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INTEGRATING DATA CONVERTERS FOR PICOAMPERE CURRENTS FROM ELECTROCHEMICAL TRANSDUCERS.

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ABSTRACT
This paper describes a current mode A/D converter designed for a maximum input current range of 5 nA and a resolution of the order of 1 pA. The converter is designed for a potentiostat for amperometric chemical sensors and provides a constant polarization voltage for the measuring electrode. A prototype chip using the dual slope conversion method has been fabricated in a 0.7 μm CMOS process. Experimental results from this converter are reported. Design problems and limitations of the converter are discussed and a new conversion technique providing a larger dynamic range and easy calibration is proposed.

1. INTRODUCTION
Electrochemical transducers are widely used in equipment such as blood analyzers. For example, an amperometric chemical sensor can be used to measure the concentration of oxygen in blood. Such a sensor provides a current proportional to the concentration if the sensor is biased with a constant voltage $V_{pol}$ of about 0.6V. However, the maximum current of the sensor is very small, in the order of a nA, so high resolution data converters (more than about 10 bits) are desirable to implement the converter as a part of an integrated transducer front-end in order to reduce size, weight, power consumption and production cost of the sensor system [1]. This implies that the converter should be designed in a standard CMOS process and that it should be designed to be insensitive to process variations and device tolerances such as offset voltages and non-linearities of components.

2. CIRCUIT DESCRIPTION
A block diagram of the prototype version of the potentiostatic A/D converter is shown in Fig. 1. It is implemented as a dual slope converter. Since the signal to be converted is essentially a (very small) dc current, an integrating converter type is advantageous because it filters out high frequency noise components. Noise is critical in an application with such small currents. However, the conversion time can be selected to be fairly long, about 200 ms, implying that stochastic noise above approximately 10 Hz is attenuated by the converter integration.

The left part of the diagram shows the potentiostatic control made by the transconductance amplifier $OTA.P$ and the pass transistor $M_{p1}$. The potentiostatic control system applies a polarization voltage $V_{pol}$ between the working electrode $WE$ and the reference electrode $RE$. For the test chip an external polarization voltage is used. The transconductance amplifier $OTA.P$ and the transistor $M_{p1}$ are connected as a voltage follower, so the input voltage to $OTA.P$ will also appear at the working electrode. This electrode is coated with a chemically sensitive membrane (CSM) and behaves as a constant current source when $V_{pol}$ is larger than about 0.4V. An optimum value of 0.6V results in a linear relationship between the current $I_{mes}$ and the solution concentration. Since the input current to $OTA.P$ is negligible, the current $I_{mes}$ flows entirely through $M_{p1}$.

The current $I_{mes}$ is charging the integration capacitor of the dual slope converter during a fixed integration time $t_{int}$. After this period the reference current $I_{ref}$ is connected to the input of the dual slope converter and the capacitor is discharged. The switching between $I_{mes}$ and $I_{ref}$ is done by a bridge of four switches $S_1$, $S_2$, $S_3$ and $S_4$. The measurement current $I_{mes}$ is switched between $S_2$ and $S_3$, keeping a constant current flow through $M_{p1}$, and a constant voltage at the drain of $M_{p1}$. The reference current $I_{ref}$ is switched between $S_1$ and $S_2$. The bridge is constructed using minimum size N-MOS transistors. The reference current is buffered by a regulated cascode $OTA.N$ and $M_{p1}$.

The current of the converter is the integrator and the comparator, followed by the digital block. The integrator is implemented by the transconductance amplifier $OTA$, the capacitor $C$ and a reset switch $S_{reset}$. The offset and the 1/f noise associated with the MOS input transistors of the integrator are compensated using an auxiliary input of reduced sensitivity in parallel with the main input of the integrator [2].

An important advantage of the dual slope ADC is its tolerance to variations and nonlinearities of most of the circuit parameters. In dual slope technique, the current to be measured $I_{mes}$ causes the integrator to ramp-up for a fixed time $t_{int} = \frac{2^N T_{clk}}{I_{ref}}$, where $T_{clk}$ is the period of the clock signal and $N$ is the resolution in bits. Then a known current of opposite polarity $-I_{ref}$ is applied until the output crosses the threshold of the comparator (Fig. 2). To achieve maximum resolution, $I_{ref}$ is selected close to the full-scale value of the current to be measured: $I_{ref} \approx I_f$. The comparator and the integrator have the same reference voltage. Assuming the ideal case, when there is no voltage error across the capacitor at the beginning and end of the conversion, that is $V_i = 0$, the ramp-down time $t_{down}$ is proportional to $I_{max}$:

$$t_{down} = I_{max} \frac{t_{up}}{I_{ref}}$$

The ramp-down time is measured using a digital timing circuit. The conversion time is given by: $t_{conv} \leq 2^{N} T_{clk}$. A clock signal
of $T_{clk} = 1\mu s$ allows 16 bits resolution within a conversion time of $t_{conv} < 200\, \text{ms}$, and still leave some time for calibration. The capacitor $C$ must have a rather big value in order to keep the integrator output voltage within the permissible output voltage range $V_{\text{range}}$ of the OTA. $C$ is limited by:

$$C \geq \frac{I_{FS} \cdot t_{up}}{V_{\text{range}}}$$

With $V_{\text{range}} = 3\, \text{V}$, $I_{FS} = 5\, \text{nA}$ and $t_{up} = 65\, \text{ms}$ a minimum capacitor value is about $100\, \text{pF}$.

