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Book Chapter

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Part of: IEEE International Symposium on Circuits and Systems (ISCAS), 2013,
pp. 2936-2939

DOI: <http://dx.doi.org/10.1109/ISCAS.2013.6572494>

ISBN: 978-1-4673-5760-9

IEEE International Symposium on Circuits and Systems, 0271-4310

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Available at: Linköping University Electronic Press
<http://urn.kb.se/resolve?urn=urn:nbn:se:liu:diva-106537>

Low-Complexity Two-Rate Based Multivariate Impulse Response Reconstructor for Time-Skew Error Correction in M -Channel Time-Interleaved ADCs

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Abstract—Nonuniform sampling occurs in time-interleaved analog-to-digital converters (TI-ADC) due to timing mismatches between the individual channel analog-to-digital converters (ADCs). Such nonuniformly sampled output will degrade the achievable resolution in a TI-ADC. To restore the degraded performance, digital time-varying reconstructors can be used at the output of the TI-ADC, which in principle, converts the nonuniformly sampled output sequence to a uniformly sampled output. As the bandwidth of these reconstructors increases, their complexity also increases rapidly. Also, since the timing errors change occasionally, it is important to have a reconstructor architecture that requires fewer coefficient updates when the value of the timing error changes. Multivariate polynomial impulse response reconstructor is an attractive option for an M -channel reconstructor. If the channel timing error varies within a certain limit, these reconstructors do not need any online redesign of their impulse response coefficients. This paper proposes a technique that can be applied to multivariate polynomial impulse response reconstructors in order to further reduce the number of fixed-coefficient multipliers, and thereby reduce the implementation complexity.

I. INTRODUCTION

Time-interleaving of analog-to-digital converters (ADCs) is a commonly used method in high-speed ADCs [1]–[3]. With this approach, the requirements on the individual channel ADCs are less stringent and helps to reduce the analog design complexity. However, the performance of the time-interleaved analog-to-digital converter (TI-ADC) depends on how well the individual channel ADCs are matched. Gain, offset, and timing mismatches between the channel ADCs, reduce the overall performance of the TI-ADC. Since a reduction of the mismatches between the channel ADCs will add more complexity to the analog design, the output of the TI-ADC is passed through a digital system which corrects the errors introduced due to mismatches. This paper considers digital correction of errors introduced due to timing mismatches in the TI-ADC. In order to ensure that the output samples are equally spaced, the sampling clocks to each channel ADC should maintain a distinct time-skew. However, mismatches between the channel ADCs will result in a static time-skew error in a channel ADC's sampling instant. Such static time-skew errors will result in a periodically nonuniformly sampled sequence at the output of the TI-ADC as illustrated in Fig. 1(b).

In [4], the reconstruction of M -periodic nonuniformly sampled signals using time-varying discrete-time FIR filters were considered in detail. Such a reconstructor can be used to correct static time-skew errors in an M -channel TI-ADC. However, the time-skew error in each channel can vary with time, and the reconstructor coefficients in [4] should be redesigned every time any channel's time-skew error changes. This implies that the reconstructor's coefficient multipliers should be implemented with general multipliers which can be updated online. Also, additional circuitry will be required to recompute the reconstructor coefficients online whenever the time-skew error of any channel changes. Several techniques have been proposed to correct

Fig. 1. (a) Uniform sampling (b) Three-periodic nonuniform sampling.

timing errors in TI-ADCs including architectures that reduce the number of filter coefficients that need online redesign [5] as well as those that do not require any online redesign [6]–[9] as long as all the channel time-skew errors are within a certain limit. While methods in [5]–[7] are specific for the two-periodic nonuniformly sampled signals, [8] and [9] can be applied to any M -periodic nonuniformly sampled signal. The method in [5] can be easily extended to the M -periodic case, but it would still require online redesign and, hence, significantly larger number of general multipliers. For a given M -channel specification, the design in [8], using the differentiator-multiplier cascade (DMC) or the DMC with reduced delay (DMC-RD) structures, require fewer fixed and general multipliers. However, due to the cascaded subfilter filter structure in [8], it is not possible to share delay elements and, compared to [9], it requires more delay elements and thereby increasing the implementation cost. The multivariate polynomial impulse response reconstructors proposed in [9] is an attractive option for reconstruction of M -periodic nonuniformly sampled signals if the maximum possible time-skew error is small, which is the case in a TI-ADC [10].

