Optical Homodyne Receiver Comprising Phase and Polarization Diversities with Digital Signal Processing

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Abstract-This paper describes a coherent optical receiver, where drifts of the carrier phase and state of polarization are estimated with digital signal processing. The complex amplitude of the signal is entirely restored with such receiver.

I. INTRODUCTION

Multi-level optical modulation formats have attracted much attention because they can increase the spectral efficiency in high capacity wavelength-division multiplexed (WDM) transmission systems. We have recently proposed and demonstrated a digital coherent receiver that can demodulate the multi-level phase-shift-keying (*M*-ary PSK) signal [1,2]: A homodyne phase-diversity receiver retrieves the in-phase and quadrature components of an optical signal without locking the phase of the local oscillator (LO). The carrier-phase drift is estimated after homodyne detection by means of digital signal processing (DSP), and the *M*-ary PSK signal is restored in a very stable manner. We have also introduced the polarization diversity scheme into our receiver. DSP-based maximum-ratio polarization combining makes the receiver sensitivity entirely independent of the state of polarization (SOP) of the incoming signal without any power penalty [3]. Together with digital phase estimation, the bit-error-rate performance of our receiver is tolerant to the phase noise and polarization fluctuation of the signal.

In this presentation, after discussing the principle of operation of the receiver, we show our experiments on demodulation of multi-level coded optical signals [2] and post-compensation for fiber dispersion [4].

II. PPRNCIPLE OF OPERATION OF THE COHERENT RECEIVER

The homodyne phase/polarization diversity receiver that was used in our experiments is shown in Fig.1. The receiver consists of a 2×8 free-space optical circuit packaged in a small metal case, where two homodyne phase diversity receivers [1] are combined with the polarization diversity configuration [3].

The incoming signal having an arbitrary SOP is separated into two polarization components with a polarization beam splitter (PBS). The *x*- and *y*-polarization components after PBS are written as

$$\begin{bmatrix} E_{sx} \\ E_{sy} \end{bmatrix} = \begin{bmatrix} \sqrt{\alpha} E_s e^{j\delta} \\ \sqrt{1-\alpha} E_s \end{bmatrix}, \tag{1}$$

where E_s denotes the complex amplitude of the incoming signal, α the power ratio of the two polarization components, and δ the phase difference between them. On the other hand, the local oscillator (LO) is split into two paths with a half mirror (HM), after its SOP is made circular by a quarter-wave plate (QWP). The *x*- and *y*-polarization components of LO after HM are written as

$$\begin{bmatrix} E_{LO,x} \\ E_{LO,y} \end{bmatrix} = \frac{1}{\sqrt{2}} \begin{bmatrix} E_{LO} \\ jE_{LO} \end{bmatrix}, \tag{2}$$

where E_{LO} is the constant amplitude of LO. The 90° optical hybrids in Fig.1 generate electric fields $E_{1,\cdots,8}$ at double balanced photodiodes PD1-PD4:

$$E_{1,4} = \frac{1}{2}\left(E_{sx} \pm \frac{1}{\sqrt{2}} E_{LO} \right), \quad E_{2,3} = \frac{1}{2}\left(E_{sx} \pm \frac{j}{\sqrt{2}} E_{LO} \right),$$
$$E_{5,8} = \frac{1}{2}\left(E_{sy} \pm \frac{1}{\sqrt{2}} E_{LO} \right), \quad E_{6,7} = \frac{1}{2}\left(E_{sy} \pm \frac{j}{\sqrt{2}} E_{LO} \right). \tag{3}$$

Photocurrents from PD1-PD4 are then given as

$$I_{PD1} \propto \sqrt{\alpha} \,|\, E_s \,|\cos(\theta + \delta), \; I_{PD2} \propto \sqrt{\alpha} \,|\, E_s \,|\sin(\theta + \delta),$$
$$I_{PD3} \propto \sqrt{1-\alpha} \,|\, E_s \,|\cos\theta, \quad I_{PD4} \propto \sqrt{1-\alpha} \,|\, E_s \,|\sin\theta, \tag{4}$$

where θ is the signal phase measured from that of LO. From Eq.(4), we find that our receiver can separately measure two polarization components of the complex amplitude E_s of the signal .

