

A 2V Rail-to-Rail Micropower CMOS Comparator

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ABSTRACT

The design and test of a rail-to-rail micropower comparator in CMOS technology is described. The circuit is intended for implantable biomedical devices powered by batteries, with a total consumption of 500nA and operation down to supply voltages of 2V. The fabricated chip has a core die area of 0.27 mm² on a 2.4µm standard analog CMOS technology with a 0.85V nominal threshold voltage. The measured typical response time is 26 µs and the maximum offset voltage is 2mV. The limitations imposed by the low supply voltage are presented. The ways of overcoming these limitations, based on an accurate sizing of the transistors for operation in the weak and moderate inversion regions are studied. An approach based on a capacitive D/A converter, for the generation of a digitally programmable reference input is also presented.

1. INTRODUCTION

The design to be presented provides a combination of a rail-to-rail comparator and a digitally programmable reference level circuit. It was designed to fit the requirements for processing cardiac signals in implantable cardiac assist devices. Its goal is to detect signals in which amplitude exceeds a programmable threshold. Coupled with a successive approximation register function it may also be used to implement a successive approximation A/D converter. In Section II we introduce the characteristics of the signals to be processed and the constraints and initial specifications imposed by the field of application. Section III discusses the reasons that led to the chosen circuit topology. Special consideration is directed to the limitations associated with the low supply voltage requirement. Section IV presents the design methodology. Finally, Section V shows the simulation and measurement results and in Section VI the conclusions are summarized.

II. CHARACTERISTICS OF THE INPUT SIGNAL AND SPECIFICATIONS

The input signals of the circuit are cardiac signals previously filtered and amplified. A standard test waveform representing the cardiac signal is a triangular wave with 2ms rise time and 13ms fall time, sometimes referred as "Tokyo signal" [1]. This signal was applied throughout the circuit development to characterize performance aspects, like the comparator delay, that may depend on the input waveform. Neither this slow varying input signal nor the low-speed digital circuit that will process the output of the comparator in the final application are much demanding on the speed of the comparator. The initial specification for the comparator maximum delay was set to 0.5ms. The offset voltage specification derives from the minimum amplitude of the amplified cardiac signal that the comparator must

III. SELECTION OF THE CIRCUIT TOPOLOGY, LOW VOLTAGE DESIGN ISSUES

The factors that determined the circuit topology were the 2V supply voltage, the rail-to-rail input operation requirement and the very low supply current available.

The first important limitation set by the first two referred factors is to preclude the utilization of switches to handle the input signal unless an on-chip clock voltage-multiplication scheme is applied. We decided to try to avoid the clock voltage-multiplication alternative for simplicity and reliability reasons. The limitation on the application of switches is illustrated in Fig. 1 that plots the on-conductance and on-resistance of a switch, implemented as the parallel connection of a n-MOS transistor driven by complementary signals, as a function of the input voltage to the switch. The EKV model ([2,3]), that gives a representation valid in all regions of operation, was applied for the transistors. Fig. 1 shows that exists a gap in the input voltage range where both transistors are virtually cut off and the on-resistance is excessively high, even for the low speed application under consideration where a relatively high RC time constant would be acceptable. To decrease the on-resistance to acceptable levels by increasing the aspect ratio of the transistors is also non practical because it would lead to a unacceptable charge injection error.

We can classify comparator topologies according to their principle of operation in 3 broad classes [4,5,6]: (i) based on a high gain

handle that is 62.5mV. To have a maximum desired error of 10% in the amplitude detection, the offset voltage must be lower than 6.25mV.

The remaining specifications are associated to the incorporation of this cell to a battery operated implantable device. The full charge voltage of a Lithium-Iodine battery, applied in these devices, is 2.8V and the circuit must be fully operative for supply voltages up to 2.0V to guarantee operation during an acceptable lapse of time from the detection of the battery low condition to its replacement. Therefore the circuit was synthesized for 2V supply voltage. The target current consumption was set to 500nA.

