Joint Sampling-Time Error and Channel Skew Calibration of Time-Interleaved ADC in Multichannel Fiber Optic Receivers

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***Abstract*— A joint sampling-time error and channel skew background calibration technique for time interleaved analog to digital converters (TI-ADC) is presented. The technique is aimed at applications in dual-polarization QPSK/QAM receivers for coherent optical communications at high data rates (e.g., 40Gb/s and beyond). Unlike previous proposals, the calibration algorithm introduced here is used to *jointly* compensate for sampling-time and channel skew errors. Estimates of the gradient of the mean squared error (MSE) or the bit error rate (BER) with respect to the sampling phases of the different signal lanes and interleaves are computed and used to iteratively minimize a cost function (i.e., MSE or BER). Computer simulations demonstrate the excellent behavior of the proposed compensation technique. The calibration algorithm can be implemented with minimal hardware requirements and with a slow clock. This allows power dissipation in a CMOS VLSI implementation to be minimized.**

I. Introduction

Optical communication technology in long-haul and metropolitan links is experiencing a transition to coherent techniques and high spectral efficiency modulation formats such as dual-polarization (DP) QPSK, DP-QAM and OFDM. The combination of coherent demodulation and digital signal processing (DSP) allows costly optical signal processing hardware used to compensate fiber optic impairments such as chromatic dispersion (CD) and polarization-mode dispersion (PMD) to be replaced by DSP-based techniques [1]. A key factor in the performance of these receivers is the analog front-end (AFE), which typically encompasses four analog to digital converters (ADC), sampling the four components of the input signal vector (see Fig. 1). These are the in-phase (I) and quadrature (Q) components of the horizontal (H) and vertical (V) polarizations. The required resolution is approximately 6-7 bits and the sampling rate is typically twice the symbol rate. Because of the forward error correction (FEC) coding overhead, the symbol rate for a 40 Gigabits per second (Gb/s) receiver may be as high as 12.5 Gigabauds (GB) which results in an ADC sampling rate of 25 Gigasamples per second (Gs/s). Similarly, for a 100Gb/s receiver the symbol rate could be as high as 32GB, resulting in an ADC sampling rate of 64Gs/s.

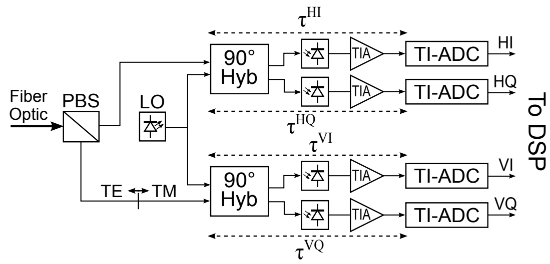

Fig. 1. Optical/analog front-end for a coherent optical receiver.

Time interleaved analog to digital converters (TI-ADC) are required to achieve these high sampling rates.

It is well known that mismatches of the sampling-time, gain, offset and frequency response among the interleaves of a TI-ADC (see Fig. 2) limit the performance of the converter unless they are compensated. This problem has motivated a large body of research and many calibration techniques have been proposed. See [2]–[7] and references therein for a review of these techniques. These investigations tend to focus on the problem of sampling-time error estimation, which is typically more difficult than the estimation of the other errors.

A problem closely related to sampling-time error calibration in TI-ADCs arises in multichannel receivers where the timing-skew (TS) among channels may severely limit the receiver performance [8]. In this case the sampling-time error occurs among different TI-ADCs instead of the interleaved ADCs of a single TI-ADC, and it may be present in the input signals before they reach the receiver AFE, as well as be augmented by the phase skews of the ADC sampling clocks. In the case of coherent optical receivers, the main source of the TS among channels is the inevitable imperfections of the optical demodulators. It can be shown that a 4-D feedforward equalizer (FFE) can compensate the skews between polarizations, but not the skew between the I and Q components of each polarization [1]. Although there have been some studies of the TS compensation problem in the technical literature [9], these studies address the compensation of the skews using DSP

