Experimental demonstration of error detection-driven nonlinearity compensation for optical fiber communication systems

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Abstract: We demonstrate how the error-detection capabilities of the inner code in an experimental rate-adaptive concatenated FEC scheme can be employed to reduce the computational complexity of the nonlinearity compensation block. Total complexity savings between 17% and 95% are reported depending on the target operating point.

1. Introduction

Advanced forward error correction (FEC) and nonlinearity compensation (NLC) are two major enablers of high spectral efficiency optical fiber communications. Concatenated FEC is currently state of the art for high coding gain solutions of practical complexity, e.g. in the ZR400 standard [1], which defines an inner soft decision (SD) Hamming code and an outer hard decision (HD) staircase code. The purpose of the SD code is to reduce the bit error rate (BER) to the decoding threshold of the outer code. On the other hand, many different flavors of NLC [2] have been proposed in the literature for compensation of (primarily) the intra-channel nonlinear interference due to the Kerr effect. The NLC method relying on modeling nonlinear propagation in the fiber using perturbation theory [3] is one of the lower complexity choices for future communication systems. In its standard form, a hard decision is made on the symbol sequence after linear equalization, which is then used to estimate the nonlinear interference. The method can be substantially improved by exploiting the FEC, which can provide better decisions. Repeating the FEC and NLC blocks iteratively is known as turbo equalization.

Previously [4], a variant of turbo equalization was proposed which exploits the error-detection capabilities of the inner code in order to select a small subset of codewords for which to perform NLC, substantially reducing the overall computational complexity. In this paper, we report the first experimental demonstration of the method proposed in [4] for improved performance at lower complexity w.r.t. conventional HD perturbation-based NLC using 16-quandrature amplitude modulation (QAM) and 64QAM for varying distances and data rates.

2. Methods and experimental data

The assumed coded modulation system is a conventional bit-interleaved coded modulation (BICM) with a concatenated FEC. The data are assumed to be encoded by an outer code, then interleaved, encoded by an inner code, then interleaved again, and mapped to symbols. The experimental data generated in [5] is processed and the well-accepted all-zero codeword method from [6] is used to evaluate the proposed FEC. A block diagram of the system is given in Fig. 1. The QAM symbols are first deinterleaved and decoded using the assumed FEC. The resulting bit sequences are treated as scrambling sequences for the all-zero codeword, which is then assumed to have been generated by the target FEC. These scrambling sequences are applied at the receiver after de-mapping to log-likelihood ratios, the FEC decoder is run, and the BER is estimated against the all-zero codeword.

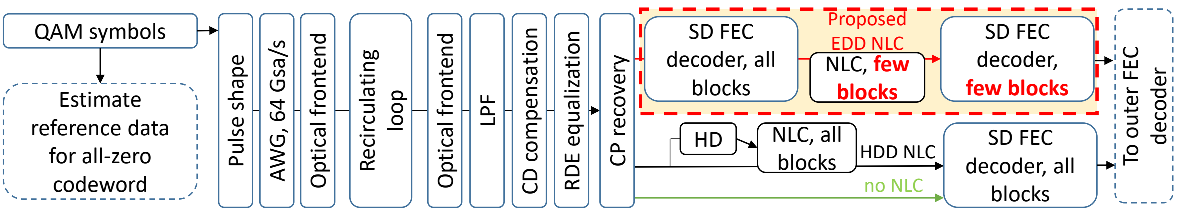

Fig. 1. Block diagram of the experimental setup and the studied NLC techniques.

The experiment is based on 16QAM and 64QAM symbol sequences (which we reasonably assume to be random due to the applied interleaver), 32 Gbaud, 5 channel wavelength division multiplexing system on a 50 GHz grid, with square root raised cosine pulse shaping with a roll-off of 0.005. The channel is realized using a recirculating loop of 2 spans of standard single mode fiber of 70 km each, followed by erbium doped fiber amplifiers (EDFAs) compensating span losses, as well as an EDFA compensating the loop switching losses. The data are captured for the central channel using a digital storage oscilloscope with 80 GSa/s. At the receiver, conventional coherent optical communication digital signal processing is applied, including chromatic dispersion compensation, 5% pilot-assisted radial decision-directed equalization (RDE), and carrier phase recovery using the digital phase locked loop algorithm [5].

