Calibration technique for new-structure, two-channel hybrid filter banks ADC

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Abstract: Frequency-interleaved ADC (FIADC) is less sensitive of the channel mismatches than the Time-interleaved ADC (TIADC). However, the filter bank mismatches and channel mismatches such as offset, gain and timing mismatches among FIADC still degrade the performance of the FIADC system. This paper presents a new structure of Two-channel Hybrid Filter Banks (HFB) based on power complementary pair, which only need Single-channel filtering. Based on the analysis of channel mismatches and filter bank mismatches, calibration is carried out for FIADC system. Average distortion error of \(1.5639 \times 10^{-5}\) db and average aliasing error of \(-93\) db can be achieved after calibration by using Third-order analog filter and 64-length digital filters.

Keywords: hybrid filter banks, power complementary pair, single-channel filtering, channel mismatches, filter bank mismatches

Classification: Circuits and modules for electronic instrumentation

References

[1] J. Matsuno, et al.: “All-digital background calibration technique for time-interleaved ADC using pseudo aliasing signal,” IEEE Trans. Circuits Syst. I, Reg. Papers 60 (2013) 1113 (DOI: 10.1109/TCSI.2013.2249176).

[2] B. Yu, et al.: “A mixed sample-time error calibration technique in time-interleaved ADCs,” IEICE Electron. Express 10 (2013) 20130882 (DOI: 10.1587/elex.10.20130882).

[3] X. Liu, et al.: “An efficient blind calibration method for nonlinearity mismatches in M-channel TIADCs,” IEICE Electronics Express 14 (2017) (DOI: 10.1587/elex.14.20170468).

[4] Y. X. Zou and X. J. Xu: “Blind timing skew estimation using source spectrum sparsity in time-interleaved ADCs,” IEEE Trans. Instrum. Meas. 61 (2012) 2401 (DOI: 10.1109/TIM.2012.2192337).

[5] C. Yuan, et al.: “A modulated hybrid filter bank for wide-band analog-to-digital converters,” Journal of Multimedia 9 (2014) 569 (DOI: 10.4304/jmm.9.4.569-575).

[6] W. Wang and Z. J. Zhang: “Design of digital synthesis filters for hybrid filter bank A/D converters using semidefinite programming,” J. Networks 9 (2014) 99 (DOI: 10.4304/jnw.9.5.1325-1332).

[7] Z. Liu and M. Lin: “Design of two-channel hybrid analog/digital banks
satisfying near-perfect reconstruction,” International Conference on Signal Processing, Proc. ICSP IEEE 1 (2004) 407 (DOI: 10.1109/ICOSP.2004.1452668).

[8] C. Lelandais-Perrault, et al.: “Wideband, bandpass, and versatile hybrid filter bank A/D conversion for software radio,” IEEE Trans. Circuits Syst. I, Reg. Papers 56 (2009) 1772 (DOI: 10.1109/TCSI.2008.2008289).

[9] S. R. Velazquez, et al.: “A hybrid filter bank approach to analog-to-digital conversion,” IEEE-SP International Symposium on Time-Frequency and Time-Scale Analysis IEEE Xplore (1994) 116 (DOI: 10.1109/TFSA.1994.467350).

[10] D. E. Goldberg, et al.: “Genetic algorithm is search optimization and machine learning,” xiii.7 (1989) 2104.

[11] L. Qu, et al.: A Filter Bank Mismatch Calibration Technique for Frequency-Interleaved ADCs (Birkhauser Boston Inc., 2016).

[12] L. Guo, et al.: “Analysis of channel mismatch errors in frequency-interleaved ADC system,” Circuits Syst. Signal Process. 33 (2014) 3697 (DOI: 10.1007/s00034-014-9828-z).

[13] T. H. Tsai, et al.: “Correction of mismatches in a time-interleaved analog-to-digital converter in an adaptively equalized digital communication receiver,” IEEE Trans. Circuits Syst. I, Reg. Papers 56 (2009) 307 (DOI: 10.1109/TCSI.2008.2002114).

[14] S. R. Velazquez, et al.: “Design of hybrid filter banks for analog/digital conversion,” IEEE Trans. Signal Process. 46 (1998) 956 (DOI: 10.1109/78.668549).

