Capacitance Mismatch Evaluation for Low-power Pipeline ADC Design

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Abstract: It is known that the capacitance mismatch affects the nonlinearity of pipeline analog-to-digital converters (ADCs) and that larger capacitances tend to have a smaller mismatch. On the other hand, small capacitances are desirable because they reduce the current consumption. Therefore, to achieve a low-power, high-resolution ADC, it is necessary to perform capacitance mismatch evaluation and determine the appropriate capacitance.

In this letter, a method for capacitance-mismatch measurement and the experimental results are described. The measurement results show that the mismatch is sufficiently small at the capacitance determined based on the noise requirement for 0.13- μm CMOS ADCs.

Keywords: Analog-to-digital converter, ADC, capacitance mismatch

Classification: Integrated circuits

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1 Introduction

Pipeline architecture is often adopted for high-speed and medium-resolution analog-to-digital converters (ADCs). An advantage of pipeline ADCs is the large tolerance to the comparator offset. The dominant causes of the non-linearity are the gain error of the residue amplifiers and the errors of the internal digital-to-analog converters (DACs) [2]. In many cases, a switched-capacitor multiplying digital-to-analog converter (MDAC) is employed as the residue amplifier and the internal DAC. If the operational amplifier in the MDAC is assumed to be ideal, the error can be attributed to the mismatch between the capacitors in the MDAC.

One of the easy ways to reduce the mismatch is to make the capacitances large. For example, if C_U is the unit capacitance and σ_u is the standard deviation of the unit capacitances on a chip, the standard deviation of a capacitor that consists of two parallel unit-capacitors is $\sqrt{2}\sigma_u$, and the mismatch $\sqrt{2}\sigma_u/2C_U$ is smaller than σ_u/C_U .

However, a larger capacitance requires a larger drive current, and this results in larger power dissipation. Consequently, identification of the optimum capacitance value for MDACs is an important issue in low-power and high-resolution pipeline ADC design, and capacitance mismatch evaluation is indispensable.

In this letter, a capacitance mismatch evaluation method using a switched capacitor circuit and the experimental results are described.

2 Mismatch detection circuit

Our aim is to design a 10-bit or higher resolution ADC using CMOS technology, and the mismatch requirement is $2^{-10} \simeq 0.1\%$ or smaller. In some reported 10-bit ADCs, 1 pF or smaller capacitors are used [5, 6] and effective number of bits (ENOB) of more than 9.0 bits have been achieved. This means that a sufficiently small noise and sufficient linearity have been obtained with such small capacitors. To achieve a lower-power ADC, mismatch measurements at smaller capacitances are required.

At such small capacitances, however, accurate measurement is difficult because the parasitic capacitances on pads have similar values. Some techniques to measure capacitance mismatch have been reported [3, 4], but the techniques need precision measurement equipment.

Fig. 1 shows the proposed capacitance-mismatch detection circuit, which requires an ordinary clock source and an ordinary sampling oscilloscope, but no high precision equipment. The circuit consists of 4 capacitors to be tested, an operational amplifier with feedback capacitors, and switches.

To simplify the description, we make some assumptions. The validity of the assumptions is discussed in the next section. The assumptions are:

1. the offset of the amplifier is negligible,
2. the gain of the amplifier and the common mode rejection ratio (CMRR) are high enough,

Fig. 1. Schematic of the proposed mismatch detection circuit: (a) in reset mode (b) in mismatch detection mode.

3. the average of C_1 , C_2 , C_3 , and C_4 is C_0 and the standard deviations σ_i for $\Delta C_i = C_i - C_0$ ($i = 1, 2, 3$, and 4) are the same for all,
4. the feedback capacitors have no mismatch, i.e. $C_{F1} = C_{F2} = C_F$, and the deviation of the capacitance ratio C_F/C_0 is negligible,
5. all parasitic capacitances are negligible, and
6. the switches are ideal; i.e., there is no clock feed-through effect.

Under these assumptions, the circuit operation is considered to be composed of the following two steps.

