

Chopper Modulation Improves OTA Information Transmission

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Abstract—We have investigated information transmission in operational transconductance amplifiers (OTA) using chopper modulation. Previous work showed that the optimal frequency bandwidth for an OTA was much higher than typical operating frequencies. Here we analyze the information transfer rates for a folded cascode amplifier and a chopper modulated folded cascode amplifier using the principles of information theory. The frequency transfer characteristic and intrinsic physical noise source of each device is modeled using process dependent noise parameters and the waterfilling technique is applied to determine the capacity as well as information rates for low frequency signals. Simulations are experimentally verified using circuits fabricated in a commercially available 3-metal, 2-poly 0.5 μm CMOS process.

I. INTRODUCTION

Operational transconductance amplifiers (OTA) are a basic building block for modern micro and nanoscale sensors. However, as feature sizes scale down, the noise contribution of each MOS device in the OTA increases and the OTA performance suffers from reduced signal to noise ratio. This is an extremely important factor in designing sensors where the input signal is small compared to the circuit noise, a common problem for lab-on-a-chip systems and other densely integrated systems. Examples of these types of integrated sensors are bioamplifiers which amplify weak extracellular signals, fluorescence (image) sensors based on fluorophores with weak emission intensities and capacitive sensors which sense the weak capacitive coupling between a cell and substrate [1]–[3].

In previous work we investigated the information transmission of OTAs by modelling input-referred noise and determining the channel capacity for transmission of analog signals through an OTA using an information theoretic approach in which the circuit was treated as a Gaussian communication channel with colored noise [4] [5]. The technique was applied to simple and wide range OTAs. The signal frequency band for optimal use of the OTA was found to be around the second pole of the frequency response, much higher than the typical operating range below the first pole. This suggests that strategies such as signal modulation might provide more efficient use of an OTA from the perspective of information transmission. Here we extend the previous work to analyse the information capacity and efficiency of an amplifier that employs chopper modulation.

By considering the circuit as a communication channel with

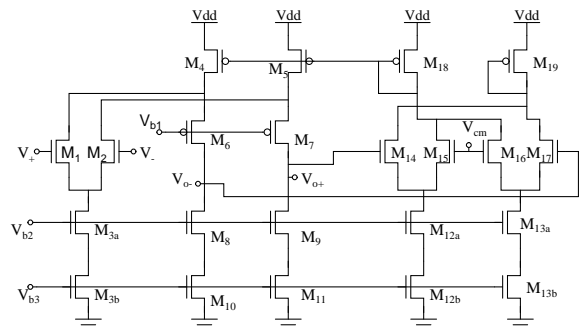


Fig. 1. Folded Cascode OTA: all nmos: 15/5; M_6 and M_7 : 70/5; M_4 and M_5 : 105/5

additive Gaussian noise the capacity of the channel can be found easily using the water-filling technique in the spectral domain. Circuits are considered to be composed of a cascade of analog stages each with its own transfer function and equivalent input noise sources. The input signal power is allocated to the frequency bands with lowest noise spectral density such that the total spectral density of *signal + noise* is constant over the allocated frequency range. In this work we consider two types of MOSFET noise, thermal and flicker. Flicker noise is caused by fluctuation in the number of charge carriers as they are trapped and released from traps located at the silicon-oxide interface. Thermal noise is caused by thermal agitation of charged carriers and is a white noise process independent of voltage.

Section II gives an overview of the amplifier design and the chopper modulation technique. Section III computes the theoretical noise and information capacity for the amplifier alone and for the amplifier using chopper modulation. Section IV presents measurement results, and Section V summarizes the work.

II. AMPLIFIER DESIGN AND CHOPPER STABILIZATION TECHNIQUE

The amplifier considered in this work is a fully differential folded cascode operational transconductance amplifier (Fig 1). This amplifier typically has a high open loop gain which implies that it has a low input-referred noise.

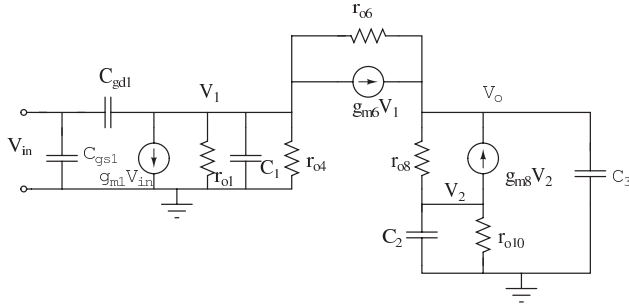


Fig. 2. Simplified Small Signal Model

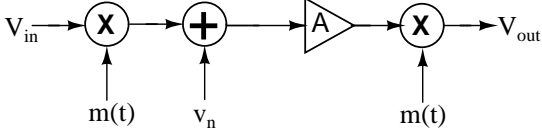


Fig. 3. Generalized schematic of a chopper amplifier

A. Fully Differential Folded Cascode OTA

For differential signals the small signal circuit model for this amplifier can be simplified to the form shown in Fig 2 where the capacitors C_1 , C_2 and C_3 represent several parasitic capacitances within the circuit.