This paper proposes a method based on the two-rate approach, to reduce the number of fixed-coefficient multipliers in the multivariate polynomial impulse response reconstructor and is organized as follows. In Section II, a brief background on M -periodic nonuniform sampling and reconstruction is provided. This is followed by Section III, which reviews the multivariate polynomial impulse response reconstructor as well as the basic two-rate approach. Section IV introduces the two-rate based multivariate polynomial impulse response reconstructor and elaborates on the design procedure. The savings obtained using the proposed structure is illustrated with the help of a design example in Section V and Section VI concludes the paper.

II. M -PERIODIC NONUNIFORM SAMPLING AND RECONSTRUCTION

The output sequence of an ideal M -channel TI-ADC can be represented as

$$x(n) = x_a(nT) \quad (1)$$

where $x_a(t)$ is the continuous-time signal applied as input to the TI-ADC and $t = nT$ represents the uniformly spaced sampling instants at the output of the TI-ADC. To obtain uniformly spaced samples at the output of the TI-ADC, the sampling clocks of any two adjacent sub-ADCs should have a time-skew of T . However, in practice, due to mismatches in the sub-ADCs and clock routing networks, the time-skew between adjacent channels will not be uniform, resulting in a nonuniformly sampled signal, $v(n)$, at the output of the TI-ADC. If $\varepsilon_n T$ represents the difference between the uniform and nonuniform sampling instants, the nonuniformly sampled output can be written as

$$v(n) = x_a(nT + \varepsilon_n T). \quad (2)$$

The time-skew error, $\varepsilon_n T$, in an M -channel TI-ADC will be M -periodic with $\varepsilon_n = \varepsilon_{n+M}$, as shown in Fig. 1(b) for a three-channel TI-ADC ($M = 3$).

In principle, the nonuniformly sampled output, $v(n)$, can be passed through a perfect reconstructor to generate the uniformly sampled sequence, $x(n)$. However, it is not practically feasible to attain perfect reconstruction and, hence, the reconstructor output, $y(n)$, only approximates the uniformly sampled sequence, $x(n)$. Design of minimum-order reconstructors using time-varying discrete-time FIR system was considered in [4]. For such a reconstructor, with a time-varying impulse response, $h_n(k)$, the reconstructed output, $y(n)$, is given by

$$y(n) = \sum_{k=-N}^N v(n-k)h_n(k) \quad (3)$$

where it is assumed, for simplicity in the design, that the reconstructor is an even-order noncausal FIR system. It was shown in [4] that, for an M -periodic nonuniformly sampled sequence, the impulse response of the reconstructor will also be M -periodic such that $h_n(k) = h_{n+M}(k)$. Designing such a reconstructor involves determining the coefficients of $h_n(k)$ that minimizes the error $e(n) = y(n) - x(n)$. For a continuous-time signal, $x_a(t)$, bandlimited to $\omega_0 T < \pi$ and nonuniformly sampled with a time-skew error, $\varepsilon_n T$, perfect reconstruction requires that

$$A_n(j\omega T) = 1, \quad \omega T \in [-\omega_0 T, \omega_0 T] \quad (4)$$

where

$$A_n(j\omega T) = \sum_{k=-N}^N h_n(k)e^{-j\omega T(k-\varepsilon_n T)}. \quad (5)$$

Practical reconstructor design involves determining the coefficients of $h_n(k)$ that minimizes the error between $A_n(j\omega T)$ and 1 for a given bandwidth $\omega_0 T$ and time-skew error $\varepsilon_n T$.

