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MIXED SIGNAL LETTER

A frequency-translation technique for low-noise ultra-low-cutoff lowpass filtering

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Abstract Presented is a frequency-translation technique for compact realization of a low-noise lowpass filter (LPF) for biopotential acquisition systems. It is by chopper-stabilizing a bandpass filter (BPF) to obtain an ultra-low-cutoff lowpass response. This technique not only removes the BPF's flicker noise and dc-offset, but also adds clock-based gain-bandwidth tunability and saves chip area because of highly relaxed time constants. A 1.4 to 15-Hz 2nd-order OTA-C ladder LPF designed in a 90-nm CMOS process verifies the merits of the technique with respect to the prior art.

Keywords Lowpass filter · Bandpass filter · Frequency translation · CMOS

1 Introduction

A low-noise ultra-low-cutoff lowpass filter (LPF) is a crucial building block for portable biomedical systems. The amplitudes of the biopotential signals are in the order of tens of μV to tens of mV and the frequency span from DC to a few kHz. Among such a low frequency range, an ultra-low-cutoff LPF cannot be designed in a simple manner as the fabrication cost of the chip will be increased when large time constants are required for integrated circuit implementation. The state-of-the-art [1–3] reduces the silicon area by applying different circuit structures for achieving an ultra-low transconductance. Yet, lowering the transconductance leads to substantial noise degradation. Further, due to the low-frequency characteristic and μV level of bio-potential signals, the 1/*f* noise of the measuring devices must be concerned.

In order to acquire, area-efficiently, the weak biopotential signals using an ultra-low-cutoff LPF while achieving low passband noise, a frequency-translation technique is introduced. It reuses the chopper stabilization [4], originated for flicker noise and dc-offset removals, to convert a high-Q bandpass filter (BPF) into a LPF with an ultra-low-cutoff. The relaxed time constants translate into significant area savings because of smaller capacitor sizes. The design example is based on a 2nd-order operational transconductance amplifier–capacitor (OTA-C) ladder BPF. The OTA is realized as a Nauta cell [5] that involves only CMOS inverters, being very suitable for realizing a high-Q BPF with small power and area.

2 Proposed LPF

Figure 1 shows the operating principle of the proposed LPF, which consists of a BPF with input and output choppers. A tunable clock generator offers the modulation signals to both choppers. The input chopper modulator (M_{IN}) will, first, frequency-translate the input biopotential signal and the corresponding contaminating signals from spectrum S_1 to S_2. The chopper frequency (f_{chop}) should be much larger than the 1/*f* noise corner frequency of the BPF, to minimize the noise contribution of the choppers to the total output noise power spectral density (PSD). Thus, f_{chop} is fixed in a range of few kHz. As depicted in spectrum S_2, the input signal of the IA is split into the upper and lower

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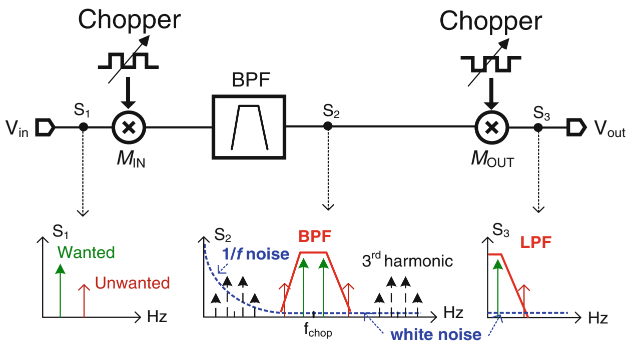

Fig. 1 Operating principle of the proposed LPF using a chopper-stabilized BPF

sidebands. The BPF with a center frequency equals to f_{chop} will select the desired signal, while rejecting the unwanted interferers, noise and odd-harmonic components at the high frequency bands. As a result, the time constant for the implementation of the BPF can be much relaxed, and the 1/*f* noise of the BPF can be suppressed concurrently. Finally, the output chopper modulator (M_{OUT}) will frequency-translate the wanted signal to the passband (spectrum S_3).