$$C_1 = C_{gd4} + C_{gs6} \quad (1)$$

$$C_2 = C_{gd10} + C_{gs8} \quad (2)$$

$$C_3 = C_{gd8} + C_{gd6} + C_{gs14} + C_{gs18} + C_L \quad (3)$$

The transfer function is derived using nodal equations based on Fig 2. The low frequency gain is given by $g_{m1}R_o$ where R_o is the equivalent resistance looking into the drains of M_6 and M_8 . The full transfer function has the general form:

$$\frac{V_{out}}{V_{in}} = A_o \frac{\left(1 - \frac{s}{z_1}\right) \left(1 - \frac{s}{z_2}\right)}{\left(1 - \frac{s}{p_1}\right) \left(1 - \frac{s}{p_2}\right) \left(1 - \frac{s}{p_3}\right)} \quad (4)$$

with the dominant pole given by $1/R_o C_3$. The first zero z_1 is given by g_{m1}/C_{gd1} , the second zero z_2 can be approximated as g_{m8}/C_2 . The higher poles are approximated as g_{m8}/C_2 and g_{m6}/C_1 .

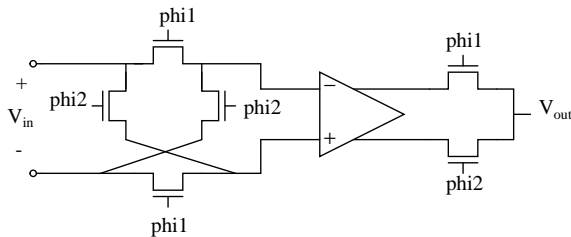


Fig. 4. Detailed schematic of chopper amplifier

B. Chopper Amplifier Architecture

The chopper modulation technique shifts the input signal to a higher frequency, amplifies it and then demodulates it back to the baseband. Fig 3 shows the principle of chopper modulation where the output voltage is given by [6]

$$V_{out} = (V_{in}m(t) + v_n) Am(t) \quad (5)$$

where $m(t)$ is the modulating signal alternating between 1 and -1 with frequency f_{chop} , and v_n and A are the noise and gain of the amplifier respectively. This equation can be simplified to

$$V_{out} = V_{in}A + v_n m(t)A \quad (6)$$

The Fourier representation of $m(t)$ is

$$m(t) = \sum_{n=-\infty}^{\infty} \frac{2}{j(2n+1)\pi} e^{j(2n+1)f_{chop}t} \quad (7)$$

This leads to a power spectral density of the output noise given by

$$S(f) = \left(\frac{2}{\pi}\right)^2 \sum_{n=-\infty}^{\infty} \frac{1}{(2n+1)^2} S_N(f - (2n+1)f_{chop}) \quad (8)$$

where S_N is the noise spectral density of the OTA alone. If the 3dB cutoff frequency is much larger than the chopping frequency and the input noise is a $1/f$ source, then the $1/f$ pole is moved to the odd harmonics of the chopping frequency and can be approximated by a white noise component at lower frequencies. For an input noise that is white, the input-referred noise of the chopper amplifier is also white and of a lower magnitude than the original amplifier noise, and it asymptotically approaches the original input noise as the ratio of the cutoff to the chopping frequency increases. While this analysis neglects the flicker noise contributions of the modulator and demodulator, it can be used to determine a lower bound on the input referred noise of the amplifier.

III. THEORETICAL NOISE AND INFORMATION CAPACITY

A. Theoretical Noise

Output noise is modelled by considering the flicker noise and thermal noise models [8] given by

$$S_{I_d} = \gamma 4KTg_m + \frac{K_f I_d^{A_f}}{f^{E_f} C_{ox} L_{eff}^2} \quad (9)$$

where K_f , A_f and E_f are process dependent constants, C_{ox} , g_m and L_{eff} are the usual MOSFET parameters and γ changes with the transistor operating region. The current spectral density of the noise sources for each transistor in the amplifier is modelled according to (9) and then referred back to the input of the amplifier by dividing by the transfer function from the input to that point in the circuit. The noise contributions derived in this way are shown in Fig 5 which compares the input-referred noise of the OTA and the OTA with chopper modulation under the same bias conditions at two chopping frequencies. For ideal feedback the equivalent

