

## 27.6 A 0.7pF-to-10nF Fully Digital Capacitance-to-Digital Converter Using Iterative Delay-Chain Discharge

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Capacitance sensors are widely used to measure various physical quantities, including position, pressure, and concentration of certain chemicals [1-6]. Integrating capacitive sensors into a small wireless sensor system is challenging due to their large power consumption relative to the total system power/energy budget, which can be as low as a few nW [4]. Typical capacitance-to-digital converters (CDCs) use charge sharing or charge transfer between capacitors to convert the sampled capacitance to voltage, which is then measured with an ADC [1-6]. This approach requires complex analog circuits, such as amplifiers and ADCs, increasing design complexity and often increasing power consumption. Moreover, the initial capacitance to voltage conversion essentially limits the input capacitance range because of output voltage saturation. This paper presents a fully digital CDC that is based on the observation that when a ring oscillator (RO) is powered from a charged capacitance, the number of RO cycles required to discharge the capacitance to a fixed voltage is naturally linear with the capacitance value. This observation enables a simple, fully digital conversion scheme that is inherently linear. As a result, the proposed CDC performs conversion across a very wide capacitance range from 0.7pF to over 10nF with < 0.06% linearity error. The CDC senses 11.3pF input capacitance with 35.1pJ conversion energy and 141fJ/c-s FoM.

Figure 27.6.1 explains the proposed conversion method. The top node of sensed capacitor  $C_T$  is directly connected to the supply node of a ring oscillator. This node is initially charged to  $V_{HIGH}$ , and is then discharged gradually as the inverter RO oscillates. As transitions occur in the RO signals, the RO draws some charge from  $C_{SENSE}$ , gradually lowering  $V_{CT}$ . As a result, the RO propagation delay increases, which is compared to a constant delay reference. A counter records the RO transition count until the period delay becomes longer than the reference delay, which becomes the output code  $D_{OUT}$ .

Since the RO delay only depends on  $V_{CT}$  (neglecting noise initially),  $D_{OUT}$  is equal to the number of RO transitions while  $V_{CT}$  is discharged from  $V_{HIGH}$  to some constant voltage,  $V_{LOW}$ . During conversion, at any particular  $V_{CT}$  value the amount of charge withdrawn per RO transition only depends on  $V_{CT}$  at that time. Therefore, the number of transitions required to reduce  $V_{CT}$  by a certain small voltage is proportional to input capacitance  $C_{SENSE}$ . As this is true at any  $V_{CT}$  level, the output code  $D_{OUT}$ , the sum of the transition counts across all continuous small intervals from  $V_{HIGH}$  to  $V_{LOW}$ , is also proportional to  $C_{SENSE}$ . As the RO draws charge directly from  $C_{SENSE}$  without initial capacitance to voltage conversion, the CDC input capacitance range is essentially unlimited, constrained only by the counter size. This is desirable when the  $C_{SENSE}$  range is uncertain at design time. Furthermore, energy used to charge  $C_{SENSE}$  is reused to oscillate the RO, reducing overall power consumption.

Figure 27.6.2 shows the detailed implementation of the CDC circuit and its operation. Here an inverter chain is used in place of an RO to discharge  $C_{SENSE}$ ; it is a 16-stage chain that is identical to the reference delay generator. Because of the identical structures, conversion stops when  $V_{CT}$  drops below  $V_{LOW}$ . The number of stages in the inverter chain is chosen for optimal SNR per conversion energy, where the energy to charge  $C_{SENSE}$  is balanced with the energy consumed by other blocks. The two propagation delays are compared by three delay comparators, which have a similar structure to an RS latch. The bottom comparator compares the propagation delay of falling edges, and the middle one compares the rising edges. Whenever the reference delay is shorter than the  $C_{SENSE}$  discharge delay chain, the comparator output pulses once, increasing counts stored in the *sub1* and *sub2* counters. A third counter tracks the main oscillation triggering signal. After each comparison, the next edge generator block triggers the next discharge and delay comparison, maintaining oscillation. All blocks except the  $C_{SENSE}$  delay chain operate at  $V_{LOW}$ , and a level converter drives the two delay chain inputs with  $V_{HIGH}$ .

