A Novel Technique for Improving Temperature Independence of Ring-ADC

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Abstract

A new temperature compensation technique for ring-oscillator-based ADC is proposed in this paper. It employs a novel fixed-number-based algorithm and a CTAT current biasing technology to compensate the temperature-dependent variations of the output, thus eliminates the need of digital calibrations. Simulation results prove that, with the proposed technique, the resolution under the temperature range of 0°C to 100°C can reach a 2-mV quantization bin size with an input voltage span of 120mV, at the sampling frequency f_s=100KHz.

1. Introduction

Windowed analog-to-digital converters (ADCs) are widely used in digital DC-DC converters due to the fact that the output voltage is regulated in the vicinity of the reference voltage. In recent years, synthesizable windowed ADCs based on delay-line [1, 2, 3] or ring-oscillator [4, 5] structures have been reported. These structures have the advantages of low-power, low-area and high-resolution so that they are very suitable for digital DC-DC applications. However, one severe problem for them is that the digital output data is temperature-dependent since the delay of the delay-cells can’t be precisely controlled against the temperature variations. Digital calibrations can be accomplished in a number of ways. A simple strategy is to subtract a reference value from the converted output [1] or to obtain a correction ratio to a reference value [6]. Since the variations of the digital output due to temperature variation are not pure dc offset or gain error, these types of calibrations may be imprecise. A look-up table can improve the precision. However, it will take a large chip area. Besides, an additional ADC is necessary to convert the measured analog temperature value into a digital one, which will further increase the circuit’s complexity and the power consumption.

In this paper, we propose a novel temperature compensation technique for ring-oscillator-based ADCs. We put forward a fixed-number-based algorithm and a Complementary-to-Absolute-Temperature (CTAT) current biasing scheme. With our technique, a ring-ADC can achieve dramatic temperature independency without the complex digital calibrations.

The proposed ring-ADC has been simulated with a typical 0.35μm CMOS mixed signal process (Charted). For comparison, a traditional ring-ADC that employs a normal fixed-time-based algorithm [5] and conventional biasing current source is also simulated. Results reveal that, with the temperature range of 0°C to 100°C, the input voltage span of 120mV and the sampling frequency of 100KHz, the proposed ring-ADC can reach a 2-mV quantization bin size while the traditional ring-ADC only 16-mV.

2. System Architecture

The principle of the proposed temperature compensation technique is explained with a ring-ADC system shown in Figure 1. It consists of a CTAT biasing current source, a ring oscillator pair and a synthesizable digital block.

The CTAT biasing current source is used to generate a temperature compensated current for the two ring-oscillators. The error voltage between $V_{io}$ and $V_{ref}$, $V_e$, brings on error current between two branches, $I_{bias,1}$-$I_{bias,2}$, thus different frequencies at the two ring-oscillators are generated. The two frequencies are captured by two counters, then our fixed-number-based algorithm is applied to derive the temperature-independent output.
3. Fixed-Number-Based Algorithm

The right part of Figure 1 can also be used to explain the fixed-time algorithm [5] if we replace the module fixed-number counting controller with a fixed-time one. The counters are reset at the beginning of each sampling cycle and stopped after a fixed period of time. Then, one counter’s output is subtracted from the other. The result, $C_e$, is given by [5]:

$$C_e = T_{ADC} \times K_f \times (I_{bias,1} - I_{bias,2})$$  \hspace{1cm} (1)

where $T_{ADC}$ is the ADC sampling period, $K_f$ is the constant characterizing the ring oscillator frequency sensitivity to its biasing current. $I_{bias,1}$ and $I_{bias,2}$ are the biasing currents of the two branches. The digitized error voltage can be calculated by scaling $C_e$. Actually, the error frequency, $f_e = f_1 - f_2 = K_f \times (I_{bias,1} - I_{bias,2}) \propto V_e$ \hspace{1cm} (2)

where $V_e$ is the error input voltage. Then,

$$C_e = T_{ADC} \times (f_1 - f_2) = T_{ADC} \times f_e$$ \hspace{1cm} (3)

When temperature migrates from $t$ to $t'$, assume $I_{bias,1}, I_{bias,2}, f_1, f_2$ change to $I_{bias,1}', I_{bias,2}'$, $f_1', f_2'$, respectively. The following discussion will show that in our solution, the CTAT current biasing will make $I_{bias,1}/I_{bias,2}$ independent of the temperature. That is to say, when temperature migrates from $t$ to $t'$, $I_{bias,1}/I_{bias,2}$ is constant. Consequently, $f_1'/f_2' = f_1/f_2$.

Then, according to Figure 2, we have:

$$C_e|_{\text{fixed-time}} = x - y = T_{ADC} \left( f_1 - f_2 \right) = T_{ADC} \times f_e$$ \hspace{1cm} (4)

$$C_e|_{\text{fixed-time}} = x' - y' = T_{ADC} \left( f_1' - f_2' \right) = T_{ADC} \times f_e$$ \hspace{1cm} (5)

Obviously, $C_e$ changes with the temperature. It means that, with the fixed-time-based algorithm, $C_e$ varies with the temperature and can’t reflect $V_e$ correctly.

