A CMOS CONTINUOUS-TIME ACTIVE BIQUAD FILTER FOR GIGAHERTZ-BAND APPLICATIONS

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ABSTRACT

A general high-Q biquad filter architecture capable of operating in the GHz range is proposed and analyzed. This filter, which is usable in bandpass and lowpass applications, utilizes two effective Q-enhancement techniques to circumvent the low-Q characteristics inherent in the circuit. Simulation results employing standard 0.5μm CMOS technology have successfully verified that the center frequency tuning and hybrid Q-tuning techniques operate between 1GHz to 2.3GHz center frequencies with Q ranging from 4.8 to over 1000. A tunable bandpass filter with a center frequency at 1.5GHz and a lowpass filter with a center frequency at 2.07GHz each having Q equal to 31 have been designed to have 63dB and 44dB input dynamic range, and power dissipation equal to 30.5mW and 27.8mW, respectively.

1. INTRODUCTION

Scaling down the dimensions of the transistors and achieving the maximum integration level by incorporating as many circuits as possible in a chip are inevitable trends for radio frequency (RF) circuit design. Recently the submicron CMOS technology in RF circuit implementation has received intense attention as it would allow integration of other digital signal processing circuits, providing a low cost single-chip solution [1]. In communication applications, gigahertz-band filters with high quality factor (Q) are demanded, but are difficult to produce in standard digital CMOS process. Under such circumstance the filtering is usually provided by discrete components such as ceramic, surface acoustic wave (SAW), or LC filters. A dominant integrated high-speed continuous-time filter implementation, called the Gm-C filter [2], uses an open-loop integrator as a basic building block, whose high impedance output port is terminated with an integrating capacitor. Such filters with frequency range from a few kHz up to low hundred MHz have been reported. However, attempts to reach higher operating frequencies have met difficulties due to excess phase problems [3].

Considerable research has been conducted for years in developing high performance monolithic filters to extend the operating frequency range. One approach is to employ silicon monolithic inductors on chip [4]. The shortcomings of passive monolithic inductors are well known as larger required area and low inductor Q due to resistive loss and capacitive coupling. Using a negative resistor generated by a positive feedback configuration can compensate for the low-Q inductor characteristic [5],[6]. Another approach is to exploit high frequency attributes of the transistors as the filter elements to simulate RLC passive filters [7].

Based on the second approach, this paper describes a general active biquad filter capable of operating in the gigahertz frequency range with reasonably high Q. Emphases are placed on detailed design topologies including transfer functions and hybrid Q-enhancement techniques.

2. BANDPASS FILTER IMPLEMENTATION

Figure 1 shows a proposed biquad bandpass filter. In this figure, transistors M1 and M2 comprise the input amplifier cascode stage. This configuration offers several advantages that include reducing the Miller Effect at the gate terminal of the transistor M1, increasing the isolation between the input stage and the following stage, and improving the circuit stability. The core components of the filter are formed by transistors M3, M6, M7, M8, and M10. This negative feedback configuration constrains the gate of the transistor M3 to be a very low impedance node at low frequencies. When the operating frequency increases, the feedback is reduced and the impedance at this node becomes inductive. The impedance level will drop eventually at higher frequencies due to the complex poles generated by the feedback circuit.

Assuming $g_m >> g_{ds}$ for all transistors and ignoring all non-dominant high-order terms, the bandpass filter transfer function can be expressed as