As shown in the timing diagram of Fig. 27.6.2, conversion starts by precharging  $C_{SENSE}$  to  $V_{HIGH}$ . This is followed by *Sense* rising, triggering the first edge to propagate through the two delay chains. The top comparator takes in a slightly delayed version of the reference delay and determines when to finish the overall conversion, which occurs when  $V_{CT}$  becomes lower than  $V_{LOW}$  by some margin. As  $V_{CT}$  approaches  $V_{LOW}$ , the bottom two delay comparators pulse  $CK_1$  and  $CK_2$ .

They initially pulse sporadically, due to noise, and then more frequently as  $V_{CT}$  crosses  $V_{LOW}$ . Just before conversion finishes, these two comparators pulse every cycle. When the top comparator pulses *Finish*, *Sense* is turned off and oscillation stops. Final  $D_{OUT}$  is the total count of comparator outputs for which  $V_{CT} > V_{LOW}$ , and is calculated as  $2 \times D_{MAIN} - (D_{SUB1} + D_{SUB2})$ .

The use of three comparators is designed to increase SNR by averaging noise over many comparisons when  $V_{CT}$  is near  $V_{LOW}$ . Comparing both rising and falling edges doubles the number of comparisons. By extending the conversion to where  $V_{CT}$  falls some margin below  $V_{LOW}$ , comparisons are performed through the whole noisy region around  $V_{LOW}$ , whereby false " $V_{CT} < V_{LOW}$ " decisions above  $V_{LOW}$  are stochastically compensated by false " $V_{CT} > V_{LOW}$ " decisions below  $V_{LOW}$ . Simulation shows that energy increases by 3% compared to the standard approach of stopping conversion immediately after the first comparison triggers, while overall conversion noise is square-rooted. In addition, the distribution of  $D_{OUT}$  using this scheme is centered at the number of exact counts from  $V_{HIGH}$  to  $V_{LOW}$ , thereby improving output code linearity.

The CDC measures the capacitance between one input node and ground, but some applications require sensing the capacitance value between two input nodes, excluding parasitic capacitance to ground. We accomplish this through three conversions, as shown in Fig. 27.6.3. First, node B is connected to ground and the capacitance between node A and ground is measured, which includes parasitic capacitance  $C_{PA}$ . Second, nodes A and B are flipped and  $C_{SENSE} + C_{PB}$  is measured. Finally, both A and B nodes are connected to  $V_{CT}$  to measure  $C_{PA} + C_{PB}$ . By adding the first two codes and subtracting the third, the parasitic capacitance is canceled. While this requires three conversions, parasitic capacitance typically remains unchanged or changes slowly (Fig. 27.6.5, bottom) and the parasitic cancelation can be performed infrequently, amortizing its overhead.

The output code varies as temperature or supply voltage changes. This code deviation is removed by one-point calibration (Fig. 27.6.3). In a calibration phase,  $V_{CT}$  is connected to an internal reference capacitor with known capacitance  $C_{REF}$ , and the ratio of  $C_{REF}$  to corresponding  $D_{OUT}$  is stored. In subsequent normal conversion, digital output codes are converted to actual capacitance values by multiplying the code and the stored ratio. If the supply voltage changes sufficiently slowly, this calibration can be re-done occasionally.

The CDC is fabricated in 40nm CMOS and tested with  $V_{HIGH}=1.0V$  and  $V_{LOW}=0.45V$ . Core circuit area without testing circuits and internal capacitors is  $0.0017mm^2$ . Figure 27.6.4 shows the test chip has a very wide input capacitance range from 0.7pF to 10nF with a small linearity error of < 0.06%. Measured output noise percentage reduces as  $C_{SENSE}$  increases due to noise averaging. At 11.3pF, the CDC has 0.109% resolution, 35.1pJ total conversion energy (including both  $V_{HIGH}$  and  $V_{LOW}$ ), and 141fJ/c-s FoM. FoM increases monotonically with the sensed capacitance. Figure 27.6.4 also shows output code sensitivity to temperature improves by 145x (from 2247ppm/°C to 15.5ppm/°C) due to calibration. Results with an actual pressure sensor (Fig. 27.6.5) demonstrate 1.39mmHg resolution with parasitic cancelation. Figure 27.6.6 summarizes CDC performance and compares it with prior work.

