10.7-MHz Fully Balanced, High-Q, Wide-Dynamic-Range Current-Tunable Gm-C Bandpass Filter

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ABSTRACT

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 typical high and constant value of a common-emitter current gain ($\beta$) and is, for the first time, 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 low total output noise of 5.303 $\mu$V rms, the 3rd-order intermodulation-free dynamic range (IMFDR₃) of 74.45 dB and the wide dynamic range of 87.45 dB at 1% IM₃. The center frequency is current tunable over 3 orders of magnitude. Comparisons to other 10.7-MHz Gm-C approaches are also included.

Keywords: 10.7 MHz, fully balanced, high-Q, wide dynamic range, 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 from 10 to 55, high total noise from 226 to 707 $\mu$V rms, and limited dynamic ranges from 58 to 68 dB. 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 high-Q, wide-dynamic-range current-tunable Gm-C bandpass filter is introduced using two fully balanced devices, i.e. an adder and a low-Q-based bandpass filter. The high-Q factor is equal to a common-emitter current gain $\beta$. For the first time, 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 and are not affected by the Q factor or variables. Possible solutions for good stability of the Q factor with temperature are also suggested. The technique is demonstrated through an example at 10.7 MHz. Temperature compensations for both the center frequency and the Q factor are summarized. Other 10.7-MHz Gm-C approaches are also compared.

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 three fully balanced components, i.e. a two-input adder $A_D$, an amplifier $A_G$ and a low-Q-based bandpass filter $A_{LQ}(s)$.

The transfer function of the low-Q-based bandpass filter $A_{LQ}(s)$ can be written as
Fig.1: Proposed system realization of a high-Q bandpass filter.

\[ A_{LQ}(s) = \frac{a_1(\frac{\omega_o}{Q_{LQ}})s}{s^2 + (\frac{\omega_o}{Q_{LQ}})s + \omega_o^2} \]  

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 \( A_{LQ}(s) \). Consequently, a closed-loop gain \( A_{HQ}(s) = v_o/v_{in} \) is given by

\[ A_{HQ}(s) = \frac{A_D A_{LQ}(s)}{1 - A_D A_G A_{LQ}(s)} \]  

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

\[ A_{HQ}(s) = \frac{A_D a_1(\frac{\omega_o}{Q_{LQ}})s}{s^2 + (\frac{\omega_o}{Q_{LQ}})s + \omega_o^2} \]  

where the quality factor \( Q_{HQ} \) is given by

\[ Q_{HQ} = \frac{Q_{LQ}}{1 - A_D A_G a_1} \]  

It can be seen from (4) that \( Q_{HQ} \) may ideally approach infinite if the denominator (1 - \( A_D A_G a_1 \)) approaches zero. In other words,

\[ A_G \rightarrow \frac{1}{a_1 A_D} \]  

In practice, the denominator of (4) may be made relatively small, i.e. \( A_G \) is in the proximity of \( 1/(a_1 A_D) \), 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-input adder \( (A_D) \) and a low-Q-based bandpass filter \( (A_{LQ}) \) whilst \( A_G \approx 1 \) (i.e. a direct connection), using matched transistors T1 to T10. In this case, equation (5) suggests that the gain of the adder \( A_D \approx 1 \) and \( a_1 \approx 1 \). Firstly, the adder \( A_D \) is a modified version of an existing adder [18] and consists of a differential pair \( (T1, T2) \), a common-collector pair \( (T3, T4) \) and two current sinks \( I_1 \). The 1st small-signal input voltage of \( A_D \) is \( v_{AB} \) between the bases of T1 and T2 (or nodes A and B). The 2nd small-signal input voltage of \( A_D \) is \( v_{CD} \) between the bases of T3 and T4 (or nodes C and D). A small-signal output voltage of \( A_D \) is \( v_{EF} \) between the emitters of T3 and T4 (or nodes E and F).