Finally, the global architecture of the circuit where this comparator will be applied requires to handle rail-to-rail input signals. The target process is a standard 2.4µm analog CMOS process with double poly and double metal. This process is intended for 5V power supply and has nominal threshold voltages of nMOS (pMOS) transistors of 0.85V (-0.85V) with a minimum and maximum specified values of, respectively, 0.7V (-0.7V) and 1.0V (-1.0V). The fulfillment of the requirements in such a process, instead of a low-voltage specialized process with lower threshold voltages, enables a broader and cheaper range of possible target processes and foundries. However, the rail-to-rail operation with 2V power supply in such a process presents various challenges that are presented in the next section as well as the means we propose to overcome them.

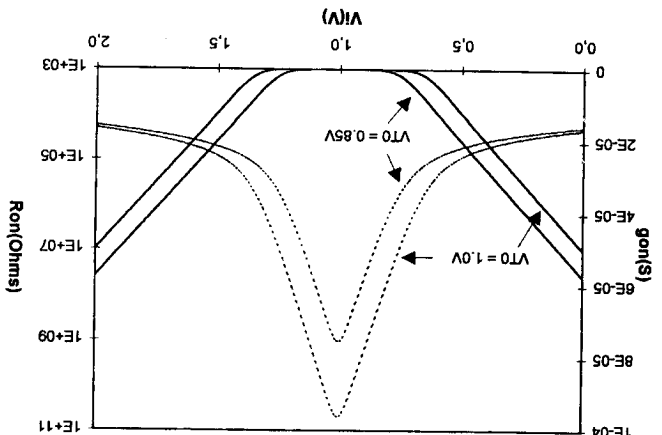


Fig. 1. On-conductance (solid lines) and resistance (dashed lines) of CMOS switch vs. input voltage for 2V power supply with nominal threshold voltage ($V_{T0} = 0.85V$) and maximum threshold voltage ($V_{T0} = 1.0V$).

amplifier, (ii) regenerative or positive feedback based and (iii) other switched-capacitor based architectures. Let's briefly describe each of these alternatives and evaluate them in light of the above requirements. Topology (i) is typically an open loop op-amp without internal compensation, sometimes in a multistage configuration to increase the speed. The second class of topology takes advantage of the positive feedback of a flip-flop structure to speed up the comparison. Although exceptions exist [4], these circuits usually apply one of the following principles: 1) a switch forces both outputs of the flip-flop to its state of unstable equilibrium and then lets it evolve to the final state according to the difference between the comparators inputs, or 2) with the flip flop structure turned off, a couple of switches force the outputs to be equal to the comparator inputs (or their amplified version) and then turns on the flip flop that will evolve to the final state. Therefore, this structure is discarded because it requires a switch operating in the whole power supply range. The topologies corresponding to the third class as well as the switched-capacitor techniques available to cancel out the comparator offset cannot be applied for the same reason.

Therefore, an op-amp based comparator topology was chosen. The input stage of the op-amp was determined by the rail-to-rail input requirement. The parallel combination of a n and a p differential pair was selected [7,8]. As will be shown in the next section, the low speed required as well as the high gain achievable operating in weak inversion allow the application of a single-stage structure. Therefore the amplifier topology shown in Fig. 2 was considered. It is an adaptation of the well known "symmetrical OTA" ([9]) structure to the rail-to-rail input stage.

Following the amplifying stage a minimum sized buffer (to minimize the amplifier load capacitance) and a latch are included. The latch may be also set to transparent mode to have a non-latched output. To guarantee the rail-to-rail input operation the following condition must be fulfilled:

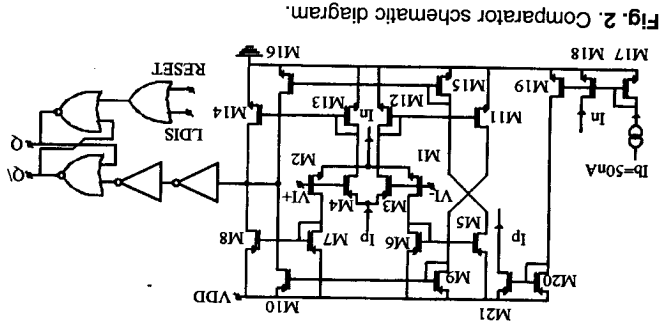


Fig. 2. Comparator schematic diagram.