### Acknowledgements:

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### References:

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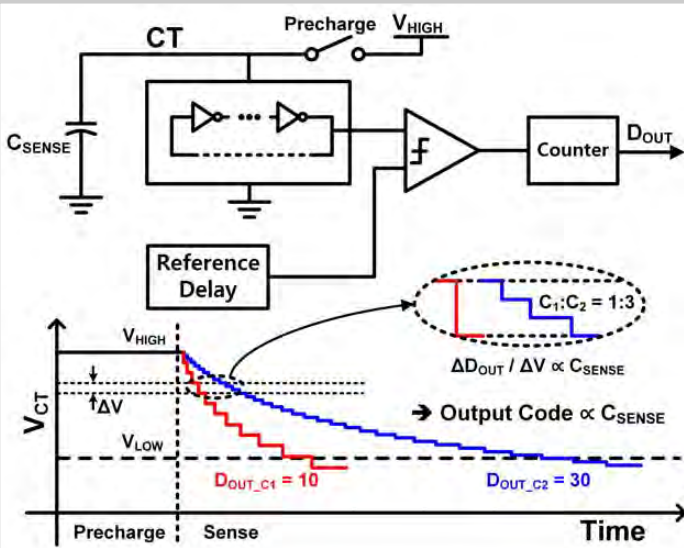


Figure 27.6.1: Basic structure and operation scheme of the proposed CDC.

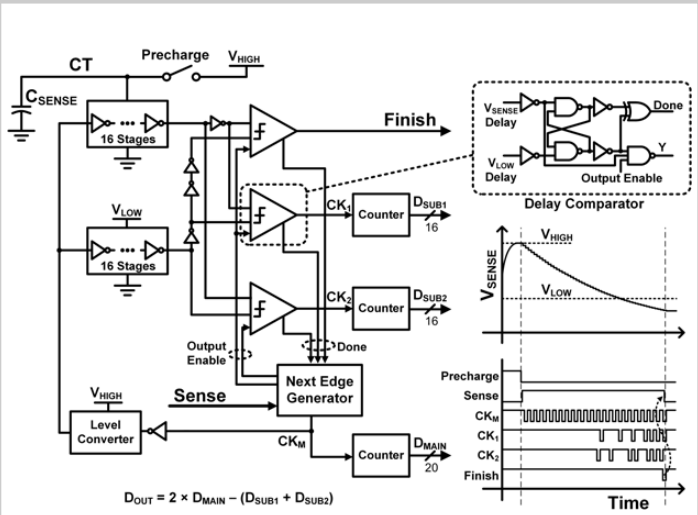


Figure 27.6.2: Detailed implementation of the CDC.

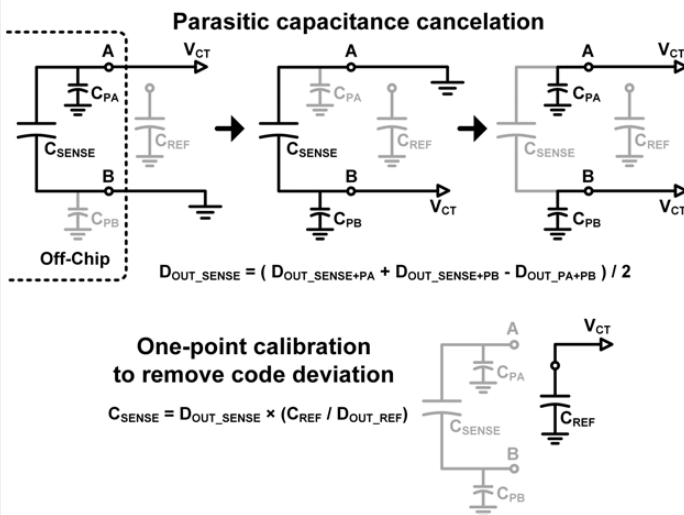


Figure 27.6.3: Techniques for removing parasitic capacitance (top) and code deviation due to voltage and temperature sensitivity (bottom).

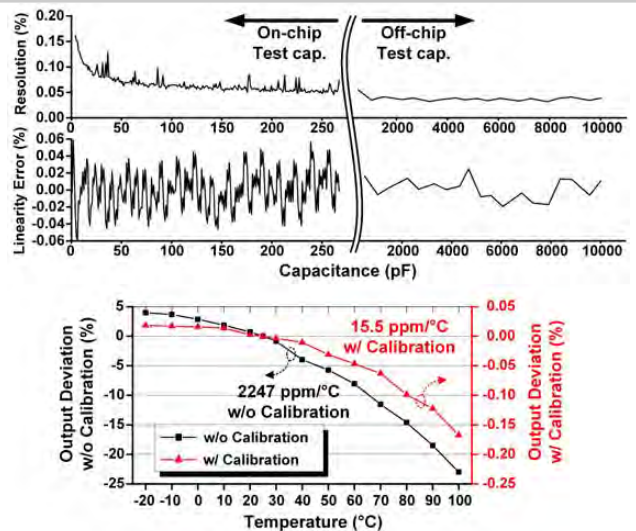


Figure 27.6.4: CDC resolution and linearity error (top), and its output temperature sensitivity (bottom).

