An efficient chopper amplifier, using a switched Gm-C filter technique

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ABSTRACT
In this paper, a chopper amplifier built with G_m-C filters that are alternatively turned on and off will be introduced. First, a theoretical framework to estimate the output of a generic switched continuous-time filter in the frequency domain is presented. This study is later employed to determinate the behaviour of the chopper amplifier: transfer function, and noise. Final conclusions and preliminary measurements are presented at the end. Due to its simplicity, the circuit is suitable for example in the field of implantable nerve signal amplification, where ultra low noise and minimal power consumption are mandatory.

Categories and Subject Descriptors
B.7.1 [**Integrated Circuits**]: Types and Design Style – *Input/output circuits, VLSI.*

General Terms
Design.

Keywords
Analog design, CMOS, low-power.

1. INTRODUCTION
In recent years there has been a considerable research effort in the development of integrated amplifiers for electro-neuro-graph (ENG) signal recording. ENG signals range from a few Hz to a few kHz and require an input referred noise of a few $\mathrm{nV}/\sqrt{\mathrm{Hz}}$, and a high CMRR. To reduce flicker noise, several of these circuits use the chopper amplifier technique [1,2]. A simple CMOS chopper circuit was proposed by Oswald et al. in 1984 [3] just by switching input branches of a Miller-like amplifier. But CMOS choppers have evolved since then to improve usual chopper benchmarks: reduce offset, but also noise and power consumption. A topology that became popular is to use a band-pass amplifier in between chopper modulators to reduce high-frequency spikes that introduce residual offset [4,5]. Further, in ref [6] nested-choppers are used to obtain an instrumentation amplifier with less than 100nV offset. But ENG amplification do not require such an extremely low input offset. In Table 1, there are compared together three works from 2004 (IEEE-ISCAS) in the field of ENG amplification. Two of them use the band-pass chopper topology to reduce flicker noise, while the third uses a non-chopped continuous time amplifier.

For a non-chopped amplifier, input referred noise is a result of power consumption and area constrains [8]:

$$v_{n\,rms} \approx \frac{\mathrm{A}}{g_m} + \frac{\mathrm{B}}{W.L} \qquad (1)$$

Where $v_{n\,rms}$ is the rms noise voltage, g_m, W, L are the input transistors transconductance, width, and length respectively. A, B, are constants depending on frequency range, and technology parameters. The first term in (1) is thermal noise contribution while the second is associated to flicker noise. As pointed in [7,8] low input noise requires: a large bias current for a large g_m to reduce thermal noise; and a large *W,L* for flicker noise reduction. Chopper amplifier technique is suitable to preserve a reduced area for example in the case of multi-channel nerve signal recording[1], because pushes flicker noise out of the band of interest. However, a complex chopper topology increases also area, and power consumption. For example, band-pass filter tuning [4] adds an extra circuit demanding a considerable silicon area. Since ENG applications do not require an extremely low residual offset, it would be better to return to a simple chopper like the one in [3]: suitable to remove flicker noise but simple. In Fig.1, our proposed chopper amplifier is shown. The input signal V_{In} is chopped at a frecuency f_{ch} - period $T_{ch} = f_{ch}^{-1}$ - and then amplified on each phase through two independent G_m-C low pass filters. The intermediate signals V_{out1}, V_{out2} are further amplified by means of a low-pass filter (LPF) to remove residual frequency components above the band of interest. V_{Out} is the output signal, and a square wave $m(t)$ switches the modulators. V_n represent a parasitic voltage source like for example input noise or offset. The topology is a modification of the one proposed by Bakker & Huijsing in [9], but sample & hold output is substituted by a current integration scheme embedded in the self chopper. The circuit is extremely simple: just two transconductors, two capacitors, and an output low pass filter. Even G_{m2} can be substituted by an integrated resistor, currents can be integrated in a single capacitor, and the output low-pass is not essential (it was placed just to provide a 'clean' output signal). Intermediate voltages V_{out1}, V_{out2} can be estimated assuming that are the result of applying a lowpass filter to the input voltage of the form:

SBCCI'05, September 4–7, 2004, Florianópolis, Brazil.