3 Circuit implementation

The involved BPF uses a ladder structure for its insusceptibility to component variation, especially in their passband. Although a typical ladder BPF can be easily deducted with the help of the filter handbook, it requires large grounding capacitors or inductors that are not easy to be realized by active devices. Here, a modified 2nd-order *RLC* ladder filter topology is proposed as shown in Fig. 2. It is customized from a 3rd-order Butterworth bandpass filter by removing the central grounding *LC* circuit. It shows that the bandpass response will have weaker stopband attenuation than the typical one as the capacitance value of C_1 and C_2 are fixed to a very small value of 100 fF. As discussed in the previous section, f_{chop} should be much higher (i.e., 4 kHz) than the 1/*f* noise corner frequency (i.e., ~400 Hz) of the IA for noise minimization. According to the *LC* resonant equation, the center frequency (f_{center}) of the bandpass filter equals to $1/2\pi(LC)^{1/2}$, the inductance value can be calculated to be 15.83 kH, where $f_{center} = f_{chop}$ is set for fulfilling the frequency-translation condition.

Figure 3 shows the actual implementation of the complete LPF based on the OTA-C ladder structure. Thanks to the Nauta cell [5] no common-mode feedback circuit is required. The overall CLF circuit consists of two grounding resistors (g_{m0} and g_{m6}) for realizing R_1 and R_2, four series capacitors (C_1 and C_2), and two gyrators (**a**) and (**b**) which are exploited to implement equivalently the inductors, L_1 and L_2, respectively. The required capacitor sizes are highly reduced as listed in Fig. 3.

Considering the front stage of the LPF before the buffer, the voltage transfer function (V_1/V_{in}) is given by,

$$\frac{V_1}{V_{in}} = \frac{j\omega C_1/g_{mOTA}}{1+j\omega C_1/g_{mOTA}-\omega^2 C_1 C_{L1}/g_{mOTA}^2} = \frac{j\omega/Q_F\omega_0}{1+j\omega/Q_F\omega_0-\omega^2/\omega_0^2} \qquad (1)$$

by choosing the OTA's transconductance $g_{mOTA} = g_{m1} = g_{m2}$, we obtain the filter parameters,

$$Q_F = \sqrt{\frac{C_{L1}}{C_1}} \text{ and } \omega_0 = \frac{g_{mOTA}}{\sqrt{C_1 C_{L1}}} \qquad (2)$$

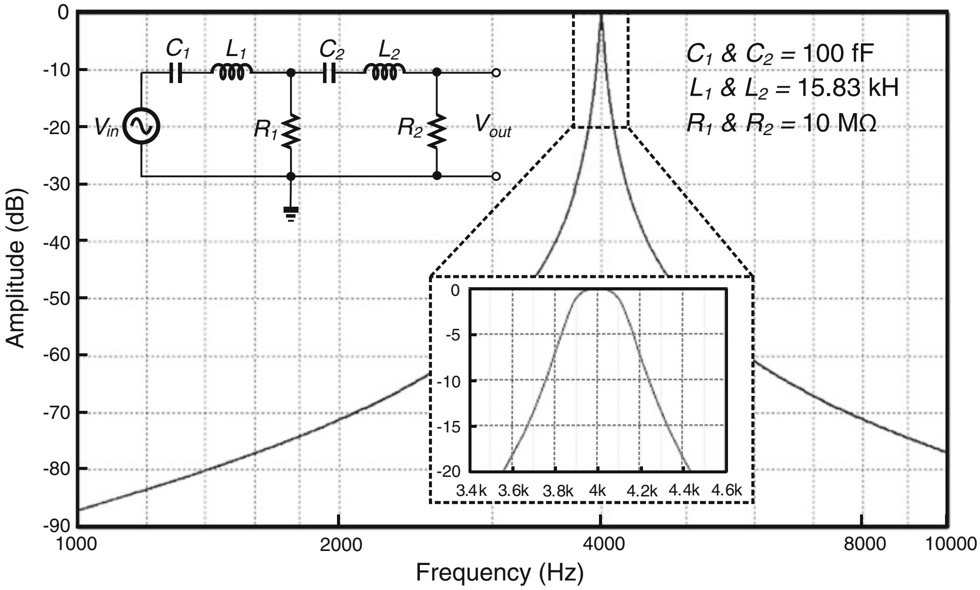

Fig. 2 Modified 2nd-order BPF prototype and its frequency response with a center frequency of 4 kHz