### 3. PROBLEM AREAS

Some of the important problems related to the design of the dual slope converter are:

1. Input offset voltages of comparator and integrator;
2. Speed of comparator;
3. Charge injection from switches;

The input offset voltages of the comparator and the integrator (or rather the difference in their offset voltages) cause the capacitor voltage $V_f$ at the end of the conversion to be different from the capacitor voltage $V_i$ at the start of the ramp-up period: $V_{off} = V_f - V_i$. For an ideal reset switch $V_i$ is equal to 0V. If the offset voltages cause the comparator output state to be low (counter disabled) after the reset, the conversion has to proceed for a time $t_{off} = CV_{off}/I_{mes}$ before the comparator changes state and the counter starts counting the time $t_{up}$. With offsets in the mV-range and $C \approx 100\, \text{pF}$ this delay is unacceptably long for small values (order of pA) of $I_{mes}$, so the comparator and OTA must be offset compensated or designed to have an offset that ensures $V_f < V_i$. If $V_f < V_i$ the deintegration time $t_{down}$ will be too long, the error being $CV_{off}/I_{ref}$. However, this is a constant error which can be eliminated by calibration.

The comparator delay has a similar effect as the offset voltage in the sense that a delayed response from the comparator causes the counter to add a number of counts equal to $t_{delay}/T_{clk}$. The delay $t_{delay}$ depends on the comparator design and the slope of the discharge, essentially being inversely proportional to the square root of the slope [3]. Simulations show delay times in the order of less than $1\, \mu s$ with a slope of $5\, \text{nA}/100\, \text{pF} = 50\, \text{pV}/\text{ps}$, implying an error of less than 1 LSB's with a 1MHz clock. Furthermore, this is a con-

![Figure 1. Block diagram of the prototype circuit.](image1)

![Figure 2. The voltage across the integration capacitor.](image2)
The noise at the comparator input and the integrator input are uncorrelated and add to form an equivalent noise (noise) of the comparator threshold. Both high frequency and low frequency equivalent input noise is important. The lower frequency limit can be controlled by the reset cycle (offset compensation cycle) and the high frequency limit can be assumed to be related to the comparator delay, being of the order of $t_{delay}$. With these assumptions we find a total equivalent rms noise from the comparator and the OTA of about 200pV. This corresponds to an uncertainty of 4 LSB's and assuming a Gaussian amplitude distribution of the noise voltage amplitude the maximum uncertainty caused by the comparator and OTA input noise is about 12 LSB's.

4. EXPERIMENTAL RESULTS

The test chip has been designed for a 0.7μm CMOS process from MIETEC. A full custom layout has been used. The circuit is designed for a supply voltage of 5V. Separate power and ground are used to the analog part of the chip to isolate the more sensitive analog circuitry from the digital switching noise. The comparator is a track-and-latch type [4]. The most critical performance limiting factor has been found to be the equivalent noise at the comparator input. This is around 500pV which is somewhat higher than expected from the simulations. The reason for this has not yet been fully resolved but some noise is believed to be caused by the external test setup and some noise is believed to be introduced from the digital part of the chip.

The converter has been tested with specially designed current sources $I_{ref}$ and $I_{max}$ and from the reset switch injects an error charge into the integration capacitor. For a small size transistor, the charge injected into the source or drain terminal in our case is approximately 2.3fC, which can be compared to the charge corresponding to 1 LSB, i.e., 50pV·100pF=5fC. Apparently, the injected charge is of the same order of magnitude as one LSB, so a means of cancellation of the injected charge is desirable. This is accomplished by using the bridge NMOS switch between $I_{max}$ and $I_{ref}$ and by using a switch with a dummy transistor for the reset.

Noise is a fundamental performance limitation of the converter. It comes from different sources: an equivalent noise from the reference current and the polarization voltage, an equivalent input noise from the integrator OTA and the comparator, and switching noise induced from the digital components in the system. The noise from the cascode transistor $M_{P2}$ and $M_{N2}$, and the switches is not important because of the impedance levels at these transistor terminals. For the test chip the reference current is supplied from an external current source. For a fully integrated version, the reference current can be supplied by a PMOS transistor which is advantageous compared to an NMOS transistor with respect to 1/f-noise. Likewise, for the test chip, the polarization voltage is applied externally. Noise and offset voltage at the input of $O T A\_P$ will appear as a small voltage across the high impedance output of the measuring electrode, resulting in a change in the current $I_{max}$ which is the noise and offset voltage divided by the output impedance of the measuring electrode.