1 Introduction

In the signal sampling system of modern information equipment such as radar, there is increasing demand for both the resolution and the sample rate of ADC. The ability of single ADC can not satisfy these two conditions at the same time.

To design High-speed and High-resolution ADCs, the architecture of multi-channel is widely adopted in many implementations. Multi-channel Time-interleaved ADC (TIADC) is firstly studied. Theoretically, multiple channel structure increases the sampling rate proportionally with the number of channels. Nevertheless, the mismatching between channels makes the performance worse than expected in practice. TIADC is extremely sensitive to the time skew, which make it not suitable for high-resolution, and the condition will be worse as the number of channels increasing. Although there are many algorithms proposed to calibration these mismatches, the results after calibration are still not satisfactory [1, 2, 3, 4].

Another approach for multi-channel ADC is called Frequency-interleaved ADC (FIADC), whose prototype architecture is shown in Fig. 1(a). Define $\Omega$ as the analog angular frequency and $\omega = \Omega T_s$ is the digital angular frequency. The analog analysis filter bank divides the overall input signal frequency band symmetrical, shown in Fig. 1(b), where $\Omega$ is maximum frequency of the input signal. The allocated subbands are sampled by ADCs in different channels, which is followed by interpolation. Finally, the digital synthesis filter bank synthesizes signals for all channels and reconstructs the digitized signal. The Hybrid Filter Bank (HFB) architectures overcome the Time-interleaved architecture’s serious defect of timing error sensitiveness [5, 6]. But there are many challenges in constructing an HFB.
One of them is to design analog and digital filters in the filter bank to provide adequate channel separation and accurate reconstruction of the converted signal. To minimize the filter bank mismatch errors (i.e., distortion error caused by the unsatisfactory amplitude-frequency characteristic of analog analysis filters and aliasing error caused by the mirroring produced by the sampling and interpolation), proper design of the filters should be done to provide good filter characteristics like sharp cutoff and large stopband attenuation. For FIADC system, the design of the analog analysis filters becomes increasingly difficult as the filter order or the quality factor becomes higher. Besides, because of passband ripples and no constant group delay of analog analysis filters, the order of digital synthesis filters is usually required to be high to achieve an accurate signal reconstruction. These all make the design become difficult.

There is another structure of Two-channel hybrid filter bank satisfying near-perfect reconstruction which based on the pairwise power complementary [7]. But there is no effective channel mismatch calibration method of this structure. This paper will give an improved structure of Two-channel HFB based on power complementary pair which only need single analog filter. The corresponding system analysis and calibration methods is given out.

\[ X(t) \xrightarrow{H_0(j\Omega)} \xrightarrow{ADC_0} M \uparrow \xrightarrow{F_0(e^{j\omega})} \]
\[ H_1(j\Omega) \xrightarrow{ADC_1} M \uparrow \xrightarrow{F_1(e^{j\omega})} \]
\[ \vdots \]
\[ H_n(j\Omega) \xrightarrow{ADC_n} M \uparrow \xrightarrow{F_n(e^{j\omega})} \]
\[ \vdots \]
\[ H_{M-1}(j\Omega) \xrightarrow{ADC_{M-1}} M \uparrow \xrightarrow{F_{M-1}(e^{j\omega})} \]

\[ Y(n) \]

\( X(t) \xrightarrow{H_0(j\Omega)} \xrightarrow{H_1(j\Omega)} \xrightarrow{H_2(j\Omega)} \cdots \xrightarrow{H_{M-1}(j\Omega)} \)

\[ \Omega \frac{M}{M} \quad 2\Omega \frac{M}{M} \quad 3\Omega \frac{M}{M} \quad \Omega \]

**Fig. 1.** (a) M-channel FIADC structure. (b) The frequency distribution of M-channel analysis filter bank.
2 Two-channels HFB design based on power complementary pair

The output of two-channel HFB can be expressed as [8]:

\[ Y(e^{j\omega}) = \sum_{k=0}^{1} X \left( j\frac{\omega}{T_s} - j\frac{\pi k}{T_s} \right) T_k(e^{j\omega}) \]  

(1)

\[ T_k(e^{j\omega}) = \frac{1}{2T_S} \sum_{m=0}^{1} H_m \left( j\frac{\omega}{T_s} - j\frac{\pi k}{T_s} \right) F_m(e^{j\omega}) \]  

(2)

where \( 2T_S \) is sampling period of single ADC.

