Data-aided single-carrier coherent receivers

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Data-Aided Single-Carrier Coherent Receivers

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Abstract Data-aided algorithms for coherent optic receivers are discussed as an extension of existing non-data aided methods. The concept presents a scalable approach with low implementation complexity and limited overhead for higher-order modulation formats.

Introduction

The first generation of resurgent 40G and 100G coherent systems was dominated by single-carrier polarization-multiplexed (PolMux) QPSK^{1,2}. The estimation of channel and synchronization parameters was typically done without the use of known sequences (non-data aided, NDA), solely relying on the statistical signal properties of the signal². Future 400G or 1T systems will require higher-order modulation, where blind estimation algorithms do not necessarily offer the best solution. For initial channel acquisition, the constant-modulus algorithm (CMA) can still be applied to QAM³. However, it is clearly suboptimal, with a decreased adaptation speed and higher bit-error rate (BER) before switching to decision-directed (DD) algorithms. Furthermore, blind demultiplexing can become increasingly complex in presence of polarization-dependent loss (PDL)⁴.

Data-aided (DA) algorithms offer fast acquisition, better estimation performance at a low BER, and scalability for next generation coherent systems⁵. In order to decrease the overhead defined by the training sequence length and its repetition rate, a semi-blind approach can be introduced as shown in Fig. 1. After initial convergence, with a data-aided acquisition of the frequency offset and the equalizer taps, the receiver continues to track in NDA mode e.g. using the least-mean square (LMS) algorithm⁶ in combination with a DD feedback carrier phase recovery. In the case of optically uncompensated links, chromatic dispersion (CD) is compensated using blind estimation⁷, in order to keep the resulting training sequence short.

Fig. 1: Data-aided receiver with blind tracking

MMSE vs. LMS

Equalizer coefficients can e.g. be computed using the minimum-mean square error solution (MMSE)⁶, which is optimal for linear equalizers, or with a stochastic gradient algorithm like the LMS. Fig. 2(a) compares the adaptation complexity of the MMSE and the LMS per

parallelization path. Here, the complexity of the LMS can be significantly reduced using the sign-update⁶, with only two complex multiplications required for the correction of the feedback carrier phase. Furthermore, it is assumed that the MMSE is implemented only once and not in every parallelization path, since the computation of the equalizer coefficients has to be processed for the training sequence only. Fig. 2(b) shows an adaptation speed comparison of the MMSE and LMS. Here, polyphase sequences⁸ were used for the MMSE to give optimum acquisition speed, while the LMS was adapted in a DD-mode with known symbol decisions in a serial implementation with no additional delay. While the MMSE achieves virtually instantaneous acquisition, the LMS has a clearly longer adaptation time, with the LMS solution approaching MMSE for a large number of training symbols. 112Gbit/s PolMux-16QAM and 13 tap FIR T/2-spaced filtering are exemplary assumed.

Fig. 2: (a) Complexity of MMSE, LMS in terms of real multiplications per parallelization path. Degree of parallelization $p=32, 128$. (b) Acquisition speed comparison.

Parallelized LMS adaptation

Despite the limited adaptation speed, the LMS has a clear advantage over the MMSE in terms of complexity. Since the acquisition time constraint in fiber optic transmission is less critical, the following analysis is based on the LMS. Fig. 3 shows a simplified parallelized design of a finite-impulse response (FIR) filter with feedback update. A binary PN-sequence is used as training sequence. Despite the non-ideal autocorrelation properties compared to polyphase sequences, the modulation and possible cross-correlation are significantly simplified. For each training sequence, the equalizer update is computed separately for each parallelization path and then summed up. It is assumed that the feedback delay is shorter than the spacing between two training sequences, so that the equalizer taps

Fig. 3: Parallelized equalization and update on the example of a 3-tap FIR filter and $p=4$.

are updated before the equalization of the next training sequence. Fig. 4 shows the convergence speed for 112Gbit/s PolMux-16QAM at an OSNR=19dB. A maximum value of $\mu=0.01$ for the LMS-update factor was chosen for the fastest stable convergence under the assumed degree of parallelization. The update speed is evaluated for a back-to-back configuration and for CD=4000ps/nm, where channel memory begins to exceed the 13 tap filter length. Two training sequence types are compared, consisting of a single binary PN-sequence with 63 bit and two concatenated PN-sequences with 31 bit length. The latter arrangement can be used to estimate the frequency offset, as described in the next section. In the worst case scenario, less than 10 iterations are required to achieve an acceptable BER in order to switch to blind DD-tracking mode.

Fig. 4: Convergence speed of 112Gbit/s PolMux-16QAM at OSNR=19dB with varying header composition and chromatic dispersion ($\mu = 0.01$)

Synchronization

MMSE equalization only requires prior frame synchronization with a rough frequency alignment, since a single header is sufficient to achieve an acceptable BER. DA-LMS is applied over several headers, requiring precise synchronization of the carrier phase. The frequency offset can be estimated using two subsequent training sequences, as in orthogonal frequency-division multiplexing (OFDM)⁹. The estimator is given by

$$\hat{\nu} = \frac{1}{LT_s} \arg \left\{ \sum_{k=1}^L r[k+L] \cdot r^*[k] \right\} \quad (1)$$

where $r[k]$ is the received symbol at instant k , T_s is the symbol duration, and L is the length of the single training sequence. While short training sequences increase the frequency acquisition range, the estimation error can be minimized using long sequences. In order to keep the overhead minimal, the frequency offset can be estimated in two steps. After an initial estimate, (1) can be applied to two subsequent headers. Here, the maximum spacing between two headers is limited by the frequency acquisition range that is defined by the estimation precision of the initial computation. For the given example, an overhead in the range of 3 – 4% can be realized.

Framing and timing synchronization algorithms depend on the amount of residual distortion in the channel, where a higher tolerance usually is desired. Framing can e.g. be performed using an OFDM method⁹, detecting the maximum of the estimator given by

$$M[k] = \left| \sum_{k=1}^L r[k+L] \cdot r^*[k] \right|^2 \quad (2)$$

During equalizer convergence, a certain timing frequency offset can be tolerated, so that the timing phase error can be corrected after the convergence of the equalizer. In presence of high CD, the equalizer taps can be used for precise timing phase recovery¹⁰. For the x-polarization, the timing phase error estimator is given by

$$\beta_x[k] = \beta_{xx}[k] + \beta_{xy}[k], \quad (3)$$

with

$$\beta_{xx}[k] = - \sum_{i=1}^{M-1} |w_{xx}^k[i]|^2 + \sum_{i=M+1}^{N_{taps}} |w_{xx}^k[i]|^2, \quad (4)$$

where w_{ij} are the filter coefficients and M is the central tap of the filter. Instead of the equalizer taps, the channel estimate can also be used, given by the cross-correlation of the transmitted and received training sequences.

Conclusion

A data-aided receiver concept for higher-order modulation formats was discussed as an extension of typical blind receivers based on CMA/LMS. Filter coefficient adaptation can be implemented using LMS with low complexity assuming a prior correction of synchronization parameters. Blind tracking of channel and synchronization parameters is required in order to keep the overhead acceptably low.

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