GHOST PULSE SUPPRESSION IN QUASI-LINEAR OPTICAL DATA TRANSMISSION SYSTEMS VIA CONSTRAINED CODING

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ABSTRACT

Nonlinear effects play an essential role in modern high-speed dispersion-managed optical links. In this paper, a popular class of systems limited by the so-called "ghost pulses" resulting from intrachannel four-wave mixing (IFWM) is considered. In previous work, a potential data rate improvement of about 50% was demonstrated through the application of constrained coding. Here, we show that the choice of the coding scheme is important. Since IFWM is highly pattern-dependent, codes with certain statistical properties reduce the ghost pulse formation and yield a lower probability of error. Our qualitative conclusions are confirmed by numerical simulations.

1. INTRODUCTION

Fiber nonlinearity is known to be a major performance limiting factor in long-haul fiber-optic communication systems operating at bit rates of 40 Gb/s per channel or higher. An interesting approach to mitigating this channel impairment is a system designed in the so-called quasi-linear regime [1]. Sometimes this regime of pulse propagation is also referred to as "highly dispersed pulse transmission" [2, 3]. More specifically, after being launched into the fiber, optical pulses quickly disperse and spread over tens, hundreds or even thousands of bit slots. Then, with the help of dispersion compensation, individual pulses can be compressed back to their initial shape, and the original waveform can be reproduced with only slight degradation. The key idea is to make the intensity pattern change so rapidly that the effect of fiber nonlinearity would average out [2].

It is widely recognized that quasi-linear systems, even multichannel ones, are mostly limited by intrachannel nonlinear effects. In particular, intrachannel four-wave mixing (IFWM) results in energy transfer between marks and creation of ghost (or shadow) pulses in spaces. Assuming on-off keying signalling format, return-to-zero pulses used for transmission of '1's also experience timing jitter caused by intrachannel cross-phase modulation (IXPM). While IXPM is usually reduced by clever dispersion management [3, 4], suppression of IFWM has been a more controversial research topic. A number of techniques have been proposed including phase coding [5, 6], unequally spaced pulses [7], alternate polarizations [8], optical phase conjugation [9] and constrained coding [10–12].

The last approach is the focus of this paper, which may be considered an extension of [12]. The main idea is based on the following observation. The intensity of IFWM-induced perturbations depends on various system parameters. Most importantly, it is proportional to $E_0^3\tau_0/T_B^4$, where E_0 is the pulse energy, τ_0 is the pulse width and T_B is the bit slot duration [2]. This means that shorter pulses are less affected by IFWM, and a low duty cycle is preferable. Certainly, such systems may have a lot of "white space" between '1's and are less attractive from the spectral efficiency point of view.

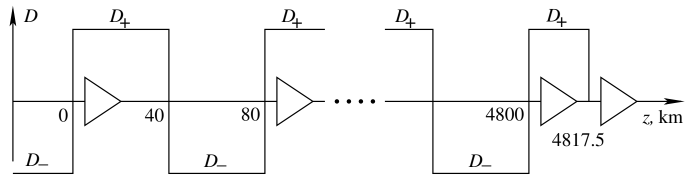

Fig. 1. Dispersion profile of the link.

In [12] we showed how the application of constrained coding can turn this seemingly redundant white space into extra information. More specifically, when transmitted sequences obeyed the so-called (2, ∞) constraint, the gain in data rate of up to 50% was demonstrated.

The coding scheme used [13] was originally designed to combat bit shifts, an important error-generating mechanism in optical systems employing dispersion-managed solitons for data transmission [14]. However, bit shifts are usually of much less importance in quasi-linear systems. In this paper, we propose an alternative (2, ∞) constrained code that turns out to be superior to the one in [13] from the point of view of ghost pulse suppression. We present experimental results as well as their intuitive explanation.

