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CORRECTION OF SAMPLE-TIME ERROR FOR TIME-INTERLEAVED SAMPLING SYSTEM USING CUBIC SPLINE INTERPOLATION

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Abstract

Sample-time errors can greatly degrade the dynamic range of a time-interleaved sampling system. In this paper, a novel correction technique employing a cubic spline interpolation is proposed for inter-channel sample-time error compensation. The cubic spline interpolation compensation filter is developed in the form of a finite-impulse response (FIR) filter structure. The correction method of the interpolation compensation filter coefficients is deduced. A 4GS/s two-channel, time-interleaved ADC prototype system has been implemented to evaluate the performance of the technique. The experimental results showed that the correction technique is effective to attenuate the spurious spurs and improve the dynamic performance of the system.

Keywords: digital correction, sample-time error, time-interleaved sampling system, cubic spline interpolation.

1. Introduction

Modern radar and communication systems are accelerating the digital signal processing circuits approaching to the front end, which brings an increasing demand for high speed Analog-to-Digital converters. Parallel architecture for ADCs (Analog-to-Digital Converters) is one of the solutions to obtain high sampling rates [1]. Under such architecture, high-speed sampling operation can be achieved by converting the analog input signals in parallel using identical ADCs. The significant advantage of the parallelism is the relaxation of sampling speed on each channel. However, the performance of a time-interleaved sampling system can be degraded by a number of factors such as offsets, gains, and sample-time errors in ADC channels [2], which can reduce the achievable linearity and limit the possibility of employing this technique. As a result, several methods have been introduced to compensate for these types of errors [3–5]. For example, offset mismatch errors can be easily calibrated by averaging the output data [6] and gain mismatch errors which is multiplicative noise to the input signal can be corrected by the FFT method [7] or blind method [8]. The sample-time error can have a more significant impact on the dynamic performance of the time-interleaved sampling system since it is often input dependent and the mismatch signal power increases as input frequency increases.

Many effective algorithms have been developed for digital sample-time error correction [9–10]. Blind equalization method is a promising way of calibrating the sample-time errors [11]. However, it is limited by the compromised accuracy and potentially has high computational cost. The fractional delay filter delays the input signal a fraction of the sampling period time [12], however, the fractional delay filters require the signal be oversampled with a factor of two or almost two compared to the Nyquist rate[13]. Also, the

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sample-time errors can be compensated by many other interpolation techniques. Neville's iteration method was adopted by Jin and Lee [14] for the digital interpolation compensation, but the algorithm requires high computational complexity which makes the method hard to be implemented.

In this paper, a novel compensation technique based on cubic spline interpolation is introduced to correct the sample-time errors. The proposed compensation scheme detects sample-time error from the actual sampling data by employing a least square estimation method. The obtained sample-time error is used to calculate the coefficients of the compensation filter based on cubic spline interpolation. The low computational complexity structure of the compensation filter in the form of a finite-impulse response (FIR) filter is deduced, which is suitable for real-time implementation.

The paper is organized as follows. The sample-time error mechanism of the time-interleaved sampling system is analyzed in Section II. Section III describes the proposed compensation method based on cubic spline interpolation and the efficient implementation structure in the form of an FIR filter is also introduced. Section IV presents the implementation of the prototype and experimental results. Finally, Section V concludes the paper briefly.

2. Sample-time error analysis of time-interleaved sampling system

Fig. 1 shows the diagram of time-interleaved sampling system. The analog input signal is applied to M ADCs in parallel, each operating at a sampling frequency of f_s / M. The mth sub-ADC, where $m = 1,...,M$, is sampling with clock phase ϕ_i that is delayed with $1/f_s$ respect to the following one. Slower ADCs sampling at f_s / M can thus be interleaved to obtain a faster ADC with an aggregate sampling rate of f_s. At the output port, a multiplexer recombines each ADC output to get a digital output y_n at the data rate of f_s.

