A 6-bit 500-Ms/s Digital Self-Calibrated Pipelined Analog-to-Digital Converter

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ABSTRACT

A 6-bit 500-Ms/s digital self-calibrated pipelined A/D converter is presented. Employing open-loop amplifiers in one bit conversion per stage architecture, the circuit operates in high speed and low power consumption. Using constant- g_m biasing technique, the open loop amplifier is process-and-temperature insensitive. Comparator offset and full-scale error are removed by digital self-calibrated correction mechanism. Designed by a 0.18- μm technology, the A/D converter operates at 500-MHz clock rate while dissipating 47 mW. The FFT simulation result shows that the SNDR is 33.52 dB at 51-MHz input frequency and 500-MSPS conversion rate.

Keywords: digital self-calibration, pipelined A/D converter

I. INTRODUCTION

High speed A/D conversion plays an important role in many applications, such as automated test equipment, oscilloscopes and digital read channel systems [4, 5]. However, low power consumption and small area are more important than higher conversion rate and input bandwidth for these purposes [5]. The pipelined architecture offers power and/or speed advantages over flash or multistep approaches due to concurrent processing of analog signals [2]. Conventional pipelined ADCs mostly use electronic feedback to achieve highly linear and drift insensitive transfer characteristics in their conversion stages. Especially, the converter front-end amplifiers with large open-loop gain are needed to achieve the desired accuracy. Due to stringent design trade-offs between low noise and high bandwidth requirements, precision amplifiers dominate the power dissipation in most high speed designs [6]. A major limitation in all algorithmic converters is given by the inherent matching of the passive components, which are hard to control and have a significant effect on yield. Self-calibration and correction techniques have been developed to reduce the effects of component mismatches [2].

A 1-b-per-stage design is particularly amenable because each stage is very simple and fast [1]. In this paper, we employ open-loop amplifiers, operating in high speed and low power consumption to the A/D converter in 1-bit conversion stage. Designed with constant- G_m bias technique, the gain of the open-loop

amplifier is determined by aspect ratio and is independent of process and temperature variation. The digital self-calibration architecture used here is a simple but effective mechanism, and tolerant of many nonidealities contributing to DNL and INL.

II. ADC ARCHITECTURE

Recently, the benefits of using open-loop structures in pipelined ADCs have been recognized and demonstrated. Using voltage-mode open-loop amplifier, the 12-bit ADC reported in [6] achieves 60% residue amplifier power savings over a conventional implementation. However, the amplification gain of open-loop differential pair depends significantly on process and temperature variation. In this work, we use an open-loop amplifier with a constant- G_m bias topology. This bias circuit can track the $g_m R$ constantly. Combining differential pair with source degeneration resistors, the voltage gain of the amplifier is linearized and can be expressed as

$$\text{gain} = \frac{g_m R_{LOAD}}{1 + g_m R_{SS}} \quad (1)$$

While the $g_m R$ is hold constant, the gain of the amplifier can be determined by aspect ratio.

In pipelined A/D converter design, power and speed optimization considerations favor the 1-bit conversion per stage approach [1]. In this case, the interstage gain is two and only one transition boundary occurs at the middle level of input in the transfer characteristic.

Fig. 1. Transfer characteristic for the 1-bit converter stage.

Figure 1 shows that the transfer characteristic of 1-bit converter stage is suffered from process and circuit nonidealities. Including comparator offset, full-scale error, amplifier offset and component mismatches, they lead to linearity error in the ADC. The key feature of digital self-calibration is that the errors at the carry transitions are directly measured under the same condition as during the normal conversion. The missing decision levels cannot be removed alone but can be eliminated by using gain less than 2 and extra stages of pipeline which gives enough redundancy in the analog decision levels.

For each calibrated stage, the calibration is performed as follows: force an analog input value of 0 V (differential) and force the digital decision to the left and to the right of the transition. Then, the height of the analog transition in the transfer characteristic (Fig. 1.) is estimated in the digital domain with the rest of the pipeline, assumed calibrated. The calibration coefficient representing the height of that transition is stored with the corresponding raw data bit.

The architecture of the ADC is shown in the Fig. 2. The ADC consists of six stages with calibration mechanism and four extra stages for compensation.

Fig. 2. The block diagram of ADC.

III. BUILDING BLOCKS

The architecture of a 1-bit conversion stage is shown in Fig. 3. With high input common voltage, only the pMOS pass gates with compensation dummy transistor is

Fig. 3. 1-bit conversion stage architecture.

inserted to operate as the T/H circuit. The comparator senses the input polarity and generates a 1-bit digital code, driving the reference circuit to settle the reference voltage to be added or subtracted by the open-loop amplifier. Logic gates are built in both of the T/H circuit and reference circuit to generate the valid control signal to calibrate the conversion stage. This design is compatibility with digital CMOS process since no MIM-caps are required.

A. Constant- G_m Bias Circuit

The Fig. 4 shows the constant- g_m bias topology. With the identical dimensions of transistors M_{S3} and M_{S4} , the PMOS devices require that current $I_{OUT} = I_{REF}$. We can write $V_{GS,S1} = V_{GS,S2} + I_{D,S2}R_S$, or

$$\sqrt{\frac{2I_{OUT}}{\mu_n C_{OX}(W/L)_N}} + V_{TH,S1} = \sqrt{\frac{2I_{OUT}}{\mu_n C_{OX}K(W/L)_N}} + V_{TH,S2} + I_{OUT}R_S. \quad (2)$$

Because transistor M_{S2} is a deep N-well device, the body effect of transistor M_{S2} can be neglected and hence

$$I_{OUT} = \frac{2}{\mu_n C_{OX}(W/L)_N} \frac{1}{R_S^2} \left(1 - \frac{1}{\sqrt{k}}\right)^2 \quad (3)$$

Thus, the transconductance of M_{S1} is equal to

$$g_{m,S1} = \frac{2}{R_S} \left(1 - \frac{1}{\sqrt{K}}\right). \quad (4)$$

This equation is independent of temperature, process and supply voltage. The gain, $g_{m,S1}R_S$, is equal to unity at $k = 4$.

Fig. 4. Constant- g_m bias topology.

B. Proposed Open-Loop Circuit

The proposed open-loop amplification circuit is shown in the Fig. 5. Two differential pairs with degeneration resistors constitute the open-loop amplifier. The nMOS transistors M_3 , M_4 , M_7 and M_8 in the deep N-well are immune from the body effect. In the input differential pair, the bias transistors $M_{1,2}$ and input transistors $M_{3,4}$ are all chosen as $(W/L)_{1,2,3,4} = 2(W/L)_{S1}$, where $(W/L)_{S1}$

Fig. 7. The simulation result of the INL and DNL with and without calibration.

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Fig. 8. The FFT spectrum with 51-MHz input frequency and 500-MSPS.

Fig.9. The layout of the ADC.

Table I. The Summary of the ADC

Technology	CMOS 0.18 1P6M
Resolution	6 bits
Conversion Rate	500 MSPS
Input Range	800 mV
Supply Voltage	1.8 V / 2.5 V
Power Consumption	
w/o clock buffer	34 mW
with clock buffer	47 mW
Chip Active Area	0.19 mm ²
DNL / INL	<0.9 LSB / <1.1 LSB
SNDR (500 MSPS)	
51 MHz	33.52 dB
244 MHz	27.30 dB