IV. DESIGN METHODOLOGY

The sizing of the transistors was done through the method presented in [11,12]. This method based on the relation between the transconductance over drain current ratio (g_m/I_D) and the normalized current $I_D/(W/L)$, allows a unified treatment of all regions of operation of the MOS transistors. The application of this method, coupled with the EKV model with a set of parameters extracted from measurements for the target process, allows an accurate sizing of the transistors.

Fig. 4 shows the calculated and measured plots of g_m/I_D vs. $I_D/(W/L)$.

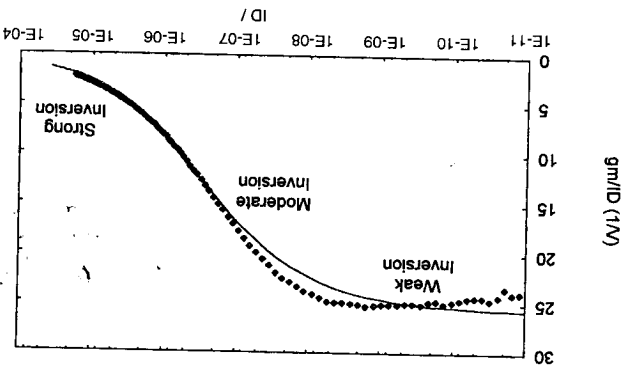
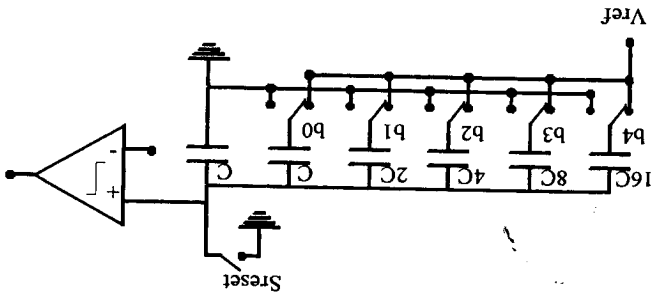


Fig. 4. Calculated (with the EKV model, solid line) and measured g_m/I_D vs. $I_D/(W/L)$ curve.

Fig. 3. Digitally programmable reference level circuit.



A digitally programmable reference level circuitry, compatible with the low power consumption required, was implemented based on the charge redistribution principle ([10]). Its schematic diagram is shown in Fig. 3. It is worth to note that in this structure all the switches are connected either to ground or to the power supply. Hence they do not have the above referred problem.

It seems difficult to satisfy condition (1) for $V_{DD} = 2V$ and the nominal V_{T0} of 0.85V, even more for the worst case condition of V_{T0} equal to 1.0V. However, the low V_{GS} values reached when operating in the moderate and weak inversion regions, together with the low saturation voltages allow to satisfy condition (1) even for V_{T0} of 1.0V. The next section describes the methodology followed to size the transistors to achieve the desired performance and to satisfy condition (1).

are not saturated does not exist.

tail pairs overlap near $V_{DD}/2$ and a gap where both M18 and M21 ensure they are saturated and operate as a current source. This condition ensures that the regions of operation of the n and p differential sources at the source of the differential pairs (M18 and M21) to the minimum voltage of the p and n input pair transistors and V_{DSAT} is where: V_{DD} is the supply voltage, V_{GSP} and V_{GSN} are the gate

$$V_{DD} \geq |V_{GSP}| + V_{GSN} + 2 \cdot V_{DSAT} \quad (1)$$

The synthesis procedure followed was:

1. From the total consumption specification and the circuit topology of Fig. 2 the current through each transistor was determined.

2. A g_m/I_D value is chosen for each transistor. From this value, proceeding as explained in [11,12], an $I_D/(W/L)$ value is obtained from the g_m/I_D vs. $I_D/(W/L)$ relation and then using the I_D value deduced in 1, (W/L) is obtained. The selection of g_m/I_D is based on different trade-offs depending on the considered transistor. Let us illustrate the approach with the case of the input differential pairs and of the current sources of the differential pairs.

