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# Monitoring inter-channel nonlinearity based on differential pilot

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ABSTRACT

We modify and simplify the inter-channel nonlinearity (NL) estimation method by using differential pilot. Compared to previous works, the inter-channel NL estimation method we propose has much lower complexity and does not need modification of the transmitter. The performance of inter-channel NL monitoring with different launch power is tested. For both QPSK and 16QAM systems with 9 channels, the estimation error of inter-channel NL is lower than 1 dB when the total launch power is bigger than 12 dBm after 1000 km optical transmission. At last, we compare our inter-channel NL estimation method with other methods.

## 1. Introduction

The nonlinearity (NL) of optical fiber becomes the main obstacle to overcome for the long-haul transmission of the next generation [1]. Through digital backward propagation or perturbation method, the intra-channel NL can be efficiently compensated [2,3]. It becomes more important to analyze and monitor the effect of inter-channel NL during dense wavelength-division multiplexing transmission. When monitoring NL in optical fiber transmission, it is necessary to differentiate between amplified spontaneous emission (ASE) noise and nonlinear noise [4]. There are mainly two ways to separate NL from ASE. The first way is utilizing the statistical difference between ASE noise and nonlinear noise. This method lacks mathematical deductions and it is only useful in some specific links [5,6]. Recently the method of artificial neural networks (ANN) is also used to monitor NL [7]. The ANN method in [7] is also based on statistical models of NL in [5]. After training process with various systems, the ANN method also achieves high accuracy. After extensive verifications, the ANN method could be a promising way for parameter estimation. The other way to calculate the power of NL is by the distribution difference of ASE and NL for special pilot [4,8–10].

Recently a cross phase modulation (XPM) power estimation method is proposed by differential pilot (DP) [9]. The method in [9] assumes that the XPM noise is Gaussian distributed after carrier phase recovery (CPR). Then the power of XPM noise can be calculated after removing the angular direction noise. The XPM power estimated by [9] underestimated the XPM noise because CPR can remove part of the phase change induced by XPM effect. To judge the performance of XPM power estimation, the reference power of XPM is obtained by blind phase search (BPS) algorithm in [9]. Although the BPS algorithm can compensate some NL phase change, it cannot compensate all the phase noise caused by NL [11,12]. Therefore the reference power of XPM is also underestimated in [9].

In order to estimate inter-channel NL accurately, we proposed a novel inter-channel NL estimation method by fractional Fourier transformation (FrFT) of linear-frequency modulation (LFM) signal in [10]. This method utilizes the NL distribution difference between frequency domain and fractional domain to monitor the inter-channel NL. Compared to the BPS method in [9], the reference power of inter-channel NL is obtained by minimum mean-square error (MMSE) (the same effect as Wiener filter), which does not underestimate the nonlinear noise. However, our method in [10] needs modification of the transmitter because LFM pilot has to be produced. The DP method uses normal symbols to generate the pilot and does not need further modification of the transmitter.

In this paper, we modify the DP method to estimate inter-channel NL. Our proposed method also simplifies the procedures of inter-channel NL estimation. Next we simulate the inter-channel estimation performance for quadrature phase-shift keying (QPSK) and 16-quadrature amplitude modulation (16QAM) systems. Then we analyze the stability and the length of the DP. At last, the LFM method in [10], the previous DP method in [9] and the modified DP method are compared.

