A novel frequency response measurement method for wideband ADCs’ system

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The parallel analogue to digital converter (ADC) structure with mixers such as digital bandwidth-interleaving ADC (DBI-ADC) is a practical structure for realising the wideband data acquisition system. The system calibration relies on accurate system frequency response measurement. However, existing measurement methods require complex circuits or are susceptible to noise. This paper proposes a new frequency response measurement method, which can be realised by a simple circuit without any synchronisation measures. At the same time, frequency response measurement can ensure sufficient accuracy. Simulation and experiment results show that the method is effective and feasible.

Introduction: In high-speed data acquisition systems, various parallel ADC structures are usually used to overcome the performance limitations of a single ADC [1, 2]. Among these commonly used structures, DBI-ADC is easy to implement and does not require too high component and circuit performance. It is an attractive method to implement high-speed wideband data acquisition systems [3].

The accurate correction of the DBI-ADC system depends on the accurate measurement. References [4] and other DBI-ADC system calibration methods developed based on FI-ADC require accurate measurement of each part of the system, which is unrealistic in engineering. In contrast, references [5, 6] that only require measurement of subchannels or the entire system are more practical in engineering.

However, the current wideband ADC system measurement method is not enough to achieve a high-precision measurement of the DBI-ADC system, especially the phase frequency (group delay) response. Since the phase itself is a relative value, the measurement phase must have a reference value, so the input signal used for phase-frequency measurement must be multi-tone. There are two main phase-frequency response methods. One is to measure the device under test (DUT) with single-tone or dual-tone, which do not need mismatch. Multi-harmonic signal measurement usually uses periodic sinusoidal signals (dual-tone) or a strict synchronisation relationship between the DUT and the reference instrument (single-tone an dual-tone).

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The measurement frequency resolution is

\[ \omega = \frac{\omega_i}{N} \]

where \( M \) is the number of measurement frequency points.

Measurment method: As shown in Figure 2, the test frequency resolution is \( \omega_i \), \( x_1(t) \) and \( x_2(t) \) are the sine signals output by the high-precision radio frequency source. Assuming that the bandwidth of the DUT is \( \omega_i \). The measurement frequency resolution is \( \omega_i = \omega_i/N \). N is the number of measurement frequency points.

\[ x_1(t) = \cos(M\omega t + \phi_1) \]

where \( M \) is constant, it determines the interval between the two spectra analysed at one time.

There is no synchronisation between \( x_1(t) \) and \( x_2(t) \), so during measurement, \( \phi_1 \) and \( \phi_2 \) are independent.

Assuming that the sampling time of the DUT is \( t_s \). The spectrum of the measured signal input to the DUT is:

\[ X_{DUT}(\omega; k) = X_1(\omega_s)\delta(\omega - \omega_s) + X_{M}(\omega_s - \delta)\delta(\omega - \omega_s - \delta) \]

\[ X_{DUT}(\omega; k) = H_{add}(\omega_s)\delta(\omega - \omega_s + \delta) \]

where \( H_{add}(\omega) \) is the frequency response of mixer at \( \omega \), \( H_{add}(\omega) \) is the frequency response of adder at \( \omega \).

Because there is no synchronised, the difference of sampling time between the reference instrument and DUT is \( \Delta t \). The spectrum of the measured signal input to the reference instrument is

\[ X_{DUT}(\omega; k) = X_{DUT}(\omega + \Delta t) \]

where \( X_{DUT}(\omega; k) \) is the frequency spectrum of DUT, \( Y_{DUT}(\omega) \) is the frequency spectrum of DUT, \( Y_{DUT}(\omega) \) is defined as:

\[ Y_{DUT}(\omega; k) = Z_{DUT}(\omega)X_{DUT}(\omega; k) \]
Assuming that the reference instrument is an ideal instrument, its frequency response is $e^{-j\omega t}$. Define $U_{REF}(\omega)$ as output signal spectrum of the reference instrument. Similar to Equation (3), we get:

$$U_{REF}(\omega) = e^{-j\omega t}X_{REF}(\omega; k) = e^{j\omega t}A_i \phi(k)X_{OUT}(\omega; k)$$

(4)

Equations (3) and (4) show that for any one measurement, only three frequency points ($\omega_{0-M}, \omega_{0-M+1}, \omega_{0-M+2}$) in the outputs are valid. Using these points for calculations can avoid the interference of spurious signals.