2a. For the input differential pairs a higher value of g_m/I_D will give, on one hand, higher g_m (and hence speed) for a given I_D ; higher gain, lower gate-source, saturation and offset voltages; on the other hand a higher g_m/I_D requires lower values of $I_D/(W/L)$ and, therefore, wider transistors and parasites for a given current. Moreover, due to the flat characteristic of the g_m/I_D vs. $I_D/(W/L)$ curve near the weak inversion region (see Fig. 4), in this region a small increment of g_m/I_D requires a large increment of (W/L) . By exploring the design space through the g_m/I_D method we can choose the best compromise between performance and area. This approach led to a value of $g_m/I_D = 24V^{-1}$ for the input differential pairs transistors that corresponds to (W/L) of 29 for the MOS transistors and 69.5 for the pMOS transistors. The corresponding V_{GS} value for these sizes of the transistors is 0.7V for $V_{T0}=0.85V$ and 0.85V for $V_{T0}=1V$. These values of V_{GS} are compatible with condition (1) with a 2V power supply and V_{DSAT} corresponding to operation near weak inversion between 0.1 and 0.15V. It is interesting to note that a g_m/I_D value of $25V^{-1}$ (fully in the weak inversion region) leads to (W/L) values of 198 and 477 for the n and p transistor respectively, offering a negligible increase in gain or speed and V_{GS} at V_{T0} equal to 1.0V is only reduced from 0.85V to 0.77V.

2b. A higher g_m/I_D of the current mirror transistors implementing the current sources of the differential pair is reflected on a lower saturation voltage, but on poorer noise and matching properties and higher (W/L) values, that in the case of a current source is critical because non minimum length transistors are used to increase the output impedance. The chosen value of $22V^{-1}$ corresponds to (W/L) equal to 25 for the p transistors

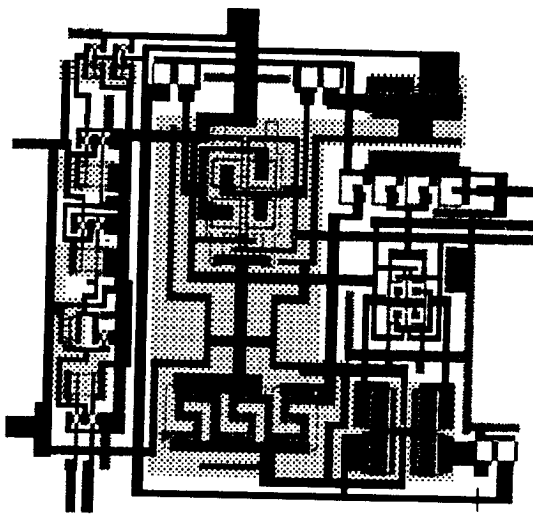
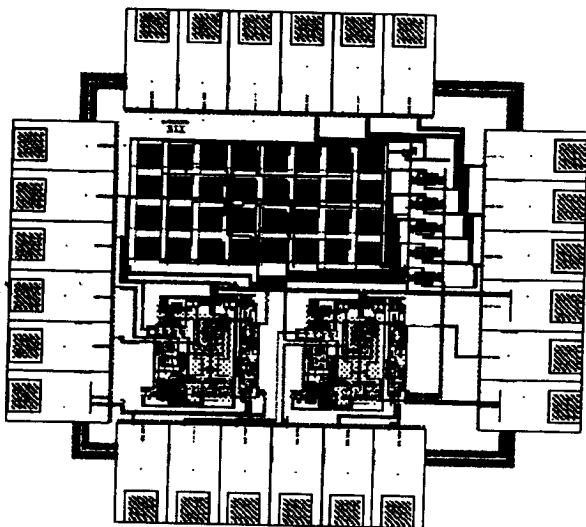


Fig. 5. Layout of comparator cell and complete chip.



The main circuit estimated and measured characteristics are summarized on Table 2. It is worth to briefly discuss the procedure followed to predict the offset voltage. Representative matching data of the standard deviation of the gain factor β ($\sigma_{\beta} = \sigma(\Delta\beta/\beta) = 0.2\%$) and threshold voltage V_{T0} ($\sigma_{T} = 2mV$) from references [13] and [14] were considered. Based on these data, the standard deviation of the current mismatch of a current mirror was calculated with equation (2).

V. SIMULATED AND MEASURED PERFORMANCES

| Transistor | W/L (μm) | g_m/I_D | Transistor | W/L (μm) | g_m/I_D |
|------------|-------------------|-----------|------------|-------------------|-----------|
| M1,2 | 278/4 | 24 | M17, M19 | 62.5/12 | 22 |
| M2,3 | 116/4 | 24 | M18 | 125/12 | 22 |
| M6-10 | 50/24 | 18 | M20 | 150/12 | 22 |
| M11-16 | 21/24 | 18 | M21 | 300/12 | 22 |

Table 1. Transistors dimensions. All dimensions are drawn dimensions that are shrunk by a 0.8 factor before fabrication.

