A 10.7-MHz Fully Balanced, Q-of-267, 103-dB-Dynamic-Range Current-Tunable Gm-C Bandpass Filter

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 three fully balanced components, i.e. an adder, a low-Q-based bandpass filter and a differential amplifier. The high-Q factor is possible through a tunable bias current. As a simple example at 10.7 MHz, the paper demonstrates the high-Q factor of 267, the low total output noise of 2.089 Vrms, the 3rd-order intermodulation-free dynamic range (IMFDR3) of 82.59 dB and the wide dynamic range of 103 dB at 1% IM3.

Streszczenie. Zaprezentowano stratę prądową filtr zrównoważony 10.7 MHz pasmowy Gm-C. Technologia bazuje na trzech elementach – sumatorze, filtrze pasmowym i wzmacniaczu różnicowym. Osiągnięto dużą dobrość dzięki stworzeniu prądowemu. Przedstawiono przykład filtru i porównano z innymi filtrami. (Strojony filtr pasmowy Gm-C o częstotliwości 10.7 MHz dobroci 267 i dynamicznie 103 dB)

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

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 (f₀) 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) [3-6], 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 μVrms and limited dynamic ranges from 58 to 68 dB.

In this paper, a 10.7-MHz fully balanced, high-Q, wide-dynamic-range current-tunable Gm-C bandpass filter is introduced using three fully balanced devices, i.e. an adder, a low-Q-based bandpass filter (ALQ) and a differential amplifier. The high-Q factor is possible through a tunable bias current. The technique is demonstrated through an example of a fully balanced high-Q current-tunable Gm-C bandpass filter. The center frequency is current tunable over 3 orders of magnitude. Comparisons to other 10.7-MHz Gm-C approaches are also included.

The proposed high-Q wide-dynamic-range bandpass filter

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, a low-Q-based bandpass filter A_LQ(s) and a differential amplifier A_G. The transfer function of the low-Q-based bandpass filter A_LQ(s) can be written as

\[ A_{LQ}(s) = \frac{\omega_o}{Q_{LQ} s} \]

(1)

The pass band gain of (1) is \( A_{LQ} = 1 \) at \( s = j\omega_o \) and \( Q_{LQ} \) is a relatively low-Q factor of \( A_{LQ}(s) \). Consequently, a closed-loop gain \( A_{LO}(s) = V_o/V_{in} \) is given by

\[ A_{LO}(s) = \frac{\omega_o}{Q_{LO} s + \omega_o s} \]

(2)

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

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

(3)

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

\[ Q_{LO} = \frac{Q_{LQ}}{1 - A_D A_o} \]

(4)

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

\[ A_D \rightarrow \frac{1}{A_o} \]

(5)

In practice, the denominator of (4) may be made relatively small, i.e. \( A_G \) is in the proximity of \( 1/(A_D) \), resulting in a relatively high quality factor \( Q_{LO} \).

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 three fully balanced components, i.e. a two-input adder (A_D), a low-Q-based bandpass filter (A_LQ) and a differential amplifier (A_G), using matched npn transistors T1 to T10 and matched pnp transistors T11 and T12. In this case, equation (5) suggests that the gain of the adder \( A_D \geq 1 \). Firstly, the adder \( A_D \) is a modified version of an existing adder [14] and consists of a differential pair (T1, T2), a
common-collector pair (T3, T4) and two current sinks I2. The 1st small-signal input voltage of A0 is vAB between the bases of T1 and T2 (or nodes A and B). A small-signal output voltage of A0 is vEF between the emitters of T3 and T4 (or nodes E and F). Secondly, the low-Q-based bandpass filter ALQ is a modified version of an existing low-Q bandpass filter [15] and consists of a differential pair (T5, T6), two capacitors C1 and 2C1, two current sinks I2 and four diode-connected transistors T7 to T10. A small-signal input voltage of A0 is vEF between the bases of T5 and T6 (or nodes E and F) and is obtained from the output vGH of A0. A small-signal output voltage of A0 is vGH between the emitters of T7 and T8 (or nodes G and H). Finally, the transfer function of the high-Q bandpass filter is AHQ = vO / vin where vO = vAB and vO = vGH. It can be seen from Fig. 2 that the circuit is fully balanced.