$$A_{V\_BP}(s) = \frac{V_{out\_BP}(s)}{v_s(s)}$$

$$\approx -\frac{g_{m1}\,(c_2+c_3)}{c}\,\frac{s+D}{s^2+s\,A+B} \quad (1)$$

where

$$\begin{aligned}
A &= \frac{g_{m6}\,(g_{m3}\,c_2+g_{m7}\,c_3)-g_{m3}\,g_{m7}\,c_4}{g_{m6}\,c}\\
B &= \frac{g_{m3}\,g_{m7}}{c}\\
D &= \frac{g_{ds}}{c_2+c_3}\\
c &= c_1\,c_2+c_2\,c_3+c_1\,c_3\\
c_1 &= c_{gd2}+c_{db2}+c_{gs3}+c_{gb3}+c_{sb7}+c_{gd9}+c_{db9}\\
c_2 &= c_{gs7}\\
c_3 &= c_{db6}+c_{gb7}+c_{gd7}+c_{db10}+c_{gd10}\\
c_4 &= c_{db3}+c_{gb4}+c_{gs4}+c_{sb6}+c_q\\
g_{ds} &= \frac{g_{ds8}\,g_{ds10}}{g_{m10}} \,||\, \frac{g_{ds6}\,g_{ds3}\,(g_{ds4}+g_{ds5})}{g_{m4}\,g_{m6}}
\end{aligned}$$

c_1, c_2, and c_3 represent accumulated parasitic capacitances in the circuit, and c_4 represents an extra physical Q-enhancement poly-silicon capacitor c_q in shunt with the parasitic capacitances of the other transistors existing at the drain of the transistor M3. By definition, the center frequency of the bandpass filter is given by

$$\omega_o = \sqrt{\frac{g_{m3}\,g_{m7}}{c}} \quad (2)$$

and the quality factor of the filter is therefore equal to

$$Q = \frac{g_{m6}\,\sqrt{c\,g_{m3}\,g_{m7}}}{g_{m6}\,(g_{m3}\,c_2+g_{m7}\,c_3)-g_{m3}\,g_{m7}\,c_4} \quad (3)$$

Eq. 2 reveals that the center frequency is determined by the transconductance of the transistors M3, M7, and c, the parasitic capacitive effect exhibiting in the circuit. Here the transconductance of the transistor M7 is chosen as a prime candidate to be tuned by varying Vbias3 while Vbias1, Vbias2, and the input DC bias voltage of the transistor M1 remain constant. An observation that can be made from Eq. 3 is that the Q can be independently tuned by varying the capacitance c_4 without altering the center frequency ω_o. Simulations show that Q increases from 2.45 to 980 at the center frequency at 1.5GHz by exclusively including a capacitor c_q equal to 0.6pF, and the frequency deviation is 15.7MHz. The low frequency zero entailed by g_{ds} in Eq. 1 is similar to a passive LC filter when the series resistance of realistic inductor is included, and this zero can be minimized by a cascode stage formed by transistors M8, M10, and another feedback gain stage constructed by transistors M4, M5, and M6.

Further insight can be gained by considering other high frequency parasitic poles and zeros implicitly present in the circuit, but not shown in Eq. 1 due to analytical complexities. However, assuming the core feedback loop is broken, increasing the values of the capacitor c_q and Vbias2 generates more phase shift at the output unity-gain frequency since the parasitic poles and/or zeros tend to move to lower frequencies, thereby decreasing the phase margin of the open-loop circuit and boosting the quality factor of the close-loop circuit [8]. Consequently, the Q-enhancement techniques result from two approaches: physically adding a capacitor c_q and electrically tuning Vbias2. In this paper, we compromise between both approaches by choosing a capacitor value c_q equal to 0.38pF and Vbias2 tuning together to reach a wide Q tuning range at widely spaced center frequencies. Simulations depicted in Figure 2 demonstrate the potentially high Q ($\gg 500$) this bandpass filter can achieve at center frequencies between 940MHz and 2.17GHz by varying Vbias2 and Vbias3. It also is shown in Figure 3 that independent quality factor variations from 4.8 to 1017 at 1.5GHz can be obtained by tuning Vbias2 from 2V to 3.385V, and the deviation of the center frequency is only 15MHz. If the low Q characteristic is required, then the value of c_q must be reduced. Q as low as 1.2 can be obtained in the frequency of interest.

The linearity of a narrow band filter can be determined by performing a two-tone test so that the third-order intermodulation products fall into the passband of the filter without much attenuation by the circuit frequency response. We applied two 0.39mV_{rms} sinusoids at 1490MHz and 1500MHz, and observed the output spectrum with 1% IM_3. The total input-referred noise voltage in the passband of the filter is equal to 0.29μV_{rms}. The dynamic range is therefore 63dB and the power dissipation is equal to 30.5mW. The simulation results are listed in Table 1.