Secondly, the low-Q-based bandpass filter \( A_{LQ} \) is a modified version of an existing low-Q bandpass filter [19] and consists of a differential pair \( (T5, T6) \), two capacitors \( C_1 \) and \( 2C_1 \), two current sinks \( I_2 \) and four loading diode-connected transistors T7 to T10. A small-signal input voltage of \( A_{LQ} \) is \( v_{EF} \) between the bases of T5 and T6 (or nodes E and F) and is obtained from the output \( v_{EF} \) of \( A_D \). A small-signal output voltage of \( A_{LQ} \) is \( v_{CD} \) between the emitters of T7 and T8 (or nodes C and D). Finally, the transfer function of the high-Q bandpass filter is \( A_{HQ} \) and is obtained through superposition, i.e. \( v_{HQ} = v_{IQ1} + v_{IQ2} \). The voltage \( v_{IQ1} \) is the output \( v_{EF} \) of \( A_D \) when the 1st-input \( v_{AB} \) of \( A_D \) is activated, i.e. \( v_{AB} = v_{in} \), but the 2nd-input \( v_{CD} \) of \( A_D \) is tem-
considered by substituting \( s = j \).

The quality factor of (7) is

\[
\omega_{HQ} = \frac{I_2}{4C_1V_T\sqrt{\frac{\beta}{\beta+1}}} 
\]  

(9)

2.4 A High Quality Factor

The quality factor of (8) is \( Q_{HQ} = (\alpha^{1/2})/(1 - \alpha) \) and therefore

\[
Q_{HQ} \approx \beta 
\]

(10)

The quality factor \( Q_{HQ} \) of the proposed technique is approximately equal to a typically high (> 100) and constant value of the current gain \( \beta \) 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 \( \beta \) 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 \( \beta \) has been reported as a function of temperature up to 300 °C \([20]\), or through the use of \( Al_xGa_{0.52-\chi}In_{0.48\chi}/GaAs \) HBTs where good stability of \( \beta \) with collector current and temperature has been demonstrated for \( X = 0.18 - 0.30 \) \([21]\).

2.5 Sensitivities

Generally, a sensitivity of \( y \) to a variation of \( x \) is given by \( S_y^x = [\partial y/\partial x][x/y] \) where \( y \) is a parameter of interest and \( x \) is a parameter of variation. Table 1 shows the sensitivities \( S_y^x \) where \((x, y) = (\beta, Q_{HQ}), (C_1, \omega_{HQ}), (V_T, \omega_{HQ}), (I_2, \omega_{HQ}) \) or \((\beta, \omega_{HQ}) \). The thermal voltage \( V_T \) also represents effects of temperature on the centre frequency \( \omega_{HQ} \) whilst the current gain \( \beta \) also represents effects of temperature on the quality factor \( Q_{HQ} \). For a relatively large value of \( \beta \), the last sensitivity \( S_{\omega_{HQ}}^\beta \) is not only relatively small (e.g. \( S_{\omega_{HQ}}^\beta = 0.0041 \) at \( \beta = 120 \)) but also relatively constant if the variation of \( \beta \) is comparatively smaller than its value (e.g. \( S_{\omega_{HQ}}^\beta = 0.0035 \) at \( \beta = 140 \)). Consequently, it can be seen from Table 1 that the sensitivities of both \( Q_{HQ} \) and \( \omega_{HQ} \) 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.
2.6 Dynamic Ranges

Dynamic ranges (DRs) of either a specific biquad or an optimized high-Q biquad in a general way have been presented [22]. An expression for the dynamic range of a second-order Gm-C biquad in a general way is given by [22]:

$$DR = \frac{v_{\text{max}}^2}{v_{\text{noise}}^2} = \frac{v_{\text{max}}^2}{kT\xi Q\left(\frac{1}{C_a} + \frac{1}{C_b}\right)}$$

(11)

where \(v_{\text{max}}\) is the maximal signal level (at the input or output of a system), \(v_{\text{noise}}\) is the mean squared noise voltage at the same point, \(C_a\) and \(C_b\) are two capacitors in the filter, \(k\) is the Boltzmann’s constant, \(T\) is the absolute temperature, \(\xi\) is the noise factor of the transconductor (Gm) and \(Q\) is the quality factor.