When the reconstructed signal is only a scaled, delayed, and sampled version of the input signal, the perfect reconstruction is achieved. Therefore, the perfect reconstruction (PR) condition of the two-channel FIADC is [8]:

\[ T_k(e^{j\omega}) = \begin{cases} 
  c e^{-j\omega d}, & k = 0; c,d \in R \\
  0, & k = 1 
\end{cases} \]  

(3)

where \( c \) is nonzero constant, referred to as the scale factor and \( d \) is the system delay. \( T_0(e^{j\omega}) \) is called the distortion function, and \( T_p(e^{j\omega}), 1 < p \leq (M-1) \), is called the aliasing function [9]. Eq. (3) indicates that the ideal distortion function \( T_0(e^{j\omega}) \) should be scaled and delayed version of the input signal, while the ideal aliasing function \( T_p(e^{j\omega}) \) should be zero. Based on Eq. (3), we define average aliasing error and average distortion error as:

\[
D_{ave} = \frac{1}{2\pi} \int_{-\pi}^{\pi} |T_0(e^{j\omega})|d\omega \\
A_{ave} = \frac{1}{2\pi} \int_{-\pi}^{\pi} |T_1(e^{j\omega})|d\omega
\]  

(4)

Where \( D_{ave} \) is average distortion error and \( A_{ave} \) is average aliasing error, the average distortion/aliasing error is an important system metrics.

Eq. (3) can be transformed into (assume \( 2T_s = \frac{1}{c} \)):

\[
\begin{cases} 
H_0 \left( j\frac{\omega}{T_s} \right) F_0(e^{j\omega}) + H_1 \left( j\frac{\omega}{T_s} \right) F_1(e^{j\omega}) = e^{-j\omega d} \\
H_0 \left( j\frac{\omega}{T_s} - j\frac{\pi}{T_s} \right) F_0(e^{j\omega}) + H_1 \left( j\frac{\omega}{T_s} - j\frac{\pi}{T_s} \right) F_1(e^{j\omega}) = 0 
\end{cases}
\]  

(5)

Solving the system of simultaneous equations formed by Eq. (5) for \( F_0(e^{j\omega}) \) and \( F_1(e^{j\omega}) \), the Fourier transform of the synthesis digital filters which provide perfect reconstruction can be expressed as [9]:

\[
F_0(e^{j\omega}) = e^{-j\omega d} \frac{H_1 \left( j\frac{\omega}{T_s} - j\frac{\pi}{T_s} \right)}{H_0 \left( j\frac{\omega}{T_s} \right) H_1 \left( j\frac{\omega}{T_s} - j\frac{\pi}{T_s} \right) - H_1 \left( j\frac{\omega}{T_s} \right) H_0 \left( j\frac{\omega}{T_s} - j\frac{\pi}{T_s} \right)} \\
F_1(e^{j\omega}) = e^{-j\omega d} \frac{H_0 \left( j\frac{\omega}{T_s} - j\frac{\pi}{T_s} \right)}{H_1 \left( j\frac{\omega}{T_s} \right) H_0 \left( j\frac{\omega}{T_s} - j\frac{\pi}{T_s} \right) - H_0 \left( j\frac{\omega}{T_s} \right) H_1 \left( j\frac{\omega}{T_s} - j\frac{\pi}{T_s} \right)}
\]  

(6)
Eq. (6) is transfer function of Digital filters which can carry out signal reconstruction and calibration.

If the digital synthesis filters can be designed as delayed conjugations of the two analog analysis filters, namely $F_0(e^{i\omega}) = H_0\left(-j\frac{\omega}{T_S}\right)e^{-j\omega d}$ and $F_1(e^{i\omega}) = H_1\left(-j\frac{\omega}{T_S}\right)e^{-j\omega d}$. While $-\pi < \omega < \pi$, from Eq. (5) we can get power complementary condition,

$$\begin{align*}
\left|H_0\left(j\frac{\omega}{T_S}\right)\right|^2 + \left|H_1\left(j\frac{\omega}{T_S}\right)\right|^2 &= 1 \\
H_0\left(-j\frac{\omega}{T_S}\right)H_0\left(j\frac{\omega}{T_S} - j\frac{\pi}{T_s}\right) + H_1\left(-j\frac{\omega}{T_S}\right)H_1\left(j\frac{\omega}{T_S} - j\frac{\pi}{T_s}\right) &= 0
\end{align*}$$

