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Electronic post-compensation of WDM transmission impairments using coherent detection and digital signal processing

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Abstract: A universal post-compensation scheme for fiber impairments in wavelength-division multiplexing (WDM) systems is proposed based on coherent detection and digital signal processing (DSP). Transmission of 10 × 10 Gbit/s binary-phase-shift-keying (BPSK) signals at a channel spacing of 20 GHz over 800 km dispersion shifted fiber (DSF) has been demonstrated numerically.

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1. Introduction
The degree to which fiber impairments are compensated determines the transmission capacity of fiber optic transmission systems. Dispersion compensating fiber (DCF) is commonly used to compensate chromatic dispersion [1]. WDM systems suffer from both intra- and inter-channel nonlinearities such as cross-phase modulation (XPM) and four-wave mixing (FWM). These effects can be suppressed using dispersion management [2, 3]. Compensation of nonlinear impairments in fiber has become the next logical step in increasing capacity of WDM systems. A few optical nonlinearity compensation schemes have been demonstrated such as lumped compensation of self-phase modulation [4] and optical phase conjugation for the compensation of both chromatic dispersion and Kerr nonlinearity in fibers [5].

Electrical dispersion compensation (EDC) and electrical nonlinearity compensation (ENLC) have received significant attention in recent years. Several electrical pre-compensation schemes have been demonstrated to compensate chromatic dispersion or nonlinearity in single-channel or WDM system [6-9]. These schemes pre-distort the transmitted signals using arbitrary waveform generators. Pre-distortion is calculated using optical phase conjugation or backward propagation, i.e., the signal is distorted by virtual fiber while compensated through the real fiber transmission. Post-compensation using coherent detection and DSP has been shown to be very effective in chromatic dispersion compensation [10] and intra-channel nonlinearity compensation [11]. Post-compensation offers great flexibility since adaptive compensation can be incorporated in this scheme. However, there have been no reports to date on post-compensation of inter-channel nonlinearities in WDM transmission.

In this paper, a universal post-compensation scheme is proposed to compensate all impairments in fiber using backward propagation. EDC and ENLC are realized simultaneously with coherent detection and DSP. This paper is organized as follows. Section 2 describes the proposed post-compensation scheme based on backward propagation for fiber dispersion and nonlinearity compensation. Section 3 presents an approach for real-time implementation of backward propagation using coherent detection and DSP. Section 4 contains results of simulation for simultaneous compensation of fiber dispersion and nonlinearity in a WDM system.

2. Post-compensation based on backward propagation

We consider transmission of WDM signals (total number of channel is $C$) and post-compensation of fiber impairments. The schematic of the transmission system is shown in Fig. 1 where post-compensation is performed in the digital domain after coherent detection. The WDM signals are transmitted over multiple amplified fiber spans. After transmission, the received signals are mixed in a 90° optical hybrid with a set of local oscillators (LOs), of
which C LOs are aligned at the center of the WDM channels. Additional LOs on the both sides are aligned with other FWM components outside the WDM signal. The in-phase and quadrature components of each WDM channel are obtained by balanced photodetectors. Analog-to-digital (A/D) conversion is followed by DSP to achieve post-compensation and data recovery.

![Diagram of backward propagation for a multi-span fiber link. L: span number; N: step number per span.](image)

The backward propagation compensation scheme for a multi-span fiber link is shown in Fig. 2. $G$ is the optical gain of the linear optical amplifiers in the fiber link, $A(t)$ and $9(t)$ are the electric fields of the received signal and the compensated signal, respectively. The nonlinear Schrödinger equation (NLSE) governing the backward propagation in each span can be written in the form 

$$\frac{\partial A}{\partial t} = \left( \hat{N}^{-1} + \hat{D}^{-1} \right) A,$$

where $\hat{N}^{-1}$ and $\hat{D}^{-1}$ are inverse nonlinear and differential operators, respectively, and given by [12]:

$$\hat{N}^{-1} = i\gamma |A|^2, \quad \hat{D}^{-1} = \frac{i\beta_2}{2} \frac{\partial^2}{\partial T^2} + \frac{\beta_3}{6} \frac{\partial^3}{\partial T^3} - \frac{\alpha}{2}$$

where $\gamma$, $\beta_2$, $\beta_3$, and $\alpha$ are fiber nonlinear, first- and second-order chromatic dispersion, and loss coefficient. When these parameters are chosen to be exactly the negative of the values for transmission fiber, nonlinearity and dispersion could be compensated through backward propagation. The NLSE is most commonly solved using the split-step Fourier method. In this approach, each span of fiber is divided into $N$ sections and $\hat{N}^{-1}$ and $\hat{D}^{-1}$ are implemented iteratively. To facilitate real-time implementation, the dispersion or differential operator ($\hat{D}^{-1}$) can be realized using a finite impulse response (FIR) filter instead of Fourier transform, which has been shown to achieve acceptable accuracy [13].

3. DSP implementation for backward propagation

To implement backward propagation in the digital domain in real time, two technological constraints must be addressed. First, backward propagation requires the optical field at the end of the transmission fiber be sampled with sufficient temporal resolution. Second, the limit of DSP speed requires parallelization of the post-compensation scheme to cover the entire spectrum of interest.

In our approach, each WDM channel is translated to the baseband using coherent detection. The LOs for all WDM channels must be phase-locked for the post-compensation to function properly. Each complex field (at baseband) is sampled at two samples per symbol. In the method proposed here, the received signal corresponding to all WDM channels is considered as a whole for backward propagation, rather than on a channel by channel basis. The baseband signals are up-sampled in the digital domain to ensure sufficient temporal resolution for backward propagation. Details of up-sampling and parallel implementation are discussed using a 10 Gbaud WDM system with a channel spacing of 20 GHz as an example.
3.1 Signal up-sampling

Parallel composition of this signal from the obtained 20 Gsa/s baseband streams is achieved as follows. Designating the sampling rate of each A/D converter as \( F_s \), a set of \( N_c \) coherent receivers may cover a total bandwidth of \( N_c F_s \). Up-sampling to a bandwidth of \( M F_s \) (\( M \) is an integer and \( M > N_c \)) can be achieved in the digital domain using the following generalized formula:

\[
 r_n = t \left( \frac{t}{N_c F_s} = \sum_{i=0}^{N_c-1} s_{i,k(n)} \cdot e^{j 2 \pi \frac{i}{M}} \right) \tag{2}
\]

where \( s_{i,k} \) is the sample at time \( t = k/F_s \), of the \( i \)th frequency band and \( k(n) = \lfloor n/M \rfloor \cdot M / F_s \), where \( \lfloor x \rfloor \) stands for the nearest integer smaller than or equal to \( x \). The above formula effectively zero-pads the spectrum to obtain up-sampling. Note that each re-sampled data point is calculated independently, making this spectral stitching technique highly compatible with parallel implementation. If the DSP speed is 10 GHz, 32 streams of 10 Gsa/s are required to cover the 320 GHz total spectrum. Recall that sampling is at 20 GHz so two identical sets of resampling modules, operating independently, should be used for up-sampling. To align the outputs from the two up-sampling modules, the output from the first module may be delayed by 50 ps.

3.2 Parallel implementation for backward propagation

![Fig 3. Block diagram of parallel implementation for backward propagation using DSP technique. The hollow arrows represent multiple inputs or outputs and the solid arrows represent single input or output.](image)

Figure 3 illustrates the block diagram of parallel implementation for backward propagation, showing the feasibility of real-time operations. \( N_t \times N_c \) is the total step number in the backward propagation. We assume the processing rate of DSP is \( R_P \). In our simulation, \( R_p \) is assumed to be 10 GHz, same as the symbol rate. However, the processing rate can be reduced using time-division DEMUX technique, which requires more processing units but helps the real-time operation at a lower rate [14].