The dynamic range of the proposed technique can be improved by not only increasing \(v_{\text{max}}\), but also reducing \(v_{\text{noise}}\) of (11) as follows.

On the one hand, it is known that, the maximal signal level \(v_{\text{max}}\) of a fully balanced circuit is typically twice the maximal signal level \(v_3\) of a single-ended circuit [22], i.e. \(v_{\text{max}} \approx 2v_3\). In other words, the magnitude \(v_{\text{max}}\) of (11) may be double through the use of a fully balanced circuit. On the other hand, the mean squared noise voltage can be reduced through the use of a shunt positive feedback configuration providing enhanced current gain and thereby improving the overall noise [23]. Table 2 summarizes values of \(C_a, C_b\), and dynamic ranges (DRs) of the proposed Gm-C techniques and other existing Gm-C approaches [22, 24].

It can be seen from Table 2 that if \(v_{M1} = v_{M2} = v_{M3}\) and \((KT\xi Q)_1 = (KT\xi Q)_2 = (KT\xi Q)_3\), then \(DR_1 > DR_2 > DR_3\). The proposed Gm-C fully-balanced technique can therefore enable a higher dynamic range \(DR_1\), especially when \(v_{\text{noise}}^2\) is also additionally reduced. In particular, as the quality factor \(Q\) in (11) becomes \(Q_{HQ}\) which is no longer a function of variables such as a center frequency, the dynamic range \(DR_1\) is therefore, unlike existing approaches [22, 24], no longer strongly affected by those variables previously associated in the Q factor. As an example, it can be expected from Table 2 that \(DR_1 = 88.28\) dB if \(v_{\text{max}} = 2v_{M1} = 127\) mV (i.e. -5 dBm through a 50 – Ω load), \(v_{M1} = 63.5\) mV, \(kT\xi = 2 \times 10^{-23}\) [24], \(Q = 120\) and \(C = 150\) pF.

### Table 1: Sensitivity \(S^y_x\) where \((x, y) = (\beta, Q_{HQ}), (C_1, \omega_{HQ}), (V_T, \omega_{HQ}), (I_2, \omega_{HQ})\) or \((\beta, \omega_{HQ})\).

<table>
<thead>
<tr>
<th>(S^y_x)</th>
<th>(S^x_{C_1})</th>
<th>(S^x_{V_T})</th>
<th>(S^x_{I_2})</th>
<th>(S^x_{\omega_{HQ}})</th>
<th>(S^x_{Q_{HQ}})</th>
</tr>
</thead>
<tbody>
<tr>
<td>1.0</td>
<td>-1.0</td>
<td>-1.0</td>
<td>1.0</td>
<td>1/2((\beta + 1))</td>
<td></td>
</tr>
</tbody>
</table>

### Table 2: Summaries of dynamic ranges (DRs) using second-order Gm-C techniques.

<table>
<thead>
<tr>
<th>Refs</th>
<th>Capacitors</th>
<th>(v_{\text{max}})</th>
<th>DR = ((KT\xi Q)\left(\frac{1}{C_a} + \frac{1}{C_b}\right))</th>
</tr>
</thead>
<tbody>
<tr>
<td>This paper (fully balanced)</td>
<td>(C_a = C) and (C_b = 2C)</td>
<td>2(v_{M1})</td>
<td>(DR_1 = 2.67)</td>
</tr>
<tr>
<td>[24] (single ended)</td>
<td>(C_a = C) and (C_b = C)</td>
<td>(v_{M2})</td>
<td>(DR_2 = 0.50)</td>
</tr>
<tr>
<td>[22] (single ended)</td>
<td>(C_a = C/2) and (C_b = C/2)</td>
<td>(v_{M3})</td>
<td>(DR_2 = 0.25)</td>
</tr>
</tbody>
</table>

### Fig.3: A measured frequency response at the centre frequency \(f_0 = \omega_{HQ}/(2\pi) = 10.7\) MHz and the quality factor \(Q_{HQ} = 121\).
Table 3: Summaries of related noise parameters obtained from Fig. 5.