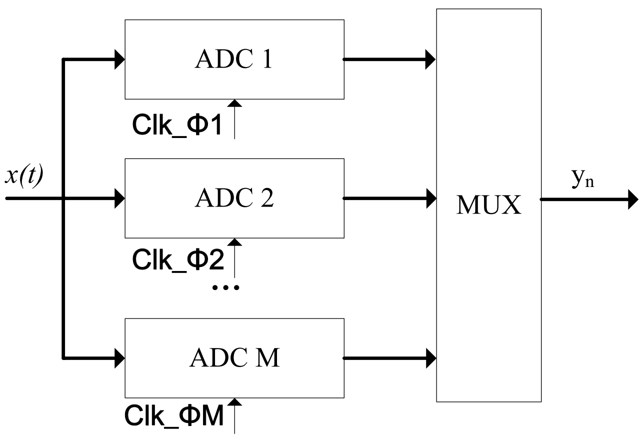


Fig. 1. Diagram of time-interleaved sampling system.

The main concern in designing time-interleaved sampling systems is to avoid mismatches, such as offset, gain and sample-time errors, in which sample-time error is the dominating mismatch. In practice, sample-time errors are caused by different signal delays among the clock paths and the channel paths. As shown in Fig. 2, T denotes the sampling period. r_m is the sample-time error of the mth sampling channel which is given as a fraction of the sampling period T. Assuming that the mth actual sampling instant is t_m, given by:

$$t_m = (m-1)T + r_m T, \tag{1}$$

where r_m denotes the sample-time error of mth channel, *m is the channel number and T* denotes the sampling period.. The sample-time error is periodic with M as shown in Fig. 2, thus we have

$$r_m = r_{m+M}. \tag{2}$$

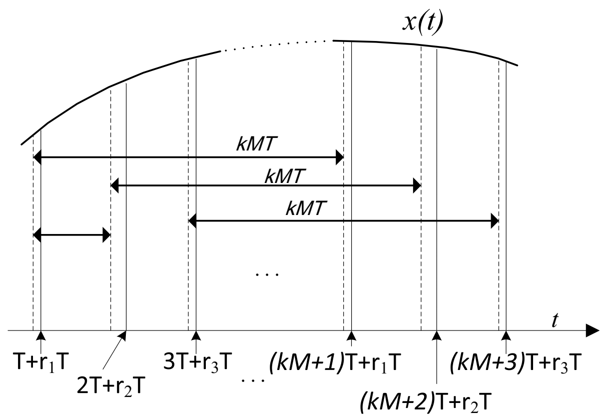


Fig. 2. Illustration of sample-time error between time-interleaved sampling channels.

3. Sample-time error compensation using cubic spline interpolation

Our proposed digital compensation scheme is shown in Fig. 3. The offset mismatch and gain mismatch are primarily compensated to avoid the influence of offset and gain mismatches on sample-time error estimation and compensation [15]. The sample-time error compensation block consists of two functional blocks. Firstly the errors are measured by the sine-fitting method. And then, the coefficients of the online compensation filter are calculated by using the cubic spline interpolation method proposed in this paper.

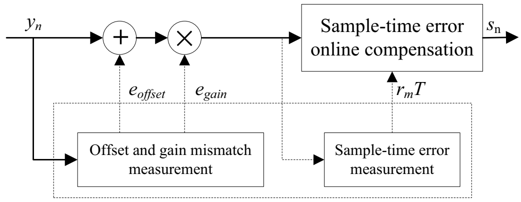


Fig. 3. The diagram of proposed compensation scheme.

3.1. Cubic spline interpolation

Cubic spline interpolation has certain similarities in regard to other interpolation methods, such as piecewise linear interpolation and Newton interpolation. Compared with other interpolation methods, cubic spline interpolation has more smooth curves, which represents the input signal more accurately [16], thus the recovery of the sampling data can be more precise. Basically, the cubic spline interpolation polynomial for the sampling data (t_1, y_1), (t_2, y_2), ..., (t_n, y_n) can be described as [17]

$$S_i(t) = y_i + b_i(t-t_i) + c_i(t-t_i)^2 + d_i(t-t_i)^3,\ t \in [t_i, t_{i+1}], \tag{3}$$

where b_i, c_i, d_i are the cubic spline coefficients. Interpolation function S should satisfy