After up-sampling, \( N_b \) sampling points are generated in parallel and output simultaneously in \( N_b \) branches every period \( T \), where \( N_b = M F_s / R_P \) is the number of parallel processing branches, and \( T = 1/R_P \) is the clock cycle in each branch. One-symbol delay latches for some branches are used to obtain additional outputs for the parallel implementation of the following FIR filter. The number of additional outputs is \( N_r - 1 \), where \( N_r \) is the FIR filter length in tap number. \( A_{k,i} \) is the \( k \)th sampling point in the \( i \)th symbol and is processed in the \( k \)th branch. Sampled data from all the branches are then sent into a number of cascaded modules to realize backward propagation using the split-step FIR method [13]. Each module performs one step...
in the backward propagation, compensating loss, dispersion and nonlinearity of a small segment of fiber. The number of modules equals to the step number. Each step contains $N_b$ sub-units to perform backward propagation for each branch respectively. $M_k$ is the sub-unit in the $k$th branch of each step. $A_{k,n}(i)$ is the output of the $k$th branch in the $n$th step and $i$ is the symbol index.

Fig 4. Block diagram of the sub-unit $M_k$ in the $k$th branch in backward propagation. The inputs and outputs terminated by a circle may interface with other branches. The dashed lines only apply to some modules. $pT$ is the delay of FIR filter, and $qT$ is the delay of inverse nonlinear operator $1$.

The block diagram of the sub-unit is shown in Fig. 4. The coefficients of the FIR filter used for dispersion compensation are designated by $a_i$, where $i = 0\ldots N_i - 1$. We renamed the inputs and outputs from Fig. 3 to simplify the illustrations. The sub-unit is designed according to the symmetric split-step scheme with two iterations when solving the NLSE [12]

$$A(z + h, T) = \exp \left( \frac{h}{2} D^{-1} \right) \exp \left( \int_{z}^{z+h} \hat{N}^{-1}(z')dz' \right) \exp \left( \frac{h}{2} D^{-1} \right) A(z, t)$$

(3)

To improve the accuracy and thus increase the step size for the split-step FIR method, we used trapezoidal rule in our simulation to calculate the nonlinearity and approximate the integral by

$$\int_{z}^{z+h} \hat{N}^{-1}(z')dz' = \frac{h}{2} [\hat{N}^{-1}(z) + \hat{N}^{-1}(z + h)]$$

(4)

It is necessary to follow an iterative procedure that is initiated by replacing $\hat{N}^{-1}(z + h)$ by $\hat{N}^{-1}(z)$, then use (3) to estimate $A(z + h, T)$ which in turn is used to calculate the new value of $\hat{N}^{-1}(z + h)$. As shown in Fig. 4, three FIR filters are used for dispersion and loss compensation and two inverse nonlinear operators for nonlinearity compensation in a sub-unit. The inverse nonlinear operator 1 performs $\exp \left( h\hat{N}^{-1}(z)/2 \right)$ and the inverse nonlinear operator 2 performs $\exp \left( h\hat{N}^{-1}(z) + \hat{N}^{-1}(z + h)/2 \right)$. $Y_{k,i}$ is the output of the first stage and $Z_{k,i}$ is the output of the second stage. The FIR filter is implemented in a parallel configuration, which has multiple inputs instead of one input combined with a series of delay latches. Therefore, each unit operates at a speed of $R_P$ although the overall bandwidth is $N_b R_P$ [7].
Since the FIR filters need multiple inputs, all the branches have to interface with adjacent ones. This is represented by the ports terminated with a circle in Fig. 4. Similar to the Up-sampling module, each of the \( N_t - 1 \) sub-units (from the \( (N_b - N_t + 2) \)th to the \( (N_b) \)th) has two outputs to the next step, while the other sub-units only have one output. The additional outputs in the \( N_t - 1 \) sub-unit are required for the inputs of the FIR filters in the next step. Also, additional interfaces with adjacent branches are needed in these \( N_t - 1 \) sub-units. All the additional outputs and modules required in some sub-units are indicated by the dashed lines in Fig. 3. Since each module processes only signals that are already available, the scheme can be carried out in real time.