Parameters α1, α2, ..., α11 and β12 are the small-signal emitter resistance of transistors T1, T2, ..., T11 and T12, respectively, where (α1 = α2) = Vr / R1, (α3 = α4) = Vr / R2, (α6 = α7) = Vr / R3, (α9 = α10) = Vr / R4. As vO1 is temporary deactivated or connected to an ac ground, i.e. vCD = 0. Therefore, vO1 of A0 enables a small-signal emitter current ie1 = vO1 / 2R1 passing through the emitters of T1 and T2. The resulting small-signal collector current of T1 and T2 is ie1 = vO1 / 2R1. Most of ie1 passes through a loading impedance Z1 = 2R2 formed by T3 and T4. As vO1 = ie1Z1, therefore vO1/vin = 1. Consequently, vO1 = vin. On the other hand, vO2 can be found at vCD = 0. Therefore, vin = 0. The result of the common-collector pair (T3, T4) is vO2 / VCD = 1, or vO2 = VCD. Consequently, vEF = vO1 + vO2, i.e. the gain of the adder A0 = 1. As vin = vO1 and VCD = vO2, therefore

\[ v_{EF} \approx v_{in} + v_{CD} \] (6)

Secondly, the low-Q-based bandpass filter ALQ is considered. The input vEF of ALQ enables a small-signal emitter current ic1 = vEF / 2R1 passing through the emitters of T5 and T6. As vO1 = vEF, therefore ic1 = ie1 = ie2. Therefore, ic1 and ic2 are all the same. Most of ic1 passes through a loading impedance Z2 = 4R2 formed by T7 to T10 where Z2 = 2R2 and therefore ic1 = ic2. The resulting output of ALQ is vGH = ic2R2, therefore ALQ = vGH / vEF = vGH / vEF represents a low-Q-based bandpass filter ALQ of the form

\[ A_{LQ} = \frac{v_{GH}}{v_{EF}} = \frac{2s\alpha}{s + (1 + \alpha)s + \alpha} \] (7)

The quality factor of (7) is Q_{LQ} = (\alpha/2) / (1+\alpha) which is a relatively low value. The center frequency of (7) is \( f_{0LQ} = (\alpha/2)\tau_\alpha \). At s = j\omega_0, the passband gain of (7) is A_{LQ} = 2\alpha / (1 + \alpha) = 1.

Finally, the high-Q bandpass filter AHQ can be considered by substituting vEF in (7) with (6) and substituting vCD in (7) with (8), therefore A_{HQ} = vO / vin = A_{LQ} / (1 - A_{HQ} A_{LQ}), i.e.

\[ A_{HQ} = \frac{v_{in}}{v_{O}} = \frac{2s\alpha}{s + (1 + \alpha - 2\alpha A_{HQ})s + \alpha} \] (9)

The center frequency of (9) is \( f_{0HQ} = (\alpha/2)\tau_\alpha = g_m d/2RC_1 \) where the transconductance \( g_m = \alpha / r_e \). The center frequency \( f_{0HQ} \) is current tunable by I2 of the form

\[ f_{0HQ} = \frac{1}{4C_1Vr}\sqrt{\frac{\beta}{\beta + 1}} \] (10)
The quality factor of (9) is $Q_{HQ} = \left(\frac{\alpha^2}{\gamma^2} + 1 + \alpha - 2\alpha A_0\right)$. As $\alpha \equiv 1$, therefore $Q_{HQ}$ is current tunable by $I_3$ of the form

$$Q_{HQ} \approx \frac{1}{2(1-A_0)} \approx \frac{1}{2(1 - \frac{R/F}{V_T})}$$

(11)

It may be suggested from (11) that the quality factor $Q_{HQ}$ ideally approaches infinite at $I_3 = V_T/R_C$. In practice, however, $Q_{HQ}$ should be current tunable to a relatively large value through $I_3$ where $I_3$ is in the proximity of $V_T/R_C$. As an example, it can be expected from (11) that $Q_{HQ}=267$ if $A_0=0.9982$, $R_C=50 \ \Omega$, $V_T=25 \ \text{mV}$ and $I_3=499 \ \mu\text{A}$. At $s = j\omega_0$, the passband gain of (9) is ideally (i.e. without loading effect and $\alpha=1$) $A_{HQ} = 1 / (1-A_0) = 2Q_{HQ}$ which is much greater than the passband gain of (7) where $A_{LO} = 2\alpha / (1+\alpha) \approx 1$ at $s = j\omega_0$.