<table>
<thead>
<tr>
<th>Noise Parameters</th>
<th>Values</th>
<th>Units</th>
</tr>
</thead>
<tbody>
<tr>
<td>(1) Resolution bandwidth (RBW)</td>
<td>100</td>
<td>kHz</td>
</tr>
<tr>
<td>(2) Noise Density</td>
<td></td>
<td></td>
</tr>
<tr>
<td>( P_{N1} = 10 \log(P_{N2}/1mW) )</td>
<td>-142.5</td>
<td>dBm/Hz</td>
</tr>
<tr>
<td>( P_{N2} )</td>
<td>5.62 \times 10^{-18}</td>
<td>W/Hz</td>
</tr>
<tr>
<td>( V_{N1}^2 = P_{N2} \times (50\Omega) )</td>
<td>2.81 \times 10^{-16}</td>
<td>( V^2/Hz )</td>
</tr>
<tr>
<td>( V_{N1} = \sqrt{V_{N1}^2} )</td>
<td>1.68 \times 10^{-8}</td>
<td>( V_{rms}/\sqrt{Hz} )</td>
</tr>
<tr>
<td>(3) Total Noise</td>
<td></td>
<td></td>
</tr>
<tr>
<td>( V_{N2} = V_{N1}^2 \times (RBW) )</td>
<td>2.81 \times 10^{-11}</td>
<td>( V^2 )</td>
</tr>
<tr>
<td>( V_{N3} = \sqrt{V_{N2}} )</td>
<td>5.303 \times 10^{-6}</td>
<td>( V_{rms} )</td>
</tr>
<tr>
<td>( P_{N3} = 10 \log(V_{N3}/1mW) )</td>
<td>-92.45</td>
<td>dBm</td>
</tr>
</tbody>
</table>

**Fig. 4:** Plots of the center frequency \( f_0 = \omega_{HQ}/(2\pi) \) and the quality factor \( Q_{HQ} \) versus the bias current \( I_2 \).

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 \( I_2 \) for three cases, i.e. the analysis, the SPICE simulations, and the experimental 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 121 and is, unlike existing approaches, independent of variables such as a center frequency. When \( I_2 > 1 \) 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 10 mA, the upper limit of the circuit prototypes has been set to 5 mA, for safe operation of the current sources.

### 3.2 Low Noise Performance

Figure 5 shows the measured output noise spectrum shaped by the transfer function of the filter, where the power noise density \( P_{N1} \) is relatively low at -142.5 dBm/Hz and the resolution bandwidth (RBW) is at 100 kHz. Table 3 summarizes resulting noise parameters in terms of (1) the resolution bandwidth, (2) the noise density and (3) the total noise. Table 3 concludes that the output noise density \( V_{N1} = 0.016 \mu V_{rms}/\sqrt{Hz} \), the total output noise \( V_{N3} = 5.303 \mu V_{rms} \) and the total noise power \( P_{N3} = -92.45 \) dBm.