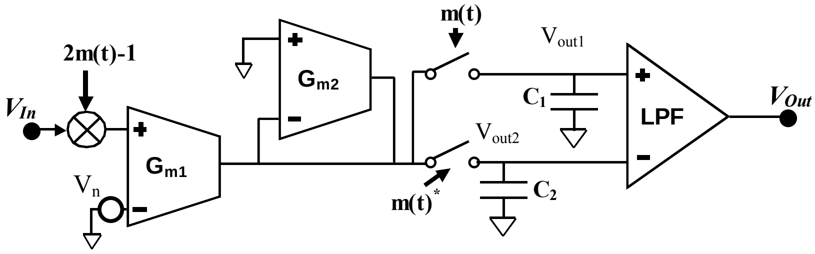

Figure 1: Topology of the proposed G_m-C amplifier. m(t) is a sqare wave signal.

$$H(\omega) = \frac{G_{m1}}{G_{m2}} \cdot \frac{1}{\left[1 + j\omega 2C/G_{m2}\right]} \qquad (2)$$

The 2 factor of the capacitor in (2) is the result of switching transconductance G_{m2} – see section below-. Because the input signal V_{in} is chopped, both branches are amplifying roughly the same signal, but with opposite sign. Their output is then subtracted, the result being $V_{Out} = H(\omega).V_{in}$. On the other hand, noise and offset voltage at the input V_n are amplified with equal sign and then subtracted. To guarantee a proper noise cancellation, the hypothesis is that the signal V_n does not change too much from one phase to another of the chopper. This is true for offset, and flicker noise (if $f_{ch} >> f$) but unfortunately not in the case of thermal noise ($V_n = V_{th}$), where $\int_{T}^{T+T_{ch}/2} V_{th}.dt$, and $\int_{T+T_{ch}/2}^{T+T_{ch}} V_{th}.dt$ are completely non-correlated. The exact calculation of the output signal is quite complex because we are dealing with a non time-invariant system, where also aliasing should be considered.

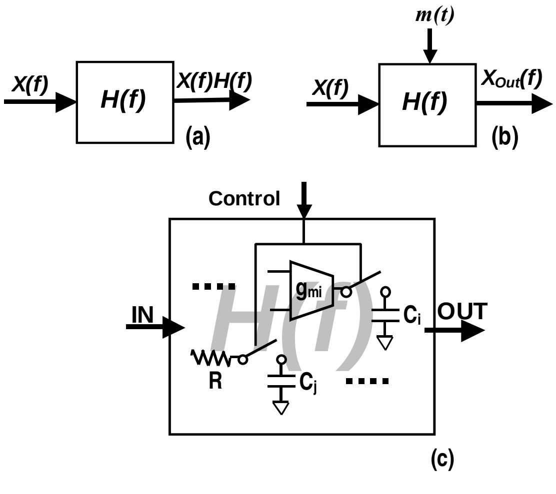

Figure 2 (a) Continuous time filter. (b) Switched continuous time filter (SCTF) (c) SCTF inside.

In the following section, we will introduce a general analysis of a switched continuous time filter (from now SCTF) in the frequency domain, that as far as we know has not been previously developed. Tools for SCTF output estimation will be developed; and the new chopper amplifier will be presented as a particular case of SCTF.

2. SWITCHED CONTINUOUS TIME FILTER ANALYSIS

In this section we study the operation of switched continuous time filters (SCTF), defined as continuous time filters with elements that are alternatively switched on and off in the signal path. A well known example is the use of switched resistors to multiply their value, but the universe of applications is wider. Consider a continuous time filter $H(f)$ and an input signal $x(t) \leftrightarrow X(f)$ (in the time and frequency domain respectively). The output signal will be $X(f).H(f)$ (Fig.2(a)). A SCTF operates alternating "active" time slots with "hold" time slots. During an "active" time, the filter is normally connected in the signal path. During a "hold" time, all the state variables inside the filter are kept constant. The filter may be G_m-C, R-C, combined, active or passive, etc, but the requisite is that the elements with memory inside the filter like the capacitors in Fig.2(c), preserve their condition (charge) during hold time. A control input to the filter sets it to "hold" or "active" depending on the value of a control signal $m(t)$; if $m(t)$ =1 the filter is active, if $m(t)$ =0 the filter is in hold. For the sake of simplicity we consider $m(t) = \sum_{n=-\infty}^{\infty} p(t - nT_S)$ as a pulse train with frequency f_S and width τ. We will refer to f_S as the switching frequency, $T_S = f_S^{-1}$ is the switching period, and τ the pulse width:

$$p(t) = 0 \; if \; |t| > \tau/2, \quad 1 \; if \; |t| < \tau/2 \quad \leftrightarrow \quad P(f) = \tau.\sin c(\tau f) \qquad (3)$$

The output signal $x_{Out}(t) \leftrightarrow X_{Out}(f)$ in Fig.2(b) is different from case in Fig.2(a) and we pretend to calculate $X_{Out}(f)$ in terms of $X(f), H(f), T_S, \tau$. Note that a SCTF is not a discrete time filter like switched capacitor or switched current filters; the SCTF does not work with samples although it holds the output during hold time. A SCTF also is not a time-invariant system so it is not possible to define a transfer function $H_{SCTF}(f)$. But note that state variables in the SCTF are continuous in time and they are modified only during the 'active' time. Let $\mathbf{y}$ denote the vector containing the state variables of the filter. To calculate $\mathbf{y}(t)$ the value of the input signal during hold time is not relevant.

It is only necessary to solve, on each active time interval ($nT_S - \tau/2, nT_S + \tau/2$), the differential equations of the filter assuming as the initial condition the value of the state variables at the end of the previous active time interval ($\mathbf{y}(nT_S - \tau/2) = \mathbf{y}((n-1)T_S + \tau/2)$). Therefore, to solve the equations in a single step we can "compress" all the active time intervals one beside each other to a single one like in Fig.3(a-c). We define a compressed signal $x_{Comp}(t) \leftrightarrow X_{Comp}(f)$ like in Fig.3(c), placing together the pieces of $x(t)$ corresponding to active time slots. If we define an intermediate auxiliary function, $x_I(t) = x_{Comp}(t) * h(t)$; $x_I(t)$ solves the filter's equations in all the active times and $x_{Out}(t)$ can be easily calculated by the inverse of the "compression" process like in Fig.3(e). The output follows $x_I(t)$ during active time (A) and we assume that the output of the filter is a state variable that does not change during hold time (B). The "compressed" signal $x_{Comp}(t)$ of Fig.3(c) is expressed:

$$x_{Comp}(t) = \sum_{n=-\infty}^{\infty} p(t - n\tau).x(t + n(T_S - \tau)) \tag{4}$$

The result of the Fourier transform in Eq.6 is as follows:

$$X_{Comp}(f) = \frac{\tau}{T_S}\sum_{n=-\infty}^{\infty} \operatorname{sin}c\left[\left(-f.\left(\frac{\tau}{T_S}-1\right)+nf_S\right).\tau\right]\cdot ... \\ ... \cdot X\left(f.\frac{\tau}{T_S} - nf_S\right) \tag{5}$$

For the sake of simplicity, in this equation as well as in (6),(8),(11) several detailed steps to compute the Fourier transform were ommited. These steps include the use of Fourier transform and convolution definition, time delay and scale change properties of the Fourier transform, and the sum of infinite exponential terms: $\sum_{n=-\infty}^{\infty} e^{-j2n\pi Tu} = \frac{1}{T}.\sum_{n=-\infty}^{\infty} \delta\left(u - n.\frac{1}{T}\right)$ [11].

