LETTER

Polarity-Reversing Type Photonic Receiving Scheme for Optical CDMA Signal in Radio Highway

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SUMMARY This letter newly proposes the polarity-reversing type photonic receiving scheme based on bipolar correlation for optical CDMA signal in radio highway. The proposed scheme can more improve the limitation of the number of radio base stations connected to radio highway and more reduce the peak laser power at the radio base station than the conventional unipolar type receiving scheme using prime codes.
key words: *radio highway, CDMA, optical polarity reversed correlator, bipolar code sequence*

1. Introduction

Wide-band data transmission needs a lot of frequency resources in future, and microcellular technology has many problems including signal transfer among many microcells. To solve these problems, we have proposed radio highway network [1], where microcells are connected by optical fibers among radio base station (RBS)s and a control station (CS) and radio signals are transmitted over them. As a configuration of radio highway network, the bus type or passive double star optical link is desirable from the viewpoint of low cost implementation and easy addition of access point.

As a multiple access method of radio highway network, CDMA will be a strong candidate because of its asynchronous access property, flexibility and transparency for various radio air interfaces. In radio highway network, the optical CDMA is more suitable than the conventional electrical CDMA [2] because the high process gain can be gained by using the broad bandwidth of optical devices when radio signals are multiplexed in optical domain.

Up to this time, various optical CDMA methods have been studied for digital signals in LAN. In optical CDMA methods using the phase modulation of optical signals, there are two methods in the configuration of correlator such as the optical coherent detection [3] and the optical phase modulator with optical direct detection (DD) [4]. However, the optical coherent detection is very complicated and the correlator using phase modulators with optical DD needs very fine narrowband optical filters that are very sensitive to temperature changes. On the other hand, the optical CDMA method using fiber delay lines can not be applied to analog radio signals, because its spreading and despreading are performed by using optical pulses [5]. So we have newly proposed direct optical switching (DOS)-CDMA method, in which the on-off switching at the optical switch (OSW) provides the spectrum spreading of radio signals directly intensity-modulated (IM) by the laser diode (LD) [6]. Thus any types of radio signal can be converted into IM/CDMA signals. In this method illustrated in Fig. 1, only optical orthogonal codes such as prime codes can be applied to obtain a desired process gain. However prime codes suffer a limit in the number of distinct code sequences, which results in the limitation of the number of RBS's connected to the radio highway. Therefore we should consider a new type of DOS-CDMA method to which we can apply bipolar codes such as Maximal length codes and Gold codes used in radio system because those codes are generally superior in the number of distinct code sequences compared with prime codes.

For digital network using optical CDMA method, sequence inversion keyed (SIK) direct sequence (DS) CDMA's that have been proposed in Refs. [7] and [8], require specially balanced bipolar codes. In order to allow the use of any unbalanced bipolar codes, the power splitting ratio of the power divider at the optical correlator has been controlled [9], and the two channels transmission using two wavelengths or two orthogonal polarizations has been proposed [10]. In SIK CDMA method, however, binary digital data are encoded and transmitted with the positive polarity and the negative polarity of bipolar codes, thereby, its correlating scheme at the receiver can not be applied to DOS CDMA signals that are converted from radio

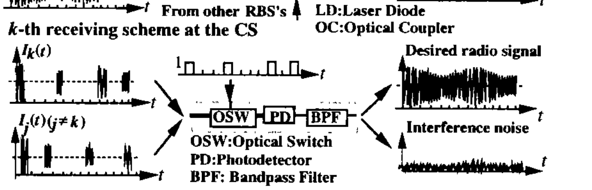

Fig. 1 Conventional unipolar type receiving scheme for DOS CDMA [6].

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signals and spectrum-spread by on-off switching.

In this letter, we newly propose polarity-reversing type photonic receiving scheme to apply bipolar codes to DOS CDMA [11], [12]. The polarity-reversing type photonic receiving scheme is realized with only a Mach-Zehnder Interferometer (MZI) and a bandpass filter (BPF), while the correlator for SIK CDMA requires two optical switches [8], [9], or four matched filters [10].