In the proposed fixed-number-based algorithm, the counters are also reset at the beginning of each sampling cycle but stopped when the first counter’s output reaches a fixed-number, say, $m$, in Figure 2. Then, (4) and (5) can be rewritten as:

$$C_e|_{\text{fixed-number}} = m - n = T_{samp} \left( f_1 - f_2 \right) = T_{samp} \times f_e$$ \hspace{1cm} (6)

where

$$T_{samp} = \frac{m}{f_1}$$ \hspace{1cm} (7)

$$C_e|_{\text{fixed-number}} = m - n = T_{samp} \left( f_1' - f_2' \right) = T_{samp} \times \frac{f_1'}{f_1} \left( f_1 - f_2 \right)$$ \hspace{1cm} (8)

Substituting (9) into (8), we get:

$$C_e|_{\text{fixed-number}} = T_{samp} \left( f_1 - f_2 \right) = T_{samp} \times f_e$$ \hspace{1cm} (9)

Thus, a temperature-independent $C_e$ can be obtained.

**Conclusion 1** With the Fixed-Number-Based algorithm, the output of the frequency-ADC will be temperature-independent if the ratio of the currents of the two branches keeps a constant all over the considered temperature range.

**Lemma 1** With the Fixed-Number-Based algorithm, for different input voltages, the currents ratios may be different. It doesn’t matter as long as the ratio is temperature-independent for a given input voltage. Further more, we don’t need to know the value of the ratio.

4. Requirements for $I_{CTAT}$

The above conduction is based on the assumption that when the temperature migrates from $t$ to $t'$, $I_{bias,1}/I_{bias,2}$ keeps unchanged, as is not the case without our CTAT current biasing technique. Given a circuit of a ring oscillator, for example, the left part of Figure 1, and an input ($V_{in}$-$V_{ref}$), we can always derive a relationship between the current of the two branches:

$$I_{bias,2} = A I_{bias,1}$$ \hspace{1cm} (11)

Generally $A$ is the function of both the input ($V_{in}$-$V_{ref}$) and the temperature. We found that we can properly design the biasing current, say, to be a CTAT, to guarantee that $A$ is independent of the temperature, but may still be the function of the input. Starting from the assumption that $A$ is only the function of the input, we inversely conduct the
requirements for the biasing current, $I_{\text{ICTAT}}$. From Figure 1, we obtain [7]:

$$\left(V_{\text{in}} - V_{\text{ref}}\right)^2 = \frac{2}{\mu_p C_m (W/L)_{M1,2}} I_{\text{ICTAT}}^2 - 2 I_{\text{bias},1} I_{\text{bias},2}$$  \hspace{1cm} (12)

By substituting (11) into (12) and using $I_{\text{bias},1} + I_{\text{bias},2} = I_{\text{ICTAT}}$, $I_{\text{ICTAT}}$ should be:

$$I_{\text{ICTAT}} = \frac{1}{2} \left(\frac{1 + A}{1 - \sqrt{A}}\right)^2 \mu_p C_m (W/L)_{M1,2}$$  \hspace{1cm} (13)

where $\mu_p$ and $C_{OX}$ are temperature-dependent parameters. Their temperature dependencies are given by [8]:

$$\mu_p = \mu_{p0} \left(\frac{T}{T_0}\right)^{\lambda_p}, \quad C_m = C_{m0} \left[1 + \alpha_{Cm} (T - T_0)\right]$$  \hspace{1cm} (14)

where $l$ is a positive constant ($l = 2$), $\alpha_{Cm}$ is negative. Thus expression (13) can be simplified as:

$$I_{\text{ICTAT}} = I_{\text{ICTAT}_0} \left[1 - B (T - T_0)\right]$$  \hspace{1cm} (15)

where

$$I_{\text{ICTAT}_0} = \frac{1}{2} \left(\frac{1 + A}{1 - \sqrt{A}}\right)^2 \mu_{p0} C_{m0} (W/L)_{M1,2}$$

$$B = l \left(\frac{T}{T_0}\right)^{\lambda_p} - \alpha_{Cm} > 0$$

The above conduction can be inversely applied. That means, if we set the biasing current as equation (15), then $A$ has to be temperature-independent. Notice that here we needn’t know exactly what value $A$ is (Please refer to Conclusion 1 and Lemma 1). So equation (15) is just what we expect. In reality, the best $I_{\text{ICTAT}s}$ for different input voltages are not completely the same, however, they are almost coincided, as shown in Figure 3.

Obviously, the slope of $I_{\text{ICTAT}}$ with respect to temperature should be negative, and can be supplied by $V_{BE}$ of a BJT.