### 3.3 Wide Dynamic Range

The circuit is excited with two sinuoids at frequencies \( f_1 = f_0 - 7.5kHz = 10.6925MHz \), and \( f_2 = f_0 + 7.5kHz = 10.7075MHz \). The 3rd-order intermodulation (\( IM_3 \)) products \( |2f_1 - f_2| \) and \( |2f_2 - f_1| \) are 10.6775 and 10.7225 MHz, respectively. Figure 6 shows the measured output spectrums at \( Q_{HQ} = 121 \) using the two-frequency excitation of -20 dBm at \( f_1 \) and \( f_2 \). It can be seen that the \( IM_3 \) products are approximately 40 dB down from the fundamentals and correspond to 1% (or 1% \( IM_3 \)). Through a 50 – Ohm load of the spectrum analyzer without the output buffer, Figure 7 depicts the measured output levels (dBm) of the fundamental at \( f_1 \) and the \( IM_3 \) at \( |2f_1 - f_2| \) versus the input levels (dBm). It can be seen from Fig. 7 that the noise
power $P_{N3} = -92.45$ dBm. At the input level of -35 dBm, the output level of $f_1$ is -18 dBm whilst the output level of the $IM_3$ is adjacent to $P_{N3}$ (or intermodulation free). Therefore the 3rd-order intermodulation-free dynamic range ($IMF\, DR_3$) = $(-18 \, \text{dBm}) - (-92.45 \, \text{dBm}) = 74.45 \, \text{dB}$. In addition, at the input level of -20 dBm, the output level of $f_1$ is -5 dBm, whilst the output level of the $IM_3$ is 40 dB down from $f_1$ (or 1%$IM_3$). Therefore, the wide dynamic range (at 1%$IM_3$) = $(-5 \, \text{dBm}) - (-92.45 \, \text{dBm}) = 87.45 \, \text{dB}$ which is consistent with the expected value $DR_1 = 88.28 \, \text{dB}$ predicted in Section 2.6.

3.4 Effects of Temperature on the Center Frequency

For the high-Q bandpass filter $A_{HQ}$, Figure 8 shows two cases of the measured variations of the normalized center frequency $f_0 / (10.7 \, \text{MHz})$ versus the ambient temperature (Celsius). The first case is an “uncompensated” case where the effects of temperature on $f_0$ have not been compensated. The second case is a “compensated” case where the effects of temperature on $f_0$ have been compensated.

The uncompensated case can be demonstrated by taking Fig. 2 into an oven except that the connected two current sinks $I_2$ are located outside the oven (i.e. the two current sinks $I_2$ will be independent of the ambient temperature in the oven). It can be seen from Fig. 8 that the normalized frequency of the “uncompensated” case decreases inversely with the ambient temperature (in the oven) as can be expected from (9) where effects of temperature caused by the thermal dependent voltage $V_T$ is in the denominator.

The compensated case can be demonstrated by taking Fig. 2 into an oven including the connected two current sinks $I_2$ (i.e. the two current sinks $I_2$ will also be affected by the ambient temperature in the oven). It can be seen from Fig. 8 that the normalized frequency of the “compensated” case remains relatively constant, as can be expected from (9) where effects of temperature caused by $V_T$ in the denominator of (9) can be compensated by the relatively similar effects caused by $V_T$ of $I_2$ in the numerator of (9), i.e. $I_2 \propto v_{BE}$ where $v_{BE} = V_T \ln (I_C/I_S)$. $I_C$ and $I_S$ are the collector and saturation currents of a BJT in LM334.

In the compensated case, the temperature coefficients of the normalized center frequencies decrease drastically. The measured temperature coefficients for ambient temperature ranging from $T_1 = 30 \, ^\circ\text{C}$ to $T_2 = 75 \, ^\circ\text{C}$ are approximately -30 ppm/$^\circ\text{C}$, i.e. $\mathcal{O}[f(T_2) - f(T_1)] \times 10^6/[f(T_1) \times (T_2 - T_1)] = (0.9986 - 1) \times 10^6/[(1)(75 - 30)]$. The measurements have been obtained by putting the two frequency-determining capacitors outside the oven, and the measured temperature coefficients are therefore due to the intrinsic circuit parameters only.
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3.5 Effects of Temperature on the Quality Factor

Effects of temperature on the quality factor have never clearly been reported. In a similar manner to Section 3.4, Fig. 9 shows two cases of the measured variations of the quality factor $Q_{HQ}$ versus the ambient temperature (Celsius), i.e. the uncompensated and the compensated cases. It can be seen from Fig. 9 that $Q_{HQ}$ in the “uncompensated” case increases versus the ambient temperature as can be expected from (10) where $\beta$ is proportional to temperature [20], [25]. In addition, the variation of $Q_{HQ}$ in the “compensated” case is reduced. Such variations of $Q_{HQ}$ are not only gradually but also 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].