The noise at the comparator input and the integrator input are uncorrelated and add to form an equivalent noise (noise) of the comparator threshold. Both high frequency and low frequency equivalent input noise is important. The lower frequency limit can be controlled by the reset cycle (offset compensation cycle) and the high frequency limit can be assumed to be related to the comparator delay, being of the order of $t_{delay}$. With these assumptions we find a total equivalent rms noise from the comparator and the OTA of about 200pV. This corresponds to an uncertainty of 4 LSB's and assuming a Gaussian amplitude distribution of the noise voltage amplitude the maximum uncertainty caused by the comparator and OTA input noise is about 12 LSB's.

5. FOLDED DUAL SLOPE CONVERTER

Clearly, a design target of 16 bits resolution has not been met with the experimental design. The most severe problem appears to be the equivalent noise at the input of the comparator. In order to reduce this problem it would be highly desirable to have a substantially larger slope of the voltage ramp. If we were able to increase the slope with a factor $n$, the voltage change during comparisons on the deintegrating ramp would also increase with a factor $n$, thus increasing the comparator resolution with that factor. However, for the same full scale current $I_{ref}$ this would require a larger dynamic output range of the integrator and this is not possible with a low supply voltage. Instead, a new multi-slope technique [5] is proposed in which the up/down integrations are folded into the available integrator output range. Given an allowed conversion time, with this technique the integration capacitor is reduced by a factor of $n$, and $n$ up/down integration cycles are performed for each conversion without resetting the integrator between these cycles.

Fig. 4 shows the integrator output. Clearly, the integration slope is increased by a factor of $n$ compared to a standard dual slope conversion within an integrator output range of $V_{range}$. The output of the converter is measured with N-bits resolution in a similar way as a dual slope converter, i.e. by subtracting the known total ramp-up time $t_{up}$ (i.e., the sum of the $n$ ramp-up times $t_{up}/n$) from the total conversion time $t_{conv}$.

It is important to realize that the exact time occurrences for the changes between integrating and deintegrating periods does not affect the comparator resolution. What is important is that the total integration and deintegration times are equal to that of the equivalent dual-slope converter, and that the integrator output voltage does not exceed $V_{range}$. To see this, consider a situation where...
the end of the ramp-down period occurs at an integrator voltage of
$V_f + \Delta V_f$, i.e., the period has been $\Delta V_f/C/I_{ref}$ too short (the $\Delta V_f$
was caused by noise at the comparator input). At the end of the next
ramp-up period, the integrator output voltage would be $\Delta V_f$
higher than in the ideal case, and the next ramp-down period would
then be $\Delta V_f/C/I_{ref}$ longer, canceling the error from the previous
ramp-down. Thus, only the comparator noise from the very last
conversion will affect the total count $t_{down}$; we have increased the
converter resolution by $\log_2(n)$ bits.

A possible problem with the folded dual-slope technique is the charge
injection generated by the switches $S_1$--$S_6$, when $n$ is high.
Assume that the charge injected on $C$ during a conversion cycle
is $Q_{sw} + q_{sw}$, where $Q_{sw}$ is a constant value and $q_{sw}$ is a sto-

castic variable that is evaluated at each cycle. When scaling $C$
by $1/n$, the relative integrator output error after $n$ cycles will be
$V_{error}/nV_{range} = (nQ_{sw} + \sqrt{n}q_{sw})/(CV_{range})$, where $q_{sw}$
is the $q_{sw}$ variance, compared with $V_{error}/V_{range} = (Q_{sw} +
\sigma_{sw})/(CV_{range})$ for the normal dual-slope converter. As the first
term is a constant error, it can be eliminated by compensation; the
second term, however, shows that there is a limit on high we can
choose $n$. Because we are dealing with such small currents, even
if $q_{sw}$ is in the order of 0.1fC, this limit cannot be neglected.
Another problem with the accumulated charge injection is when mea-
suring very small currents: if $n^2Q_{sw}/C$ happen to be negative and
larger than the accumulated integrator output caused by the input
current, the output of the converter will be a zero. For this reason,
it is necessary to deliberately introduce a small positive input offset
current (which, of course will be eliminated during the compensa-
tion phase).

Apart from the above mentioned accumulated charge injections,
the problem areas, and the effects of these, for the folded dual-slope
converter are the same as for the conventional dual-slope converter
with an integrator range $nV_{range}$. The converter can easily be cali-
brated because each switch is operated a fixed number of times dur-
ing each conversion. The errors from the comparator delay, charge
injection, and capacitor nonlinearities can be made signal indepen-
dent if the same transition of the comparator controls the end of a
cycle and the beginning of the next one. An $N = 16$bit folded dual
slope converter with $n = 2^6$, and $C = 10\mu F$ offers a good solution.

6. CONCLUSION

A prototype chip of an analog to digital converter for electrochem-
ical amperometric sensors was designed and tested. Test chips
demonstrated that a resolution of about 11 bits is readily achievable
with a dual slope converter with a full range of 5mV.

In order to achieve higher resolution a new conversion method,
folded dual slope, was proposed. The folded dual slope converter
requires only very simple extra digital circuitry. Experimental re-
results combined with simulations indicate that the folded dual slope
converter can achieve 16 bits resolution in a conversion time less
than 200ms. A future version of the test chip will include simple
modifications of the digital block to implement the proposed con-
verter.

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