Note in Eq.(5) that aliasing may occur if the bandwidth of the input signal is larger than $f_S/2$. The intermediate signal $X_I(f)$ is calculated:

$$X_I(f) = X_{Comp}(f).H(f) = \\ \frac{\tau}{T_S}.\sum_{n=-\infty}^{\infty} \operatorname{sin}c\left(-f.\left(\frac{\tau}{T_S}-1\right)+n\tau f_S\right).X\left(f.\frac{\tau}{T_S} - nf_S\right).H(f) \tag{6}$$

The exact output signal $X_{Out}(f)$ is the sum of two components $x_{Out}(t) = x_{OutA}(t) + x_{OutB}(t)$ that have to be calculated separately. $x_{OutA}(t)$ corresponds to the output of the filter in the "active" time slots:

$$x_{OutA}(t) = \sum_{n=-\infty}^{\infty} p(t - nT_S).x_I(t - n(T_S - \tau)) \tag{7}$$

The transform of Eq.(7) is calculated:

$$X_{OutA}(f) = \sum_{n=-\infty}^{\infty} \operatorname{sinc}(f(\tau - T_S) + n).X_I\left(f\frac{T_S}{\tau} - \frac{n}{\tau}\right) \tag{8}$$

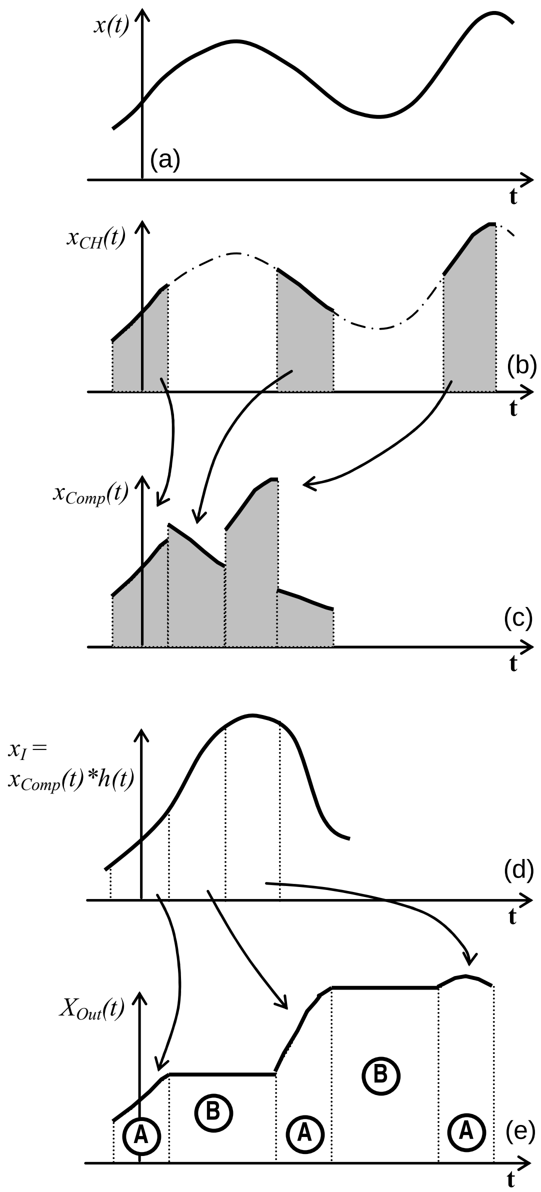

Figure 3 : Evaluation process for the output of a SCTF.
(a) input signal $x(t)$
(b) chopped signal $x_{CH}(t)$
(c) compressed signal $x_{Comp}(t)$
(d) intermediate signal $x_I(t)$
(e) output signal $x_{Out}(t)$ **for "active" time slots (A) and "hold" time slots (B).**

For the calculation of $x_{OutB}(t)$ corresponding to 'hold' time slots, we define first:

$$p'(t) = 1 \; if \; 0 < t < (T_S - \tau), \quad 0 \; elsewhere \quad \leftrightarrow \\ P'(f) = (T_S - \tau).\operatorname{sinc}((T_S - \tau)f).e^{-j\pi f(T_S - \tau)} \tag{9}$$

and

$$x_{OutB}(t)=\sum_{n=-\infty}^{\infty}\delta\left(t-nT_S-\frac{\tau}{2}\right).x_I\left(\frac{\tau}{T_S}\left(t+\frac{T_S-\tau}{2}\right)\right)*p'(t) \quad (10)$$