First, the input switch SW_X connects the input and the capacitors directly as shown in Fig. 1 (a), and switches SW_1 and SW_2 short the input and output of the amplifier. We call this status ‘reset mode’. The two input terminals of the amplifiers are virtually shorted, and the voltage at the input terminals is the same as the output common-mode voltage V_{COM} . The capacitors have charges of $Q_{1R} = C_1(V_{IN1} - V_{COM})$, $Q_{2R} = C_2(V_{IN2} - V_{COM})$, $Q_{3R} = C_3(V_{IN1} - V_{COM})$, and $Q_{4R} = C_4(V_{IN2} - V_{COM})$.

Next, as shown in Fig. 1 (b), the switches SW_1 and SW_2 are opened and the input switch SW_X turns to opposite status. The amplifier with feedback capacitors operates as a charge-to-voltage converter. We call this status ‘mismatch detection mode’. The charges on the capacitors change to $Q_{1D} = C_1(V_{IN2} - V_{COM})$, $Q_{2D} = C_2(V_{IN1} - V_{COM})$, $Q_{3D} = C_3(V_{IN2} - V_{COM})$, and $Q_{4D} = C_4(V_{IN1} - V_{COM})$, and the output voltage V_{OUT} is

$$V_{OUT} = \frac{(Q_{1R} - Q_{1D}) - (Q_{2R} - Q_{2D})}{C_{F1}}$$

$$\begin{aligned}
 & - \frac{(Q_{3R} - Q_{3D}) - (Q_{4R} - Q_{4D})}{C_{F2}} \\
 = & \frac{C_1 - C_2 - C_3 + C_4}{C_F} (V_{IN1} - V_{IN2}) \\
 = & \frac{\Delta C_1 - \Delta C_2 - \Delta C_3 + \Delta C_4}{C_F} V_{IN}, \tag{1}
 \end{aligned}$$

where $V_{IN} = V_{IN1} - V_{IN2}$. The output voltage depends on the capacitance mismatch. The detection circuit repeats these two steps and the output is observed with an oscilloscope. In our experiments, we used a sampling oscilloscope with an averaging function to reduce random noise.

From Eq. (1), the standard deviation of the output voltage σ_v is

$$\sigma_v = \frac{V_{IN}}{C_F} \sqrt{\sum_{i=1}^4 \sigma_i^2} = \frac{2V_{IN}}{C_F} \sigma_1. \tag{2}$$

Therefore, the standard deviation of the capacitance mismatch is obtained to be

$$\frac{\sigma_i}{C_0} = \frac{C_F}{2C_0} \frac{\sigma_v}{V_{IN}}. \tag{3}$$

Consequently, the deviation of the capacitance mismatch is obtained by measuring the output voltages of the samples.

3 Justification of the assumptions

In this section, the validity of our assumptions is discussed. First of all, the allowable estimation error for the standard deviation should be clarified. We need only a rough estimate, and an error of 30% is acceptable. For example, when the capacitances are determined to make 3σ within 0.5 LSB, 99.7% of capacitor pairs have mismatches of less than 0.5 LSB. Even if the real σ is 30% larger than estimated value, still 96.4% of capacitor pairs will have mismatches of less than 0.5 LSB. Therefore, an estimation error of 30% is acceptable.

Next, the nominal value in the mismatch detection circuit is discussed. The nominal value of C_F/C_0 is set as 0.5, and $V_{IN} = V_{IN1} - V_{IN2}$ is set as 0.5 V. According to Eq. (3), a 1-mV deviation at the output corresponds to a capacitance mismatch of 0.05%.

Under the above conditions, the following 6 effects are negligible.

1. Offset of the amplifier: The input offset of the amplifier is estimated to be a few mV, and is detectable when the circuit is in the ‘reset mode’. The offset effect is easily eliminated.
2. Gain and CMRR: The circuit simulation shows that the dc gain of the amplifier is more than 60 dB and CMRR is more than 80 dB. Both figures are sufficiently high, because the error caused by the finite loop gain is less than 1% and the common-mode input signal is attenuated rather than amplified.