2. BENCHMARK SYSTEM DESIGN

In long-haul repeaterless systems optical amplifiers are inserted periodically to compensate for fiber losses. The amplification process, however, is inevitably accompanied by the generation of amplified spontaneous emission (ASE) noise that accumulates over the entire link and degrades optical signal-to-noise ratio (OSNR). Therefore, certain minimum pulse energy E_0 is required to provide a satisfactory OSNR at the receiver. On the other hand, the intensity of IFWM-induced perturbations is proportional to E_0^3 if the pulse width is fixed. The key design tradeoff is to select the value of E_0 in such a way that the contributions from ASE and IFWM are somehow balanced.

Having this in mind, a 40-Gb/s "benchmark" system is designed as in [12] ($T_B = 25$ ps). The dispersion map consists of a 40-km-long segment of a standard fiber with $D_+ = 17$ ps/(nm·km) followed by a reverse dispersion fiber of the same length and exactly opposite dispersion $D_- = -17$ ps/(nm·km). To reduce IXPM-induced timing jitter, we use a precompensating fiber of length 17.5 km and dispersion D_- along with a postcompensating fiber of the same length and dispersion D_+. The average dispersion of the link is thus zero. All fibers have attenuation 0.25 dB/km and a nonlinear parameter 2.5 W^{-1}/km. Optical amplifiers are placed every 80 km at the beginning of the anomalous dispersion span, their spontaneous-

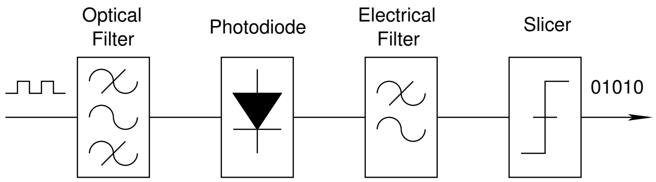

Fig. 2. Receiver structure.

emission factor being 1.4. The transmission distance is 4835 km including the compensating sections. The dispersion profile of the link is sketched in Fig. 1, where the amplifiers are represented by triangles.

Gaussian pulses of width $\tau_0 = 3$ ps, peak power P_0 and energy $E_0 = \sqrt{\pi} P_0 \tau_0$ are transmitted. Their envelope $A(t)$ may be written as

$$A(t) = \sqrt{P_0} \exp\left(-\frac{t^2}{2\tau_0^2}\right).$$

The standard deviation of the timing jitter is found to always be below 1.2 ps, which implies its negligible contribution to the probability of error p_e. The receiver has a simple structure and consists of an ideal 280-GHz optical bandpass filter, a photodiode, an integrate-and-dump electrical filter and a slicer (Fig. 2). The photodiode is modelled as a square-law detector with quantum efficiency $\eta = 0.9$. If ASE-induced amplitude jitter dominates, p_e can be computed as in [15]. To make a stringent test of our coding scheme, p_e is allowed to be as large as 10^{-2}. We assume that an error-correcting code (ECC) may be applied afterwards to achieve the target bit error rate.

Numerical simulations were carried out for 8192-bit de Bruijn pseudorandom input sequences using the split-step Fourier method with flexible step size [16]. Such a large number of bits is essential for an accurate representation of intrachannel nonlinear effects [17]. The channel capacity was estimated based on the probability of error observed. We varied P_0 in the range of 4–12 mW to find the best operating point. Equivalently, E_0 ranged from 21.3 to 63.8 fJ. When $P_0 < 4$ mW, the influence of IFWM is negligible and p_e was in a good agreement with [15]. For example, $p_e = 0.018$ when $P_0 = 4$ mW. However, the optimum P_0 was found to be around 6–8 mW (Fig. 3). The minimum p_e was about 0.007, which corresponds to the channel capacity of approximately 37.5 Gb/s. In this region of operation ghost pulses make a significant contribution to the probability of error that could be 0.002 or less in a purely linear regime.