Eq.(10) has a Fourier transform:

$$X_{OutB}(f)=\frac{T_S-\tau}{\tau}.\mathrm{sinc}\left(\left(T_S-\tau\right)f\right).\sum_{n=-\infty}^{\infty}\left(-1\right)^n.X_I\left(\frac{T_S}{\tau}.\left(f-nf_S\right)\right) \quad (11)$$

Equations(5,8,11) allow the calculation of the exact output signal of the switched continuous time filter in terms of the continuous time filter transfer function $H(f)$, the input signal $X(f)$, and the switching parameters T_S,τ. The input signal is scaled up in frequency in (5) and down (8,11); but $H(f)$ in (6) is not scaled. Roughly in (6), we are filtering a frequency scaled version of the input signal that is then downscaled. From another point of view, neglecting the effect of the modulating sinc() functions and the effect of the aliasing (n=0):

$$X_{Out}(f)\approx X(f).H\left(\frac{T_S}{\tau}.f\right) \quad (12)$$

The filter $H(f)$ has been scaled to low frequency by a factor $\frac{T_S}{\tau}$. In Eq.(12) it is clear the effect of resistor multiplication for example in the case of R-C filters, and why a 2 factor appear in eq.(2).

3. CHOPPER AMPLIFIER DESIGN, AND SIMULATIONS

The chopper amplifier of Fig.1 has been designed –but not yet fabricated- to fulfil ENG requirements. G_{m2} was substituted with a resistor R=50kΩ, and C=300pF values were selected to achieve a 3db decay at 5kHz. A chopper frequency f_{ch}=20kHz was chosen to avoid aliasing. The low frequency gain is fixed at G=85 using an input transconductor G_{m1} of 1.7mS. Transconductors and MOS switches were simulated at transistor level, using a 0.8µm CMOS technology. In the plot of Fig.4, the simulated transient response of the amplifier while applying a 1mV square wave at the input is shown. Low-pass characteristic can be observed. Owing to aliasing, a SCTF has no transfer function *H(f)*. but it is possible to define a pseudo transfer-function as the output amplitude at a frequency *f*, when applying a pure tone of requency *f* at the input. This pseudo transfer function was calculated with (8,11), using MATLAB and N=15 terms for all the sums. The calculated pseudo transfer, as well as spice-simulated results are shown in the plot of Fig.5. As expected, 3db decay is close to 5kHz. Offset, and low frequency noise cancellation, can be examined simulating the transient response of the chopper to an input noise voltage V_n as in Fig.1. In the plot of Fig.6, a 100mV_{pp} input noise tone, and chopper output, transient simulations in spice are shown. Note the output signal is a series of spikes out of the band of interest. To predict the total output flicker noise, one can also calculate a pseudo-transfer function for flicker noise (using V_n as the input in Fig.1) and then integrate in frequency. If the chopper frequency f_{ch} (f_S) is enough high, this calculation leads to a negligible flicker noise contribution in the band of interest.

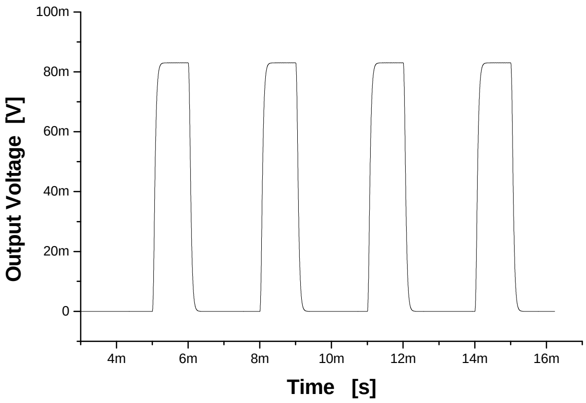

Figure 4: Transient simulation of chopper amplifier with 1mv-sqare wave input.