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## 2. Principle and system setup

### 2.1. Inter-channel NL estimation principle

First we analyze the effect of intra-channel NL. The intra-channel nonlinear noise can be expressed by (1) [3].

$$\Delta_h(k) = (j\gamma \frac{\tau^2}{\sqrt{3}\beta_2})(\sum_m \sum_n H_{m+k} H_{n+k} H^*_{m+n+k} E_1(-j\frac{mnT^2}{\beta_2 L}) + \sum_m \sum_n H_{m+k} V_{n+k} V^*_{m+n+k} E_1(-j\frac{mnT^2}{\beta_2 L})) \tag{1}$$

where $m$, $n$ and $k$ are arbitrary integers for different signal symbol indexes, $\Delta_h(k)$ is intra-channel NL perturbation pulse at time slot $k$ on h polarization, $H$ and $V$ are the transmitted symbols for h and v polarizations, $T$ is the symbol period, $L$ is fiber length, $\tau$ is the pulse width, $\beta_2$ is dispersion coefficient, $\gamma$ is nonlinear coefficient and $E_1$ is the exponential integral function. The DP signal in h and v polarization can be represented by $e^{j\omega kT}$ and $e^{-j\omega kT}$, where $\omega$ is the angular frequency. The corresponding frequency for h and v polarization is positive and negative $1/4$ of the symbol rate respectively [8,9]. Substitute the DP expression into (1), the summation part of $\Delta_h(k)$ becomes linear combination of $e^{j\omega kT}$ and $e^{j\omega 2mT}e^{j\omega kT}$. After summation of $m$, $\Delta_h(k)$ is a signal with angular frequency $\omega$, which shares the same frequency as the transmitted signal on h polarization. The intra-channel distortion on v polarization also has the same frequency as the transmitted signal. In conclusion, the intra-channel nonlinearity does not produce new frequency component for DP signal.

Second the inter-channel NL is analyzed by the cross phase modulation (XPM) model [13]. The polarization crosstalk of v polarization on h polarization $PC_{VH}$ and the phase noise of h polarization caused by other channels $PN_H$ are expressed by (2) and (3) respectively. The polarization crosstalk and phase noise are both convoluted by a XPM filter, which is depicted in (4).

$$PC_{VH} = \sum_{c=2}^{C} \sum_{s=1}^{S} jH(t - D_{c,s}, c, s)V^*(t - D_{c,s}, c, s) \otimes h(t, c, s) \tag{2}$$

$$PN_H = \sum_{c=2}^{C} \sum_{s=1}^{S} j(2|H(t - D_{c,s}, c, s)|^2 + |V^*(t - D_{c,s}, c, s)|^2) \otimes h(t, c, s) \tag{3}$$

$$H(\omega, c, s) = \frac{8\gamma(1 - exp(-\alpha_l L + j\Delta\beta_{c,s}\omega L))}{9(\alpha_l - j\beta_{c,s}\omega)} \tag{4}$$

where $c$ and $s$ are channel and span index respectively, $C$ and $S$ are the total number of channels and spans respectively, $H/V(t, c, s)$ is the transmitted symbol of channel $c$ for h or v polarization at time slot $t$ after $s$ spans. $\alpha_l$ is fiber loss per km, $\Delta\beta_{c,s}$ and $D_{c,s}$ are the group velocity difference and accumulated differential delay respectively between the interfering channel with channel index $c$ and the probe channel after $s$ spans, $h(t, c, s)$ and $H(\omega, c, s)$ are the XPM filter response in time domain and frequency domain and $\otimes$ denotes convolution. In long-haul optical transmission, $exp(-\alpha_l L)$ is much lower than 1 and $H(\omega, c, s)$ turns into $8\gamma/9(\alpha_l - j\beta_{c,s}\omega)$. Then $H(\omega, c, s)$ is a low-pass filter with cutoff frequency $\alpha_l/\beta_{c,s}$. The power spectrum distribution of DP is depicted in Fig. 1.