$$S_i(t_i) = y_i, S_i(t_{i+1}) = y_{i+1}.\ \ i = 1,2,...,n-1, \tag{4}$$

$$S'_{i-1}(t_i) = S'_i(t_i).\ \ i = 1,2,...,n-1, \tag{5}$$

$$S''_{i-1}(t_i) = S''_i(t_i).\ \ i = 1,2,...,n-1, \tag{6}$$

$$S''_1(t_1) = S''_{n-1}(t_\mathrm{n}) = 0. \tag{7}$$

where $S'(t)$ and $S''(t)$ represent the first order and second order derivatives. Equations (4) to (6) represent the continuity of spline and its first two derivatives. Equation (7) represents the natural boundary conditions. There are n-1 different cubic polynomials, each with three coefficients b_i, c_i and d_i, so there are a total of 3n-3 unknown coefficients. In order to completely determine the coefficients of the polynomials, we need 3n-3 equations. From (4) we got n-2 equations as

$$\begin{aligned} & y_2 = y_1 + b_1(t_2-t_1) + c_1(t_2-t_1)^2 + d_1(t_2-t_1)^2, \\ & \vdots \\ & y_n = y_{n-1} + b_{n-1}(t_n-t_{n-1}) + c_{n-1}(t_n-t_{n-1})^2 + d_{n-1}(t_n-t_{n-1})^2. \end{aligned} \tag{8}$$

From (5), we got n-2 equations as

$$\begin{aligned} & S'_1(t_2) - S'_2(t_2) = b_1 + 2c_1(t_2-t_1) + 3d_1(t_2-t_1)^2 - b_2 = 0, \\ & \vdots \\ & S'_{n-2}(t_{n-1}) - S'_{n-1}(t_{n-1}) = b_{n-2} + 2c_{n-2}(t_{n-1}-t_{n-2}) + 3d_{n-2}(t_{n-1}-t_{n-2})^2 - b_{n-1} = 0. \end{aligned} \tag{9}$$

From (6), we got n-2 equations as

$$\begin{aligned} & S''_1(t_2) - S''_2(t_2) = 2c_1 + 6d_1(t_2-t_1) - 2c_2 = 0, \\ & \vdots \\ & S''_{n-2}(t_{n-1}) - S''_{n-1}(t_{n-1}) = 2c_{n-2} + 6d_{n-2}(t_{n-1}-t_{n-2}) - 2c_{n-1} = 0. \end{aligned} \tag{10}$$

To simplify the equations, we defined $\delta_i = t_{i+1} - t_i, \Delta_i = y_{i+1} - y_i$. Solving (8) and (10), we got

$$\begin{aligned} d_i &= \frac{c_{i+1} - c_i}{3\delta_i}, i = 1, ..., n-1. \\ b_i &= \frac{\Delta_i}{\delta_i} - c_i\delta_i - d_i\delta_i^2 = \frac{\Delta_i}{\delta_i} - \frac{\delta_i}{3}(2c_{i+1} + c_i), i = 1, ..., n-1. \end{aligned} \tag{11}$$

where δ_i and Δ_i are the defined intermediate variable. Substituting (11) to (9) gives

$$\begin{aligned} &\delta_1 c_1 + 2(\delta_1 + \delta_2)c_2 + \delta_2 c_3 = 3\left(\frac{\Delta_2}{\delta_2} - \frac{\Delta_1}{\delta_1}\right) \\ &\vdots \\ &\delta_{n-2}c_{n-2} + 2(\delta_{n-2} + \delta_{n-1})c_{n-1} + \delta_{n-1}c_n = 3\left(\frac{\Delta_{n-1}}{\delta_{n-1}} - \frac{\Delta_{n-2}}{\delta_{n-2}}\right). \end{aligned} \tag{12}$$