The reference HD NLC algorithm performs (in order): 1) hard symbol decision; 2) nonlinear interference estimation based on perturbation-theoretic modeling of the fiber propagation; 3) NLC; 4) deinterleaving; 5) inner FEC decoding; 6) outer FEC decoding (which is exempt from this work). Other types of NLC are also directly supported by this method, e.g. digital back-propagation [2] or neural network-types [7].

The proposed error-detection driven (EDD) algorithm performs (in order): 1) deinterleaving; 2) inner FEC decoding, including error detection; 3) interleaving and mapping to symbols; 4) interference estimation and NLC *only for the codewords which are not detected to be error free*; 5) deinterleaving; 6) inner FEC decoding *only for the codewords which are not detected to be error free*; 7) outer FEC decoding.

The studied inner FEC is a polar code of length $N = 1024$ and rate K/N, which can achieve nearly continuous rate adaptation by tuning the number of information bits K [8]. In each data point, 5 independent traces were captured, each holding 80.000 QAM symbols, resulting in $> 1.4 \cdot 10^6$ information bits transmitted in the worst case (16QAM with $K = 900$). The net data rate (excluding the additional overhead of the outer code) is $\eta = K/N \cdot \log_2(M)$, where M is the size of the constellation. The polar code is decoded using successive cancellation list decoding with a list size of 32 and error-detection is based on a cyclic redundancy check of size 16 bits. Other types of error detection are also supported with this framework, e.g. using syndrome calculation for more conventional linear block codes (e.g. the ZR400-standardized Hamming [1] or for block low-density parity check (LDPC) codes).

3. Results

First, the performance using 64QAM is demonstrated in Fig. 2a) for 16-span transmission. For this specific distance, we select two options for the inner code: $K = 860$ assumed to be concatenated with a strong outer code (e.g. the staircase code of ZR400 with a rate of 0.93 with a decoding threshold of $1.25 \cdot 10^{-2}$) and $K = 815$ which is suitable for concatenation with a weaker outer code (e.g. the KP4 code with a rate of 0.95 and decoding threshold of $2.26 \cdot 10^{-4}$ [9]). The proposed method allows to reach the respective threshold, while the HD NLC fails. The BLER before the NLC at the two thresholds is ≈ 0.324 and ≈ 0.004, which correspond to the fraction of blocks which require NLC and the direct computational savings in the NLC block.

Then, the performance using 16QAM is demonstrated in Fig. 2b) for varying distances at the optimal launch power. In this case, we exemplify the cases of $K = 900, 940$ and 980, leading to $\eta = 3.51, 3.67$ and 3.82 bits/QAM symbol, respectively. The NLC gain is between 2 and 4 spans, slightly larger in the case of the proposed EDD NLC. This corresponds to ≈ 0.5 dB of gain in Q-factor. The block error rate (BLER) before the NLC is given in Fig. 2c). As discussed in [4], for $N = 1024$, the polar code decoder complexity per codeword is $\xi_{FEC} < 2 \cdot 10^6$ of real-valued additions, selections and multiplexing operations (which we treat equally). The NLC complexity per QAM symbol for the selected memory of the NLC is $\xi_{NLC}^{SYM} > 1.59 \cdot 10^6$, and per codeword is $\xi_{NLC} = \xi_{NLC}^{SYM}/N \cdot \log_2 M$ real-valued additions. The total complexity of the proposed method is thus $\xi_{TOT} = BLER \cdot \xi_{NLC} + (1 + BLER) \cdot \xi_{FEC}$, and is plotted in Fig. 2d) for the 16QAM case together with the complexity of the HD NLC, which is $\xi_{FEC} + \xi_{NLC}$. Net complexity gains are observed for almost the entire range, except the long distances and high rates, where the BLER is very high. This region is, however, not interesting for practical outer FECs.

As examples, the staircase code threshold is achieved for up to (approximately, after interpolation of the BER curves) 66, 58 and 48 spans for the three rates, respectively, which correspond to a BLER of $\approx 0.6, 0.6$ and 0.7, respectively, and total complexity savings of $\approx 35\%, 28\%$ and 17%, respectively. On the other hand, the KP4 threshold is achieved for up to (approximately, after interpolation and extrapolation of the BER curves) 55, 45 and 37 spans, where the BLER is $\approx 0.03, 0.03$ and 0.05 and the complexity savings are $\approx 95\%, 94\%$ and 93%.