We also simulated a benchmark system with additional phase coding [5] via introducing a phase shift of π for every other pulse. This type of coding suppresses the strongest ghosts by virtue of destructive interference and often improves the system performance. Our simulations confirmed that the channel capacity is somewhat higher, indeed, although IFWM eventually overwhelms the system (the upper curve in Fig. 3).

3. CODING SCHEME

A binary channel is said to be runlength-limited (RLL), or (d, k)-constrained, if channel sequences are required to have at least d but at most k '0's between any pair of adjacent '1's. Information transmission through an RLL channel implies two necessary steps. First, an arbitrary source word must be mapped unambiguously to a sequence that satisfies the channel constraints. Second, some error control must be implemented to protect the data from noise. A

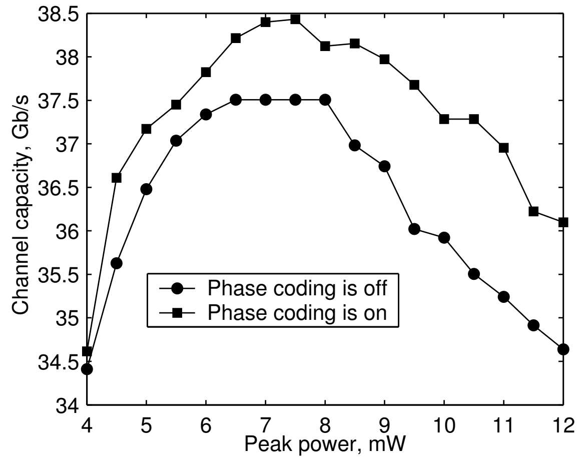

Fig. 3. Benchmark system capacity with and without phase coding.

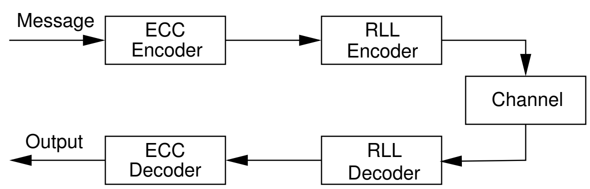

Fig. 4. Traditional ECC–RLL concatenation.

popular approach is to utilize a concatenation of an inner modulation RLL (or constrained) code with an outer ECC. A message to be transmitted is first encoded with the ECC. The result is supplied to the RLL encoder (or modulator). The modulator transforms an incoming unconstrained word to a channel sequence that satisfies the required constraints. At the receiver the decoders work in the reverse order. The received word is first demodulated by the RLL decoder (or demodulator) and the ECC decoder cleans up possible errors afterwards. (Fig. 4).

A rate 1/2 modulator used in this paper can be fully described by a finite-state machine presented in Fig. 5. At each time instance the encoder accepts one input bit and outputs two channel bits, the result depending not only on the input but also on the internal state of the machine. We assume that the encoder always starts and finishes in state "A." This can be achieved by appending a dummy '0' to any input word $u = (u_1, u_2, \ldots)$. For example, $u =$ '1110' would result in a codeword $x =$ '00010010'.

A sliding-window demodulator (SWD) decodes noiseless channel sequences with the help of Table 1. Sometimes the decoder can make a decision based on just two input bits. Sometimes, however, it has to look two bits ahead so, in general, the window size is 4

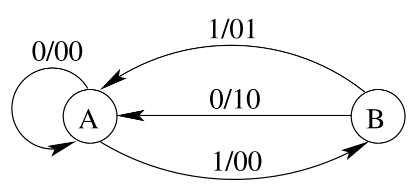

Fig. 5. Encoding machine.

Table 1. Decoding algorithm.

