A Clockless Method of Flicker Noise Suppression in Continuous-Time Acquisition of Biosignals

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***Abstract*—This paper presents a novel chopping method allowing suppression of 1/f flicker noise in continuous-time acquisition systems without the need for a fixed-frequency clock, stochastically deriving the chopping signal from the input and hence achieving completely signal-dependent power consumption. The method is analysed, its basis of operation explained and a proof-of-concept implementation presented alongside simulated results demonstrating an increase in achieved SNR of more than 8 dB during acquisition of ECG, EAP and EEG signals.**

I. Introduction

Recent years have seen a significant amount of research interest in electronic devices relying on processing and acquisition of biological signals such as electrocardiographs (ECG), electroencephalographs (EEG), local field potentials (LFPs) or extracellular action potentials (EAPs) for purposes of scientific research or construction of medical devices. A common property of the acquired signals is their small frequency - While EAPs can occupy a bandwidth as large as 10 kHz, LFPs occupy a bandwidth smaller than 300 Hz [1]. Similarly, EEG and ECG signals are considered to be of frequencies smaller than 100-150 Hz [2]. Those signals are thus typically affected by the presence of 1/f flicker noise inherent to operation of CMOS circuits. To suppress its effects, current systems presented in literature typically employ either chopper-stabilised amplifiers [2], [3] up-modulating the input signal to higher frequencies or auto-zeroing techniques [4] exploiting autocorrelation properties of 1/f noise [5]. Those techniques however typically rely on periodic switching action requiring a fixed-frequency clock generator.

The stringent requirements imposed on power consumption of many biomedical devices (e.g. 80 mW/cm^2 maximal heat dissipation in neural implants [6]) are however leading to emergence of a new class of acquisition frontends such as [7] called continuous-time (CT) (or level-crossing) analogue-to-digital converters (ADCs) operating without a fixed clock. Rather than obtaining samples of the input signal at a fixed rate as is typical in conventional discrete-time (DT) ADCs, CT ADCs record crossing time of subsequent quantisation levels thus alleviating the need for a fixed-frequency clock generator and paving a way to activity-dependent power consumption achieved by their sampling rate being correlated with instantaneous spectral content of the input signal [7]. This allows exploitation of the typical sparse nature of some biological signals (especially EAPs and ECGs) containing long periods of small information content interrupted by sporadic spikes of rich spectral content.

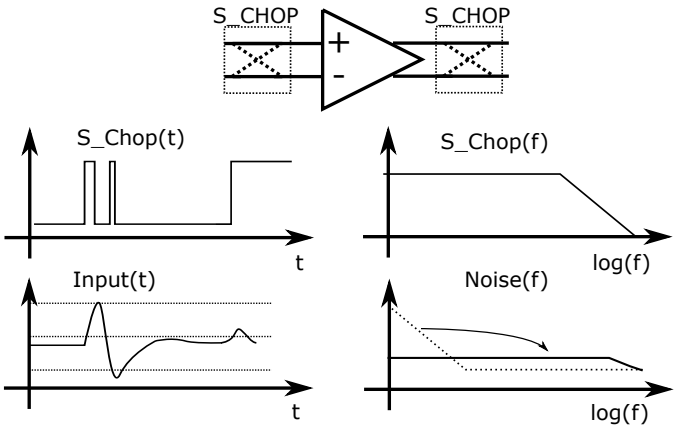


Fig. 1. Illustration of wideband input-dependent chopping signal and achieved flicker noise shaping.

While it is in theory possible to use conventional autozeroing or chopper stabilisation methods in CT ADCs, it is undesirable due to the requirement of operating a fixed-frequency clock leading to input-independent power consumption wasting energy during quiet periods of the input signal. As an alternative we are therefore presenting a novel method aimed at use in CT ADCs termed sparse chopping that leads to a reduction of flicker noise while permitting a completely input-dependent power consumption. This is achieved by replacing a fixed-frequency chopper by an input-derived wideband chopping signal that stochastically changes its state whenever the input signal crosses quantisation levels as demonstrated in Fig. 1.

This paper is organised as follows: section II explores behaviour of signal-derived chopping signals; Section III describes and quantises the achieved noise shaping; Section IV presents a proof-of-concept implementation; Section VI concludes this work demonstrating functionality of the scheme on sample biological signals.