5. Generation of $I_{\text{ICTAT}}$

A CTAT biasing current source shown in Figure 4 realizes the $I_{\text{ICTAT}}$. We can get:

$$I_{M6} = \frac{S_8}{S_7} \frac{V_{\text{in}}}{R_3}, \quad I_{M7} = \frac{S_7}{S_8} \frac{2}{\mu_p C_m R_4^2} \left(\frac{1}{\sqrt{S_7}} - \frac{1}{\sqrt{S_8}}\right)^2$$  \hspace{1cm} (17)

where

$$S_i = (W/L)_i, \quad i = 1, 2, 3...$$

As we know,

$$V_{BE} = V_{BE0} \left[1 + \alpha_{v_B} (T - T_0)\right]$$  \hspace{1cm} (18)

where the expression of $V_{BE}$ is obtained through the first-order approximation of Taylor series expansion. Then, expression (17) can be rewritten as:

$$I_{M6} = I_{M6_0} \left[1 + \alpha_{Cm} (T - T_0)\right]$$  \hspace{1cm} (19)

$$I_{M7} = \frac{S_8}{S_7} \frac{V_{BE0}}{R_3} \left[1 + \alpha_{Cm} (T - T_0)\right]$$

$$I_{M7} = \frac{S_8}{S_7} \frac{2}{\mu_p C_m R_4^2} \left(\frac{1}{\sqrt{S_7}} - \frac{1}{\sqrt{S_8}}\right)^2$$  \hspace{1cm} (20)

$$I_{M8} = I_{M6} - I_{M7}$$

$$= \left(I_{M6_0} - I_{M7_0}\right) \left[1 + \frac{I_{M6_0} I_{Cm_{\alpha}} - I_{M7_0} I_{Cm_{\alpha}}}{I_{M6_0} - I_{M7_0}} (T - T_0)\right]$$  \hspace{1cm} (21)

Since $\alpha_{v_B}<0$, $\alpha_{Cm}<0$ and $\alpha_{\alpha}>0$ ($\alpha_{\alpha} = 5.79\times10^{-4}/^\circ C$, using n’ poly resistor under charted 0.35μm CMOS mixed signal technology), $I_{Cm_{\alpha}}$ is negative and $I_{Cm_{\alpha}}$ positive.

Thus $I_{M8}$ has a negative temperature coefficient ($I_{M8} < I_{M8_0}$), which can be tuned by the size of $M4$, $M5$, $M7$ and $R4$. Finally, the required $I_{\text{ICTAT}}$ can be achieved by a current mirror:

$$I_{\text{ICTAT}} = \frac{S_{15}}{S_{10}} S_{12} I_{M8}$$  \hspace{1cm} (22)

The realized $I_{\text{ICTAT}}$ curve, obtained using the structure shown in Figure 4, is also plotted in Figure 3. It reveals that the CTAT biasing current source does provide an excellent fit to the required $I_{\text{ICTAT}}$ for whole temperature range.

6. Simulation Results

A ring-ADC using the proposed fixed-number-based algorithm together with the CTAT current biasing technique is simulated through mentor ADMS simulator, under charted 0.35μm CMOS mixed signal technology.
The fixed-number is assigned to 128 and the sampling rate is set to 10μs. Meanwhile, a traditional ring-ADC, which employs a fixed-time-based algorithm and a conventional biasing current source, is also simulated for comparison, with its fixed time set to 10μs. Both two ring-ADCs are simulated under three temperatures, 0℃, 30℃ and 100℃, respectively. Results are plotted in Figure 5 and Figure 6, with their Y-axis representing Ce. Obviously, the former reveals a great stability of the digital output Ce under different temperatures, while the latter shows severe variations, especially at large input voltage error. The proposed technique does provide an excellent temperature-independent characteristic for ring-ADCs.

If considered under a certain temperature, our temperature compensation technique leads to a decreased output linearity, as shown in Figure 5, however, if considered within the temperature range of 0℃ to 100℃, our technique achieves much better output linearity than the traditional one, resulting a much higher converted resolution. Simulation results reveal that our ring-ADC can reach a 2-mV quantization bin size while the traditional ring-ADC only 16-mV [5]. A typical high-performance digital DC-DC chip, ISL6592 [9] requires the quantization bin size of 3.125-mV.

Tab.1 provides a summary of the comparisons. It should be emphasized that all the results are obtained across the temperature range of 0℃ to 100℃. By applying the proposed technique, ring-ADCs do not need any additional digital calibration to compensate the temperature variation.

<table>
<thead>
<tr>
<th>Parameters</th>
<th>with temperature compensation</th>
<th>without temperature compensation</th>
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</thead>
<tbody>
<tr>
<td>Technology</td>
<td>Charted 0.35μm CMOS mixed signal technology</td>
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<tr>
<td>Sampling frequency</td>
<td>f_s = 100KHz</td>
<td></td>
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<tr>
<td>Input voltage span</td>
<td>120mV</td>
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<tr>
<td>Resolution</td>
<td>2mV/LSB</td>
<td>16mV/LSB</td>
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<tr>
<td>Number of bits</td>
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<td>3 bits</td>
</tr>
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References