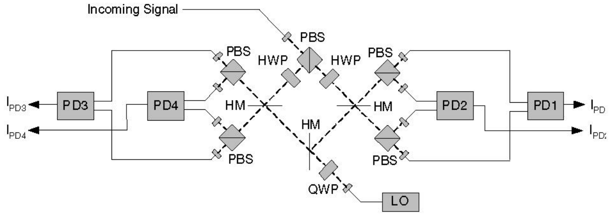

Fig.1: Schematic of homodyne receiver employing phase and polarization diversities.

III. DSP CIRCUIT FOR POLARIZATION COMBINING AND CARRIER PHASE ESTIMATION

Analogue-to-digital converters (ADCs) sample and digitize the four outputs $I_{PD1},\ldots,I_{PD4}$. The *x*- and *y*-polarization components of the complex amplitude of the signal are given from Eq.(4) as

$$E_{sx}(iT) = I_{PD1}(iT) + jI_{PD2}(iT), \; E_{sy}(iT) = I_{PD3}(iT) + jI_{PD4}(iT), \tag{5}$$

where T denotes the sampling time interval, and i the number of samples.

The left-hand side of Fig.2 shows the post-processing circuit realizing the maximal-ratio polarization combining process. We define the ratio $r(i)$ as

$$r(i) = E_{sx}(iT)/E_{sy}(iT). \tag{6}$$

Polarization parameters α and δ of the incoming signal vary much more slowly than the phase modulation. Therefore, by averaging $r(i)$ over many symbol intervals, it is possible to obtain

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accurate values of α and δ: The ratio $r(i)$ averaged over $2\ell+1$ samples is written as

$$\overline{r}(i)=\frac{1}{2\ell+1}\sum_{j=-\ell}^{\ell} r(i+j), \tag{7}$$

and α and δ are calculated from

$$|\overline{r}(i)|=\sqrt{\alpha}/\sqrt{\alpha-1},\ \arg(\overline{r}(i))=\delta. \tag{8}$$

The signal complex amplitude, independent of SOP of the incoming signal, is then reconstructed by maximal-ratio combining as

$$E_s(iT)=\sqrt{\alpha}E_{sx}(iT)e^{-j\delta}+\sqrt{1-\alpha}E_{sy}(iT). \tag{9}$$

The signal phase $\theta(iT)$ obtained from Eq.(9) contains both the phase modulation $\theta_m(iT)$ and the phase noise $\theta_n(iT)$. The next step is to estimate the phase noise and obtain the phase modulation $\theta_m(iT)$. The procedure to estimate θ_n is shown in the right-hand side of Fig.2 [1]. In the case of *M*-ary PSK signals, we take the *M*-th power of $E_s(iT)$, because the phase modulation is removed from $E_s(iT)^M$. Averaging $E_s(iT)^M$ over $2k+1$ samples constitutes a phase estimate as

$$\theta_e(iT)=\arg\left(\sum_{j=-k}^{k} E_s((i+j)T)^M\right)/M. \tag{10}$$

The phase modulation $\theta_m(iT)$ is determined by subtracting $\theta_e(iT)$ from the measured phase of $\theta(iT)$.

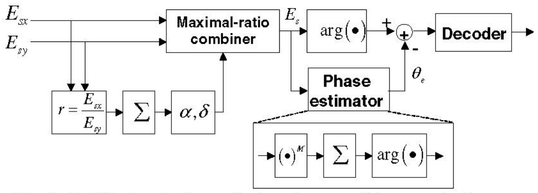

Fig.2: DSP circuit for polarization combiner and phase estimator.

IV. BIT-ERROR RATE MEASUREMENTS OF *M*-ARY PSK SIGNALS

The back-to-back BER of the BPSK, QPSK, and 8-PSK signals was measured to access the receiver performance [2]. The DFB laser output was modulated through $LiNbO_3$ phase modulators to generate PSK signals at a symbol rate of 10 Gsymbol/s. The received signal was amplified with an erbium-doped fiber amplifier (EDFA) to –10 dBm before it was detected with the coherent receiver. The linewidth of the transmitter and the LO was about 150 kHz and frequency drifts of the lasers were kept below 10 MHz. The photodiode currents were simultaneously sampled at a rate of 20 Gsample/s with analog-to-digital converters (ADCs). The collected samples were resampled to keep only one point per symbol and combined to form a 100-ksymbol-long stream. The signal was demodulated through digital signal processing described in Sec.III, and the number of bit errors was counted through off-line measurements. Fig.3 shows BERs measured as a function of the received power. Fig.4 represents the constellation map of the 8-PSK signal obtained in the error-free state.