II. Signal-Dependent Choppers

Conventional chopping techniques [5] rely on multiplication of the input signal s(t) by a periodic rectangular signal c(t) up-modulating the input above the system's 1/f noise corner f_c thus allowing its processing without flicker noise. Another

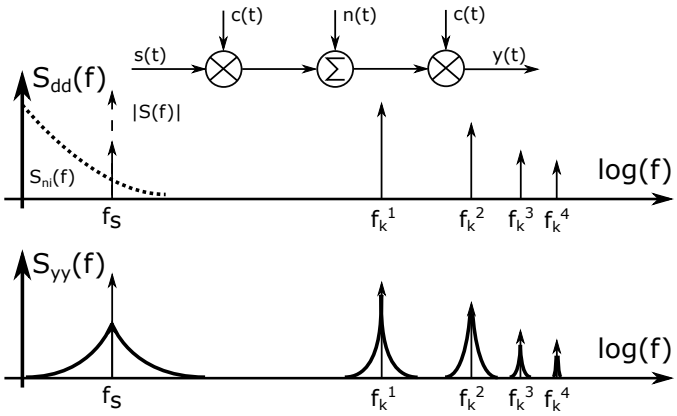


Fig. 2. Demonstration of flicker noise suppression using chopper d(t) (top) changing states whenever the input signal crosses quantisation levels. s(t) - input tone, n(t) added flicker noise, y(t) - output

multiplication of the output by the same signal c(t) leads to demodulation and recovery of the original while at the same time up-modulating the added flicker noise out of band of the input signal. As an alternative to a fixed-frequency oscillator, clock signals derived from the input signal can be employed.

A. Deterministic Chopping

Let us therefore consider a chopping signal $d(t)$ that similarly to $c(t)$ oscillates between values +1 and -1 but only inverts its state whenever the input $s(t)$ crosses the threshold between one quantisation level and another. As demonstrated by Fig. 2 such a signal contains a significant amount of spectral content occupying the same bandwidth as the original signal and its copies centered around $f_k^l = f_s \times N \times l$ where f_s is the center-frequency of $s(t)$, N the oversampling ratio governed by the amount of quantisation levels as described in [7] and l an arbitrary integer. Due to $d(t)$ having spectral content occupying frequencies of the input signal, some flicker noise is modulated on top of the input which is undesirable.

Since $d(t)^2 = 1$ we know that the input signal passes through the chopper unchanged while the output noise is multiplied by $d(t)$. This leads to the output noise power spectral density (PSD) $S_{no}(f)$ being a convolution of the input noise PSD $S_{ni}(f)$ with $S_{cc}(f)$, the PSD of the chopping signal such that $S_{no}(f) = S_{ni}(f) * S_{cc}(f)$. Since the power of $d(t)$, P_d, must be equal to 1, the total noise power at the output must be equal to the input and the noise is thus spread across a wider frequency range.

Assuming that the oversampling ratio is large enough such that the flicker noise modulated to f_k^l does not fall within the bandwidth of the input signal, the amount of flicker noise at the output is decreased by a factor of R which is at least

$$R \geq \frac{\int_0^{f_{BW}} |S_{dd}(f)|^2 df}{\int_{f_{BW}}^{\infty} |S_{dd}(f)|^2 df} \quad (1)$$

The precise number is difficult to estimate as the spectrum of d(t) is dependent on stochastic properties of the input signal and probability of its quantisation level crossings which is a problem that still remains largely unsolved [8].

B. Stochastic Chopping

As it is necessary to ensure that the chopping signal has as large a part of its spectral content at high frequencies as possible, it is desirable to explore ways of modifying the chopper $d(t)$ such that this is achieved. Let us therefore consider an alternative chopping signal that only takes values of +1 and -1 and whenever the input signal crosses a quantisation threshold either holds its current value or changes it with an equal probability.

Assuming that the random process $r(t)$ deciding whether the value of the chopper is +1 or -1 has a white spectrum and therefore its autocorrelation is $R_{rr}(\tau) = \delta(\tau)$, it can be shown that the autocorrelation of a chopper $g(t)$, $R_{gg}(\tau)$ is equal to the probability of samples $g(t)$ and $g(t+\tau)$ occurring between two consecutive quantisation level changes and hence being equal.