IN_1	IN_2	OUT
01	–	1
10	–	0
00	00	0
00	01	1
00	10	1

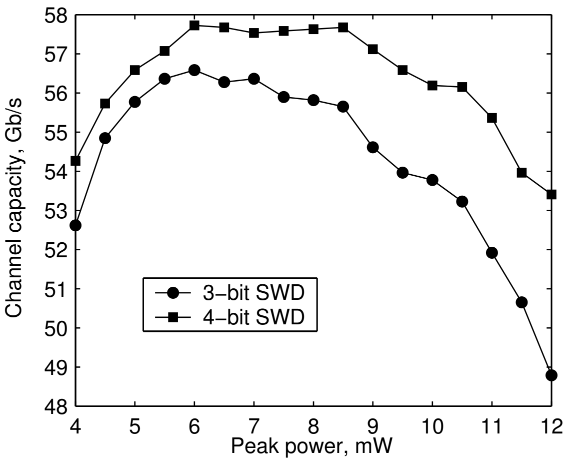

Fig. 6. Channel capacity for two coding schemes.

bits. In reality, of course, the channel output may not satisfy the (2, ∞) constraint, and the decoder has to deal with "illegal" contents of the sliding window. For simplicity, we output a '1' in this situation although better decoding strategies might be possible.

The output of the demodulator is an unconstrained binary sequence $v = (v_1, v_2, \ldots)$ with possible errors introduced by both the channel and the RLL decoder itself. Ignoring potential correlations between errors, we are interested in the crossover probabilities $a = p(v_i = 1|u_i = 0)$ and $b = p(v_i = 0|u_i = 1)$ that give us both the probability of error and a lower bound on the channel capacity. If a reasonable information-theoretic model of the physical channel can be constructed, a and b may be estimated quite accurately (e.g., [13, 14]). The combined effect of ASE and IFWM is not so easily translated into the language of information theory, however, and we obtain a and b directly from Monte Carlo simulations.

4. APPLICATION TO A REAL SYSTEM

In this section the new coding scheme will be applied to the benchmark system designed in Section 2 with $T_B = 25$ ps. Since channel sequences are (2, ∞)-constrained, T_B can potentially be reduced by a factor of three, while maintaining the same minimum pulse separation and, hence, approximately the same intensity of IFWM-induced perturbations.

If the channel were perfect, the application of the (2, ∞) constraint would give a 50% increase in data rate. To estimate the gain for the real channel, we simulated coded systems with $T_B = 8.35$ ps and varied P_0 in the same range as for the benchmark system. Simulations were carried out using 24000-bit pseudorandom constrained sequences. A lower bound on the channel capacity was calculated based on the experimental crossovers a and b obtained at the output of the RLL decoder. As in [12], we did observe a potential gain in

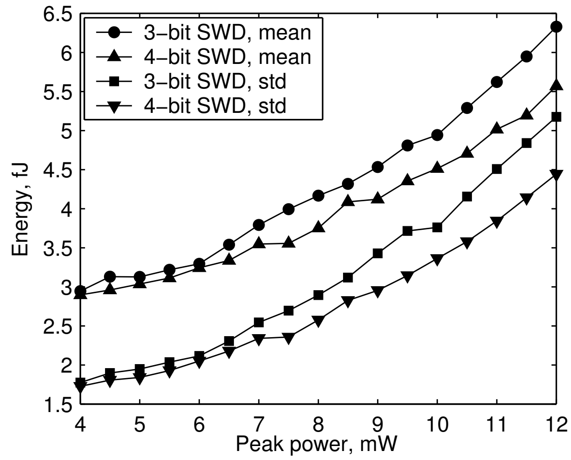

Fig. 7. Energy in spaces for two coding schemes.

data rate of about 50%.

The main purpose of the experiments, however, was to compare the coding scheme from Section 3 with the one in [12]. Since the latter uses a 3-bit sliding window decoder, it is referred to as a "3-bit SWD" in the figures. A comparison in terms of channel capacity (Fig. 6) favors the new scheme that clearly outperforms its rival throughout the entire range of P_0. It is more insightful to compare the energy accumulated in '0' bit slots at the receiver for both schemes (Fig. 7). Ideally, it must be close to 0. In a real system, both ASE and IFWM gradually pump energy into spaces so its distribution at the output of the link is far from trivial. Nevertheless, a comparison of means and variances shows consistent advantage of the new scheme from the point of view of ghost pulse suppression. At low P_0, most of the energy in '0's is due to ASE noise so the distributions are almost identical. As the contribution of IFWM grows up, the difference becomes more and more clear.