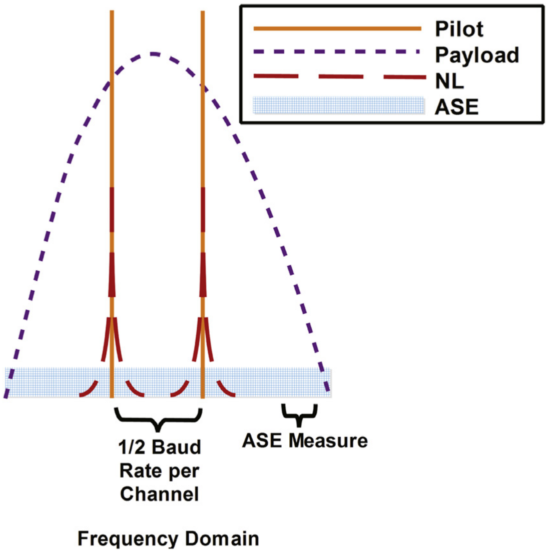

**Fig. 1.** Power spectrum distribution of DP signal after optical transmission with NL and ASE noise.

As analyzed above, intra-channel NL does not generate new frequency component and inter-channel NL has low-pass filter nature from XPM model. If the ASE is measured at the spectrum which is far away from the DP, the inter-channel NL is not included when calculating the power of ASE noise. The bandwidth of DP can be treated as infinitely close to zero and the inter-channel NL contained by the peak can be omitted. Then the peak is the combination of the pilot and intra-channel NL. The received signal with peak excluded can be seen as the ASE noise and inter-channel NL. The power of inter-channel NL $P_{inter-channel NL}$ can be obtained by (5).

$$\begin{aligned} P_{inter-channel NL} &= P_{Total} - P_{Peak} - P_{ASE} \\ &= P_{Total} - P_{Peak} - B_{Sample}N_{AM}/B_{AM} \end{aligned} \tag{5}$$

where $P_{Total}$ is the total power of the sampled signal after nonlinear transmission, $P_{Peak}$ is the peak power of received pilots, $P_{ASE}$ is the total power of ASE noise in the sampled signal, $N_{AM}$ is the power of noise in ASE measurement zone shown in Fig. 1, $B_{AM}$ is the bandwidth of ASE measurement zone and $B_{Sample}$ is the bandwidth of the sampled signal. The inter-channel NL contribution to the power of the peak itself is ignored in (5). Therefore the interference contained by the peak only comes from the ASE and intra-channel NL. After subtracting the power of ASE and the peak power of received pilots from $P_{Total}$, the power of inter-channel NL can be obtained by (5). It should be noted that Four-Wave Mixing (FWM) effect belongs to inter-channel effect and it exists randomly in frequency domain. If the bandwidth of ASE measurement zone and the pilot is narrower, the possibility that FWM exists inside the ASE measurement zone and the pilot zone becomes smaller. In most cases, there is no FWM in the ASE measurement zone and the pilot zone. The sample rate of the received pilot is the same as the payload, therefore the FWM noise which exists in the spectrum of the channel under test is also included by the received pilot. As explained above, the peak of the received signal contains the original useful DP and intra-channel NL after ignoring the power of inter-channel NL contribution to the power of the peak itself. Since the ASE estimation is probably not affected by FWM, the measured noise at the ASE measurement zone can be treated as pure ASE noise. After subtracting the power of the peak and the power of ASE noise, the residual power becomes the combination of XPM and FWM noise. Therefore the inter-channel NL can be monitored by (5) with XPM and FWM noise included.

When calculating the power of the peak of the signal after nonlinear long-haul transmission, the spectrum of the peak is surrounded by the inter-channel NL and ASE noise. The number of points ($n_{Peak}$) that the peak of DP occupies should be calibrated for different system configurations. The frequency resolution ($\Delta f$) of the sampled DP is $1/(nT)$, where $n$ is the number of DP symbols sampled at the receiver. The laser linewidth is determined by the magnitude of phase noise in laser. The variance of phase noise is positively related to the product of symbol period and laser linewidth [14,15]. For larger variance of phase noise, the bandwidth of laser output signal also becomes larger. The DP is a single frequency signal on its own polarization and hence the bandwidth of DP at the receiver is positively related to the laser linewidth. Therefore the product of frequency resolution ($\Delta f$) and the number of samples in the peak ($n_{Peak}$) will increase with the laser linewidth. In other words, the number of points in the peak of the DP