**Sensitivity**

Generally, a sensitivity of $y$ to a variation of $x$ is given by

$$S_y = \frac{\partial y}{\partial x} = \frac{\partial y}{x} \frac{x}{\partial x}$$

(12)

As $S_y$ may be written as

$$S_y = \frac{1}{x} S_x$$

(13)

where $S_x$ is the sensitivity of $x$ to a variation of $y$. Table 1 shows the sensitivity $S_y$ where $(x,y) = (C_1, \text{os}Q), (V_T, \text{os}Q), (I_2, \text{os}Q), (\beta, \text{os}Q), (R_C, \text{os}Q), (V_T, \text{os}Q)$ or $(I_3, \text{os}Q)$. It can be seen from Table 1 that the sensitivity of $\text{os}Q$ to the variations of $C_1$, $V_T$, or $I_2$ are desirably independent of parameters. In addition, the sensitivity of $\text{os}Q$ to the variation of $\beta$ is inverse proportional to $\beta$, and is a particularly low value ($<1$) when $\beta$ is large. Therefore, the sensitivity of $Q_{HQ}$ is between $-1$ and $1$. For high-$Q$ realizations ($I_3 \equiv V_T/R_C$), the sensitivities of $Q_{HQ}$ to the variations of $R_C$, $V_T$, or $I_3$ are in the same order as those given in the literature [16,17].

**Table 1. Sensitivities**

<table>
<thead>
<tr>
<th>$S_{\text{os}Q}/C_1$</th>
<th>$S_{\text{os}Q}/V_T$</th>
<th>$S_{\text{os}Q}/I_2$</th>
<th>$S_{\text{os}Q}/\beta$</th>
<th>$S_{\text{os}Q}/R_C$</th>
<th>$S_{\text{os}Q}/V_T$</th>
<th>$S_{\text{os}Q}/I_3$</th>
</tr>
</thead>
<tbody>
<tr>
<td>$-1.0$</td>
<td>$1.0$</td>
<td>$1/(2[1+\beta])$</td>
<td>$-2Q_{HQ}$</td>
<td>$2Q_{HQ}$</td>
<td>$-1.0$</td>
<td>$1.0$</td>
</tr>
</tbody>
</table>

**Dynamic ranges**

Dynamic ranges (DRs) of either a specific biquad or an optimized high-$Q$ biquad in a general way have been presented [18]. An expression for the dynamic range of a second-order Gm-C biquad in a general way is given by [18]:

$$DR = \frac{v_{\text{max}}^2}{v_{\text{noise}}^2} = \frac{2v_{\text{max}}^2}{kT\xi^2 \left(\frac{1}{C_1} + \frac{1}{C_2}\right)}$$

(14)

where $v_{\text{max}}$ is the maximal signal level (at the input or output of a system), $v_{\text{max}}^2$ is the mean squared noise voltage at the same point, $C_1$ and $C_2$ 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 $v_{\text{max}}^2$, but also reducing $v_{\text{noise}}^2$ of (14) as follows.

**Table 2. Dynamic ranges**

<table>
<thead>
<tr>
<th>Refs</th>
<th>Capacitors</th>
<th>$V_{\text{max}}$</th>
<th>$v_{\text{max}}^2$</th>
<th>$v_{\text{noise}}^2$</th>
<th>$DR$</th>
</tr>
</thead>
<tbody>
<tr>
<td>(this paper)</td>
<td>$C_1 = C$, $C_2 = 2C$</td>
<td>$V_{M1}$</td>
<td>$288 \ \text{mV}$</td>
<td>$-5 \ \text{dBm}$ through a $50-\Omega$ load</td>
<td>$105.72 \ \text{dB}$</td>
</tr>
<tr>
<td>(fully balanced)</td>
<td>$C_1 = C$, $C_2 = 2C$</td>
<td>$V_{M2}$</td>
<td>$0.5\text{v}<em>{\text{M}}$, $0.5\text{v}</em>{\text{M}}$</td>
<td>$0.5\text{v}<em>{\text{M}}$, $0.5\text{v}</em>{\text{M}}$</td>
<td>$94.38 \ \text{dB}$</td>
</tr>
<tr>
<td>(single ended)</td>
<td>$C_1 = C$, $C_2 = 2C$</td>
<td>$V_{M3}$</td>
<td>$0.5\text{v}<em>{\text{M}}$, $0.5\text{v}</em>{\text{M}}$</td>
<td>$0.5\text{v}<em>{\text{M}}$, $0.5\text{v}</em>{\text{M}}$</td>
<td>$94.38 \ \text{dB}$</td>
</tr>
</tbody>
</table>