The results obtained may be explained by the fact that the two schemes produce constrained sequences with different distributions of '0's and '1's. The first important issue is the probability q_1 that a randomly selected constrained bit is a '1'. For the new scheme, $q_1 = 1/6$ (16.67%), while $q_1 = 5/28$ (17.86%) for the scheme in [12]. Mathematically, the lower the number of '1's in the transmitted bit stream, the fewer interacting doubles and triples they create. Physically, the less energy is pumped into the fiber at the input, the weaker the nonlinear interactions among the pulses are. To validate this conjecture, we designed a third (2, ∞)-constrained code with $q_1 = 5/24$ (20.83%). The performance of that scheme was clearly worse in comparison with either one considered here.

The second critical factor is that the introduction of constrained coding makes channel words more irregular. As observed in [7], irregularity helps in ghost pulse suppression. Moreover, certain configurations of bits that lead to the creation of strong ghost pulses are now less likely to occur or entirely forbidden. For example, a benchmark system with no phase coding is known to mostly suffer from sequences like '1110111' or '11011'. The reason is that they contain a lot of pairs and triples of pulses that pump energy into the central '0' via IFWM. Equivalent (2, ∞)-constrained representations of these "worst" words are '1001000001001' and '1001001000001001001'. However, the encoding machine (Fig. 5) never delivers such output!

Indeed, a run of two '0's ('1001') can only be produced by a state sequence "BAB," the leading '1' always occupying an even

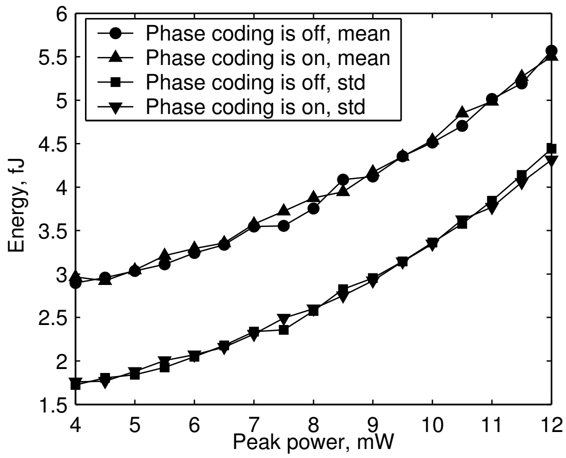

Fig. 8. Energy in spaces with and without phase coding.

position in the bit stream. After that, the encoder is always in state "*A*" and cannot immediately generate another '1'. Therefore, the block '1001001' is forbidden by the modulator, and the first worst configuration of bits is eliminated. The second bad sequence begins with a '1001' so the first '1' must be located at at even position. However, there is also another '1001' that would start at an odd position in that case. Again, since it is prohibited by the structure of the encoder, the corresponding worst sequence never occurs at the output of the modulator. It is worth mentioning that the coding scheme in [12] does not forbid either block although their probabilities of occurrence are rather low.

We also simulated constrained systems with phase coding and found no significant difference from the systems without phase coding (Fig. 8). This result is actually not surprising. The coding approach in [5] is aimed against the worst channel sequences and suppresses the strongest ghost pulses. Such sequences simply do not occur in our constrained system, i.e., the strongest ghosts have already been busted.

5. CONCLUSION

In this paper we proposed a new $(2, \infty)$-constrained code for high-speed fiber-optic communication systems operating in the quasi-linear regime and compared it with its predecessor [12]. The new coding scheme was found to further suppress ghost pulse formation induced by IFWM perturbations, while maintaining a 50% gain in data rate. Wider system margins may prove highly beneficial for multichannel systems, which always have to take into account impairments caused by interchannel crosstalk as well.

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