. Most of i_{c1} passes through a loading impedance $Z_1 = 4r_{e4}/(1+s\tau_2)$ formed by T3 to T6 where $\tau_2 = 4r_{e4}C_1= 4(2r_{e1}/\alpha)C_1 = 8r_{e1}C_1/\alpha$ therefore $\tau_2 \cong \tau_1/\alpha$. The resulting output of A_{LQ1} is $v_{01} \cong i_{c2}Z_1$, therefore $A_{LQ1} = v_{O1} / v_{AB}$ represents a low-Q-based bandpass filter $A_{LQ1} = A_{LQ}$ of the form

$$\mathrm{A}_{LQ} = \frac{2\, s\alpha / \tau_1}{s^2 + (1+\alpha)\dfrac{s}{\tau_1} + \dfrac{\alpha}{\tau_1^{\,2}}} \qquad (6)$$

The center frequency of (7) is $\omega_{LQ} = (\alpha^{1/2})/\tau_1$. The quality factor of (6) is $Q_{LQ} = (\alpha^{1/2}) / (1+\alpha) \cong 0.5$ which is a relatively low value. Similarly, the low-Q-based bandpass filter A_{LQ2} is considered. The input v_{FG} of A_{LQ1} enables a small-signal emitter current $i_{e1} = v_{AB}\,(4sC_1)/(1+s\tau_1)$ passing through the emitters of T7 and T8, where $\tau_1 = 8r_{e1}C_1$. The resulting small-signal collector current of T5 and T6 is $i_{c1} = \alpha i_{e1}$. Most of i_{c1} passes through a loading impedance $Z_1 = 4r_{e4}/(1+s\tau_2)$ formed by T3 to T6 where $\tau_2 = 4r_{e4}C_1= 4(2r_{e1}/\alpha)C_1 = 8r_{e1}C_1/\alpha$ therefore $\tau_2 \cong \tau_1/\alpha$. The resulting output of A_{LQ2} is $v_{01} \cong i_{c2}Z_1$, therefore $A_{LQ2} = v_{O2} / v_{AB}$ represents a low-Q-based bandpass filter $A_{LQ2} = A_{LQ1} = A_{LQ}$ of (6). Finally, the high-Q bandpass filter A_{HQ} can be considered by substituting A_{LQ} in (6) with (5), therefore $A_{HQ} = v_O / v_{in} \cong A_{LQ} / (1-A_{LQ})$, i.e.

$$A_{HQ} = \frac{v_o}{v_{in}} = \frac{2\alpha s/\tau_1}{s^2 + (1-\alpha)\frac{s}{\tau_1} + \frac{\alpha}{\tau_1^2}} \tag{7}$$

The center frequency of (7) is $\omega_{HQ} = (\alpha^{1/2}) / \tau_1 = g_{m3} / [(\alpha^{1/2})\, 8C_1]$ where the transconductance $g_{m3} = \alpha / r_{e3}$. it can be seen from (2) and (7) that $a_1 \cong \alpha^{1/2} \cong 1$. The center frequency ω_{HQ} is current tunable by I_2 of the form

$$\omega_{HQ} = \frac{I_2}{8C_1 V_T}\sqrt{\frac{\beta}{\beta + 1}} \tag{8}$$

2.4 A High Quality Factor

The quality factor of (7) is $Q_{HQ} = (\alpha^{1/2}) / (1-\alpha)$ and therefore

$$Q_{HQ} \cong \beta \tag{9}$$

The quality factor Q_{HQ} of the proposed technique is approximately equal to a typically high (>100) and constant value of the current gain β and is, for the first time, no longer a function of variables such as a center frequency. The typically constant quality factor Q_{HQ} results in not only a great reduction in the need for additional or complicated tunable circuits, but also a much better sensitivity of the Q factor.

Variations of Q_{HQ} with temperature may be expected, as β may depart from its typically constant value due to temperature. Such variations, however, are relatively much smaller and slower than the variations of most reported Q factors which have generally been a function of variables such as a center frequency [14, 15] or have particularly been inversely proportional to the center frequency [16, 17].

In addition, possible solutions for good stability of the quality factor Q_{HQ} with temperature could be suggested through the use of, for example, InGaP/GaAs Heterojunction Bipolar Transistors (HBTs) where a relatively constant current gain β has been reported as a function of temperature up to 300°C [19], or through the use of $Al_XGa_{0.52-X}In_{0.48}P$/GaAs HBTs where good stability of β with collector current and temperature has been demonstrated for X = 0.18 - 0.30 [20].

