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9.3 A 12b 50MS/s Fully Differential Zero-Crossing-Based ADC Without CMFB

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As intrinsic device gain and power supply voltages decrease with CMOS technology scaling, it is becoming increasingly challenging for designers of conventional opamp-based switched-capacitor circuits to meet gain and output swing targets, and to ensure stability. Zero-crossing based circuits (ZCBC) are presented in [1-3] as an alternative architecture where each opamp is replaced with a current source and a zero-crossing detector. This changes the dynamics of the system while preserving the functionality. To further improve the robustness of ZCBC designs, we present a 50MS/s, 12b ZCBC pipelined ADC with fully differential signaling and automatic offset compensation.

A simplified schematic and timing diagram of 2 adjacent fully differential ZCBC pipeline stages are shown in Fig. 9.3.1 and Fig. 9.3.2, respectively. During the sampling phase, the input voltage is sampled onto sampling capacitors C1 and C2. The transfer phase begins when the pre-charge signal q⋄ turns on devices M1, M2, and M3 to initialize the load capacitors C3 and C4 to VDD and ground, respectively. This ensures that the differential output voltage starts below the minimum full-scale range. After the pre-charge phase, current sources I⋄, IΔ, and IC, which are split to avoid series switches for improved linearity [2], begin to charge the capacitors. The resulting voltage remains at zero-crossing detector (ZCD) detects the virtual ground condition, i.e. v⋄ = vΔ. At this point, the ZCD output q⋄ turns off the differential sampling switch M⋄ to sample the desired residue voltage.

In traditional opamp-based implementations, common-mode feedback (CMFB) is essential to keep both channels of the differential signal within range. CMFB is necessary because an opamp typically has common-mode gain. The common-mode gain of a ZCBC circuit, however, is relatively small and results from ramp rate differences between the positive and negative channels. The fully-differential ZCBC reported in [3] uses 2-phase charge transfer [1] and employs a conventional CMFB circuit during the coarse phase. In this work, single-phase operation is used for higher speed, making conventional CMFB circuitry difficult to implement. Therefore, the switching scheme shown in Fig. 9.3.1 and Fig. 9.3.2 is developed to remove the need for CMFB. In a conventional opamp-based implementation, switches M⋄ are left on for the entire transfer phase and the common-mode error is sampled onto the load capacitors. In this implementation, however, since these switches are turned on only during the pre-charge phase, the inside plates of the sampling capacitors are initialized to the common-mode voltage but are left to float when the output nodes ramp. The differential sampling switch (M⋄) holds the inside plates together so they float at the same potential and ensures that C3 and C4 charge at the same rate as C1 and C2. The parasitic capacitance on the inside plates holds the common-mode error, but the only effect is a small common-mode level shift at the input of the ZCD. Since the output common-mode is reset at the output of every stage by the preset operation, there is no common-mode level shift at the input of the ZCD. Since the output common-mode is reset at the output of every stage by the preset operation, there is no

The schematic in Fig. 9.3.1 shows a power-supply feed-through asymmetry to the output nodes v⋄. Such an asymmetry limits the high-frequency power-supply noise-rejection capabilities of a fully differential circuit. To improve the power-supply noise rejection, symmetric dummy current sources (not shown in Fig. 9.3.1) that are permanently disabled are added on both the positive and negative channels to provide 1st-order parasitic capacitance matching between the power supplies and the output nodes. This ensures that high-frequency power-supply noise coupling is common-mode and does not get sampled on the differential output nodes v⋄.

The 2CD used in this design is shown in Fig. 9.3.3. A pre-amplifier consisting of differential pair devices M1 and M2 is actively loaded with a current mirror that performs a differential-to-single-ended conversion. The output of the pre-amplifier drives a dynamic threshold detector that is similar to the dynamic ZCD used in [2]. To save power, the output of the detector is also fed back to M1 to turn off the bias current in the pre-amplifier as soon as the ZCD switches. The 2CD also features a digitally programmable offset. The schematic uses iterated-instance notation to succinctly draw parallel devices M2[0:3], M3[0:3], M4[0:3], and M5[0:3]. Binary-weighted device widths in M⋄[0:3] and M⋄[0:3] can be enabled with devices M3[0:3] and M4[0:3] to modify the current gain of the mirror current and provide a digitally programmable output to the ZCD.

Offset in a ZCBC design comes from conventional sources such as device mismatch as well as the voltage ramp overshoot due to the finite delay of the ZCD. In this work both sources of error are removed by a chopping technique similar to [4] and depicted in the block diagram of Fig. 9.3.4. The input signal is modulated with the chopping signal q⋄. The ADC converts the modulated signal, and then prior to demodulation, a Chopper Offset Estimator (COE) uses the digital output of the ADC to estimate the offset. The Offset Controller (OC) adjusts the digitally programmable offset of the ZCD in each stage to null the offset in the analog domain. The COE works by computing the mean of the signal over a block of data. The OC looks at the sign of the mean and drives an up/down counter to adjust one side of the current mirror in the ZCD.

This ADC employs redundancy that samples the input voltage at all MDAC and feedback capacitors [5] to reduce the ZCD noise referred to the ADC input rather than improve the closed-loop bandwidth. Additional redundancy is further used to reduce the required output voltage range. This is implemented by using a 3.3b sub-ADC in each 4+ gain MDAC, increasing the reference voltages to be larger than the input voltage range, and placing the bit decision boundaries appropriately (see the block diagram and residue plot of Fig. 9.3.4). This improved effective output range gives more headroom to the current sources for improved linearity.

Implemented in 0.3mm 2 in a 90nm CMOS process with a power supply voltage of 1.2V, this design consumes 4.5mW at 50MS/s. The DNL, INL, and frequency response are shown in Fig. 9.3.5. The DNL is within ±0.5 LSB and the INL is within ±3 LSB on a 12b scale. The noise floor is 72.3dB below full scale, showing that this design is distortion-limited. The dominant source of distortion comes from offsets in the bit-decision comparators that make up the sub-ADC of each stage. These offsets are larger than Monte-Carlo simulation predicted and cause the cascaded current sources to leave saturation when the residue nears the positive reference. The measured ENOB to a near-Nyquist-rate input tone is 10.6b at 50MS/s and 25MS/s, respectively, and the resulting FOM is 86fJ/conversion-step and 98fJ/conversion-step. The die micrograph is shown in Fig. 9.3.7.

References:
Figure 9.3.1: Simplified schematic of 2 stages of a fully differential ZCBC pipelined ADC.

Figure 9.3.2: Timing diagram for ZCBC pipelined ADC.

Figure 9.3.3: Zero-Crossing Detector (ZCD) used in this design featuring digital offset programmability.

Figure 9.3.4: Pipelined ADC showing systematic offset cancellation technique via chopper offset estimation.

Figure 9.3.5: Measured DNL, INL, and frequency responses.

Figure 9.3.6: Summary of measured performance.
Figure 9.3.7: Die micrograph.