Advanced Digital Signal Processing for Coherent and Non-coherent Optical Transmission

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ABSTRACT

Digital signal processors (DSP) are recently introduced in coherent long-haul optical transmission systems as well as in non-coherent access networks. This contribution reviews a number of specific signal processing algorithms that are useful in such systems. We focus on advanced equalisers for linear and nonlinear distortions. We examine the Tomlinson-Harashima precoding approach as well as advanced flexible non-integer fractionally-spaced butterfly equalisers. For nonlinearity compensation Volterra based equalization is investigated.

Keywords: digital signal processing, Tomlinson-Harashima precoding, non-integer fractionally-spaced equalization, nonlinear compensation

1. INTRODUCTION

A typical DSP algorithms scheme used in optical transmitters and receivers is illustrated in Fig. 1. In case of coherent detection, the DSP will usually include the compensation of almost all transmission impairments, including I/Q imbalance, linear and nonlinear fibre impairments, any kind of filtering effects leading to inter-symbol interference (ISI), carrier frequency offset (FO) / phase noise, and timing offset / jitter. Note that the order of the compensation stages given in Fig. 1 may vary according to the type of the algorithms used. In case of non-coherent detection, the information of the phase of the received electrical field is lost due to the squaring operation of the photo-diode. On one hand, phase noise and carrier offset are no longer influencing the received signal, and therefore do not require any compensation stage in the DSP scheme. On the other hand, since the phase information is in-fact lost, any type of compensation algorithm, which requires this information, will achieve suboptimal performance, namely the equalisers of interest in this contribution. Moreover, nonlinear compensation algorithms such as digital back-propagation (DBP) will completely fail without the signal's phase information, and therefore cannot be used in the DSP scheme. In this contribution, we concentrate on the compensation of ISI and nonlinear impairments caused by the Kerr effect [1]. Assuming first a linear system, the equalization of ISI is usually done in two stages: a static frequency-domain dispersion compensation for the elimination of most of the chromatic dispersion (CD) [2], and a second stage equaliser for the compensation of residual dispersion and any other sources of ISI such as filtering effects and bandwidth limitations of devices. The second stage equaliser can be generalized to a 2×2 MIMO structure (known in the literature as a butterfly structure [3]) to compensate for PMD and to separate between the *x* and *y* polarisations in case of a dual polarisation transmission. Each one of the equalization stages can be generalized to a compensator of nonlinear impairments, at the expense of an increased computational complexity: The first stage equaliser can be designed as a DBP modelling an inverse nonlinear fibre using either the spilt-step Fourier method or a frequency-domain Volterra model [4, 5]. The second stage equaliser can be generalized as a time-domain Volterra equaliser [6].
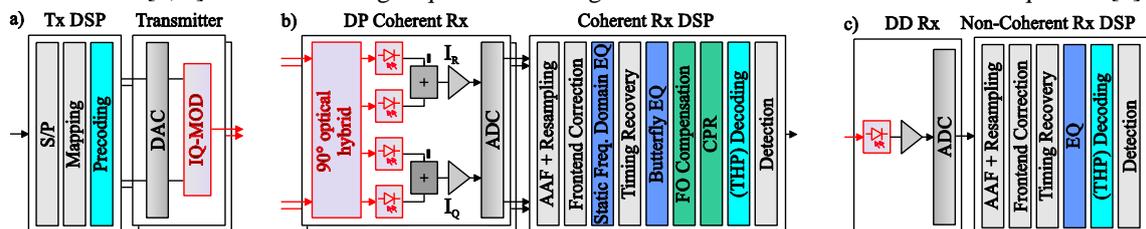

Figure 1. A general DSP scheme given as block diagrams of a) the transmitter (Tx) DSP including a simplified optical transmitter, b) a simplified dual polarisation (DP) coherent receiver (Rx) with its general DSP and c) a non-coherent receiver including direct detection (DD) and its Rx DSP. AAF stands for anti-aliasing filter.