After backward propagation, the desired WDM channel signal is down-converted to the baseband then chosen by a low-pass FIR filter. After filtering, the signal in each channel is resampled to one sample per symbol and phase estimation is performed to recover the data.

The computation efficiency and latency of this DSP processing scheme should be investigated. Most operations are required by the backward propagation because of the recursive processes for long-haul transmission. Ignoring the computation and latency caused by up-sampling and DEMUX & phase estimation, the number of required multiply-accumulate (MAC) units is

\[ N_{MAC} = N_t \times N_t \times (12 \times N_t + 9) \]

The total number of MACs is calculated as follows. Each FIR filter requires \( 4 \times N_t \) multiplications and \( 4 \times N_t - 2 \) summations; inverse nonlinear operator 1 requires 7 multiplications and 3 summations; inverse nonlinear operator 2 requires 9 multiplications and 5 summations. The latency of the backward propagation scheme is

\[ T_L = \left\lceil \log_2 N_t \right\rceil + 21 \times \frac{T}{2} \]

where \( \left\lceil x \right\rceil \) stands for the nearest integer greater than or equal to \( x \). A look-up table is used for the calculation of \( e^{jx} \). The latency is calculated as follows. It is assumed that the look-up table requires one clock cycle \( T \) and each multiplication or summation requires half of clock cycle \( T/2 \). Each FIR filter requires \( \left\lceil \log_2 N_t \right\rceil + 2 \times (T/2) \); inverse nonlinear operator 1 and inverse nonlinear operator 2 require \( 7 \times (T/2) \) and \( 8 \times (T/2) \), respectively.

4. Simulation results

In our numerical simulations, the WDM system consists of 10 × 10 Gbit/s BPSK channels distributed around 1550 nm with a channel spacing of 20 GHz. The BPSK signals are generated by driving phase modulators with de-correlated 10 Gbit/s non-return-to-zero (NRZ) pseudorandom bit sequences (PRBS). The forward transmission of WDM signals over DSF is simulated using the VPI transmission-maker. The fiber loss, dispersion, dispersion slope and nonlinearity at 1550 nm are 0.2 dB/km, 0 ps/km/nm, 0.04 ps/km/nm² and 1.8 /W/km, respectively. The EDFAs are set to power mode with noise figure of 5 dB. Both transmitter lasers and LOs have linewidth of 2 MHz. The initial relative phases of the transmitter lasers are set to zero without loss of generality. The optical demultiplexers have a 3-dB bandwidth of 19 GHz and the photodetectors have an electrical bandwidth of 9 GHz.

Using coherent detection, each 20 GHz band is translated to baseband and sampled at 20 GSa/s thus the total sampling rate of the 10 WDM channels plus two extra FWM channels is 240 GSa/s. The inclusion of two extra FWM channels, adjacent to the WDM channels, is necessary for nonlinearity compensation. Then the signal is up-sampled to a total bandwidth of 320 GHz. The DSP speed is assumed to be 10 GHz. So \( F_s = 20 \) GHz, \( N_c = 12, M = 16, N_b = 32 \) and \( T = 100 \) ps.

Using the symmetric split-step scheme for backward propagation, the step number in each DSF span is reduced to 50, corresponding to 2 km of DSF. The FIR filter used for dispersion compensation is 33 taps long. The filter coefficients are extracted by an inverse Fourier
transform of the DSF frequency response. The low-pass FIR filter used after backward propagation for demultiplexing has a 3-dB bandwidth of 10 GHz.

![Eye diagrams](image1)

![Eye diagrams](image2)

Fig. 5. Eye diagrams of the 5th WDM channel: a) at back-to-back, b) after 500 km transmission over DSF without ENLC, c) after 500 km transmission over DSF with ENLC, d) after 800 km transmission over DSF with ENLC.