The measured temperature coefficients, ranging from $T_1 = 30 \degree C$ to $T_2 = 75 \degree C$, are reduced approximately from 1.010 ppm/$\degree C$ in the “uncompensated” case to 367 ppm/$\degree C$ in the “compensated” case. As mentioned earlier, alternative solutions for good stability of the quality factor $Q_{HQ}$ with temperature should be suggested through the use of special HBTs where good stability of $\beta$ with temperature has been reported [20, 21].

4. COMPARISONS TO OTHER 10.7-MHZ GM-C TECHNIQUES

As mentioned earlier, 10.7-MHz bandpass filters are typically based on switched capacitors (SC) or Gm-C techniques. Table 4 particularly compares various results of the proposed Gm-C technique to those of existing Gm-C approaches. In an attempt to enable fair comparisons, all center frequencies are fully homogenous at 10.7 MHz. For purposes of information, irrelevant results of SC techniques as well as relevant results of Gm-C techniques that are not fully homogenous are also included in Table 4, although some comparisons may be somewhat unfair. It can be observed from Table 4 that the proposed 10.7-MHz technique offers not only the high-Q factor of 121 compared to others between 10 to 55, but also the wide dynamic range of 87.45 dB at 1% $IM_3$ compared to others between 61 to 68 dB. In addition, the total output noise is 5.303 $\mu V_{rms}$ compared to others between 226 to 707 $\mu V_{rms}$.

5. POSSIBLE ON-CHIP HIGH-Q WIDE-DYNAMIC-RANGE BANDPASS FILTER

Preferable requirements for an on-chip integrated bandpass filter include low power consumption, low silicon areas of capacitors, high dynamic ranges and high center frequencies whilst maintaining high quality factors. On the one hand, equation (9) suggests
Table 4: Comparisons of the proposed Gm-C bandpass filter and existing Gm-C approaches Switched-Capacitor (SC) techniques are also included for information.

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<td></td>
</tr>
<tr>
<td>Center freq. (MHz)</td>
<td>10.7</td>
<td>10.7</td>
<td>10.7</td>
<td>10.7</td>
<td>10.7</td>
<td>10.7</td>
<td>10.7</td>
<td>10.7</td>
<td>10.7</td>
<td>10.7</td>
<td>10.7</td>
<td></td>
</tr>
<tr>
<td>Bandwidth (kHz)</td>
<td>88</td>
<td>-</td>
<td>500</td>
<td>267.5</td>
<td>535</td>
<td>300</td>
<td>1070</td>
<td>464</td>
<td>305.7</td>
<td>-</td>
<td>1070</td>
<td>368.9</td>
</tr>
<tr>
<td>Q factors</td>
<td>121</td>
<td>-</td>
<td>21.4</td>
<td>40</td>
<td>20</td>
<td>-</td>
<td>10</td>
<td>35</td>
<td>55</td>
<td>10</td>
<td>29</td>
<td></td>
</tr>
<tr>
<td>Sampling freq. (MHz)</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td>65</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td></td>
</tr>
<tr>
<td>Sensitivity of ω₀</td>
<td>-1 to 1</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td>-</td>
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<td></td>
</tr>
<tr>
<td>Sensitivity of Q</td>
<td>1</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td></td>
</tr>
<tr>
<td>Output noise density ($\mu V/\sqrt{Hz}$)</td>
<td>0.016</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td>1</td>
<td>0.38</td>
<td></td>
<td></td>
</tr>
<tr>
<td>Total Output noise ($\mu V_{rms}$)</td>
<td>5.303</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td>295</td>
<td>-</td>
<td>226</td>
<td>707</td>
<td>240</td>
<td></td>
</tr>
<tr>
<td>Dynamic range @ 1%IM₃ (dB)</td>
<td>87.45</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td>68</td>
<td>-</td>
<td>-</td>
<td>61</td>
<td>58.4</td>
<td>-</td>
<td>68</td>
<td></td>
</tr>
<tr>
<td>@ 3%IM₃ (dB)</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td></td>
</tr>
<tr>
<td>Power consumption $P_C$ (mW)</td>
<td>60</td>
<td>16</td>
<td>6</td>
<td>108</td>
<td>220</td>
<td>-</td>
<td>9</td>
<td>16</td>
<td>23</td>
<td>17</td>
<td></td>
<td></td>
</tr>
</tbody>
</table>