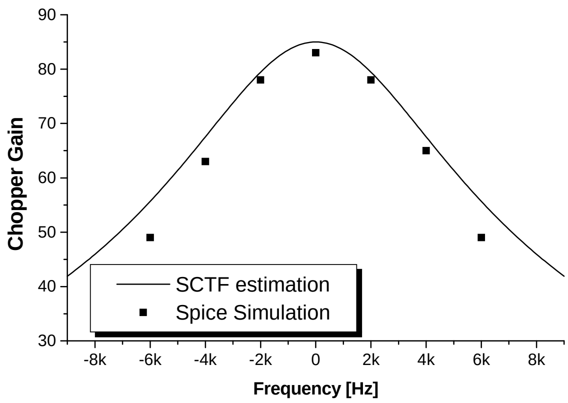

Figure 5: Simulated and predicted (SCTF theory) transfer of the chopper.

3.1 Thermal Noise Analysis

Due to its infinite-bandwidth nature that produces aliasing, applicating the pseudo transfer function technique to white noise may produce misleading results. Instead, thermal noise contribution is calculated with (4-11) on each switched G_m-C filter, substituting $X(f)$ by a constant PSD $S_n(f)$. Terms multiplying $X(f)$ in (4) to (11) are also squared. The result is very interesting and indicates that thermal input referred rms noise voltage does not increase too much using a continuous time G_m-C filter, or the chopper amplifier of Fig.1 (both are supposed to have the same bandwidth, and the same input transconductance G_{m1} introducing noise). A bit more interesting experiment is as follows: suppose the designer decides to reduce the duty cycle of each switched G_m-C to for example a 10% instead of 50%. According to SCTF theory, the 3dB decay can be preserved just by changing C to a 5 times smaller value. With a reduced duty cycle, power-hungry G_{m1} may be turned off during near 80% time where it is inactive thus reducing power consumption. Or better – in an attempt to reduce thermal noise- to increase 5 times the G_{m1} bias current to reduce input thermal noise according to (1), while

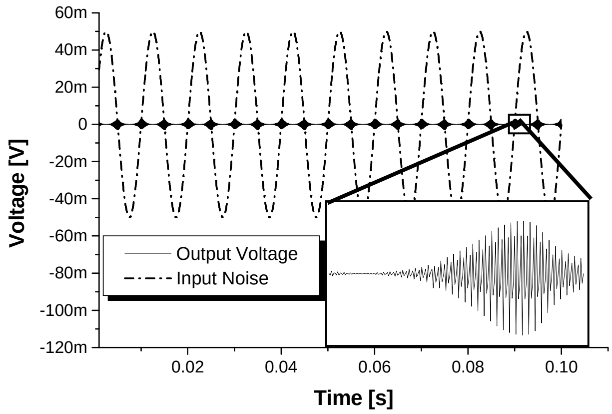

Figure 6: Transient rejection of chopper amplifier of $100mV_{pp}$-100Hz noise tome V_n.

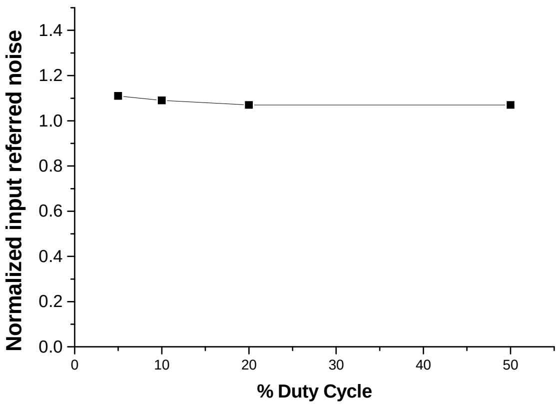

Figure 7: Theoretical input rms noise voltaje in terms of duty cycle. Values are normalized with a non-chopped filter noise .

preserving the original power consumption. Input referred rms thermal noise voltage, in terms of duty cycle on this condition is shown in Fig.7. Unfortunately the effect of aliasing compensates G_{m1} input thermal noise reduction as the duty cycle is lowered. Once again, it was not possible to defeat thermal noise.

3.2 Complete ENG filter

To evaluate an overall ENG amplifier-filter behavior, a G_m-C 100-5kHz band-pass stage was also simulated. The topology is the same employed in [10] for a piezoelectric accelerometer. It has a gain G=100, and consumes only 2µA current. This filter substitutes the low-pass one in Fig.1. The simulated complete ENG amplifier consumes 1mW power, has an estimated input referred noise of $6\,\mathrm{nV}/\sqrt{\mathrm{Hz}}$, and occupies a $0.8mm^2$ area, that is comparable to circuits in Table 1.