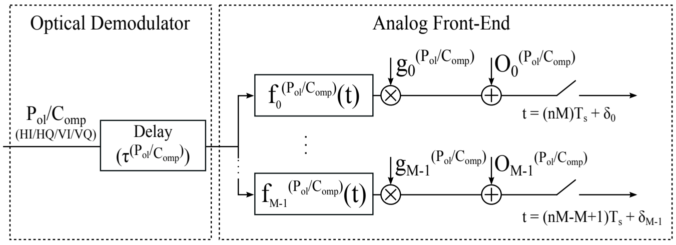

Fig. 2. Channel model of one of the four channels (HI,HQ,VI,VQ).

based techniques once the skew is known, but not the skew estimation problem. To the best of the authors knowledge no automatic skew compensation technique has been proposed so far. The TS estimation is challenging in the presence of large intersymbol interference (ISI) as typically occurs at the input of the receiver. The main sources of ISI in the optical channel are CD and PMD. A common specification for the coherent receiver is to be able to compensate the CD of several thousands kilometers of standard optical fiber, which results in ISI spanning hundreds or even thousands of symbol periods. Unfortunately, once the ISI is compensated by the equalizers the I and Q components are combined in intricate ways, which makes skew estimation even harder.

In this paper, we propose a novel technique to jointly estimate and compensate the I/Q TS and the sampling-time errors of the TI-ADC in coherent optical receivers. Estimates of the gradient of the slicer mean squared error (MSE) or the bit error rate (BER) with respect to the sampling phases of the different signal lanes and interleaves are computed and used to iteratively and jointly minimize the sampling time errors of the TI-ADC and the skew among the I and Q channels. Unlike other schemes proposed to calibrate sampling-time errors in TI-ADC [5], [6], our technique optimizes a metric directly related to the receiver performance. In this sense, our approach is closely related to the technique proposed in [10].

The proposed technique is demonstrated based on computer simulations using a complete simulation model of a DP-QPSK transceiver. It is shown that this technique can adjust the sampling phase of several interleaves (in the examples considered, 32 ADCs distributed in 4 channels) with a simple iterative round-robin phase adjust of each interleave. As mentioned before, the slicer MSE is one of the options for a calibration feedback signal. Another option takes advantage of the use of forward error correction to enable estimation of the BER of the receiver. Both the MSE and BER are direct measures of the system performance. In a practical application, the choice of one or the other should be based on optical signal to noise ratio (OSNR) at the input of the receiver. In general, MSE should be used as feedback when the OSNR is high, and BER should be used at low OSNR.

The rest of this paper is organized as follows. Section II describes the receiver architecture used in coherent receivers and discusses the impact of sampling-time and channel skew errors on the receiver performance. Section III introduces the calibration algorithm, as well as calibration convergence and simulation results. Section IV summarizes the conclusions.

II. Impact of Sampling-Time and Channel Skew Errors on the Performance of DP-QPSK Optical Receivers

A. System Model

Figure 1 shows a block diagram of an optical front-end (OFE) for a DP-QPSK coherent receiver. The optical signal at the OFE output consists of four channels (the in-phase and quadrature components (I/Q) of the two polarizations (H/V)). The resulting electrical signals (i.e., *HI, HQ, VI*, and *VQ*) are subsequently processed by the AFE. Typically, oversampled digital receivers are used to compensate the dispersion experienced in optical links (e.g., $T_s = \frac{T}{2}$ where T is the symbol period) [1]. Figure 2 shows a simplified model for one of the four channels. Each AFE stage consists of an M-parallel time interleaved ADC system. Blocks $f_0^{(P_{ol}/C_{omp})}(t)$ to $f_{M-1}^{(P_{ol}/C_{omp})}(t)$ model the independent frequency responses of each track and hold (T&H) unit of the channel P_{ol}/C_{omp}, with $P_{ol} \in \{H, V\}$ and $C_{omp} \in \{I, Q\}$. Gain errors and offsets are modeled by $g_0^{(P_{ol}/C_{omp})}$ to $g_{M-1}^{(P_{ol}/C_{omp})}$ and $O_0^{(P_{ol}/C_{omp})}$ to $O_{M-1}^{(P_{ol}/C_{omp})}$, respectively. Parameters $\delta_0^{(P_{ol}/C_{omp})}$ to $\delta_{M-1}^{(P_{ol}/C_{omp})}$ model the sampling-time errors. Parameter $\tau^{(P_{ol}/C_{omp})}$ represents a delay introduced in the channel P_{ol}/C_{omp} by the optical demodulators. Owing to imperfections in optical demodulators, these delays can be different among the channels. In particular, the time delay difference (*skew*) between components I and Q of each polarization (i.e., $\tau_s^{(P_{ol})} = \tau^{(P_{ol}/I)} - \tau^{(P_{ol}/Q)}$) result of interest because it cannot be compensated by the equalization stage. As we shall show in the following, the combination of the sampling time errors and the I/Q TS may seriously degrade the receiver performance.