3. Deviation of C_F/C_0 : The deviation is estimated to be less than a few %, i.e. the deviation is much smaller than the nominal value. The deviation may result in a σ_v error of a few %. Note that the nominal values of $\Delta C_i (i = 1, 2, 3, 4)$ are 0, and σ_i is much larger than the nominal value. This difference makes σ_i detectable, and makes the deviation of C_F/C_0 negligible.
4. Mismatch of the feedback capacitors: If $C_{F1} = C_F(1 + \alpha)$ and $C_{F2} = C_F(1 - \alpha)$, α is estimated to be less than a few %. Equation (1) can be rewritten as

$$\begin{aligned} V_{OUT} &= \frac{Q_{12}}{C_{F1}} - \frac{Q_{34}}{C_{F2}} \\ &\simeq \frac{Q_{12}(1 - \alpha) - Q_{34}(1 + \alpha)}{C_F} \\ &= \frac{Q_{12} - Q_{34} - \alpha(Q_{12} + Q_{34})}{C_F} \end{aligned} \quad (4)$$

where $Q_{12} = (Q_{1R} - Q_{1D}) - (Q_{2R} - Q_{2D})$ and $Q_{34} = (Q_{3R} - Q_{3D}) - (Q_{4R} - Q_{4D})$. Equation (4) shows that α can affect the output voltage when $|Q_{12} - Q_{34}| \ll |Q_{12} + Q_{34}|$. However, the standard deviations of $Q_{12} - Q_{34}$ and $Q_{12} + Q_{34}$ should be the same, and α is much smaller than 1. Thus the effect of α on the standard deviation is negligible.

5. Parasitic capacitances: The parasitic capacitances on the input nodes of the amplifier are negligible, because the nodes are virtually shorted. The capacitances on input switch SW_X and the output nodes may affect the frequency response of the circuit, but do not contribute to the output voltage.
6. Clock feed-through via SW_X : Even though the SW_X , which consists of MOS transistors, injects a charge into the capacitors, the injected charge is absorbed by the low-impedance voltage source. As a result, the clock feed-through via the SW_X is negligible.

As for SW_1 and SW_2 , the charge from the switches to the input nodes of the amplifier is considerable. When the injected charges are the same, the common-mode input voltage may change without affecting the differential output. When there is a mismatch, the output voltage changes.

To compensate the mismatch effect, two steps are added to the circuit operation and the four step operation is as follows:

In the first step, the circuit is set to reset-mode. The input switch SW_X connects the input and capacitors directly as shown in Fig. 1 (a), and switches SW_1 and SW_2 short the input and the output of the amplifier.

Next, switches SW_1 and SW_2 are opened, and the amplifier with feedback capacitors operates as charge-to-voltage converter. The switch SW_X remains connected as in reset mode. In this situation, the charge injection through SW_1 and SW_2 is converted to the output voltage. This mode is called the clock feed-through detection mode.

Fig. 2. Simplified output signal waveform

In the 3rd step, the circuit is set to reset mode again.

In the last step, the circuit is set to mismatch detection mode. In this mode, the switches SW_1 and SW_2 are opened, and the input switch SW_X is set to the opposite status as shown in Fig. 1 (b).

A diagram of the output signal waveform is shown in Fig. 2, and the desired output voltage V_M is obtained as the difference between the voltage in the mismatch detection mode and that in the clock feed-through detection mode.

4 Measured results

The proposed circuit was fabricated using 0.13- μm CMOS technology, and 32 samples were used. The capacitance for mismatch measurement was around 0.1 pF. The average was 0.4 mV and the standard deviation was 0.8 mV. According to Eq. (3), a standard deviation of 0.8 mV corresponds to a capacitance mismatch of 0.04%.

5 Conclusions

The measured results indicate that the capacitance mismatch is not a serious problem for 10-bit ADC design because capacitances of which 3σ is 0.5 LSB is 0.6 pF, and the capacitance of 0.6 pF is smaller than capacitance to meet the noise requirements.

The trend toward deep submicron technology is reducing the power supply voltage, which in turn reducing the acceptable limits of the signal swing. This means that the permissible noise level is decreasing and a larger capacitance is required for the same resolution ADC. On the other hand, it is difficult to increase the capacitance per unit area, and the mismatch is not changing correspondingly. Therefore, at a 10 bit resolution, the capacitance mismatch is less important for low-power ADC design than thermal noise.

Acknowledgement

This work was supported by the Telecommunication Advancement Organization of Japan (TAO).