2.5 Sensitivities

Generally, a sensitivity of y to a variation of x is given by $S_x^y = [\partial y/\partial x][x/y)]$ where y is a parameter of interest and x is a parameter of variation. Table 1 shows the sensitivity S_x^y where (x, y) = (β, Q_{HQ}), (C_1, ω_{HQ}), (V_T, ω_{HQ}), (I_2, ω_{HQ}) or (β, ω_{HQ}). The thermal voltage V_T also represents effects of temperature on the centre frequency ω_{HQ} whilst the current gain β also represents effects of temperature on the quality factor Q_{HQ}. For a relatively large value of β, the last sensitivity $S_\beta^{\omega_{HQ}}$ is not only relatively small (e.g. $S_\beta^{\omega_{HQ}} = 0.0041$ at β = 120) but also relatively constant if the variation of β is comparatively smaller than its value (e.g. $S_\beta^{\omega_{HQ}}$

= 0.0035 at β = 140). Consequently, it can be seen from Table 1 that the sensitivities of both Q_{HQ} and ω_{HQ} are relatively constant between –1 to 1. Such sensitivities are, unlike existing approaches [14, 15], no longer strongly affected by the Q factor or variables.

Table 1. Sensitivity S_x^y where (x, y) = (β, Q_{HQ}), (C_1, ω_{HQ}), (V_T, ω_{HQ}), (I_2, ω_{HQ}) or (β, ω_{HQ})

$S_{\beta}^{Q_{HQ}}$	$S_{C_1}^{\omega_{HQ}}$	$S_{V_T}^{\omega_{HQ}}$	$S_{I_2}^{\omega_{HQ}}$	$S_{\beta}^{\omega_{HQ}}$
1.0	-1.0	-1.0	1.0	1/[2(β+1)]

3. Simulation Results

As a simple example, all transistors in Fig. 2 are modeled by a simple transistor PN2222 where the average transition frequency (f_T) is 120 MHz and β is approximately 120 [21]. The bias current I_1 = 1 mA and C_1 = 150 pF. Figure 3 illustrates the measured frequency response of Fig. 2 at the center frequencies $f_0 = \omega_{HQ}/(2\pi)$ = 10.7 MHz. It can be seen from Fig. 3 that the bandwidth (BW) is 2×44 = 88 kHz and therefore the measured quality factor Q_{HQ} (=f_0/BW) is relatively high at approximately 121 which is consistent with the value of β =120.

Figure 4 shows plots of the center frequencies $f_0 = \omega_{HQ}/(2\pi)$ and the corresponding quality factor Q_{HQ} of Fig. 2 versus the bias current I for three cases, i.e. the analysis, the SPICE simulations, and the expected results. It can be seen from Fig. 4 that f_0 is current tunable over 3 orders of magnitude. As expected, Q_{HQ} essentially remains almost constant at approximately 120 and is, unlike existing approaches, independent of variables such as a center frequency. When I > 10 mA, f_0 drops with further increase of the bias current due to effects of parasitic capacitances at higher frequencies. Although the upper value of I_2 can be expected to be higher than 50 mA.

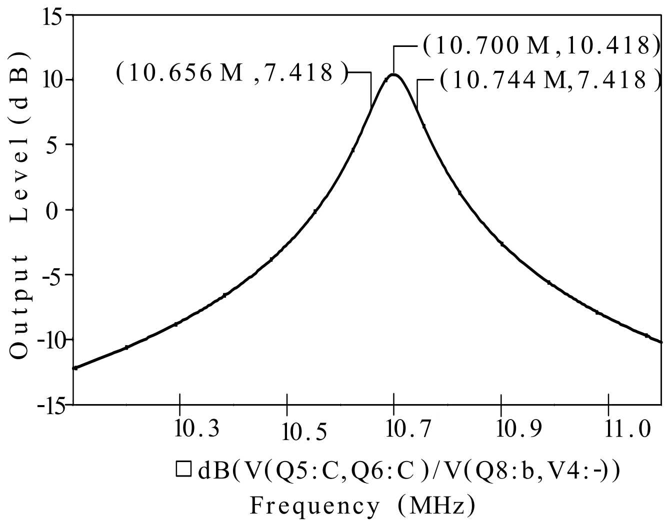

Figure 3. A measured frequency response at the centre frequency $f_0 = \omega_{HQ}/(2\pi)$ = 10.7 MHz and the quality factor Q_{HQ} = 120.