2. COMBINED TOMLINSON-HARASHIMA AND VOLTERRA EQUALIZATION

The second stage equalization can be done either with a plain feed-forward equaliser (FFE) or a combined FFE and a decision-feedback equaliser (DFE). In addition, in order to avoid its error propagation effect, and to allow a straight-forward application with channel-coding schemes, the feedback structure of the DFE can be replaced by a precoder at the transmitter side [7]: a structure referred to as *Tomlinson-Harashima precoding* (THP) [8, 9] (see Fig. 2). Restricted to joint equalization of CD and filtering-induced ISI, THP was shown to outperform FFE and to offer similar performance as of DFE. In addition, THP was shown to allow compensation without error

propagation, which for DFE was potentially catastrophic in the vicinity of the FEC limit [10, 11]. As mentioned before, all structures can be redesigned in order to be applied for compensation of Kerr nonlinearities using a Volterra nonlinear equaliser (VNLE) [12, 13], which can replace the FFE structure, or can be combined with the DFE/THP feedback structure.
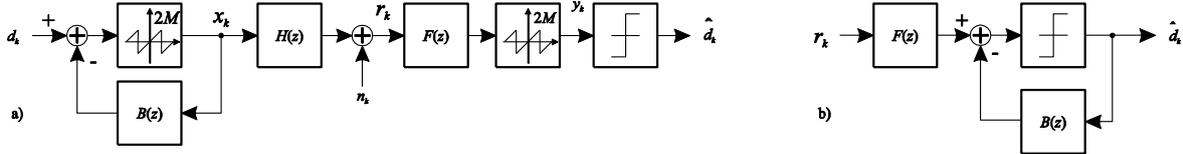

Figure 2. Block diagrams of a) a linear system with THP and an FFE and b) a combined FFE-DFE. The variables B, F and H are the feedback filter, forward filter and channel transfer functions, respectively.

2.1 Visualization of Equalised Signals for Nonlinear Transmission

In the following scenario, a 25 GBd 16-QAM signal was transmitted through a single span of 100 km of standard single-mode fibre (SSMF) at a launch power of 8 dBm. The first stage equaliser was set to compensate for 90% of the accumulated dispersion. The task of the second stage equaliser was to equalise for residual dispersion (10%), filtering effects due to receive filters and bandwidth limitation of devices such as the DAC, and of course the Kerr nonlinearities. We compared four different equalisers, denoted as FFE(3) and FFE(4), an FFE(3)-DFE(1) and a THP(1)-FFE(3), where the respective number of linear coefficients is given in brackets. Volterra equalization was applied in this work for the FFE part only. The resulting total number of equaliser coefficients (linear and nonlinear) for the above mentioned four equalisers sums up to 21, 44, 22 and 22, respectively. The equalised constellations are shown in Fig. 3. By examining the results for compensation of ISI only, we can see clearly the nonlinear behaviour of the system, as nonlinear phase distortions linger after the (partial) removal of the ISI. Notice that the FFE-DFE structure shows a tighter cluster composition than a plain FFE with the same number of coefficients. Nevertheless, it is possible to see symbols, which do not belong to any of the 16 clusters. These errors can be attributed to the DFE's error propagation effect. The positive effect of the VNLE on the received constellation is visible, and especially for the last two equaliser designs, the performance is greatly improved. In the results for the design with THP, we examine the constellation after modulo decoding. In comparison with DFE, the error propagation effect is no longer visible, as expected (see four inner constellation points). However, the constellation *before* modulo decoding exhibits larger values than the original constellation [7], which are are more susceptible to the SPM effect, since they are of greater instantaneous power. Without an additional phase correction for the outer ring, the modulo operator will falsely decode these values, as can be seen clearly in the upper left subfigure of Fig. 3. Combining now THP with VNLE, the system performance considerably improves as for the other designs. However, the fact that the signal after THP-VNLE features higher power values causes a slight degraded performance compared with VNLE-DFE. Notice that the performance of the combined feedforward and feedback structure shows the same performance as the FFE(4) realization, although having only half of the number of coefficients
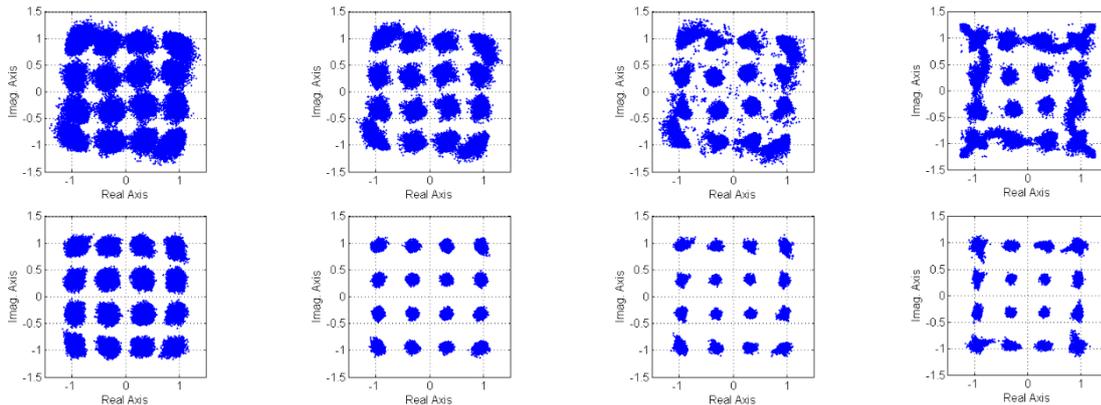