Fig. 5 shows the layout of the comparator cell and of the complete chip including the reference generation circuitry.

3. An L value is chosen for each transistor. For the differential pair transistors, minimum length can be chosen. In our case a non-minimum value was chosen based on reliability considerations. For the current mirrors, the L value is derived from a trade-off between its influence on the current mirrors output impedance (and through it on the amplifier gain and common mode rejection ratio) and on the sizes of the transistors and parasites that influence the circuit speed.
4. The gain, speed and offset are predicted and checked against the desired performance. If they are not acceptable, the chosen g_m/I_D , L values or current consumption specification must be modified.

and 10.4 for the n transistors. This value allows an acceptable behavior of the current source with V_{DSAT} of 0.15V while the estimated rms noise current (0.24nA) is negligible with respect to the bias current of 100nA.

Table 2. Circuit characteristics.

| Design values and simulations results | Measurements | Comments. |
|---------------------------------------|-------------------------------|--|
| Supply Voltage | 2 V | |
| Total Standby Current | 500 nA | See Fig. 6 |
| Delay | 24.7 µs 15.5 µs 25.6 µs | Common mode input: 1.915V Common mode input: 1.15V Common mode input: 0.115V 10mV input step with 15mV overdrive, I _b (shown in Fig. 2) = 72 nA, VDD = 2.4V. |
| Maximum Delay | 170 µs | 32.5mV input triangular signal with 1.5mV overdrive, Common mode range: 0.2V - 1.8V. |
| Amplifier gain | 59 dB | Common mode level: 0.3V |
| Amplifier gain | 61 dB | Common mode level: 1.0V |
| Amplifier transition frequency | 235 KHz | Common mode level: 0.3V, Load Capacitance: 0.42pF |
| Amplifier transition frequency | 344 KHz | Common mode level: 1.0V, Load Capacitance: 0.42pF |
| Offset voltage | 6.6 mV | max. 2mV for 0.1V > V _i > 1.9V. The estimation is based on representative matching data from Refs. [13] and [14]; β standard deviation σ(Δβ/β)=0.2% and V _{T0} standard deviation σ _T =2mV. |

$$\sigma(\Delta I_D / I_D) = \sqrt{\sigma_{\beta^2} + \left(\frac{I_D}{g_m}\right)^2 \sigma_T^2} \quad (2)$$

Once the mismatch of the current mirrors is characterized, the worst combination of these mismatches is determined and the compensating offset voltage, taking into account also the mismatch of the differential pair, is calculated.

Fig. 6 shows the evolution of the current consumption with the input common mode voltage. The regions where either one or both differential pairs are operating are clearly visible.

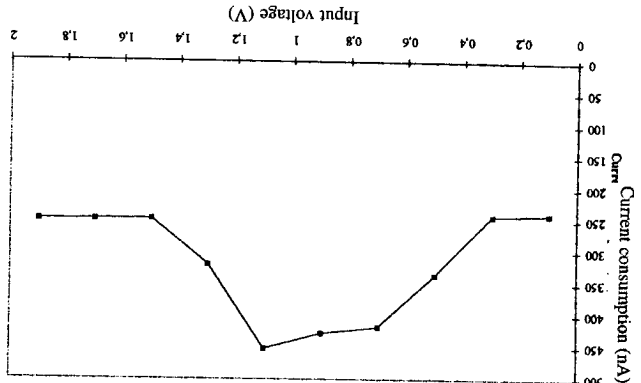


Fig. 6. Total current consumption vs. input common mode voltage.

VI. CONCLUSIONS

The design presented is compatible with the requirements for processing cardiac signals in implantable devices operating from 2V battery voltage with a total consumption of 500nA. The limitations imposed by the low supply voltage were discussed. The method followed to precisely size the transistors to operate in the weak and moderate inversion regions was presented. This approach allowed to lower the gate-source and saturation voltages, achieving rail-to-rail input operation even in the worst case condition of 1.0V maximum threshold voltage.

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