The back-to-back electrical eye-diagram of the 5th WDM channel is shown in Fig.5 a). The rails on eyes are from linear crosstalk due to small channel spacing. Fig.5 b) and c) show the eye diagrams of the received signals after 500 km transmission over DSF without and with ENLC, respectively. The $Q$-factor of the eye diagram in Fig. 5 b) is about 6. It is clearly seen that the eye diagram in Fig.5 c) are more open because of nonlinearity compensation. Fig.5 d) shows the eye diagram ($Q = 6$) after 800 km transmission over DSF with ENLC. Fig.6 shows the dependence of the $Q$-factor of the 5th WDM channel on the average launching power. After transmission of 500 km DSF the optimum launching power is increased from -11 dBm to -9 dBm due to nonlinearity compensation. When the launching power is more than -9 dBm, the performance of ENLC is limited by the nondeterministic effect, which is introduced by the amplified spontaneous emitted (ASE) noise from the optical amplifiers along the fiber chain. After transmission of 800 km and with ENLC, the $Q$-factor is similar to that after 500 km transmission but without ENLC. This implies that the transmission distance with digital nonlinearity compensation is increased by 60%. It is noted that the degradation of the $Q$-factor in long-haul transmission is also from the numerical errors in the backward propagation calculations. The system performance could be further improved by optimizing the FIR filter used for dispersion compensation.
5. Discussion and conclusions

In conclusion, a method for universal digital post-compensation for all fiber impairments in WDM transmission is proposed. Post-compensation is accomplished through backward propagation in the digital domain. The split-step FIR method and a parallel architecture have been designed to facilitate real-time implementation. The numerical simulations indicate that 2 samples per symbol hardware sampling and up-sampling in the digital domain are sufficient to achieve significant nonlinearity compensation. Simulation of WDM transmission in DSF of 10 channel BPSK signals reveals a 60% increase in transmission distance using nonlinearity compensation. Similar results for standard single-mode fiber transmission were achieved with FIR filter length 3 times larger than that used for DSF transmission.

The number of required computations is of utmost importance in the eventual realization of the post-compensation method. Taking the parameters used in our simulations for a 10 channel WDM system with a channel spacing of 20 GHz, for 8 transmission spans of 100 km DSF, the required computation and latency are $5.184 \times 10^6$ MACs and 0.78 µs, respectively. The average number of MACs required for each channel is $5.184 \times 10^5$. It should be noted that this overall computation and latency is realized using fixed step size. It is expected that employing variable step size [15] would reduce the number of operations by a factor of 4-5, bringing the average required MACs and processing speed per channel to within about one order of magnitude of the state of art field-programmable gate arrays (FPGAs). Block processing of the incoming samples coupled with fast Fourier transform implementation allows dispersion compensation in the frequency domain [16]. This may lead to further decrease in the number of operations and better compensation at longer transmission distances. Furthermore, when high-order modulation formats such as quadrature amplitude modulation (QAM) with multi-bit/symbol spectral efficiency are used, the required MACs per symbol should remain about the same and thus the required MACs per bit should decrease proportionally with spectral efficiency. The channel spacing can be further reduced to be equal to the symbol rate using the orthogonal WDM approach that we have demonstrated recently [17]. The combination of high spectral efficiency and narrow channel spacing can potentially reduce the number of operations per bit by another order of magnitude.

In the simulation, we have assumed that the phases of the LOs in coherent detection are locked. While this requirement is necessary for pre-compensation schemes, it is not absolutely necessary for post-compensation. The relative phase drift between the LOs can be monitored and factored into post-compensation using heterodyning between the LOs. Parallel polarization of the WDM channels along the entire link has been assumed in the simulations.
for simplicity. Due to its random nature, polarization mode dispersion (PMD) will be another limiting factor on the effectiveness of ENLC; however, this would only occur in high-symbol rate (e.g. 40 G symbol/s) systems.