Figure 3. Equalised constellation diagrams. In the following the number next to the equaliser denotes the number of linear filter coefficients. From left to right: FFE(3), FFE(4), FFE(3)-DFE(1), THP(1)-FFE(3).Upper row: linear equalisers, lower row Volterra-based nonlinear equalisers.

3. NON-INTEGER FRACTIONALLY-SPACED EQUALIZATION

Since the performance of a symbol-spaced equaliser depends heavily on the sampling delay [3], a timing-recovery (TR) module is needed for best performance [14, 15]. To avoid this requirement, it is possible to use a *fractionally-spaced equaliser* (FSE), i.e. an FFE, with delay taps at a *fraction* of the symbol period, at the expense of a higher sampling rate than the symbol rate. In addition, in order to allow a higher flexibility of the transceivers, which should be capable of handling a range of baud rates, a *non-integer* FSE can be used, i.e. an

equaliser processing a non-integer number of samples-per-symbol (SPS), and which possesses all of the advantages of an integer FSE. Note that by using a proper anti-aliasing filter (AAF) preceding the ADC, the number of SPS can be reduced below two, hence relaxing the ADC speed requirements [3]. A block diagram of a non-integer FSE in a butterfly structure is given in Fig. 4a. For a detailed description of the equaliser the reader is referred to [16].
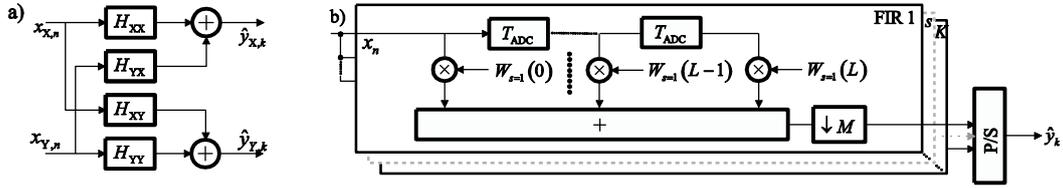

Figure 4. a) Block diagram of a butterfly equaliser. b) Block diagram of the non-integer fractionally spaced equaliser denoted as H_{io} in 4a.

3.1 System Setup

In our experimental setup shown in Fig. 5, a 10 GBd 4-QAM signal was transmitted through a re-circulating loop, where a single round-trip is 366.6 km long. After coherent detection, the signal was recorded with a real-time oscilloscope operating at 50 GSamples/s, and was then offline post-processed using Matlab.
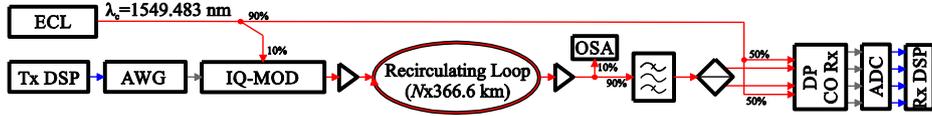

Figure 5. Experimental setup with a re-circulating loop of 366.648 km roundtrip distance.