III. REVIEW OF MULTIVARIATE POLYNOMIAL IMPULSE RESPONSE RECONSTRUCTORS AND TWO-RATE APPROACH

A. Multivariate Polynomial Impulse Response Reconstructors

The coefficients of $h_n(k)$ in [4] should be redesigned whenever the time-skew error $\varepsilon_n T$ changes which occurs in practice occasionally due to voltage and temperature changes. Hence, all the multipliers used in such reconstructors are implemented using variable coefficient multipliers. Also, such reconstructors need extra circuitry to implement the online redesign block. In [9], multivariate polynomial

Fig. 2. Realization of multivariate polynomial impulse response reconstructor.

impulse response reconstructors were introduced to reconstruct M -periodic nonuniformly sampled signals. The impulse response of the reconstructor, $h_n(k)$, is expressed as a multivariate polynomial given by [9]

$$h_n(k) = \delta(k) + \sum_{\mathbf{r} \in \mathbf{S}} \alpha_{r_0 r_1 \dots r_{M-1}}(k) \varepsilon_n^{r_0} \varepsilon_{n+1}^{r_1} \dots \varepsilon_{n+M-1}^{r_{M-1}} \quad (6)$$

with

$$\mathbf{r} = [r_0 \ r_1 \ \dots \ r_{M-1}]^T \quad (7)$$

and $\mathbf{S}$ is a set of vectors such that

$$r_0 \in [1, 2, \dots, L], \quad (8)$$

$$r_m \in [0, 1, \dots, L-1], \quad m = 1, 2, \dots, M-1, \quad (9)$$

and

$$r = \sum_{m=0}^{M-1} r_m \in [1, 2, \dots, L]. \quad (10)$$

Equations (8)–(10) imply that the combinations of r_m that are included in the multivariate polynomial expansion of $h_n(k)$ should contain at least one of $\varepsilon_n, \varepsilon_n^2, \dots, \varepsilon_n^L$. Excluding terms that do not contain ε_n is necessary to ensure that the reconstructor passes the sample as it is if, for a particular sub-ADC, the time-skew error is $\varepsilon_n T = 0$. Taking the z -transform of $h_n(k)$ to obtain the transfer function, $H_n(z)$, and using (6), it can be shown that [9]

$$H_n(z) = 1 + \sum_{p=1}^P e_p Q_p(z) \quad (11)$$

where

$$Q_p(z) = \sum_{k=-N}^N \alpha_p(k) z^{-k} \quad (12)$$

with $\alpha_p(k)$ representing the coefficients corresponding to the p th combination of $r_0 r_1 \dots r_{M-1}$ and, $e_p, p = 1, 2, \dots, P$, are the P valid combinations of $\varepsilon_n^{r_0} \varepsilon_{n+1}^{r_1} \dots \varepsilon_{n+M-1}^{r_{M-1}}$ that satisfy (8)–(10), for a given pair of L and M . For example, if $L = 2$ and $M = 4$, $P = 5$ since the possible combinations of $r_0 r_1 r_2 r_3$ that satisfy (8)–(10) are 1000, 1001, 1010, 1100, and 2000. In order to simplify the design, subfilters $Q_p(z)$ are here assumed to be noncausal with order $2N$. Figure 2 shows the realization of the multivariate reconstructor.

B. Two-Rate Approach

The basic two-rate approach is shown in Fig. 3(b). Multirate theory [11] can be used to convert the two-rate structure in Fig. 3(b) to an equivalent single-rate realization of the FIR filter $Q_p(z)$ as shown in Fig. 3(c). The filter $F(z)$ is a linear-phase half-band filter in which the filter coefficients are symmetric with every other coefficient being equal to zero. Hence, its zeroth polyphase component $F_0(z)$ is also symmetric while the polyphase component $F_1(z)$ is a pure delay equal to $z^{-(D_F-1)/2}$ where D_F is the delay of $F(z)$. The requirements on the subfilter $G_p(z)$ is relaxed resulting in a lower order filter compared to $Q_p(z)$. The sparsity and symmetry of $F(z)$ combined with the simpler $G_p(z)$, helps in reducing the overall

Fig. 3. (a) Original filter. (b) Two-rate structure. (c) Equivalent single-rate realization using polyphase components of $F(z)$ and $G_p(z)$.

Fig. 4. Realization of the proposed two-rate based multivariate polynomial impulse response reconstructor.

complexity of the structure in Fig. 3(c) compared to that in Fig. 3(a). Also, additional savings is obtained when the half-band filter $F(z)$ is shared among all the $Q_p(z)$ subfilters in Fig. 2.