Assuming the simple case of fixed time period T_s between taken samples, as would be the case in discrete-time sampling, the resulting chopper signal $g(t)$ is of a purely triangular autocorrelation $R_{gg}(\tau) = \Lambda(\tau/T_s)$. By the Wiener-Khinchin theorem it is thus implied that the spectrum $S_{gg}(f)$ of this chopper has to be $S_{gg}(f) = T_s \times \mathrm{sinc}^2(f \times T_s)$. As illustrated by Fig. 3, it could therefore be said that $S_{gg}(f)$ is approximately white for small frequencies, completely vanishes at multiples of average sampling rate $f_k 1$ and its lobes decay at a rate of 20 dB/dec. Its 3-dB point, f_c^{gg} can be found by solving $\mathrm{sinc}^2(f_c^{gg} \times T_s) = 1/2$ giving rise to $f_c^{gg} \approx 0.44/T_s = 0.44 f_s$

A typical CT sampling process however does not lead to fixed T_s and the time between samples depends on instantaneous spectral activity in the acquired signal [7]. The resulting chopper signal $q(t)$ is therefore going to have different properties from $g(t)$. As a realistic CT sampling process has a limited maximal amount of samples that can be taken in a period of time and hence has a minimal time T_s^{min} between samples, the autocorrelation of $q(t)$, $R_{qq}(\tau)$ must be linear and equal to $R_{gg}(\tau)$ for $0 \leq \tau \leq T_s^{min}$. Its shape for $\tau > T_s^{min}$ however depends on stochastic properties of the input signal and probability of period between samples. More precisely it is equal to probability of the input signal s(t) not crossing quantisation levels for a period τ.

By performing empirical experiments on samples of prerecorded ECG, EEG and neural (LFP+EAP) signals, such a probability was found to be approximately exponentially decaying for $\tau > T_s^{min}$ thus implying that $R_{gg}(\tau)$ also shows exponential decay for $\tau > T_s^{min}$. As illustrated in Fig. 3 this therefore leads to S_{qq} behaving similarly to a white noise process filtered by a single-pole low-pass filter. The position of the single pole is approximately the same as that of the -3-dB point of $S_{gg}(f)$ and can thus be approximated as $f_c^{qq} \approx 0.44/T_s$ where T_s is the average sampling period.

III. Flicker Noise Shaping

The establishment of the shape of the PSD $R_{gg}(t)$ of a stochastic chopper $g(t)$ obtained from a typical biological signal allows further investigation of its effects on flicker noise shaping. It should be noted that using a wideband stochasting chopping signal has previously been reported in [9] for a

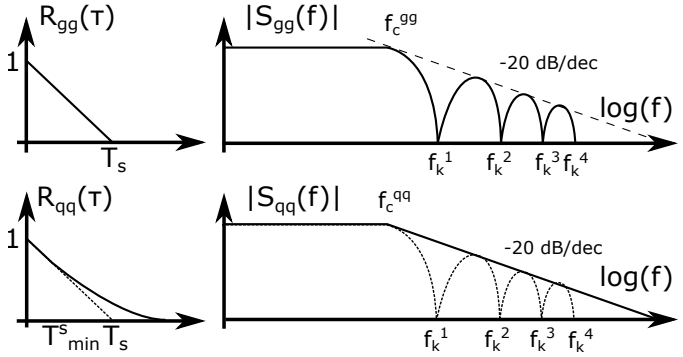


Fig. 3. Autocorrelation (left) and Power Spectral Density (PSD) (right) of a stochastic chopping signal obtained in discrete-time (top) and continuous-time (bottom) sampling process.

conventional discrete-time delta-sigma ADC as a method for suppression of inter-modulation tones commonly present in systems relying on conventional chopping schemes.

The chopper PSD can be estimated as

$$S_{qq}(f) = \frac{k_s}{f^2 + (f_c^{qq})^2} = \frac{f_c^{qq}/\pi}{f^2 + (f_c^{qq})^2} \tag{2}$$

where $k_s = f_c^{qq}/\pi$ is a constant chosen such that the power of $q(t)$ is equal to 1. The output noise PSD $S_{no}(f)$ is equal to $S_{no}(f) = S_{ni}(f) * S_{qq}(f)$ and thus, as illustrated in Fig. 4, the main effect of chopper $q(t)$ is spreading the flicker noise power over a larger range of frequencies ($\approx 0 - f_c^{qq}\ Hz$) while also decorrelating its samples and whitening its spectrum. This adds the flicker noise power to the thermal noise floor and hence leads to its increase while allowing a significant reduction of noise power in small-frequency signals.