that not only the power consumption ($P_C$) due to $I_2$ but also the silicon areas due to $C_1$, can be simultaneously reduced for the same ratio of (9). On the other hand, equation (11) suggests that the smaller values of the capacitance in the circuit, the smaller value of the dynamic range (DR). As a result, higher dynamic ranges on chip require higher power consumptions and more silicon areas of capacitors.

As an example at the center frequency $f_0 = 10.7$ MHz whilst maintaining the high quality factor $Q_{HQ}$ = 121, Fig. 10 predicts preliminary interpolation of a power consumption $P_C$ and a corresponding dynamic range (DR at 1%IM₃) versus the capacitance $C_1$. It can be seen from Fig. 10 that a higher dynamic range DR = 87.45 dB requires a higher power consumption $P_C = 60$ mW at $C_1 = 150$ pF, whilst a lower DR = 65.7 dB requires a lower $P_C = 0.4$ mW at $C_1 = 1$ pF.

High-frequency performance of the circuit will be limited by the transition frequency ($f_T$) of the transistor. Equation (9) suggests that a higher, more useful, center frequency can be expected using a smaller value of capacitor $C_1$ (e.g. using stray capacitances), a higher value of $I_2$ and a higher $f_T$ (e.g. in the region of several GHz) of better transistors. As a particular example, all transistors in Fig. 2 are modeled by a better transistor BFR90A with higher $f_T$ at 5 GHz [27], $\beta = 120$ and the bias currents $I_1 = I_2 = 1$ mA. Figure 11 shows high-frequency performance of Fig. 2 through the analysis and the SPICE simulations in terms of the center frequency and the quality factor $Q_{HQ}$. In this particular example, $Q_{HQ}$ is maintained relatively high and the upper frequency is limited at approximately 600 MHz at $C_1 = 1$ pF.

6. CONCLUSION
A fully-balanced high-Q, wide-dynamic-range current-tunable Gm-C bandpass filter has been proposed based on two simple components, i.e. the adder
and the low-Q-based bandpass filter. The Q factor is as high as the relatively constant value of a common-emitter current gain ($\beta$). For the first time, a high-Q factor is no longer a function of variables such as a center frequency resulting in not only a great reduction in additional Q-tunable circuits but also a much better sensitivity of the Q factor. Sensitivities of either the Q factor or the center frequency have been constant between -1 to 1 independent of the Q factor or variables. An example has been demonstrated at 10.7 MHz for a high-Q factor of 121, the low noise power of -92.45 dBm, the wide dynamic range of 87.45 dB at 1% $IM_{3}$ and the 3rd-order intermodulation-free dynamic range ($IMFDR_3$) of 74.45 dB. The center frequency has been current tunable over 3 orders of magnitude. The proposed technique has been compared with other 10.7-MHz Gm-C approaches and has offered a potential alternative to a 10.7-MHz high-Q wide-dynamic-range bandpass filter.

7. ACKNOWLEDGMENT

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References


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