Equation (12) is a tridiagonal system of equations. Along with the natural boundary condition (7), the entire system of equations could be represented as:

$$\begin{cases} c_1 = c_n = 0. \\ \delta_1 c_1 + 2(\delta_1 + \delta_2)c_2 + \delta_2 c_3 = 3\left(\frac{\Delta_2}{\delta_2} - \frac{\Delta_1}{\delta_1}\right) \\ \vdots \\ \delta_{n-2}c_{n-2} + 2(\delta_{n-2} + \delta_{n-1})c_{n-1} + \delta_{n-1}c_n = 3\left(\frac{\Delta_{n-1}}{\delta_{n-1}} - \frac{\Delta_{n-2}}{\delta_{n-2}}\right). \end{cases} \tag{13}$$

Equation (13)can be written in a standard matrix-vector form as

$$\mathbf{TC} = \mathbf{D}, \tag{14}$$

where each matrix or vector has a substructure as follows:

$$\begin{aligned} \mathbf{T} &= \begin{pmatrix} 1 & 0 & 0 & & \\ \delta_1 & 2\delta_1 + 2\delta_2 & \delta_2 & & \\ & \ddots & \ddots & \ddots & \\ & & \delta_{n-2} & 2\delta_{n-2} + 2\delta_{n-1} & \delta_{n-1} \\ & & 0 & 0 & 1 \end{pmatrix} \\ \mathbf{C} &= [c_1, c_2, \cdots, c_n]^T \\ \mathbf{D} &= \left[0, 3\left(\frac{\Delta_2}{\delta_2} - \frac{\Delta_1}{\delta_1}\right), \cdots, 3\left(\frac{\Delta_{n-1}}{\delta_{n-1}} - \frac{\Delta_{n-2}}{\delta_{n-2}}\right), 0\right]^T, \end{aligned} \tag{15}$$

The coefficients of the polynomial c_i could be resolved by the tridiagonal system of equations. The coefficients b_i, d_i could be calculated based on c_i from (11). When we get the coefficient of the cubic spline interpolation polynomial in $[t_i, t_{i+1}]$, the sampling value at ideal sampling instant kT could be obtained by

$$S_i(kT) = y_i + b_i(kT - t_i) + c_i(kT - t_i)^2 + d_i(kT - t_i)^3, kT \in [t_i, t_{i+1}]. \tag{16}$$

As shown in(16), the data at the ideal sampling instant kT could be recovered by the cubic spline interpolation method. However, the performance of cubic spline interpolation can be negatively affected as it is strongly computationally intensive. As shown in (11) to(15), it is highly complex to calculate the coefficients of the interpolation polynomial and obtain the interpolation result. Hence, the high computational complexity increases the difficulty of using this method directly. For the online compensation, we derived the interpolation polynomial in accordance with the characteristic of the time-interleaved sampling system with the view of low computational complexity and appropriate for real-time implementation.

3.2. FIR filter structure of cubic spline interpolation for time-interleaved sampling system

Assume that channel number M=2. Without loss of generality, we define channel 1 as the reference channel. The sample-time error of channel 1 r_1=0. The sample-time error of channel 2 is r_2, and we assume that r_2>0 since it is possible to interchange the channel if the opposite is true in the two-channel time-interleaved sampling system. For the specific sampling points $(0, y_1)$, $(T+r_2T, y_2)$, $(2T, y_3)$, we obtain the sampling value of channel 2 at ideal sampling instant T through interpolation by the polynomial

$$S_2(T) = y_1 + b_1T + c_1T^2 + d_1T^3, T \in [0, T + r_2T]. \tag{17}$$

The coefficients b_1, c_1, d_1 can be obtained through (11) to (15) as:

$$\begin{aligned} b_1 &= \frac{\Delta_1}{\delta_1} - \delta_1\left(\frac{\Delta_2}{\delta_2} - \frac{\Delta_1}{\delta_1}\right) / 2(\delta_1 + \delta_2), \\ c_1 &= 0, \\ d_1 &= \left(\frac{\Delta_2}{\delta_2} - \frac{\Delta_1}{\delta_1}\right) / 2\delta_1(\delta_1 + \delta_2). \end{aligned} \tag{18}$$