Solved by Equation (3) and (4), the amplitude frequency response of points $\omega_{0-M}$ can be calculated as:

$$|Z_{DUT}(\omega_{0-M})| = \left|\frac{Y_{DUT}(\omega_{0-M})}{U_{REF}(\omega_{0-M})}\right|$$

(5)

Define $\Delta \phi(\omega; k) = \phi_{DUT}(\omega; k) - \phi_{REF}(\omega; k)$. Similarly, the phase frequency response at points $\omega_{0-M}$ can be calculated as

$$\phi_{DUT}(\omega; 0-M) = \Delta \phi(\omega_{0-M}; k) + \omega_{M-1}(\Delta t_1 - \tau_k)$$

(6)

Removing the time variables between Equations (6), it results:

$$\phi_{DUT}(\omega; 0-M) = \Delta \phi(\omega_{0-M}; k) - k - M \frac{\Delta \phi(\omega_{0-M}; 1)}{M}$$

(7)

In Equation (7), $\phi_{DUT}(\omega; 0-M)$ is the phase frequency response at $\omega_{0-M}$. For a linear time invariant system, it should be the same for all different $k$.

Notice that in equation (7) only the $\phi_{DUT}(\omega_{0-M})$ term is unknown since $\Delta \phi(\omega(0-M); k)$ can be obtained by $\phi_{DUT}(\omega(0-M); k)$ and $\phi_{REF}(\omega(0-M); k)$.

After completing the test of all frequency points in the entire frequency band, sweeping all $k$ values, according to Group Delay definition, GD $= -\frac{\phi_{DUT}(\omega_{0-M})}{\omega_{0-M}}$, we can get its estimation using the difference method.

$$GD_{DUT}(\omega_{0-M}) = \frac{\phi_{DUT}(\omega_{0-M}+1)}{\omega_{0-M}} - \frac{\phi_{DUT}(\omega_{0-M})}{\omega_{0-M}}$$

(8)

When using (7) in (8), a constant term $-\frac{\phi_{DUT}(\omega_{0-M})}{\omega_{0-M}}$, independent of $k$, appears. According to the FI-ADC perfect reconstruction (PR) requirements, we expect the DUT to satisfy that the group delay is a constant. Rewrite the group delay $GD_{DUT}(\omega)$ as the sum of the average group delay $GD_{DUT}(\omega)$ and the group delay deviation $\Delta GD_{DUT}(\omega)$:

$$GD_{DUT}(\omega) = GD_{DUT}(\omega) + \Delta GD_{DUT}(\omega)$$

(9)

It can be seen that PR conditions are only concerned with $\Delta GD_{DUT}(\omega)$. So $GD_{DUT}(\omega)$ can be replaced by $\Delta GD_{DUT}(\omega)$. That means $-\frac{\phi_{DUT}(\omega_{0-M})}{\omega_{0-M}}$ can be ignored.

Therefore, $\phi_{DUT}(\omega_{0-M})$ can be considered as a null reference phase. Correspondingly, Equations (7) and (8) can be rewritten as:

$$\phi_{DUT}(\omega_{0-M}) = \Delta \phi(\omega_{0-M}; k) - k - M \frac{\Delta \phi(\omega_{0-M}; 1)}{M}$$

(10)

$$GD_{DUT}(\omega_{0-M}) = \frac{\phi_{DUT}(\omega_{0-M}+1)}{\omega_{0-M}} - \frac{\phi_{DUT}(\omega_{0-M}; k+1)}{\omega_{0-M}}$$

(11)

The spurious signals in the system are mainly divided into two categories, one is the spurious signal with fixed frequency; the other is the frequency change with the input signal. The spurious signal with a fixed frequency has the same influence on all measurement methods and can be corrected by measuring the spurious signal. For spurious signals whose frequency changes with the input signal, because the frequency of the input signal $x(t)$; $k$ entering the ADC in this method is controlled by the parameters $k$ and $M$, the spurious signals can be avoided by changing the values of $k$ and $M$. The influence of scattered signals. Specifically, the steps are as follows:

1. Determine the parameters $M_1$ and sequence $k_1$, $k_2$ is the value set of $k$ to make $k - M$ and $k + M$ traverse all frequency points. Calculate the frequency point set $\Omega_1$ where the input signal and the spurious signal spectrum overlap. If $\Omega_1 = \emptyset$, the measurement results using $M_1$ and $k_1$ do not need to be corrected and can be used directly.
2. Calculate the parameters $M_2$ and the sequence $k_2$, the overlapping frequency point sequence $\Omega_2$ satisfies $\Omega_1 \cap \Omega_2 = \emptyset$.
3. Measure the system according to $M_1$, $k_1$ and $M_2$, $k_2$ respectively, and record the measurement results as $Z_{\Omega_1}$, $Z_{\Omega_2}$.
4. $Z$ is the average value of $Z_{\Omega_1}$ and $Z_{\Omega_2}$ (0 for unmeasured frequency points)

As what has been discussed above, we can get DUT frequency response measurement steps of this method are:

1. Determine the appropriate $\omega_{0-M}, M_1, k_1, k_2$ to reduce the influence of the second term in (7) or (10). This way, $M_1$ and $M_2$ should be as large as possible. Establish also the number of times, $I$, that the measurements have to be repeated.
2. Connect the measurement system according to Figure 2.
3. Measure the system according to $M_1$, $k_1$. Get the sampled value and calculate the DUT frequency response according to the Equations (5) and (10), the results are recorded as $\{Z_{DUT}(\omega_{0-M}), \phi_{DUT}(\omega_{0-M})\}$. The average of $\{Z_{DUT}(\omega_{0-M}), \phi_{DUT}(\omega_{0-M})\}$ and $\{Z_{DUT}(\omega_{0-M}), \phi_{DUT}(\omega_{0-M})\}$.
4. Measure the second set according to $M_2$, $k_2$.
5. Corrected measurement result and calculate the system group delay according to Equation (11).

Simulation: In order to verify the validity and feasibility of the measurement method proposed in this paper, simulations and experiments were carried out in this section and the next section. Simulation mainly analyzes the effectiveness of the measurement method, and compares the theoretical performance with other methods.

In the simulation, a 2-subchannel DBI-ADC system has been designed in MATLAB, and the sampling rate of each subchannel is 10 GS/s. System sampling rate is $f_s = 20$ GS/s, bandwidth is $\Omega_2 = 5$ GHz. The simulation results are compared with the other three methods under different measurement errors. The measurement result of the sine frequency sweep method without noise is used as the reference’s true value, that means this result considered as true (ref $Z_{DUT}$).

The frequency response of the mixer and adder used in this method is regarded as an equivalent 5th-order elliptical low-pass filter. Set frequency are $f_s = 10$ MHz, $M_1 = 501$, $k_1 = 1, 2, \ldots, M_1 - 1$, $M_2 = 511$, $k_2 = 11, 12, \ldots, M_2 - 11$. When the system adds Gaussian white noise with $\sigma^2 = 10^{-3}$ and $\sigma^2 = 10^{-5}$.

| $\sigma^2$ | $A_{amp}$ | $B_{amp}$ | $C_{amp}$ | $D_{amp}$ |
|-----------|----------|----------|----------|----------|
| 10^{-3}   | 0.000038 | 4.257477 | 13.30664 | 0.503075 |
| 10^{-5}   | 0.000001 | 0.025853 | 0.124904 | 0.006725 |
| 0         | 6.98E-9  | 0.000413 | 2.99E-6  | 2.46E-7  |

amp: amplitude; GD: group delay deviation; $f_s$: 10 MHz, 150 ps period pulse; C: 10 MHz square wave; D: This method (uncorrected).

Table 1: M.S.E of amplitude and group delay deviation with different noise
Table 2. MSE of group delay deviation with skew mismatch

| $\mu$ | $A_{GD}$ | $B_{GD}$ | $C_{GD}$ | $D_{GD}$ | $E_{GD}$ |
|-------|----------|----------|----------|----------|----------|
| $0.01T_s$ | 1.935715 | 0.000615 | 0.015368 | 0.000235 |
| $0.001T_s$ | 0.020184 | 0.000615 | 0.015368 | 0.000189 |

Table 3. MSE of amplitude and group delay deviation with spurious

| $\alpha$ | $A_{GD}$ | $B_{GD}$ | $C_{GD}$ | $D_{GD}$ | $E_{GD}$ |
|----------|----------|----------|----------|----------|----------|
| 0.1      | 0.172672 | 0.382463 | 0.240955 | 0.182197 | 0.171922 |
| 0.01     | 0.001708 | 0.004810 | 0.002260 | 0.001787 | 0.001711 |

$\mu$: This method (corrected)

Expanding uncertainty of skew mismatch. Other methods are almost unaffected by skew mismatch.