3. LOWPASS FILTER IMPLEMENTATION

The same bandpass filter architecture can be modified to act as a lowpass filter by using a different output termination as illustrated in Figure 4. The characteristic function of the lowpass filter is similar to that of the bandpass filter due to the fact that they both originate from the same closed-loop circuit, but the zeros of the transfer function moves to very high frequency and can be reasonably neglected in this analysis. In this figure, only one transistor M8 is employed at the output port since the cascode stage will generate additional parasitic poles and zeros at high frequencies.

Similarly, assuming $g_m >> g_{ds}$ for all transistors in Figure 4 and by ignoring other non-dominant high-order terms, it can be shown that the transfer function of the lowpass biquad filter is

$$A_{V\_LP}(s) = \frac{V_{out\_LP}(s)}{v_s(s)}$$

$$\approx \frac{g_{m1}\, g_{m3}}{c} \frac{1}{s^2 + s\, A + B} \tag{4}$$

where A, B, and c are the same symbolic expressions as Eq. 1. Accordingly, the center frequency ω_o and the quality factor Q are given by Eq. 2 and Eq. 3, respectively.

Comparing Eq. 4 with Eq. 1, similar conclusions regarding ω_o and Q tunings can be made as follows:

1. The center frequency of the lowpass filter can be tuned electrically by adjusting Vbias3 while maintaining the same DC current through transistors M1, M3, and M4.

2. The Q can be adjusted by physically inserting an extra capacitor c_q and electrically tuning Vbias2.

In the following simulations, Q tuning was achieved with c_q equal to 0.35pF and Vbias2 tuning. By varying both Vbias2 and Vbias3, Q greater than 1000 with center frequencies between 1.26GHz and 2.3GHz are achieved.

A lowpass filter with a center frequency at 2.07GHz and Q equal to 31 has been designed to illustrate the feasibility of this filter architecture. The Q tuning range is between 10 to 1256 by tuning Vbias2 exclusively, and the measured deviation of the center frequency is 25MHz. The Total Harmonic Distortion (THD) is obtained by applying a 10MHz single-tone signal in this broad-band lowpass filter without significantly attenuating other dominant harmonic terms, and 18.4mV_{rms} input signal is the maximum input voltage for 1% THD. The total input-referred passband noise is $0.123mV_{rms}$. Therefore, the input dynamic range is 44dB and the power dissipation is equal to 27.8mW. The overall performance metrics are listed in Table 1.

4. CONCLUSION

A tunable CMOS biquad filter topology for both bandpass and lowpass filtering has been analyzed, designed, and simulated. This simple architecture allows a filter with a center frequency close to transistor f_T and high quality factor possible by considering the distributed high frequency parasitic capacitances and characteristics of transistors as filter elements. Other parasitic poles and zeros along with lower principal complex poles largely contribute to the quality factor in the gigahertz filter design. In the future, temperature sensitivities of center frequencies and quality factor will be further investigated and analyzed.

5. REFERENCES

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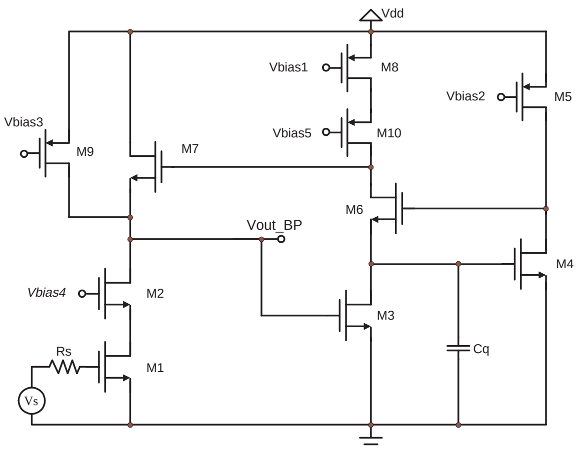

Figure 1: The bandpass filter.