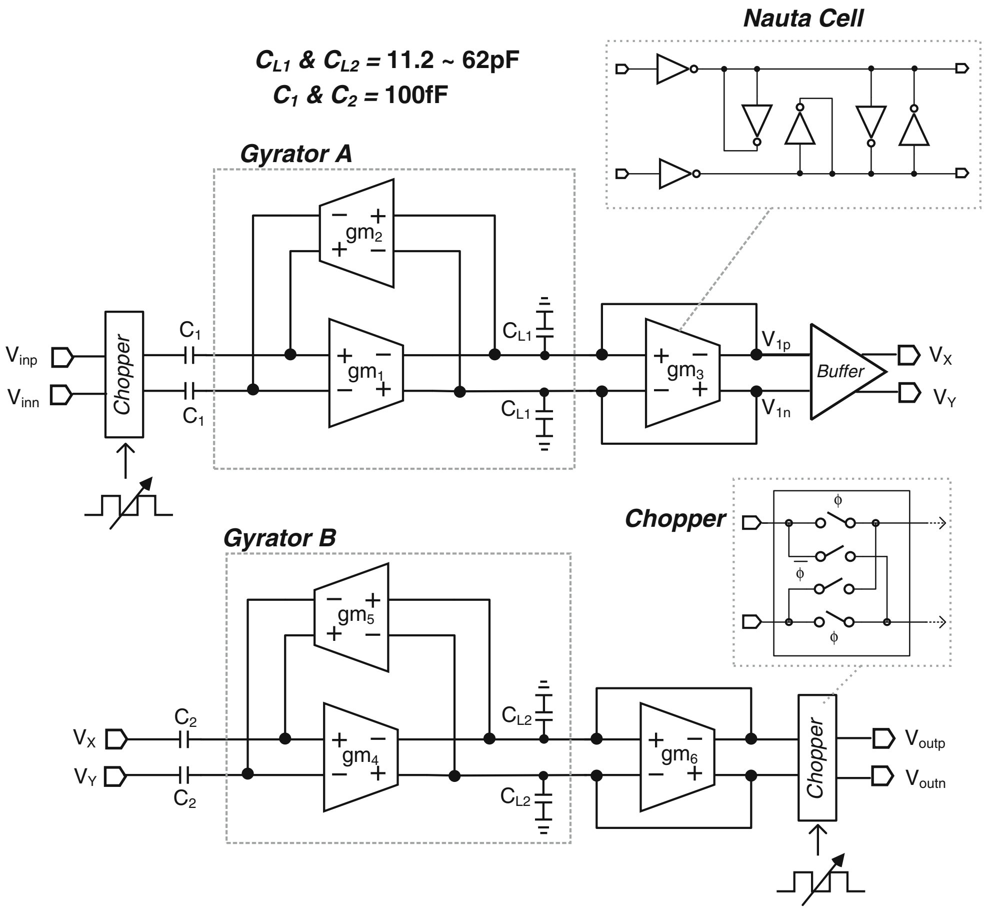

Fig. 3 Actual implementation of the 2nd-order OTA-C LPF

where Q_F is the filter's quality factor and $f_{\mathrm{center}} = 2\pi\omega_0$. Moreover, a buffer is added to separate the front and back stages of the LPF, thus the loading effect to the gain of the front stage can be neglected and the transfer function of the 2nd-order bandpass filter ($V_{\mathrm{out}}/V_{\mathrm{in}}$) can simply be derived as the square product of Eq. (1).