4. Conclusion

An error detection-driven method for reducing the complexity of nonlinearity compensation was experimentally demonstrated for coded modulation systems with concatenated FEC architecture. The method results in computational complexity savings between 17% and 95%, depending on the operating point. The method is most effective at low BLER, which corresponds to a simple outer FEC.

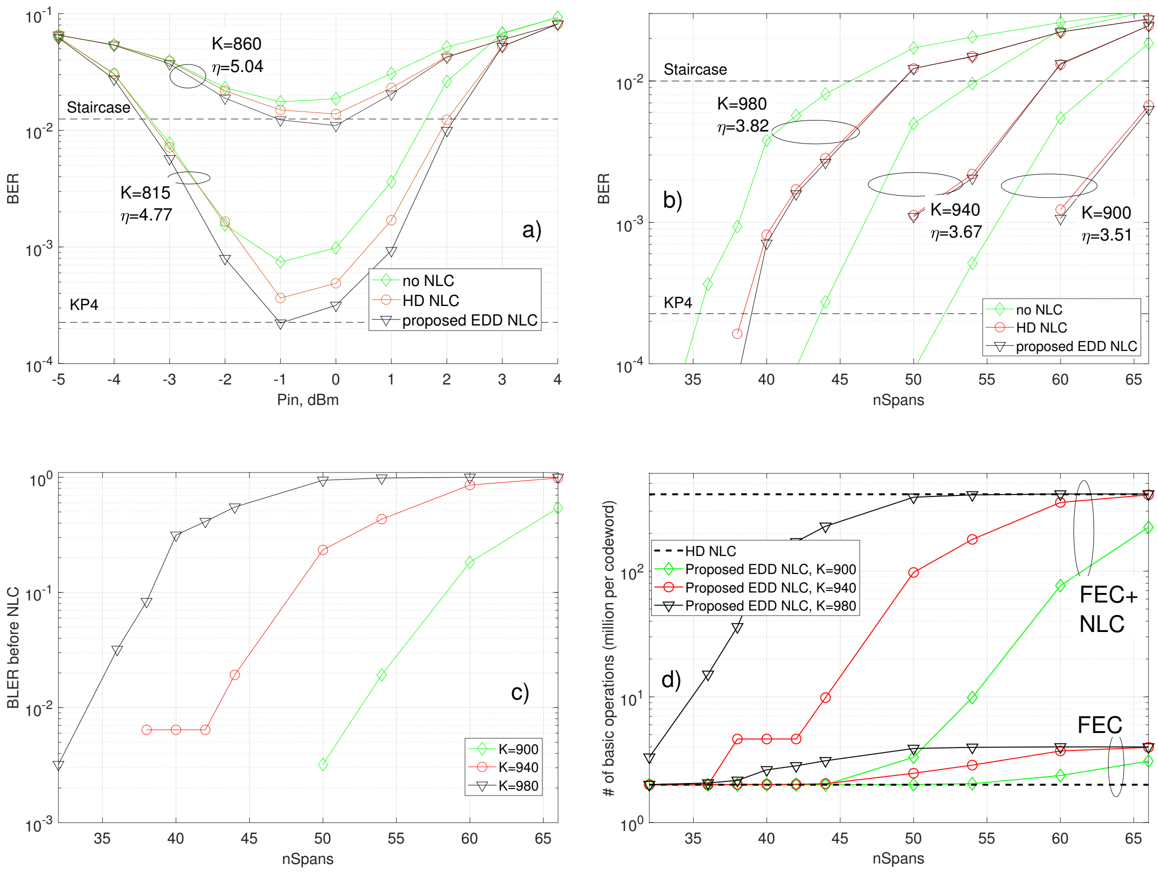

Fig. 2. Results of the experiment. **a):** BER for 64QAM at 16 spans of transmission for two different target rates. The corresponding target outer code thresholds are also given; **b):** BER for 16QAM at the respective optimal launch power of each studied method as a function of the transmission distance for three different target rates; **c):** BLER for the three studied rates, corresponding directly to the reduction of the computational complexity of the NLC and at the same time, inner FEC decoding computational complexity increase; **d):** total complexity in millions of basic operations per codeword for the proposed system and the reference HD NLC.

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