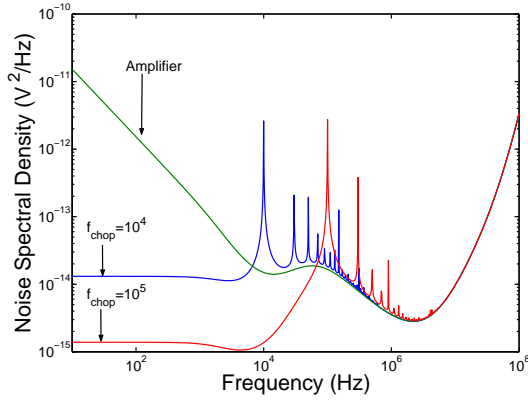


Fig. 5. input-referred noise of the amplifier alone and the chopper amplifier at two chopping frequencies

input noise generators may be moved unchanged outside the feedback loop without affecting circuit noise performance. The modulator/demodulator switches are assumed to be noiseless and any noise due to charge injection and clock feedthrough is neglected. The overall input-referred noise is cupshaped; at higher frequencies the gain decreases and the minimum input noise occurs near the second pole of the amplifier. In Fig 5, $S_{N_{chop}}$ is plotted for the first 30 terms of series of the equation in (8). At low frequencies the noise for the chopper amplifier is white and starts to increase as the frequency approaches the chopping frequency. At the chopping frequency and odd harmonics thereof, the noise is infinite since $1/f$ is transformed to $1/(f - f_{chop})$, which is replicated at odd harmonics of the chopping frequency for square wave modulation.

B. Information Capacity

The capacity of a Gaussian channel with colored noise is given by the waterfilling technique [9] as

$$C = \int_{f_1}^{f_2} \log \left(1 + \frac{\nu - N(f)}{N(f)} \right) df \quad (10)$$

where $P = (\nu - N(f))\Delta f$ is the input signal power and N is the noise spectral density. ν is the total spectral density of signal and noise over the signal bandwidth f_1 to f_2 . Applying this to the amplifier without modulation we obtain the familiar results of the waterfilling technique, where the capacity is a monotonically increasing function of signal power, with the signal allocated over an optimal frequency bandwidth. For typical amplifiers and OTAs this implies that signal power is optimally allocated at frequencies near the second pole of the amplifier.

When modulation is introduced, input signals above the chopping frequency would experience unacceptable aliasing, rendering the signals indecipherable and making the bandwidth above the chopping frequency unsuitable for signal transmission. This occurs because the modulating signal is a square wave made up of an infinite sum of sine waves. If a pure sine wave was used as the modulating signal, then in

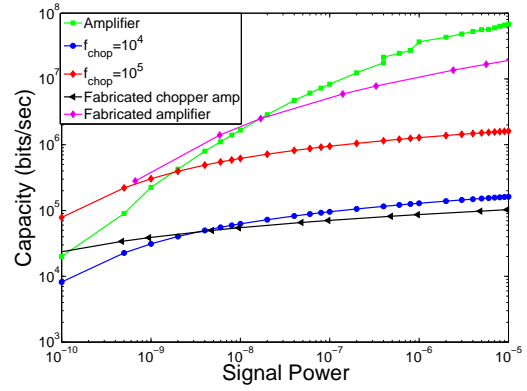


Fig. 6. Capacity as a function of Power

principle either those frequencies below or above the chopping frequency could be used (but not simultaneously).

We now examine in greater detail the information transmission under the assumption that the input-referred noise can be approximated as white noise below the chopping frequency (Fig 5) and the input signal is also restricted to be less than the chopping frequency. Clearly a frequency bandwidth, Δf , must be chosen such that it lies between some lower limit f_1 and upper cutoff f_2 . Waterfilling in this case gives

$$P(f) = (\nu - N_0) \Delta f \quad (11)$$

where ν is a constant, and allows us to explicitly solve for the information rate

$$I = \int_{f_1}^{f_2} \log_2 \left(1 + \frac{(\nu - N_0)\Delta f}{N} \right) df \quad (12)$$

Since the term inside the logarithm is constant, this expression can be simplified as

$$C = \Delta f \log_2 \left(1 + \frac{P}{N_o \Delta f} \right) \quad (13)$$

where P is the total signal power and N is the total noise power in the operating range. Increasing the signal power for the chopper amplifier causes the total *signal + noise* to increase, but not as steeply as for the linear amplifier. This implies that the information rate of the chopper amplifier is higher than for the linear amplifier. Fig 6 shows that the capacity of the simple OTA is higher than the capacity of the chopper amplifier at high signal power. This occurs because the frequency range of the chopper amplifier is restricted below the chopping frequency, whereas the optimal frequency range is unbounded for the OTA. At very low signal power the capacity of the chopper amplifier exceeds the capacity for the OTA for high chopping frequencies. In Fig 7 we restrict the allowed signal bandwidth to below the amplifier cutoff frequency and find that the information rate is consistently larger for the chopper amplifiers and that information rate increases with chopping frequency.