V. POST-COMPENSATION FOR GVD

Coherent detection can linearly recover the amplitude and phase information of optical signals. Therefore, post-processing of the received signal allows fully electronic compensation for chromatic dispersion, which is a linear transfer function operating on the optical complex amplitude. In this section, we demonstrate unrepeated transmission of 20-Gbit/s optical QPSK signals over a 200-km standard single-mode fiber (SMF), where after homodyne phase-diversity detection, digital signal processing is employed for carrier phase estimation as well as dispersion compensation [4]. A total dispersion of up to 4,000 ps/nm is compensated effectively through a simple transversal digital filter implemented in our coherent receiver.

Fig. 5 shows measured BER of the 20-Gbit/s QPSK signals, when the fiber dispersion is compensated with various filters. For comparison, the back-to-back BER is also shown in Fig. 5. We find that even a small number of taps of the transversal filter provides significant improvement of the BER performance. 39 taps are sufficient to compensate for dispersion value of 4,000 ps/nm.

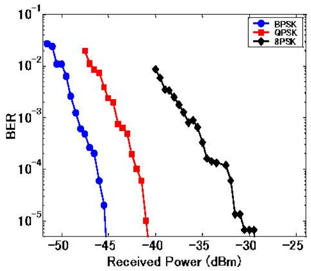

Fig.3 Back-to-back BER curves for BPSK, QPSK, and 8-PSK signals.

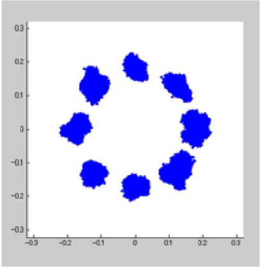

Fig.4 Constellation map for the 8-PSKsignal.

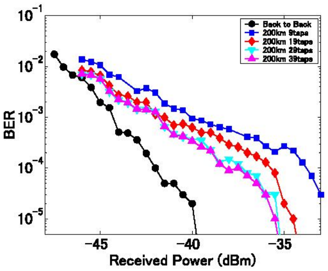

Fig.4 BERs measured as a function of the received power after transmission through a 200-km SMF.

VI. CONCLUSION

We have investigated the homodyne phase-diversity receiver, where the carrier phase and SOP are estimated with digital signal processing. We have demonstrated demodulation of *M*-ary PSK signals (*M*=2, 4, 8) at the symbol rate of 10 Gsymbol/s and post-compensation for GVD. The next step is to develop dedicated integrated circuits for real-time operation of the receiver.

REFERENCES

[1] K. Kikuchi, "Phase-diversity homodyne detection of multi-level optical modulation with digital carrier phase estimation," IEEE J. Selected Topics on Quantum. Electron., vol.12, no.4, pp.563-570 (July/Aug. 2006)

[2] S. Tsukamoto, K. Katoh and K. Kikuchi, "Coherent demodulation of optical multilevel phase shift-keying signals using homodyne detection and digital signal processing," IEEE Photon. Technol. Lett., vol.18, no.10, pp.1131-1133 (May 2006)

[3] S. Tsukamoto, Y. Ishikawa, and K. Kikuchi, "Optical homodyne receiver comprising phase and polarization diversities with digital signal processing," European Conference on Optical Communication (ECOC 2006), Cannes, France, Mo4.2.1 (24-28 Sept. 2006)

[4] S. Tsukamoto, K. Katoh and K. Kikuchi, "Unrepeated transmission of 20-Gbit/s optical quadrature phase-shift keying signal over 200-km standard single-mode fiber based on digital processing of homodyne-detected signal for group-velocity dispersion compensation," IEEE Photon. Technol. Lett., vol.18, no.9, pp.1016-1018 (May 2006)