(7)

Define analog filter transfer function $H_0(j\Omega)$ and $H_1(j\Omega)$ as:

$$\begin{align*}
H_0(j\Omega) &= \frac{E(j\Omega)}{E(j\Omega) + O(j\Omega)} \\
H_1(j\Omega) &= \frac{O(j\Omega)}{E(j\Omega) + O(j\Omega)}
\end{align*}$$

(8)

(9)

where $E(j\Omega) = \sum_{n=0}^{N} a_n(j\Omega)^n$, $n = \text{even number} \& n \leq N - 1$ is the even part of the denominator and $O(j\Omega) = \sum_{n=0}^{N} a_n(j\Omega)^n$, $n = \text{odd number} \& n \leq N - 1$ is the odd part of the denominator $D(j\Omega)$. $\Omega$ is analog angular frequency. $D(j\Omega)$ is defined as:

$$D(j\Omega) = \sum_{n=0}^{N} a_n(j\Omega)^n$$

(10)

where N is the order of filter and $a_n$ is the nth polynomial of the denominator. The analog filter can be then designed through Genetic algorithm [10] with the objective function defined as follows

$$\vartheta = \int_0^\pi \left|H_0\left(j\frac{\omega}{T_S}\right) - H_1\left(j\frac{\omega}{T_S} - j\frac{\pi}{T_s}\right)\right| d\omega$$

(11)

From transfer function of analog analysis filters, shown in Eq. (8) and Eq. (9), it is easy to observe that:

$$H_1(j\Omega) = 1 - H_0(j\Omega)$$

(12)

Then Eq. (5) can simplify as:

$$H_0\left(j\frac{\omega}{T_S}\right)\left[F_0(e^{i\omega}) - F_1(e^{i\omega})\right] + F_1(e^{i\omega}) = e^{-j\omega d}$$

(13)

$$H_0\left(j\frac{\omega}{T_S} - j\frac{\pi}{T_s}\right)\left[F_0(e^{i\omega}) - F_1(e^{i\omega})\right] + F_1(e^{i\omega}) = 0$$

(14)

It can be observed that the perfect reconstruction can be achieved only need one analog filter. The improved structure of Two-channels HFB is shown in Fig. 2. This structure can decrease mismatch errors between two analog filters.
3 Channel mismatch errors and filter bank mismatch errors calibration

For the channel mismatch errors and filter bank mismatch errors has substantial overlap [11, 12], so it is difficult to do estimation and calibration separately. Here gives a combined correction algorithm.

Error model of filter bank mismatches and channel mismatches are shown in Fig. 3. \( C_1g \) is gain ratio of the two channels and \( C_1t \) is time difference between two channels. Offset mismatches is not considered here since it can be calibrated effectively through proper algorithms [13].

Section (2) proved formula to calculate the parameter of analog filters. However, there are various realization error such as parasitic resistance and parasitic capacitance are inevitable, which cause the deviation of \( H_0(j\Omega) \). To overcome this problem, we adopt the third order analog filter whose transfer function is given by:

\[
\hat{H}_0(j\Omega) = \frac{\hat{a}_0 + \hat{a}_2(j\Omega)^2}{\hat{a}_0 + \hat{a}_1(j\Omega) + \hat{a}_2(j\Omega)^2 + \hat{a}_3(j\Omega)^3} e^{j\Delta t}
\]

(15)

Where \( \hat{a}_n = (1 + \epsilon_n) a_n \) (n = 0, 1, 2, 3) and \( \epsilon_n \) is analog deviation factor of \( a_n \) which is the coefficient of polynomials. Use \( \hat{H}_0(j\Omega) \) to substitute \( H_0(j\Omega) \) and define \( H_{i0}(j\Omega) = H_0(j\Omega)(1 + \Delta g) e^{j\Delta t} \). Therefore, we can get:

\[
H_{i0}(j\Omega) = \frac{[\hat{a}_0 + \hat{a}_2(j\Omega)^2](1 + \Delta g)}{\hat{a}_0 + \hat{a}_1(j\Omega) + \hat{a}_2(j\Omega)^2 + \hat{a}_3(j\Omega)^3} e^{j\Delta t}
\]