Submitting the coefficients to (16), we get

$$S_2(T) = y_1 + \left(\frac{\Delta_1}{\delta_1} - \frac{\delta_1 \frac{\Delta_2}{\delta_2} - \Delta_1}{2(\delta_1 + \delta_2)}\right)T + \frac{\frac{\Delta_2}{\delta_2} - \frac{\Delta_1}{\delta_1}}{2\delta_1(\delta_1 + \delta_2)}T^3. \tag{19}$$

Making a further derivation, we submit the intermediate variable $\Delta_i = y_{i+1} - y_i$ to (187), then we have

$$S_2(T) = (1 - \frac{T}{\delta_1} - \frac{T}{2(\delta_1+\delta_2)} + \frac{T^3}{2\delta_1^2(\delta_1+\delta_2)})y_1 + (\frac{T}{\delta_1} + \frac{T}{2\delta_2} - \frac{T^3}{2\delta_1^2\delta_2})y_2 + (\frac{T^3 - \delta_1^2 T^3}{2\delta_1\delta_2(\delta_1+\delta_2)})y_3. \quad (20)$$

According to the periodical characteristic of the sample-time error between sampling channels as shown in (1) to (2), we have

$$\begin{aligned} \delta_{2N-1} &= t_{2N} - t_{2N-1} = T + r_2 T, \\ \delta_{2N} &= t_{2N+1} - t_{2N} = T - r_2 T, \\ \delta_N + \delta_{N+1} &= 2T. \end{aligned} \quad (21)$$

Submitting (21) to (20), we have

$$S_2(T) = \frac{3r_2^2 + 2r_2}{4(1+r_2)^2} y_1 + \frac{-2r_2^2 + 4r_2 + 4}{4(1+r_2)^2(1-r_2)} y_2 + \frac{-r_2^2 + 2r_2}{4(1+r_2)(1-r_2)} y_3. \quad (22)$$

Similarly, for sampling points ($2T$， y_3), ($3T+r_2T$， y_4), ($4T$， y_5), the interpolation value of channel 2 at ideal sampling instant $3T$ could be gotten by

$$S_2(3T) = y_3 + b_3(3T - 2T) + c_3(3T - 2T)^2 + d_3(3T - 2T)^3. \quad (23)$$

Calculating the coefficients through (11) to (15) and submitting the coefficients to (22), we get

$$S_2(3T) = y_3 + (\frac{T}{\delta_1} + \frac{T}{2(\delta_1+\delta_2)} - \frac{T^3}{2\delta_1^2(\delta_1+\delta_2)})\Delta_3 + (\frac{T^3}{2\delta_1\delta_2(\delta_1+\delta_2)} - \frac{\delta_1 T}{2\delta_2(\delta_1+\delta_2)})\Delta_4 . \quad (24)$$

Submitting the intermediate variable $\Delta_i = y_{i+1} - y_i$ to (24), then we have:

$$S_2(3T) = \frac{3r_2^2 + 2r_2}{4(1+r_2)^2} y_3 + \frac{-2r_2^2 + 4r_2 + 4}{4(1+r_2)^2(1-r_2)} y_4 + \frac{-r_2^2 + 2r_2}{4(1+r_2)(1-r_2)} y_5. \quad (25)$$

Comparing (25) with (22), it is clear that the interpolation polynomials of channel 2, $S_2(T)$ and $S_2(3T)$ are determined by the same coefficients.

We define the coefficients of the derived polynomial which is only related to r_2 as $h_{m,g}$(m=2,g=1,2,3):

$$h_{2,1} = \frac{3r_2^{\ 2} + 2r_2}{4(1+r_2)^2},$$
$$h_{2,2} = \frac{-2r_2^{\ 2} + 4r_2 + 4}{4(1+r_2)^2(1-r_2)}, \quad (26)$$
$$h_{2,3} = \frac{-r_2^{\ 2} + 2r_2}{4(1+r_2)(1-r_2)}.$$

Submitting (26) to (25) and (22), then we get

$$S_2((2n-1)*T) = h_{2,1}y_{2n-1} + h_{2,2}y_{2n} + h_{2,3}y_{2n+1}.\ n = 1,2,... \quad (27)$$