Assuming integrated flicker noise power is P_f, the output noise PSD can be expressed as

$$S_{no} = \frac{P_f}{f_c^{qq}} + N_t \approx \frac{P_f}{0.44 \times N \times f_{BW}} + N_t \tag{3}$$

where N_t is the thermal noise floor, N the oversampling ratio and f_{BW} the bandwidth of the input signal. As explained in [7] the oversampling ratio in a CT sampling process is tightly coupled to the acquisition resolution and can be expressed as $N = 2^B$ where B is the ADC resolution in bits. Its increase by one bit therefore leads to a decrease of the up-modulated flicker noise floor by $-3\ dB$.

Once chopping is applied to the input signal, its spectrum is convoluted with the spectrum of the chopper and hence becomes scrambled, spread to a frequency range between $\approx 0 - f_c^{qq}\ Hz$ and whitened. The original is however fully recovered once chopping is applied again at the output.

IV. Circuit Implementation

An example practical implementation of the proposed chopping scheme based on circuit presented in [7] is illustrated in Fig. 5. Potential of the last quantised sample is held at capacitor C_B which is compared to the current input using a comparator. Once the calculated difference exceeds a certain threshold V_{thr2} or V_{thr1} determined by inverter transistor sizes a pulse is generated acting as an output of the circuit indicating crossing of a quantisation level as well as a trigger of a current

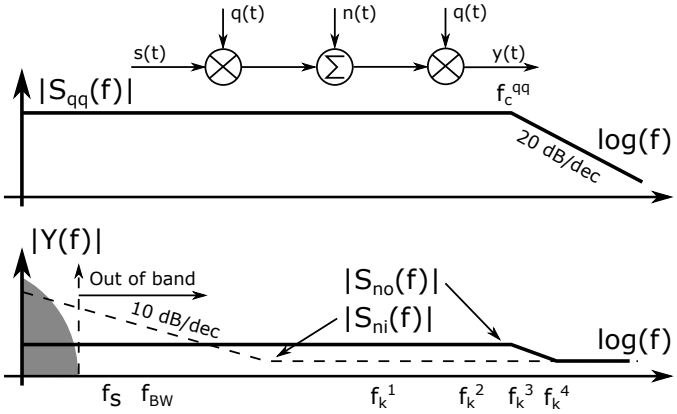


Fig. 4. Demonstration of flicker noise suppression using stochastic chopper q(t). Top - PSD of q(t). Bottom - PSD of input noise $S_{ni}(f)$ (dashed) and output noise $S_{no}(f)$ (solid).

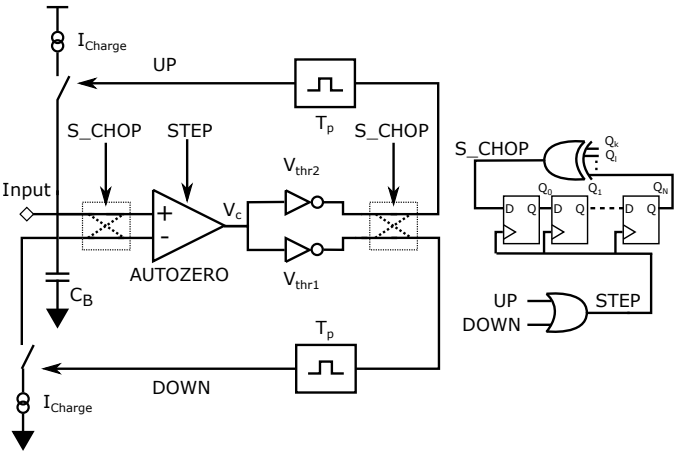


Fig. 5. Example implementation of a CT-ADC employing clockless chopping.

source adding or removing charge packet from capacitor C_B bringing it back to the potential of the input and returning the circuit back to its steady state.

Chopping is applied before the comparator such that the input signal is spread before any flicker noise is added. The signal then passes through inverters adding additional gain and acting as threshold crossing detectors. The effect of chopping at the input to the comparator is equal to switching polarity of its inputs and output signals `UP` and `DOWN` indicating respectively whether a higher or lower quantisation level was entered. The reconstruction (demodulation) chopping circuit is therefore operating completely in digital domain and can be implemented using digital gates such that

$$UP = S_CHOP \cdot A + \overline{S_CHOP} \cdot B \tag{4}$$

$$DOWN = S_CHOP \cdot B + \overline{S_CHOP} \cdot A \tag{5}$$

A. Comparator

In order to ensure efficient operation of the circuit it is critical to make sure that the used comparator has as small input offset as possible. While, similarly to traditional chopping schemes, stochastic chopping upmodulates any DC offset to out-of-band frequencies, doing so leads to generation of additional output pulses detrimental to the overall power consumption of the circuit.