**Experimental Results**

As a simple example, all transistors in Fig. 2 are modeled by a simple transistor 2N2222 and 2N2907 where the average transition frequency ($f_T$) is $120 \ \text{MHz}$ and $\beta$ is approximately $120$ [21]. All current sinks are LM334 [22]. The bias current $I_1=I_2=1.2 \ mA$, $I_3=0.5 \ mA$, $R_C=50 \ \Omega$ and $C_1=150 \ \text{pF}$. Figure 3 illustrates the measured frequency response of Fig. 2 at the center frequencies $f_0 = \frac{1}{2\pi\text{res}} = 10.7 \ \text{MHz}$. It can be seen from Fig. 3 that the bandwidth (BW) is $2\times20 = 40 \ \text{kHz}$ and therefore the measured quality factor $Q_{HQ}$ ($=f_0/BW$) is relatively high at approximately 267.
constant at approximately 267 and is, unlike existing approaches, independent of variables such as a center frequency. When \( I_2 > 1 \text{ mA} \), \( f_0 \) 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.

![Image](attachment:image1.png)

**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 –153.6 dBm/Hz and the resolution bandwidth (RBW) is at 200 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 \( V_{N1} = 0.0046 \mu V_{rms}/\sqrt{Hz} \), the total output noise \( V_{N3} = 2.0893 \mu V_{rms} \) and the total noise power \( P_{N3} = -100.59 \text{ dBm} \).

![Image](attachment:image2.png)

**Wide dynamic range**

The circuit is excited with two sinusoids at frequencies \( f_1 = f_0 - 7.5 \text{ kHz} = 10.6925 \text{ MHz} \) and \( f_2 = f_0 + 7.5 \text{ kHz} = 10.7075 \text{ MHz} \). The 3rd-order intermodulation (IM3) products \( |2f_1-f_2| \) and \( |2f_2-f_1| \) are 10.6775 and 10.7225 MHz, respectively.

![Image](attachment:image3.png)

Figure 6 shows the measured output spectrums at \( Q_{HQ} = 267 \) using the two-frequency excitation of -20 dBm at \( f_1 \) and \( f_2 \). It can be seen that the IM3 products are approximately 40 dB down from the fundamentals and correspond to 1% (or 1% IM3). Through a 50-\( \Omega \) 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 IM3 at \( |2f_1-f_2| \) versus the input levels. It can be seen from Fig. 7 that the noise power \( P_{N3} = -100.59 \text{ dBm} \). At the input level of –45 dBm, the output level of \( f_1 \) is –18 dBm whilst the output level of the IM3 is adjacent to \( P_{N3} \) (or intermodulation free). Therefore the 3rd-order intermodulation-free dynamic range (IMFDR₃) = (–18 dBm) – (–100.59 dBm) = 82.59 dB. In addition, at the input level of –20 dBm, the output level of \( f_1 \) is 2.2 dBm, whilst the output level of the IM3 is 40 dB down from \( f_1 \) (or 1% IM3). Therefore, the wide dynamic range (at 1% IM3) = (2.2 dBm) – (–100.59 dBm) = 102.79 dB which is

**Table 3.** Summaries of related noise parameters obtained from Figure 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>200 kHz</td>
<td></td>
</tr>
<tr>
<td>2 ( P_{N3} )</td>
<td>10 log ( (P_{N3}/1\text{mW}) )</td>
<td>-153.6 dBm/Hz</td>
</tr>
<tr>
<td>2 ( P_{N3} )</td>
<td>-</td>
<td>4.3652×10⁻⁷ ( \text{W/Hz} )</td>
</tr>
<tr>
<td>Noise density</td>
<td>( V_{N2} \times (50 \Omega) )</td>
<td>2.1826×10⁻⁷ ( V_{rms}/\sqrt{Hz} )</td>
</tr>
<tr>
<td>Noise density</td>
<td>( V_{N2} )</td>
<td>4.6718×10⁻⁹ ( V_{rms}/\sqrt{Hz} )</td>
</tr>
</tbody>
</table>