Referring to Eq. (2) the bandwidth (f_{passband}) of the bandpass filter can be given by,

$$f_{\mathrm{passband}} \approx \frac{\omega_0}{2\pi Q} = \frac{g_{\mathrm{mOTA}}}{2\pi C_{\mathrm{L1}}} \tag{3}$$

Equations (2) and (3) hint the way to control f_{passband} without affecting the Q_F is by changing g_{mOTA}. This operation is equivalent to control the lowpass cutoff ($f_{\mathrm{cut\text{-}off}}$) of the LPF. However, since the input CM voltage of the OTA is fixed and the increment of g_{mP} and g_{mN} call for more power, adjusting g_{mOTA} to control f_{passband} may not be that efficient. Thus, the cutoff tuning is achieved by changing the size of the gyrator's loading capacitors (C_{L1} and C_{L2}). Changing C_{L1} and C_{L2} will also alter f_{center} although the effect is less severe than using g_{mOTA}. As a result, the LPF should include a variable clock signal generator to track f_{center} with f_{chop}.

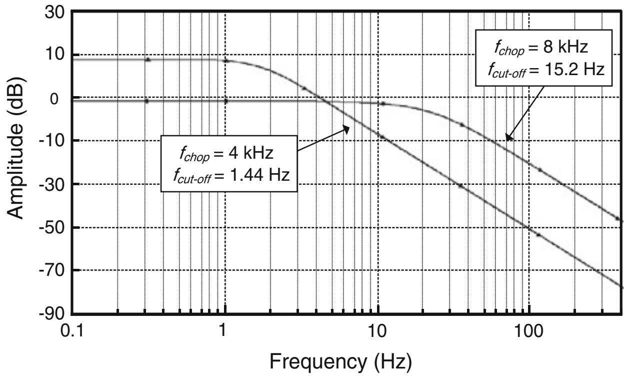

Fig. 4 Frequency responses of the proposed LPF after demodulation with different f_{chop} and $f_{\mathrm{cut\text{-}off}}$

4 Simulation results

The design and verification are based on a 90-nm CMOS process. Figure 4 shows the simulated frequency responses under two different sizes of C_L and f_{chop}, showing the flexibility of both bandwidth and gain adjustments. The cutoff is not exactly half of the f_{passband}, as the bandpass

Table 1 Performance benchmarks

	This work	[1]	[2]	[7]
Technology	90 nm CMOS (thick-oxide MOS)	0.35 μm Bipolar-CMOS-DMOS	0.8 μm CMOS	0.35 μm CMOS
Filter order	2nd-order	2nd-order	6th-order	5th-order
Supply voltage	3 V	3.3 V	±1.5 V	3 V
Supply current	8.37 μA	50–500 μA	3.33 μA	9.3 μA
Cut-off frequency	1.44–15.2 Hz	1.5–15 Hz	2.4 Hz	2.4 Hz to 10 kHz
Total output noise amplitude (peak-to-peak)	88.28 $\mu V/\sqrt{Hz}$	900 $\mu V/\sqrt{Hz}$	50 μV	48 $\mu V\sqrt{Hz}$ @ 2.4 Hz
DC gain	−1 to 8 dB	0 dB	−10 dB	−6 dB
Dynamic range	49.5 dB	60 dB	60 dB	68 dB

response is not the ideal shaping and the demodulated residual odd harmonic components may also affect the final lowpass response. However, the result confirms that a proper lowpass response with 40 dB/decade stopband attenuation can be achieved, providing that the BPF's quality factor is higher than ten, which is reasonable to achieve in practice at such a frequency range. The maximum tuning range of the lowpass cutoff is around 15 Hz.

Comparing with the prior works [1, 2, 7] in Table 1, the proposed solution attains tunable and ultra-low cutoffs with small power and output noise due to the techniques of frequency-translation and a more reasonable transconductance value based on the Nauta cell OTA. On the other hand, the stopband attenuation in this work is fixed to 40 dB/decade without involving a resonant zero to limit the attenuation at high frequency as in [1]. Nevertheless, this work still shows less dynamic range than [1] due to the existence of chopper spikes. The dynamic range can be improved by adding a spike filter after the LPF [6].

5 Conclusion

A frequency-translation technique for low-noise ultra-low-cutoff lowpass filtering is presented. Taking advantages from the chopper stabilization, which has been the technique of flicker noise and dc-offset reduction, the time constant for realization an ultra-low-cutoff LPF can also be significantly relaxed. These concurrent noise and area reduction features particular suit biopotential acquisition systems.

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