B. Numerical Results

Next we analyze the impact of the sampling-time errors, $\delta_i^{I(Q)}, \quad i = 0, 1, ..., M-1$, and the I/Q TS, τ_s, on the performance of a $T/2$ DP-QPSK coherent receiver[1]. Sampling-time errors $\delta_i^{I(Q)}$ are assumed random variables uniformly distributed in the interval $[-\Delta_{max}, +\Delta_{max}]$. The number of interleaved ADCs per channel is $M = 8$. We consider a typical fiber link of $1000Km$ with $100ps$ of differential group delay (DGD) and $4000ps^2$ of second-order PMD. Gain, offset, and T&H frequency response mismatches are neglected. The ADC outputs are processed by a digital signal processor (DSP), which implements the main receiver functions such as compensation of CD and PMD, timing and carrier recovery stages, FEC decoder, etc. (see Fig. 4). More details of digital coherent receivers can be found in [1] and references therein.

Figure 3 shows the OSNR penalty at $BER = 10^{-3}$, as a function of the maximum sampling-time error Δ_{max} and the I/Q timing-skew τ_s. Note that $\Delta_{max}/T < 0.03$ and $\tau_s/T < 0.05$ could be required to achieve an OSNR penalty lower than $0.2dB$. At ultra high transmission rates (e.g., 100 Gb/s and beyond), the use of efficient calibration techniques is mandatory to satisfy these specifications.

[1]For simplicity of notation, in the rest of this work we focus the analysis on only one polarization (i.e., the polarization index is omitted).

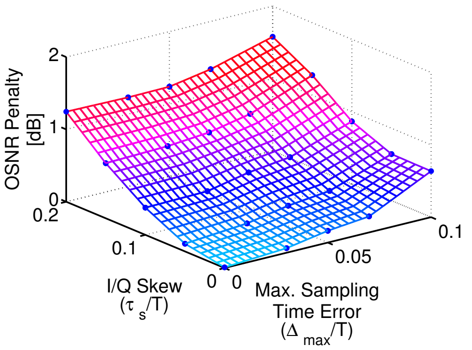

Fig. 3. OSNR penalization at $BER = 10^{-3}$ as a function of the maximum sampling-time error (Δ_{max}) and the I/Q timing-skew.

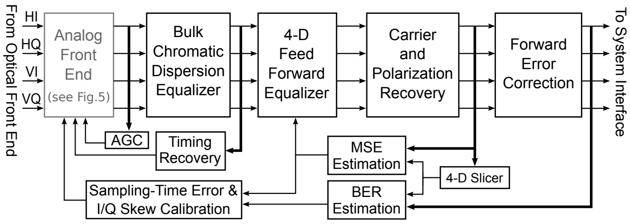

Fig. 4. Digital signal processor architecture for a DP-QPSK coherent receiver.

III. Joint Calibration of Sampling-Time and Channel Skew Errors

A. Calibration Principle

In new-generation optical receivers the BER estimation is possible since powerful FEC codes are typically available. Taking advantage of this fact, we propose a calibration technique designed to minimize the BER (denoted as *M-BER*). Furthermore, we also propose a second calibration technique designed to minimize the MSE measured at the FFE slicer (denoted as *M-MSE*) that avoids the use of FEC decoder information.