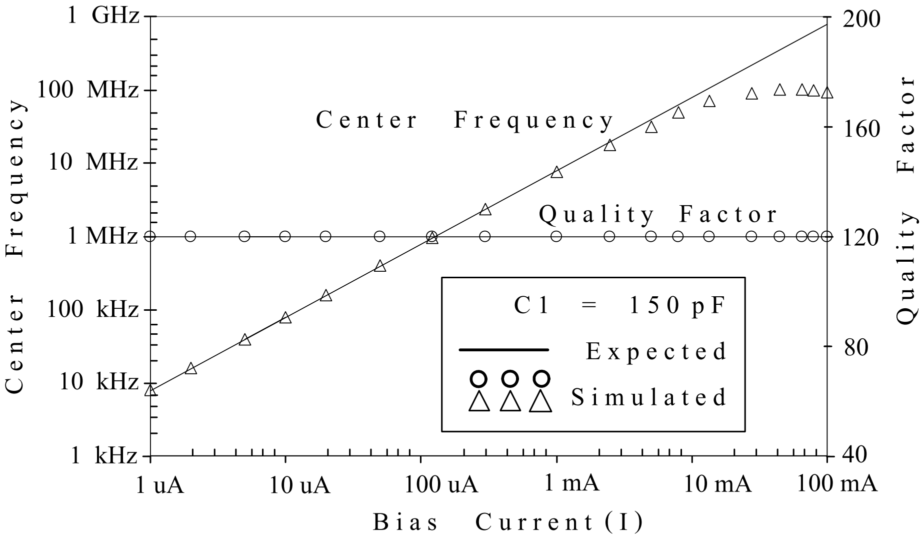

Figure 4. Plots of the center frequency $f_0 = \omega_{HQ}/(2\pi)$ and the quality factor Q_{HQ} versus the bias current I

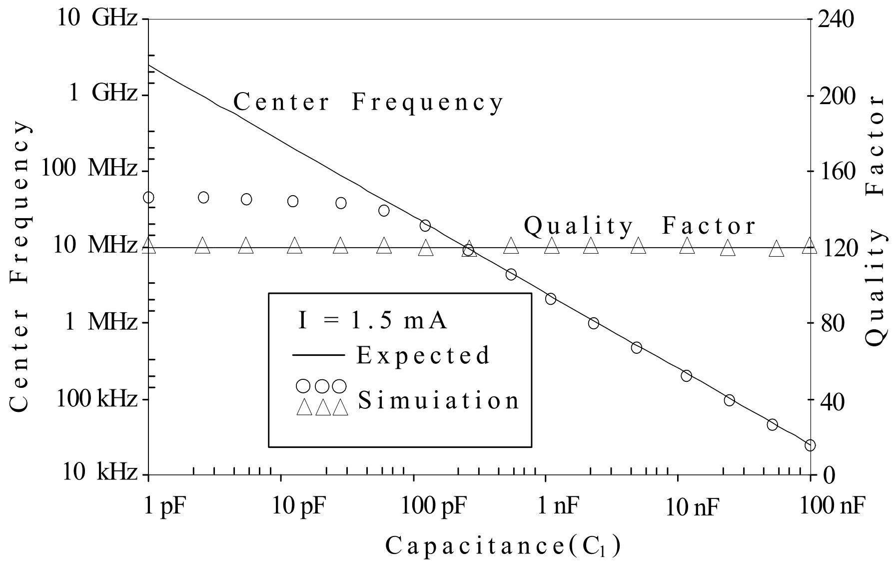

Figure 5. An example of the center frequencies f_0 and the quality factor Q_H versus capacitance C_1 with fixed bias currents I = 1.5 mA.

Figure 5 shows high-frequency performance of Fig. 1 through the analysis and the PSPICE simulations in terms of the center frequency and the quality factor QHQ. In this particular example, QHQ is maintained relatively high and the upper frequency is limited at approximately 81 MHz at C1 = 1 pF.

4. Conclusions

A 10.7-MHz fully balanced, high-Q, low-sensitivity current-tunable Gm-C bandpass filter has been proposed. The technique is relatively simple based on two fully balanced components, i.e. the output adder and the low-Q-based bandpass filter. The Q factor is approximately equal to a typically high and constant value of a common-emitter current gain (β) and is, for the first time, independent of variables such as a center frequency. Possible solutions for good stability of the Q factor with temperature have been suggested. Not only can the need for additional Q-tunable circuits be greatly reduced, the sensitivity of the Q factor can be greatly improved. Sensitivities of either the Q factor or the center frequency are constant between –1 to 1 and are no longer strongly affected by the Q factor or variables. As a simple example at 10.7 MHz, the paper has demonstrated the high-Q factor of 120. The center frequency is current tunable over three orders of magnitude. Comparisons to other 10.7-MHz Gm-C approaches have been included. The proposed technique offers a potential alternative to a 10.7-MHz high-Q bandpass filter.

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