Representing $h_{m,g}$ in vector form, we have the compensation vector of each channel as:

$$H_1 = [h_{1,1}\ \ h_{1,2}\ \ h_{1,3}]^T,$$
$$H_2 = [h_{2,1}\ \ h_{2,2}\ \ h_{2,3}]^T. \quad (28)$$

Because the sample-time error of the reference channel is 0, there is no compensation required for the reference channel. Hence we have

$$H_1 = [0\ 1\ 0]^T. \quad (29)$$

According to (27), we obtained a general expression of cubic spline interpolation polynomial for two-channel time-interleaved sampling system as:

$$S_n = S_m(t_i) = Y_i^T H_m, m = 1,2, i = 1,2,3,... , \quad (30)$$

where S_n is the compensated output, Y_i is the input sampling data vector, H_m is the *m*th cubic spline interpolation coefficient vector. Since the reference channel does not need compensation, the whole computational complexity can be greatly reduced. Meanwhile, the sample-time error discussed in this paper is a time-invariant delay which means the time error will keep stable in limited periods. Although compared to the online correction method, the proposed method needs to update the compensation coefficients after a certain timespan, the proposed method will no longer need to calculate the coefficient of the interpolation polynomial iteratively as shown in paper [13]. As a result, the computational overhead could be further reduced.

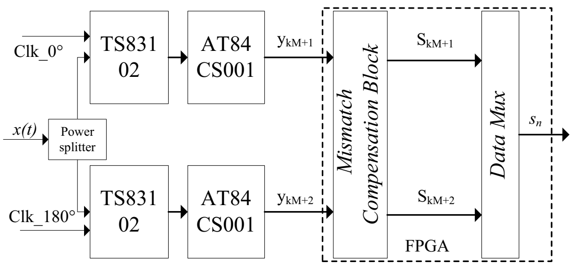
3.3. Sample-time error measurement

Sample-time error can be estimated blindly by comparing the statistical information of the digital output between the mismatched sub-ADC channels and the reference sub-ADC channel. However, reliable blind estimation needs a large data sample size and computational cost [18–19]. In the experiment sample-time errors are extracted by the sine-fitting method [20] for simplicity and lowest resource utilizations. The average value of the measuring results, which is tested at different frequencies through multiple tests, is used to calculate the compensation coefficients. Firstly, a sine wave with a frequency of f_{in} is applied to the time-interleaved ADCs. And then, the output samples of each channel are fitted by sine wave function of $x_n = A_n \cos(2\pi f_0 t_n) + B_n \sin(2\pi f_0 t_n)$ by using the least square criterion. The sample-time errors are calculated by computing the relative difference between the sub-ADC channels, which are given by

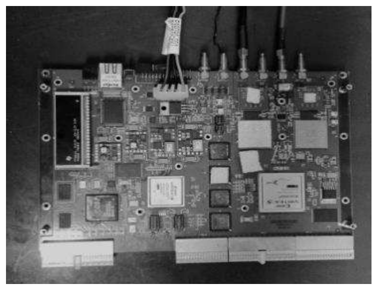
$$r_m T = [\tan^{-1}(\frac{B_m}{A_m}) - \tan^{-1}(\frac{B_1}{A_1})] / 2\pi f_{in} - (m+1)T, \tag{31}$$

where $r_m T$ is the sample-time error of channel m and $\tan^{-1}$ is the standard inverse tangent operator. A_1 and B_1 are the estimated parameters of the reference channel, while A_m and B_m are estimated parameters of the mismatched channel, respectively. f_{in} is the frequency of the input training sinusoid wave.

4. Experimental results



a) Block diagram,



b) Photograph of the ADC prototype.

Fig. 4. Implemented two-channel parallel ADC prototype.

To verify the proposed compensation method, we have developed and performed sample-time error compensation on a two-channel time-interleaved ADC prototype, as shown in Fig. 4. 4GS/s of overall sampling rate could be achieved with 10-bit resolution by interleaving two 10-bit ADC chips [21] sampling at 2GHz maximally. The digital data outputs of the time-interleaved sampling system were demultiplexed by two demultiplexer chips [22], and then the demultiplexed data were captured and compensated in FPGA. In the experiment, the clock frequency was set to 2GHz. First, we obtained the sample-time error r_2 at different frequency by the aforementioned least square estimation method. The measured sample-time error is r_2=0.0282 and the corresponding compensation filter coefficients were computed from (26).