The proposed techniques (i.e., M-BER and M-MSE), unlike previous work, can be used to simultaneously compensate *both* the sampling-time errors of TI-ADC and the I/Q TS. Figures 4 and 5 show an implementation example of these mixed-signal calibration techniques. In this example, the AFE (Fig. 5) includes a set of digitally programmable time-delay cells [5], [6]. These $4 \times M$ time-delay cells are controlled by the calibration block in order to set the sampling phase as the algorithm requires.

B. Calibration Algorithm

Let ζ_i be the sampling phase of the i-th TI-ADC. In our case, note that the total number of ADCs is $4 \times M$. We also define $\mathcal{C}(\zeta_0, \zeta_1, ..., \zeta_{4\times M-1})$ as the cost function to be minimized (i.e., BER or MSE). Then, the *gradient algorithm* could be used to iteratively adjust the sampling phase in order to minimize the cost function $\mathcal{C}(.)$, that is,

$$\overline{\zeta}(n+1) = \overline{\zeta}(n) - \mu \nabla_{\overline{\zeta}(n)} \mathcal{C}, \tag{1}$$

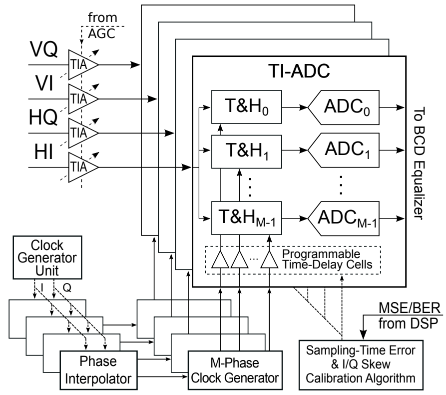

Fig. 5. Detailed block diagram of proposed Analog Front-End based in four TI-ADC with mixed signal calibration.

where n is the number of iteration, $\overline{\zeta}(n) = [\zeta_0(n), \zeta_1(n), ..., \zeta_{4\times M-1}(n)]^T$ is the sampling phase vector at the n-th iteration (symbol ^{T} denotes transpose), $\nabla_{\overline{\zeta}} \mathcal{C} = \left[\frac{\partial \mathcal{C}(\zeta_0,...,\zeta_{4\times M-1})}{\partial \zeta_0}, ..., \frac{\partial \mathcal{C}(\zeta_0,...,\zeta_{4\times M-1})}{\partial \zeta_{4\times M-1}}\right]^T$ is the gradient of the cost function $\mathcal{C}(.)$, while μ is the step size. Unfortunately, it is very difficult to derive simple closed-form expressions for BER or MSE as a function of the sampling phase of TI-ADC and I/Q skew. This fact precludes the use of a well known minimization technique such as (1). Therefore, we introduce here the following iterative method to adjust the ADC sampling phases:

1: Set to zero the $4 \times M - 1$ sampling phases (i.e., $\zeta_i = 0$ ($i \in [0, 1, ..., 4 \times M - 1]$). Moreover, set to zero the ADC index k.
2: Estimate the initial cost function $\mathcal{C}$ (BER or MSE).
3: Move the sampling phase of the k-th ADC in a *positive* direction, that is,

$$\zeta'_k = \zeta_k + \mu_s, \tag{2}$$

where μ_s the calibration step of the corresponding programmable time-delay cell (e.g., 1% of the baud period).
4: Re-estimate the cost function for the new sampling phase, $\mathcal{C}'$.
5: Adjust the sampling phase of the k-th ADC according to

$$\zeta_k = \zeta_k - \mu_s sign\left(\mathcal{C}' - \mathcal{C}\right). \tag{3}$$

6: Select a new ADC and repeat steps 2 through 5. Use a round-robin selection mode to chose the next ADC (e.g., for $M = 8$, the order results $k = 0, 8, 16, 24, 1, 9, 17, 25, 2, ..., 23, 31$).
7: After all ADCs have been adjusted by a time step μ_s, repeat steps 2 to 6.