(16)

The sampling results of two channels satisfy:

\[
\begin{align*}
X_0(e^{j\omega}) = \frac{1}{2T_s} & \sum_{k=-\infty}^{\infty} X\left(j\frac{\omega}{T_s} - j\frac{\pi k}{T_s}\right) H_{i0}\left(j\Omega - j\frac{\pi k}{T_s}\right) \\
X_1(e^{j\omega}) = \frac{1}{2T_s} & \sum_{k=-\infty}^{\infty} X\left(j\frac{\omega}{T_s} - j\frac{\pi k}{T_s}\right)
\end{align*}
\]

(17)

Assume that the input signal \( x(t) = \sin(\Omega_0 t) \left( \Omega_0 \leq \frac{\pi}{2T_s} \right) \), according to Parseval’s Theorem, take out \( N_h \) point data, we can get:
Where \( x_m(n) \) (m = 0, 1) is time domain expression of \( X_m(j\Omega) \). The larger the \( N_h \) is, the more accurate the result of Eq. (18) will be. Import five various cosine signals whose frequencies are different respectively to solve Eq. (18), we can get the filter bank mismatch errors \( \varepsilon_n \) and gain mismatch error \( \Delta g \). Next, we estimate the timing mismatch error.

Similarity, assume that the input signal \( x(t) = \sin(\Omega_o t) \left( \Omega_o < \frac{\pi}{2T_s} \right) \). From Eq. (17), we can get:

\[
\begin{align*}
x_0[n] &= G_0 \sin[\Omega_o(n - \Delta t)T_s + \angle H_0(j\Omega)] \\
x_1[n] &= G_1 \sin(2n\Omega_o T_s)
\end{align*}
\]

Where \( G_0 \) and \( G_1 \) are gains of two channels and \( \angle H_0(j\Omega) \) is phase delay of \( H_0(j\Omega) \), the product of \( x_0[n] \) and \( x_1[n] \) can be expressed by:

\[
G(n) = \frac{1}{2} G_0 G_1 \cos(\Omega_o T_s \Delta t - \angle H_0(j\Omega)) - \frac{1}{2} G_0 G_1 \cos[\Omega_o T_s (4n - \Delta t) + \angle H_0(j\Omega)]
\]

Take out \( N_h \) sample values, the average of \( G(n) \) is given by:

\[
\frac{1}{N_h} \sum_{n=0}^{N_s-1} G(n) = k_0 \cos(\Omega_o T_s \Delta t - \angle H_0(j\Omega)) - R
\]

where \( k_0 = \frac{1}{2} G_0 G_1 \), \( R = \frac{1}{N_h} \sum_{n=0}^{N_s-1} k_0 \cos[\Omega_o T_s (4n - \Delta t) + \angle H_0(j\Omega)] \), and \( R \) satisfies:

\[
\lim_{N_s \to \infty} \frac{1}{N_h} \sum_{n=0}^{N_s-1} k_0 \cos[\Omega_o T_s (4n - \Delta t) + \angle H_0(j\Omega)] = 0, \quad \Omega_o \neq \frac{p\pi}{2T_s}
\]

Where \( p = 0, 1, 2 \ldots \), then we can get:

\[
\lim_{N_s \to \infty} \frac{1}{N_h} \sum_{n=0}^{N_s-1} G(n) = k_0 \cos(\Omega_o T_s \Delta t - \angle H_0(j\Omega))
\]

In the same way, we can get:

\[
\lim_{N_s \to \infty} \frac{1}{N_h} \sum_{n=0}^{N_s-1} G^1(n) = \lim_{N_s \to \infty} \frac{1}{N_h} \sum_{n=0}^{N_s-1} x_0(n)x_1(n - 1)
\]

\[
= k_0 \cos[\Omega_o T_s (\Delta t - 2) - \angle H_0(j\Omega)]
\]

The value of \( k_0 \) and \( \Delta t \) can be obtained from Eq. (23) and Eq. (24), and then \( H_0(j\Omega) \) could be estimated. Consequently, the estimation of \( F_0(e^{j\omega_0}) \) and \( F_1(e^{j\omega_0}) \) are got according Eq. (6). Finally, signal calibration is carried out in digital domain.
4 Simulation result

To demonstrate the effectiveness of proposed calibration method, a two-channel 10 GS/s 12-bit FIADC system is designed as an example.