Skew mismatch $<0.01T_s$ can only be achieved by TI-ADC now, and in other ADC structures will be larger. This means that in practice, skew mismatch has a greater impact on the sine sweep method.

Table 3 is the MSE of amplitude and group delay deviation with spurious by different measurement methods. Assuming that the spurious signals existing in the system have the same amplitude, $a$ is their attenuation relative to the input signal. It can be seen from Table 3 that our method after correction (in column E) is obviously superior to square wave and periodic pulse, and achieves the accuracy similar to sine sweep.

Experiment: The experiment system is a 2-subchannel DBI-ADC system without correction. It has a sampling rate of 20 GSa/s and a bandwidth of 5.5 GHz. The frequency response measurement methods are the fast edge signal shown in Figure 1, and this method.

In experiment system, the reference instrument is WaveMaster 813Zi-B oscilloscope. The power splitter/combiner of the system is ZX10R-14-S+. The mixer is LTC5562. Set frequency are $f_0 = 10$ MHz, $M_1 = 551$, $k_1 = 1, 2, \ldots, M_1 - 1$, $M_2 = 561$, $k_2 = 11, 12, \ldots, M_2 - 11$.

The system amplitude-frequency response measurement value is shown in Figure 3, the group delay deviation measurement value is shown in Figure 4.

According to Figures 3 and 4, the measurement errors of this method are significantly smaller than that of the fast-edge signal. One reason is that the fast edge signal is greatly affected by noise, especially the high frequency part. The other reason is that the energy of the fast side signal between the low frequency lobes drops sharply, which leads to the increase of group delay measurement error.

Conclusion: This paper proposes a new frequency response measurement method for wideband ADC acquisition systems such as DBI-ADC. This method combines the advantages of the multi-harmonic signal method and the sine frequency sweep method. The accuracy of the measurement results can reach the level of the sine frequency sweep method. Simulation and experiment verify the effectiveness of this method.

Acknowledgment: This work is supported partly by the National Natural Science Foundation of China under Grant (No.61801092, No. 61701077 and No. 61671114), Fundamental Research Fund for the Central University of China (No. ZYGX2015KYQD074).

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Fig. 3 DBI-ADC system amplitude frequency response

Fig. 4 DBI-ADC system group delay deviation

References
1 Yang, K., Shi, J., Yang, H.: A TIADC mismatch calibration method for digital storage oscilloscope. In: 13th IEEE International Conference on Electronic Measurement & Instruments, pp. 379–383. IEEE, Piscataway, NJ (2017)
2 Lei, Q., Zheng, Y., Siek, L.: Analysis and design of high performance frequency-interleaved ADC. In: 2013 IEEE International Symposium on Circuits and Systems, pp. 2022–2025. IEEE, Piscataway, NJ (2013)
3 Pupalaikis, P.J.: An 18 GHz bandwidth, 60 GS/s sample rate real-time waveform digitizing system. In: IEEE/MTT-S International Microwave Symposium, pp. 195–198. IEEE, Piscataway, NJ (2007)
4 Song, J., et al.: Digital correction of frequency-response errors in bandwidth-interleaved ADCs. Electron. Lett. 52(19), 1596–1598 (2016)
5 Lamarche, F., Joshi, L., Sureka, A.K.: Method and apparatus for a high bandwidth oscilloscope utilizing multiple channel digital bandwidth interleaving. US Patent 795,793,8B2, 7 June 2011.
6 Zhou, Y., Ye, P., Yang, K.: A novel wideband high-speed data acquisition system correction method. In: 2019 IEEE International Instrumentation and Measurement Technology Conference, pp. 1–6. IEEE, Piscataway, NJ (2019)
7 Liu, T., et al.: Analysis and design of M-channel frequency-interleaved ADC with analog filter estimation. Analog Integr. Circ. Signal Processing 81, 173–180 (2014)
8 Pupalaikis, P.J., Lamarche, F., Digital group delay compensator. US Patent 705,091,8B2, 23 May 2006
9 Gao, J., et al.: An adaptive calibration technique of timing skew mismatch in time-interleaved analog-to-digital converters. Rev. Sci. Instrum 90(2), 025102 (2019)