The gradient direction of the cost function, $sign\left(\mathcal{C}' - \mathcal{C}\right)$, for the M-BER algorithm can be estimated by the error counting arisen from the comparison between FEC and FFE slicer outputs. On the other hand, the M-MSE cost function difference estimation can be carried out using well-known estimation techniques [11]. The minimum number of MSE

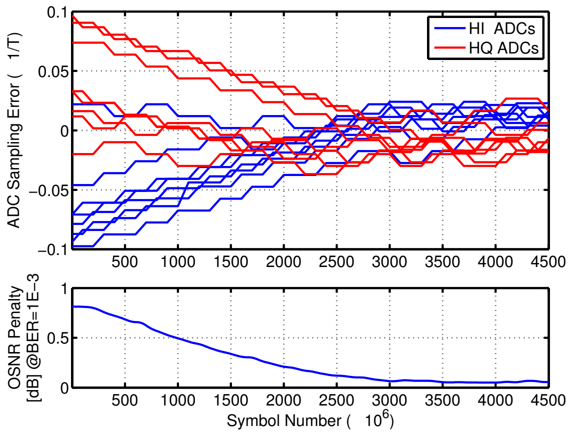

Fig. 6. Calibration convergence from an initial random set of sampling-time error and I/Q TS. M-MSE calibration with $OSNR = 13dB$. Top: ADC phase errors for polarization H. Bottom: OSNR penalty at $BER = 10^{-3}$

samples averaged for an accuracy gradient direction estimation is:

$$N_s > \left(\frac{2n_\sigma (4 \times M)}{k \cdot OSNR} \cdot \frac{T^2}{\mu_s \delta_i} \right)^2 \tag{4}$$

where δ_i is the phase error of the ADC under calibration, μ_s is the calibration time step and $4 \times M$ is the total number of ADCs in the AFE. The sampling-time error noise [12] depends on the channel response and it is represented by k (between 12 to 15 for dispersive channels). Finally, n_σ is the estimation confidence interval of the gradient direction in terms of standard deviations from the mean ($n_\sigma \in [1, 3]$).

C. Performance Evaluation

The capability of the proposed techniques to compensate the sampling time errors and the I/Q TS is analyzed in this section. Figure 6 shows the sampling-time error for all ADCs of polarization H (8 per channel). In this example, we set the same system parameters used in Section II with a sampling-time error bounded by $\Delta_{max} = \pm 0.07T$, an I/Q timing-skew $\tau_s = 0.1T$ and input $OSNR = 13dB$. We observe that the phase errors of the ADCs are converged together around the mean of sampling-time error while the OSNR penalty at constant $BER = 10^{-3}$ is minimized from $0.8dB$ to $0.05dB$. Owing to space limitations, we consider here the M-MSE calibration approach.

Figure 7 presents a comparative pre/post calibration performance (for $\Delta_{max} = \pm 0.1T$ and $\tau_s = 0.1T$). Furthermore, the post-calibration performance of the receiver is near optimal despite the residual sampling-time error after convergence.

IV. Conclusions

We have proposed a new joint sampling-time error and channel skew background calibration technique for TI-ADC. The method is aimed at applications in next-generation AFE of coherent optical receivers. Simulation results have demonstrated the excellent behavior of the technique to mitigate both undesirable effects (i.e., the sampling-time errors and the I/Q skew). The proposed calibration algorithm operates in background, after DSP convergence, and it can periodically compensates sampling phase variations during normal operation (e.g., due to temperature variation). Moreover, it requires low extra implementation complexity and it can works effectively on highly dispersive channels typical of long-haul and ultra long-haul optical links.

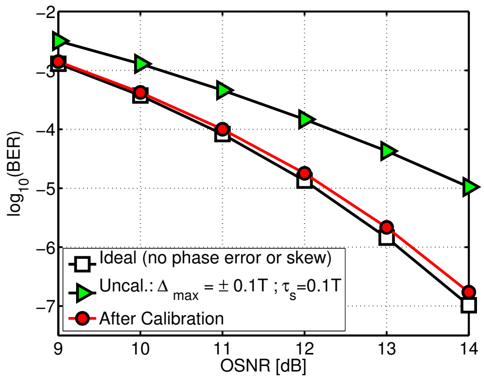

Fig. 7. Performance comparison of pre and post calibration with M-MSE.

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