![Image](attachment:image4.png)

**Table 4.** Values used in the calculations.

<table>
<thead>
<tr>
<th>Noise Parameters</th>
<th>Values</th>
</tr>
</thead>
<tbody>
<tr>
<td>( V_{N2} )</td>
<td>3.652×10⁻¹² ( V^2 )</td>
</tr>
<tr>
<td>( V_{N3} )</td>
<td>( \sqrt{V_{N2}} )</td>
</tr>
<tr>
<td>( P_{N3} )</td>
<td>10 log ( \frac{V_{N3}^2}{(50\Omega)(1\text{mW})} )</td>
</tr>
</tbody>
</table>
consistent with the expected value $\text{DR}_1 = 105.71 \, \text{dB}$ predicted in section of dynamic range.

**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 the center frequency $f_0$ 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 (10) where effects of temperature caused by the thermal dependent voltage $V_T$ is in the denominator of (10).

**Fig. 8. Normalized centre frequencies versus ambient temperature for the uncompensated and compensated cases.**

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 (10) where effects of temperature caused by $V_T$ in the numerator of (10) can be compensated by the relatively similar effects caused by $V_T$ of $I_2$ in the numerator of (10), 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 $-29 \, \text{ppm}/^\circ\text{C}$, i.e. $\equiv \frac{[f(T_2) - f(T_1)] \times 10^6}{([f(T_1)] \times (T_2-T_1))} = \left(0.9987-1\right) \times 10^6 / \left(\left([1](75-30)\right)\right)$. 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.

**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 D, 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 (11) where effects of temperature caused by the thermal dependent voltage $V_T$ is in the denominator of (11).

**Fig. 9. The quality factor $Q_{HQ}$ versus ambient temperature for the uncompensated and compensated cases.**

Unlike the two cases in Fig. 8 where the temperature dependent capacitors are located outside the oven, both cases in Fig. 9 have been obtained by including the temperature dependent resistors $2R_c$ inside the oven. It may be observed from both cases in Fig. 9 that the uncompensated effects of the ambient temperature due to the resistor $R_c$ in the numerator of the ratio $R_cI_3/V_T$ in (11) remain evident.

**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 (10) suggests 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 (10). On the other hand, equation (12) suggests that the smaller the values of the capacitance in the circuit, the smaller the 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 $f_0 = 10.7 \, \text{MHz}$ whilst maintaining the high quality factor $Q_{HQ} = 267$, 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 $DR = 103 \, \text{dB}$ requires a higher power consumption $P_C = 90 \, \text{mW}$ at $C_1 = 150 \, \text{pF}$, whilst a lower $DR = 81 \, \text{dB}$ requires a lower $P_C = 0.6 \, \text{mW}$ at $C_1 = 1 \, \text{pF}$.

High-frequency performance of the circuit will be limited by the transition frequency ($f_T$) of the transistor. Equation (10) 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 [23], $\beta = 120$ and the bias currents $I_1 = I_2 = 1 \, \text{mA}$. 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 500 MHz at \( C_1 = 1 \) pF.

Fig. 10. Preliminary interpolation of the power consumption (P\(_C\)) and the dynamic range (DR at 1% IM\(_3\)) versus \( C_1 \) at \( f_0 = 10.7 \) MHz and \( Q_{HQ} = 267 \).

Fig. 11. An example of the center frequencies \( f_0 \) and the quality factor \( Q_{HQ} \) versus capacitance \( C_1 \) with fixed bias currents \( I_1 = I_2 = 1 \) mA, \( I_3 = 500 \) uA.

**Conclusion**

A fully-balanced high-Q, wide-dynamic-range current-tunable Gm-C bandpass filter has been proposed based on three simple components, i.e. the adder, low-Q-based bandpass filter and differential amplifier. The high-Q factor is possible through a tunable bias current. An example has been demonstrated at 10.7 MHz for a high-Q factor of 267, the low noise power of -100.59 dBm, the wide dynamic range of 103 dB at 1% IM\(_3\) and the 3\(^{rd}\)-order intermodulation-free dynamic range (IM\( FDR\)) of 82.59 dB. The center frequency has been current tunable over 3 orders of magnitude. The proposed technique has offered a potential alternative to a 10.7-MHz high-Q wide-dynamic-range bandpass filter.

**REFERENCES**


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