Make a reasonable assumption that the analog filter transfer functions satisfy:

\[
\left| H_0\left( \frac{\pi}{2T_s} \right) \right|^2 = \frac{1}{2}
\]

\[
\left| H_0\left( \frac{\pi}{T_s} \right) \right|^2 = 0
\]

Using genetic algorithm, we get the analog filter expressed as:

\[
H_0(j\Omega) = \frac{101.5355 + 10.2877(j\Omega T_s)^2}{101.5355 + 50.9470(j\Omega T_s) + 10.2877(j\Omega T_s)^2 + (j\Omega T_s)^3}
\]

The amplitude-frequency property of \( H_0(j\Omega) \) and \( H_1(j\Omega) \) are show in Fig. 4.

![Fig. 4. The amplitude-frequency curve of analog filters.](image)

According to practical experience, we assume the parameter errors: \( \epsilon_0 = \epsilon_2 = 0.012, \epsilon_1 = \epsilon_3 = 0.015, \Delta g = 0.02, \Delta t = 3 \times 10^{-12} \). Using estimate method illustrated in section 3, we get the estimation results as shown in Table I. Obviously, the proposed method is effective for error estimation. Then design digital synthesis filters using Inverse Fast Fourier Transform to do calibrate. The method of designing digital filters describe in [14]. Using \( Q \) samples of the mth-channel synthesis filter Fourier transform \( F_m(e^{j\omega})|_{\omega=2\pi q}, q = 0, 1, \ldots, Q - 1, Q \) should be large enough that the impulse response has sufficiently decayed so that the time aliasing is negligible (e.g., \( Q = 1024 \)), therefore, the definition of the \( Q \)-point Inverse Fast Fourier Transform is:

\[
f_m^{(Q)}(n) \approx \frac{1}{2\pi} \int_{-\pi}^{\pi} F_m(e^{j\omega}) e^{j\omega n} d\omega
\]

However, given that the resulting \( Q \)-point impulse response is too long, we truncate it with rectangle window function (length L)

\[
\hat{f}_m(n) = f_m^{(Q)}(n) \omega(n)
\]

where the rectangle window function is
\[ \omega(n) = \begin{cases} 1, & 0 \leq n \leq L - 1 \\ 0, & L \leq n \leq Q \end{cases} \] 

(29)

Then adjust the system delay \( d \) and the window length \( L \), until the joint error function is minimized.

\[ e = \sum_{m=0}^{1} \sum_{n=L}^{Q} (f_m^{(Q)}(n))^2 \] 

(30)

This truncation operation brings digital synthesis filter errors which increase the distortion and aliasing errors. The digital synthesis filter length \( L \) should be smallest value that still yields the desired distortion and aliasing errors.

We finally designed two 64-length FIR digital filters and the system delay \( d = 32 \) with the joint error less than \( 1.24 \times 10^{-12} \). The distortion error before calibration is shown in Fig. 5(a) and the aliasing error before calibration are shown in Fig. 5(b). The aliasing error and the distortion error after calibration are shown in Fig. 6. The error is reduced obviously so that the calibration method is useful.

![Distortion Error](image1)

![Aliasing Error](image2)

Fig. 5. (a) Distortion error before calibration. (b) Aliasing error before calibration.
can calculate average aliasing error and average distortion error shown in Eq. (4), average distorting error is $1.5639 \times 10^{-5} \text{db}$ and average aliasing error is $-93 \text{db}$. It is enough to design 10 GS/s 12-bit FIADC system.

We can find there still have errors in Fig. 6. The main errors come from two parts. One is the estimation deviation of parameter in the analog filters. The other is digital synthesis filter errors due to truncate operation.
5 Conclusion

This paper introduces an improved structure of HFB which only need single analog filter, besides, an effective estimation and calibration technique of mismatch errors is proposed for FIADC, which is a competitive two-channel structure in High-speed and High-resolution applications. And the simulation results show that the technique has excellent effect.

Further researches are realizing this system in practical circuit, increasing number of channels to improve sampling rate, and looking for a better method for designing digital filters.