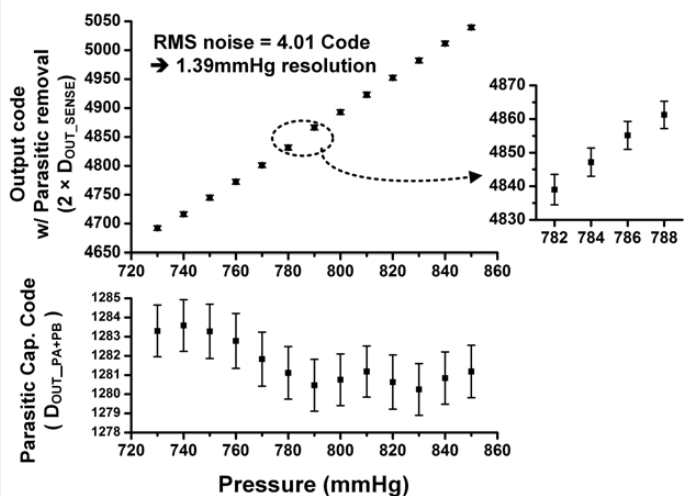


Figure 27.6.5: Measured result with capacitive pressure sensor with parasitic cancellation.

	[1] JSSC 09	[2] ISSCC 14	[3] VLSI 14	[5] JSSC 13	[6] JSSC 12	This work
Technology	1.5μm CMOS	0.18μm CMOS	0.18μm CMOS	0.16μm CMOS	0.35μm CMOS	40nm CMOS
Method	CDS + Cyclic ADC	CDS + SAR	SAR + ΔΣ	ΔΣ	Period Modulation	Iterative Delay-Chain Discharge
Input range	N/R	2.5 – 75.3pF	0 – 24pF	0.54 – 1.06pF	N/R	0.7pF – 10nF
Resolution	75aF	6.0fF	0.16fF	70aF	N/R	0.109% <sup>1</sup> (12.3fF)
Meas. Time	0.5ms	4ms	230μs	0.8ms	7.6ms	19.06μs <sup>1</sup>
Power	36μW <sup>5</sup>	160nW	33.7μW	10.3μW <sup>5</sup>	211μW <sup>5</sup>	1.84μW <sup>1</sup>
Conversion Energy <sup>2</sup>	18nJ	640pJ	7.75nJ	8.26nJ	1.61μJ	35.1pJ <sup>1</sup>
FoM <sup>3</sup> (fJ/c-s)	22000	181	175	3900	139000	141 <sup>1,4</sup>

1 Measured when sensing 11.3pF capacitance w/o parasitic cancellation or calibration.  
 2 Conversion Energy = Power \* (Meas. Time)  
 3 FoM = (Conversion Energy) / 2<sup>(20 log (Input range / 2 Sqrt(2)) / Resolution) - 1.76) / 6.02  
 4 Input range is assumed to be 0.7pF – 11.3pF for this calculation  
 5 Estimated number from the paper  
 N/R: Not reported</sup>

Figure 27.6.6: Performance summary and comparison.

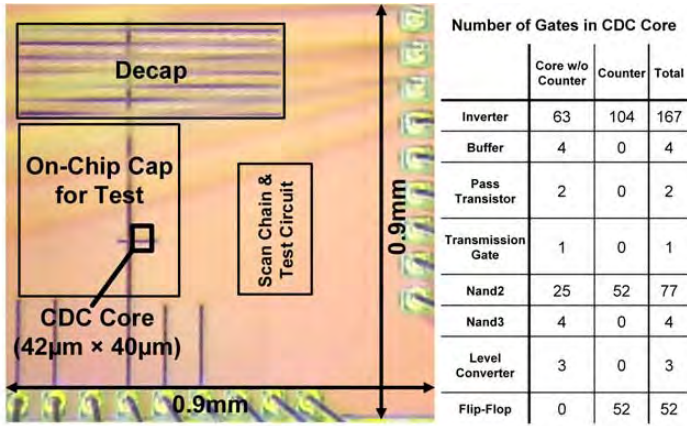


Figure 27.6.7: Die micrograph of the 40nm CMOS test chip.