IV. PROPOSED TWO-RATE BASED MULTIVARIATE POLYNOMIAL IMPULSE RESPONSE RECONSTRUCTOR

The realization of the proposed two-rate based multivariate impulse response reconstructor is shown in Fig. 4. In this reconstructor, the two-rate approach is used to reduce the number of multipliers in the fixed $Q_p(z)$ subfilters. Also, additional savings are obtained since the polyphase components of the half-band filter, $F_0(z)$ and $F_1(z)$, are shared between all the subfilters $Q_p(z)$. In order to simplify the design, the delay term z^{-1} in Fig. 3 is propagated into $F_1(z)$ block in Fig. 4, and hence, $F_1(z)$ represents a delay of $z^{-(D_F+1)/2}$. If $-\varepsilon_{\max} \leq \varepsilon_n \leq \varepsilon_{\max}$, the symmetry properties imposed on the coefficients of $Q_p(z)$ in the multivariate reconstructor [9] in Fig. 2 can also be applied to the coefficients of $G_{p0}(z)$ and $G_{p1}(z)$ in Fig. 4.

A. Reconstructor Design

The coefficients of $F_0(z)$, $G_{p0}(z)$, and $G_{p1}(z)$ are determined such that the same coefficients can be used for all values of $\varepsilon_n \in [-\varepsilon_{\max}, \varepsilon_{\max}]$. Whenever the time-skew error $\varepsilon_n T$ changes, the coefficients used in the variable multipliers can be updated directly with the new ε_n . Since the reconstructor coefficients are designed offline, the coefficients can be designed using minimax optimization. In order to design an L th-order multivariate reconstructor for an M -channel TI-ADC, the reconstructor design problem can be stated as follows:

Given the order of the subfilters $F_0(z)$, $G_{p0}(z)$, and $G_{p1}(z)$, $p = 1, 2, \dots, P$, as well as $-\varepsilon_{\max} \leq \varepsilon_n, \varepsilon_{n+1}, \dots, \varepsilon_{n+M-1} \leq \varepsilon_{\max}$, for a given bandwidth, $\omega_0 T < \pi$, determine the impulse response coefficients of these subfilters and a parameter δ , to minimize δ subject to

$$|A_n(j\omega T) - 1| \leq \delta, \omega T \in [0, \omega_0 T]. \quad (13)$$

The requirement is satisfied if, $\delta \leq \delta_e$, after the optimization, where δ_e is the maximum reconstruction error that can be tolerated.

The computation of $A_n(j\omega T)$ is simplified if it is evaluated using matrix multiplications. Since $F(z)$ is a half-band filter, its order will be $4K+2$ which means that the length of the impulse response of the zeroth polyphase component, $F_0(z)$, will be $2K$. Hence, the impulse response of $F_0(z)$ can be represented as

$$\mathbf{f}_0 = [f_0(-K) \ f_0(-K+1) \ \dots \ f_0(K-1)] \quad (14)$$

The impulse response of the delay, $F_1(z)$, can also be considered to be of length $2K$ and is denoted as

$$\mathbf{f}_1 = [f_1(-K) \ f_1(-K+1) \ \dots \ f_1(K-1)] \quad (15)$$

where

$$f_1(k) = \begin{cases} 1, & k = 1 \\ 0, & k \neq 1 \end{cases}. \quad (16)$$

Assume that $\mathbf{g}_{p0}$ and $\mathbf{g}_{p1}$ are the impulse response vectors of $G_{p0}(z)$ and $G_{p1}(z)$, respectively, such that

$$\mathbf{g}_{p0} = [\alpha_{p0}(-R) \ \alpha_{p0}(-R+1) \ \dots \ \alpha_{p0}(R)] \quad (17)$$

and

$$\mathbf{g}_{p1} = [\alpha_{p1}(-R) \ \alpha_{p1}(-R+1) \ \dots \ \alpha_{p1}(R)] \quad (18)$$

where both $\mathbf{g}_{p0}$ and $\mathbf{g}_{p1}$ are assumed to have a length of $2R+1$ to simplify the derivation.