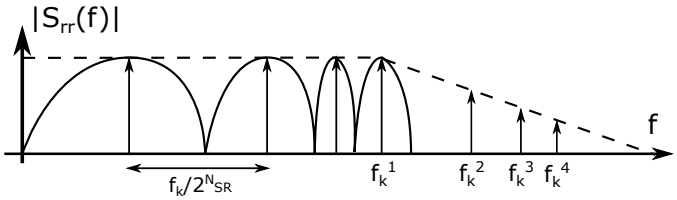


Fig. 6. Spectrum of a chopper obtained using an LFSR-generated pseudo-random m-sequence of length $2^{N_{SR}}$.

This can be solved by employing an autozeroing comparator. To save power this can be constructed such that the offset is sampled only when an output pulse is generated hence obtaining activity-dependent autozeroing. This unfortunately comes at a price of increasing the bandwidth of the input comparator and hence its power consumption. If the circuit is designed for a minimal time between samples T_s^{min}, the comparator's output must be able to settle twice per period - once during the offset sampling phase and once during the amplification phase. The required bandwidth is thus doubled.

B. Pseudo-Random Sequence Generator

As the operation of the chopper is dependent on obtaining a random bitstream it is necessary to employ a circuit that allows its generation. While one possibility is using a noise-based random bit generator such as described in [10], this requires use of circuits dissipating static power consumption which is not desirable.

An alternative is using a linear feedback shift register (LFSR) as illustrated in Fig. 5. This has an advantage of having purely dynamic power consumption. Since it is desirable to achieve minimal correlation between chopper samples, the LFSR should be configured to generate a maximal length code (m-sequence).

Using N_{SR} shift registers generates a code which has an autocorrelation of 1 for multiples of $2^{N_{SR}}$ samples and 0 for all other delays. Thus the autocorrelation of the generated time signal $m(s)$ is $R_{mm}(\tau) = \sum_{l=-\infty}^{\infty} \delta(\tau - l \times 2^{N_{SR}}/f_k)$ and its spectrum $S_{mm}(f) = \sum_{l=-\infty}^{\infty} \delta(f - l \times f_k/2^{N_{SR}})$ - a train of pulses separated by a frequency of $\Delta f = f_k/2^{N_{SR}}$.

Since the spectrum of the newly created chopper $r(t)$, $S_{rr}(t)$ is a convolution of the earlier introduced S_{cc} and S_{mm}, as illustrated in Fig. 6, it is composed of images of the input signal dominated by images derived from f_k^1. Since those images are separated by $\Delta f = f_k/2^{N_{SR}}$ and it is desirable to obtain a near-white spectrum of $S_{rr}(f)$ it should be ensured that $\Delta f = 2f_{BW}$ where f_{BW} is the input signal bandwidth. However, since $f_k = N \times f_{BW} = 2^B \times f_{BW}$ we obtain $N_{SR} = B$ and therefore the LFSR length should be chosen to match the bit resolution of the ADC.

V. Simulated Results

To assess the functionality and performance improvement of the proposed chooping scheme, its functional model was implemented in Matlab and Verilog-A. The amount of noise at the output was then quantised using a signal to noise ratio (SNR) defined as $SNR = E(s_q(t))/E(s_q(t) - s(t))$ where $E(\cdot)$ is the energy operator and $s(t)$ and $s_q(t)$ the input and quantised signal respectively. The CT-ADC was configured to quantise the input signal with an 8-bit precision and the LFSR composed of 8 registers was used in accordance with findings of subsection IV-B. The comparator was modeled such that it added a combination of white and flicker noise of f_c=1 kHz of power leading to SNR of ≈6 dB without a chopping scheme employed. To remove noise upmodulated to higher frequencies, the output was filtered using a sharp cut-off digital low-pass filter of a corner frequency of 10 kHz. Once the chopper was enabled, an increase in SNR by ≈7-8 dB was observed during simulated acquisition of prerecorded LFP+EAP, ECG as well as EEG signals. ECG and EEG samples were obtained from [11] and [12] using [13].

VI. Conclusion

A novel method allowing suppression of flicker noise in CT-ADCs while preserving fully signal-dependent power consumption was presented. To the best knowledge of the authors this is currently the only method presented in literature to do so. Simulated results show a typical achieved SNR improvement of the order of 10 dB during acquisition of typical biological signals.

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