3.2 Digital Signal Processing

The DSP scheme is similar to that shown in Fig. 1. To emulate a slower ADC the recorded signal with 50 GSamples/s was filtered with an AAF and then resampled using interpolation to sampling rates between 10 GSamples/s and 25 GSamples/s. This was done to investigate the impact of different ADC speeds and different numbers of SPS on the equaliser performance. In the proposed butterfly equaliser we compare different numbers of SPS, namely 1, 5/4, 3/2, 7/4, 2, 9/5 and 5/2. Since the symbol-spaced equaliser (SSE) with 1 SPS is heavily depending on timing errors, a timing-recovery algorithm had to be used before equalization to assure the best performance [17]. Since the number of equaliser coefficients has different impact on the performance for each equaliser, an optimized filter length not exceeding 15 coefficients was used in the experiment.

3.3 Experimental Results

The performance of flexible non-integer FSE for various SPS without TR was compared with SSE at the optimum sampling point for optical transmission systems reaching up to 5500 km at the optimal launch power of -2 dBm into each fibre segment. In order to have a fair comparison, the AAF bandwidth and the number of averaging filter taps for CPR were optimized for each OSNR. Using FSE, the generated 10 GBaud QPSK signal was successfully transmitted over 3666 km, before exceeding the hard decision (HD) FEC limit of $3.8 \cdot 10^{-3}$, as shown in Fig. 6a. The Q-factor performance is estimated from the BER for each measurement. The Q-factor gain of all FSE compared to SSE at the optimum sampling point with a DSP realization working at 1 SPS is depicted in Fig. 6b. In all realizations, the EDC and the butterfly equaliser operate at f_{ADC}, using a sufficient number of equaliser coefficients for CD and PMD compensation. All FSE realizations show significantly better
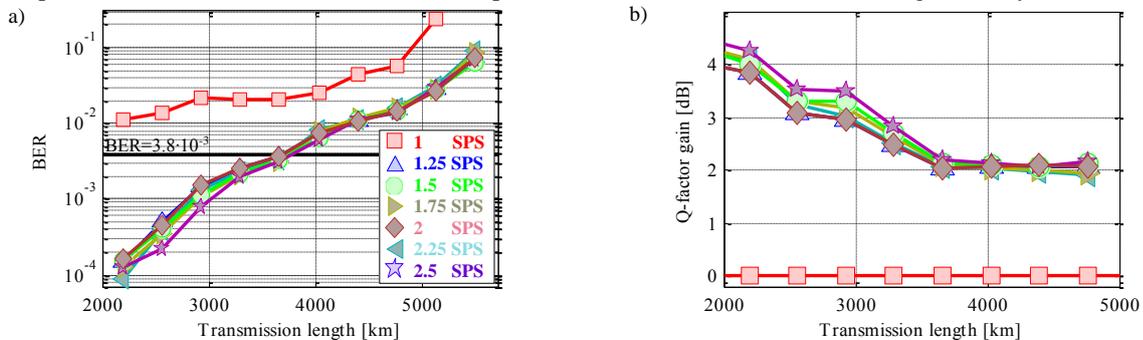

Figure 6. a) Measured BER vs. transmission length from 2000 up to 6000 km for various SPS. The HD-FEC limit of $3.8 \cdot 10^{-3}$ is marked as a reference. b) FSE gain in comparison to SSE with TR (Q-factor gain was calculated directly from BER).

performance than the SSE (entirely working at 1 SPS). A FSE operating with more than 1 SPS shows a Q-factor gain of 2 dB after 5132 km and up to 4 dB after 2199 km of transmission, even if the ADC sampling rate is limited below 2fS up to a minimum of 1.25 SPS.

4. CONCLUSIONS

In this contribution we have reviewed DSP schemes for coherent as well as non-coherent fibre-optic transmission, where emphasis of two compensation methods, namely a combined THP and Volterra equalization, and non-integer fractionally spaced equalisers. In the first method, the combined feed forward and feedback structure was shown to outperform the plain feed forward structure, or to offer a comparable performance, with a drastic reduction of the number of filter coefficients. Non-integer FSE was shown to outperform the symbol-spaced equaliser without the requirement of a TR circuitry, at the expense of a moderate increase in the ADC sampling rate. In addition, the element of transceiver flexibility was demonstrated as well, allowing full functionality of the proposed equalisers for a verity of ADC sampling rates and data rates ratios.

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