Then, $A_n(j\omega T)$ in (13) can be computed using

$$A_n(j\omega T) = e^{j\omega T \varepsilon_n} + \mathbf{e} \mathbf{Q}^T \mathbf{a} \quad (19)$$

where

$$\mathbf{e} = [e_1 \ e_2 \ \dots \ e_P], \quad (20)$$

$$\mathbf{Q} = \mathbf{F}_1 \mathbf{G}_1, \quad (21)$$

and

$$\mathbf{a} = [\mathbf{a}_1 \ \mathbf{a}_1]^\dagger \quad (22)$$

with $\mathbf{X}^T$ and $\mathbf{X}^\dagger$ representing transpose and conjugate transpose of the matrix $\mathbf{X}$, respectively. Further, e_p , $p = 1, 2, \dots, P$, in (20) are the P valid combinations of $\varepsilon_n^{r_0} \varepsilon_{n+1}^{r_1} \dots \varepsilon_{n+M-1}^{r_{M-1}}$ that satisfy (8)–(10).

The matrix $\mathbf{F}_1$ in (21) is defined as

$$\mathbf{F}_1 = [\mathbf{F}_{10}^T \ \mathbf{F}_{11}^T] \quad (23)$$

with

$$\mathbf{F}_{10} = [\mathbf{F}_{t10}^T \ \mathbf{Z}_{2K+2R,2R+1}^T] \quad (24)$$

and

$$\mathbf{F}_{11} = [\mathbf{Z}_{2K+2R,2R+1}^T \ \mathbf{F}_{t11}^T] \quad (25)$$

where $\mathbf{Z}_{r,q}$ is an $r \times q$ zero matrix and $\mathbf{F}_{t1r}$ is a $(2K+2R) \times (2R+1)$ Toeplitz matrix with first row $[f_{1r}(-K) \ \mathbf{Z}_{1,2R}]$ and first column $[\mathbf{f}_{1r} \ \mathbf{Z}_{1,2R}]^T$.

The matrix $\mathbf{G}_1$ in (21) is defined as

$$\mathbf{G}_1 = [\mathbf{g}^{(1)} \ \mathbf{g}^{(2)} \ \dots \ \mathbf{g}^{(P)}] \quad (26)$$

where

$$\mathbf{g}_p = [\mathbf{g}_{p0} \ \mathbf{g}_{p1}]^T \quad (27)$$

for $p = 1, 2, \dots, P$.

The vector $\mathbf{a}_1$ in (22) is defined as

$$\mathbf{a}_1 = [a_1(-N) \ a_1(-N+1) \ \dots \ a_1(N-1)] \quad (28)$$

with $a_1(k) = e^{-j\omega T(k-\varepsilon_{n-k})}$ for $k = -N, -N+1, \dots, N-1$ and $N = 2K + 2R$.

The design steps to identify the coefficients of $F_0(z)$ and of the subfilters $G_{p0}(z)$ and $G_{p1}(z)$, $p = 1, 2, \dots, P$, are as follows:

- 1) Determine the order, $\hat{N}_F$, of a standard half-band linear-phase FIR filter, $F(z)$, with passband edge at $\Omega_c = \omega_0 T/2$, stopband edge at $\Omega_s = \pi - \Omega_c$, and with the maximum ripple in the passband and stopband being δ_e .
- 2) For all combinations of N_F (in the vicinity of $\hat{N}_F$) and N_{G_1} , the order of the subfilters $G_p(z)$ (in the vicinity of $\hat{N}_{G_1} = 6$ which has been determined experimentally), identify the coefficients of $F_0(z)$ and of the subfilters $G_{p0}(z)$ and $G_{p1}(z)$, $p = 1, 2, \dots, P$, by solving the optimization problem in (13) for all the combinations of $-\varepsilon_{\max}$, 0, and $\varepsilon_{\max}$ for $\varepsilon_n, \varepsilon_{n+1}, \dots, \varepsilon_{n+M-1}$. If δ obtained from the optimization routine is smaller than δ_e , save the results.
- 3) From all the results in Step 2 that satisfied the requirement, choose the one with lowest complexity as the final solution.