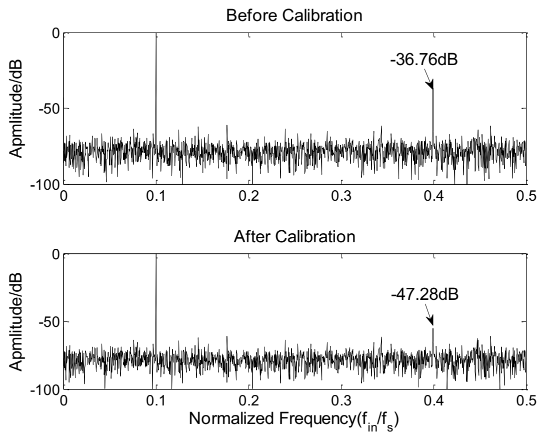


Fig. 5. FFT plot of ADC output before and after compensation (fin=400MHz).

Fig. 5 shows t he time-interleaved sampling system when a sinusoidal signal with a frequency of 400 MHz is sampled at 4 GS/s. The amplitude of the sample-time error spur is measured at -35.76 dB, and attenuated to -47.28 dB after compensation.

Fig.6 shows improvement of the spurious-free dynamic range (SFDR) and signal-to-noise and distortion ratio (SNDR) with the proposed compensation method. From Fig.6a, we can observe that the SFDR could be improved approximately 15dB in average after compensation in the first Nyquist zone. The measured SNDR as a function of the input frequency is plotted in Fig.6b with and without sample-time error compensation. Also, the SNDR of the ADC with compensation can be improved efficiently by the proposed cubic spline interpolation method.

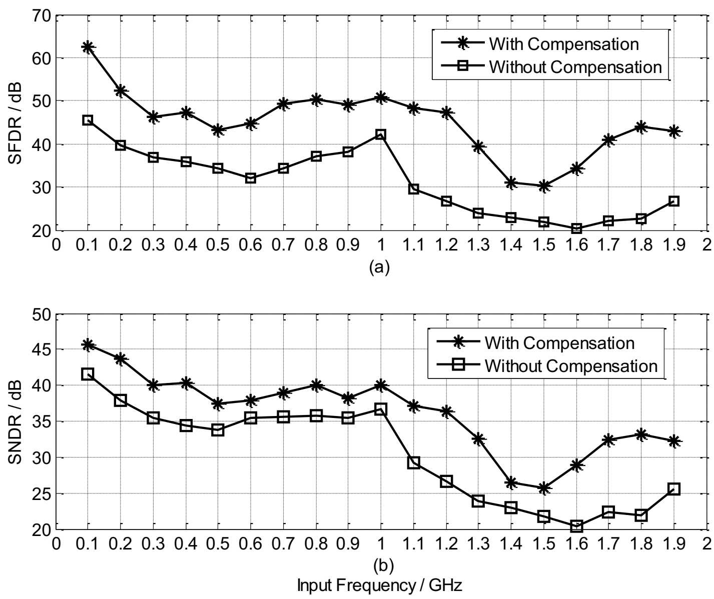


Fig. 6. SFDR and SNDR performance with and without compensation.

5. Conclusion

In this paper, a sample-time error compensation technique based on cubic spline interpolation for parallel ADC systems is presented. Through a detailed derivation, an efficient implementation architecture for the cubic spline interpolation method is demonstrated. The proposed technique has been implemented in a two channel 4GS/s 10-bit time-interleaved ADC prototype, and experimental results show that the cubic spline interpolation method provides decent compensation results and the proposed efficient implementation architecture of the compensation filter is feasible to compensate the sample-time error in real time. Sample-time error spur could be attenuated 15dB in average by the proposed correction technique and the dynamic performance could also be improved considerably.

Acknowledgments

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