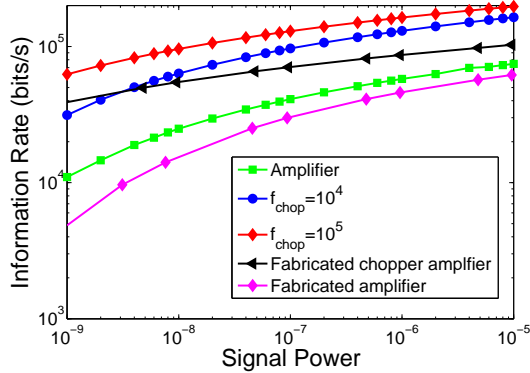


Fig. 7. Information Rate assuming input signal is restricted to below $10kHz$

IV. MEASUREMENT RESULTS

We fabricated a test amplifier along with modulator/demodulator circuits in a commercially available $0.5\mu m$ 3-metal, 2-poly CMOS technology. A photomicrograph of the fabricated amplifier is shown in Fig 8, the OTA is $54.6\mu m \times 80.4\mu m$ and the power consumption is $60\mu W$. The modulators are implemented as cross coupled switches and the clock inputs are non-overlapping (Fig 4). Transmission gates are used to reduce the effects of charge injection and clock feedthrough. The transfer function and output noise of the amplifier with and without the modulators/demodulators was measured using network/spectrum analysers (Agilent 4395A and 4396B) at a bias current of $10\mu A$. The measured experimental output noise was divided by the measured transfer function and the resulting input-referred noise is shown in Fig 9.

The input-referred noise of the amplifier exhibits the typical cup shape and the noise of the chopper amplifier exhibits residual $1/f$ noise at low frequencies. Due to limited bandwidth resolution in the measurement, the peaks in the noise above the chopping frequency are not well resolved. The chopping frequency was set equal to the corner frequency of the amplifier at $10kHz$. The experimentally derived capacity is plotted in Fig. 6 as a function of signal power. While the amplifier has a higher capacity since its bandwidth is unrestricted, the chopper amplifier has a higher capacity for very low signal powers and a higher information rate when frequencies are limited below f_{chop} .

The optimal frequency range of operation for each amplifier is also drastically different. The optimal frequency for the OTA by itself is around $100kHz - 1MHz$, at least a decade higher than the cutoff frequency at $10kHz$ and at approximately the second pole of the system. This implies that there would be significant attenuation of any input signal regardless of how efficiently the amplifier is operating. In theory the optimal frequency range of the chopper amplifier is from DC to just below the chopping frequency, but in practice occurs at frequencies close to the chopping frequency due to residual $1/f$ noise at low frequencies. The residual flicker noise is most likely due to noise and charge injection in the modulator and demodulator.

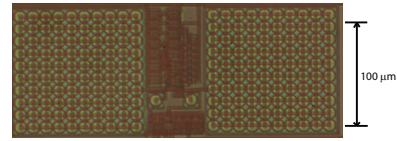


Fig. 8. Photomicrograph of fabricated amplifier

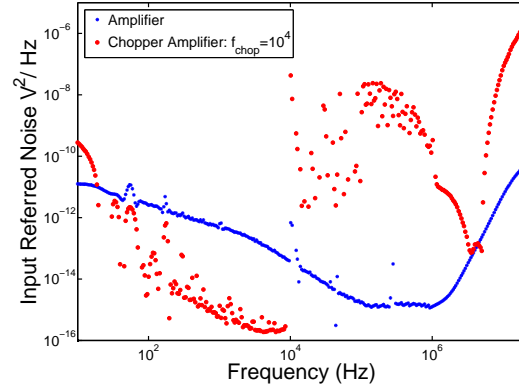


Fig. 9. Input-referred Noise of the Folded Cascode OTA and chopper amp

V. CONCLUSION

We have analyzed the capacity and information transmission rate of a folded cascode OTA and a chopper modulated folded cascode OTA. Chopper modulation provides for efficient use of signal power as it minimizes the device noise at low frequencies. This implies that chopper modulation is an efficient way to implement sensors that transduce very weak input signals. We have shown experimental results for the capacity and the information rate that track theoretical predictions.

ACKNOWLEDGMENT

We would like to thank MOSIS for chip fabrication; this chip was developed as part of an undergraduate class project. We would like to thank A. Banes, P. Hurtado, E. Arvelo, S. Sahand and Z. Bekka. This material is based upon work supported by the National Science Foundation under Grant No. 0238061 and the Laboratory for Physical Sciences.

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