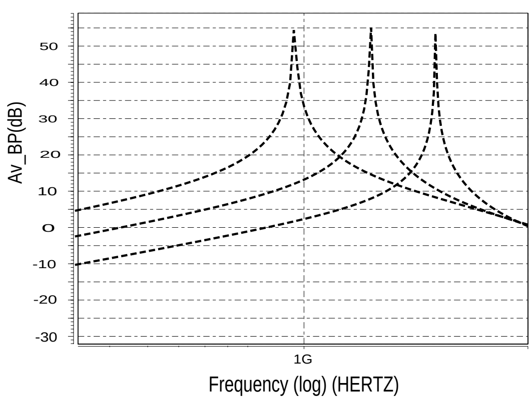

Figure 2: The bandpass filter with Q larger than 500 at 940MHz, 1.5GHz, and 2.17GHz.

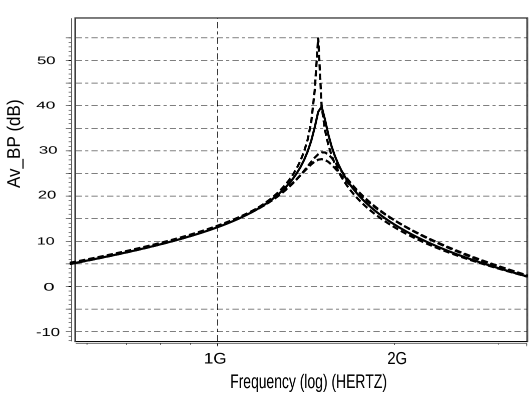

Figure 3: The independent Q tuning of the bandpass filter with a center frequency at 1.5GHz.

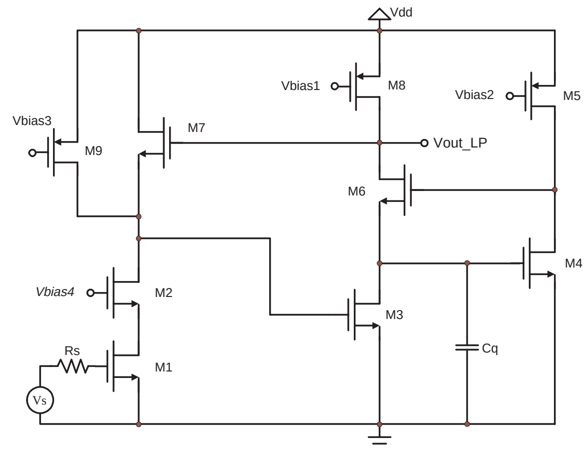

Figure 4: The lowpass filter.

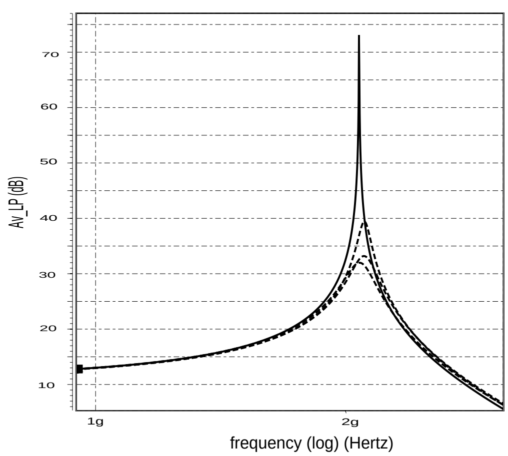

Figure 5: The independent Q tuning of the high-Q lowpass filter with a center frequency at 2.07GHz.

Table 1: The performance metrics of bandpass and lowpass filters with a designated Q=31.

	BP filter	LP filter
Capacitor Cq	0.38pF	0.35pF
Center frequency	1.5GHz	2.07GHz
Q-tuning range	4.8-1017	10-1256
Maximum input swing for 1% IM_3 in BP for 1% THD in LP	0.39mV_{rms}	18.4mV_{rms}
Total passband noise	0.29μV_{rms}	0.123mV_{rms}
Dynamic range	63dB	44dB
Power dissipation	30.5mW	27.8mW