Since $-\varepsilon_{\max}$, 0, and $\varepsilon_{\max}$ for $\varepsilon_n, \varepsilon_{n+1}, \dots, \varepsilon_{n+M-1}$ and as all combinations of $\varepsilon_n, \varepsilon_{n+1}, \dots, \varepsilon_{n+M-1}$ are considered in Step 2, it is sufficient to determine the subfilter coefficients that satisfy the minimization problem in (13) for any one channel.

V. DESIGN EXAMPLE

In this section, the reconstruction of nonuniformly sampled output from a four-channel TI-ADC ($M = 4$) is considered with $L = 2$, $\varepsilon_{\max} = 0.02$, $\omega_0 T = 0.8\pi$, and with a maximum reconstruction error of $\delta_e = -80$ dB. In this case, the reconstructor consists of five subfilters $Q_p(z)$ corresponding to the five possible combinations of $r_0 r_1 r_2 r_3$ that satisfy (8)–(10) namely 1000, 1001, 1010, 1100, and 2000. The coefficients of $F_0(z)$, $G_{p0}(z)$, and $G_{p1}(z)$, $p = 1, 2, \dots, P$, are identified such that the maximum error, δ , in (13) is less than $\delta_e = -80$ dB. The design which satisfies the maximum reconstruction error requirement, and which gives the minimum number of distinct filter coefficients, is selected. Applying symmetry constraints for the subfilters $G_{p0}(z)$ and $G_{p1}(z)$ helps to further reduce the number of distinct multipliers. The impulse response coefficients of the subfilters, $G_{p0}(z)$ and $G_{p1}(z)$, corresponding to $r_0 r_1 r_2 r_3 = 1000$, are anti-symmetric. For $r_0 r_1 r_2 r_3 = 1010$ and $r_0 r_1 r_2 r_3 = 2000$, the corresponding $G_{p0}(z)$ and $G_{p1}(z)$ impulse responses are symmetric. The subfilters corresponding to $r_0 r_1 r_2 r_3 = 1001$ and $r_0 r_1 r_2 r_3 = 1100$ have the same coefficients but in the reverse order. The reconstructor that satisfies the requirement in this example require 12 distinct coefficients for $F_0(z)$ and 28 distinct coefficients for realizing the five $G_{p0}(z)$ and $G_{p1}(z)$ subfilters in transposed direct-form FIR structures. Hence, the proposed architecture requires 40 fixed-coefficient multipliers and 5 general multipliers. This means that more savings in terms of the number of fixed coefficient multipliers is achieved compared to the structure in [9], which requires 53 fixed-coefficient and 5 variable multipliers. To implement the above specification, the DMC reconstructor [8] using transposed direct-form FIR structure requires only 33 fixed and 3 variable multipliers but the reconstructor needs 66 delay elements. However, due to its structure, the proposed reconstructor realized using direct-form FIR structures require only 31 delay elements with 40 fixed multipliers. It should be noted that, a comparison of the complexity between [8] and the proposed method is delicate, since [8] designs a less stringent reconstructor compared to that designed in this paper where the requirement is satisfied in the minimax sense. Figure 5 shows the reconstruction error for all the four channels while using the subfilter coefficients designed to meet the given

Fig. 5. Magnitude of $A_n(j\omega T) - 1$, $n = 0, 1, 2, 3$, for the two-rate based multivariate polynomial impulse response reconstructor in the design example, for $\varepsilon_0 = -0.02$, $\varepsilon_1 = 0.015$, $\varepsilon_2 = -0.01$, and $\varepsilon_3 = 0.009$.

specification and when $\varepsilon_0 = -0.02$, $\varepsilon_1 = 0.015$, $\varepsilon_2 = -0.01$, and $\varepsilon_3 = 0.009$.

VI. CONCLUSION

In this paper, we outlined a two-rate based multivariate polynomial impulse response reconstructor for an M -channel TI-ADC. It was shown, with the help of a design example, that the proposed structure provides further savings in the number of fixed-coefficient multipliers compared to the original multivariate polynomial impulse response reconstructor.

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