4. PRELIMINARY MEASUREMENTS

Although the proposed chopper has not been fabricated, some preliminary measurements were obtained in a low frequency band with a previously fabricated OTA, and external 100nF capacitors and switches, using f_{ch}=15kHz. The topology is shown in Fig.8

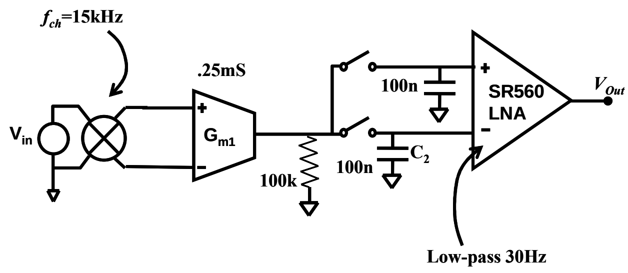

Figure 8: Measured chopper amplifier.

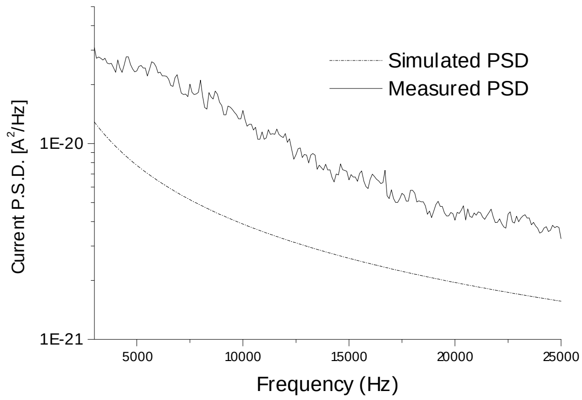

Figure 9: Measured input referred noise of G_{m1} (non-chopped). Simulated PSD was obtained using flicker noise model in [8].

with a resistor R=100kΩ instead of G_{m2}. G_{m1} is a .25mS standard symmetrical OTA, fabricated in a 0.8µm technology. G_{m1} measured input referred offset is 14mV, and measured input referred noise voltage is $90\mu V_{rms}$ integrated in a band of interest from 0.1 to 30Hz. In Fig.9, measured noise PSD for non-chopped G_{m1} is shown. A 30Hz low-pass filter was placed at the output. With the chopper amplification scheme, measured input referred offset was reduced to 200µV and noise to $11\mu V_{rms}$. Although these values are greater than expected –probably due to the use of discrete elements -, the operation of the G_m-C chopper amplifier has been demonstrated.

5. CONCLUSIONS

A very simple chopper amplifier intended for implantable ENG applications has been presented. This circuit can be examined as a switched continuous time filter (SCTF). A set of equations have been introduced to examine generic SCTFs in the frequency domain. The tool, as well as spice simulations, were used to examine the chopper amplifier behaviour regarding noise, and output transfer function. Using the proposed circuit, and G_m-C filter techniques, complete ENG systems can be implemented with specifications close to previously reported works.

Table 1: A comparative survey of 3 ENG amplifiers presented in ISCAS-2004.

	Filter Band	Supply	Area	Noise	Power	
[7] Oses et al	**100-5kHz**	**5V**	**1.1mm^2**	**5.1nV/Hz$^{-1/2}$**	**1mW**	**Continuos Time**
[2] Uranga et al	**LP 3kHz**	**5V**	**2.7mm^2**	**6.6nV/Hz$^{-1/2}$**	**1.3mW**	**Chopper**
[1] Gosselin et al	**100-5kHz**	**1.8V**	**reduced**	**30nV/Hz$^{-1/2}$**	**25uW**	**Chopper**

6. ACKNOWLEDGEMENT

The author would like to thank Prof.Alicia Fernandez, and Rafael Grompone for helpful suggestions; also to the Microelectronics Group at Universidad de la República where part of this work has been developed.

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