2. Principle of Polarity-Reversing Type Photonic Receiving Scheme for DOS-CDMA

Figure 2 illustrates the configuration of the transmitter and the polarity-reversing type photonic receiver for DOS-CDMA using the bipolar code $c_k(t)$ having the value of 1 or -1. At the k-th RBS, LD is directly modulated by the radio signal, $g_k(t)$, with the optical modulation index of 1. DOS CDMA is performed at OSW driven with the biased bipolar code $1 \times c_k(t)/2$, which has the code length L, the frame period T_F and the chip width T_C. Then, the intensity of optical signal from the k-th RBS is written by

$$I_k(t) = P_S(1+g_k(t))\frac{1+c_k(t)}{2}, \tag{1}$$

where P_S is the peak transmitted laser power and $g_k(t)$ is the radio signal with its bandwidth B_{rf} and carrier frequency f_{rf}. At the receiver, many optical signals from many RBS's are correlated, photodetected and filtered at the optical polarity reversed correlator (OPRC), which is composed of an optical polarity reversing divider (OPRD) and a balanced mixing photodetector (BMPD). The OPRD is realized by a MZI which comprises two beam splitter (BS)s and two optical phase shifter (OPS)s. The OPRD is driven by two switching voltages, v_{p1} and v_{p2} proportional to $c_k(t)$. During the positive polarity period of $c_k(t)$, the phase difference between two OPS's is zero and thus,

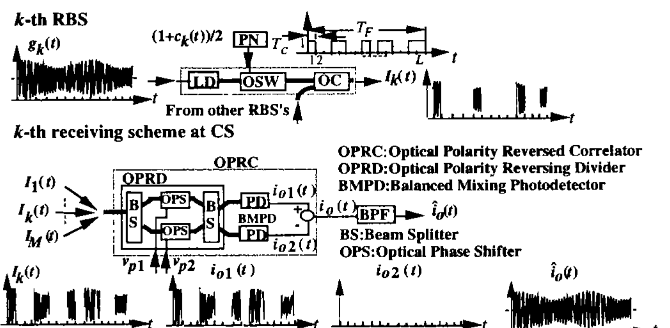

Fig. 2 Configuration of the transmitter and the polarity-reversing type photonic receiver for DOS-CDMA using bipolar codes.

the upper port of OPRD outputs the optical signal. During the negative polarity period of $c_k(t)$, the phase difference between two OPS's is π and thus, the lower port of OPRD outputs the optical signal. Output currents of the BMPD are expressed as

$$i_{o1}(t) = \alpha\sum_{j=1}^{M} P_r g_j(t)\frac{1+c_j(t)}{2}\frac{1+c_k(t)}{2} + i_{n1}(t), \tag{2}$$

$$i_{o2}(t) = \alpha\sum_{j=1}^{M} P_r g_j(t)\frac{1+c_j(t)}{2}\frac{1-c_k(t)}{2} + i_{n2}(t). \tag{3}$$

where α and M are the responsivity of BMPD and the number of connected RBS, $i_{n1}(t)$ and $i_{n2}(t)$ are additive noise currents, respectively. And P_r is the received optical power given by P_s/F_{loss}, where F_{loss} is the optical loss between each RBS and OPRC. We assume that an optical amplifer (OA) is equipped at each RBS and the gain of OA is equal to F_{loss}, thus P_s is equal to P_r.

The input current to the BPF is given by

$$i_o(t) = i_{o1}(t) - i_{o2}(t) = i_{S_k}(t) + i_I(t) + i_N(t). \tag{4}$$

where $i_{S_k}(t)$, $i_I(t)$ and $i_N(t)$ are the desired signal, the interference and the additive noise, respectively. $i_{S_k}(t)$ and $i_I(t)$ are expressed as

$$i_{S_k}(t) = \alpha P_r g_k(t)\frac{1+c_k(t)}{2}, \tag{5}$$

$$i_I(t) = \sum_{j=1, j\neq k}^{M} \alpha P_r g_j(t)\frac{1+c_j(t)}{2} c_k(t). \tag{6}$$

The output currents, $i_{o1}(t)$ and $i_{o2}(t)$, show that for the desired k-th signal the positive polarity of $c_k(t)$ matches with the code of $I_k(t)$ at the upper port of OPRC, on the other hand the k-th signal is obstructed at the lower port of OPRC, which results in no detection output current. For the interference signal $I_j(t)$ ($j \neq k$), the upper or lower port of OPRC generates the interference current only when the positive or negative polarity of $c_j(t)$ coincides with the same polarity of $c_k(t)$. Thus, the OPRC acts as the matched-filter for IM/CDMA signals. The desired signal, $i_{S_k}(t)$, is the pulse amplitude modulation (PAM) signal generated from the radio signal, $g_k(t)$. Since the positive polarity of bipolar code sequence drives the OSW periodically with the sampling period T_F at the RBS transmitter, each signal pulse in one code sequence period has the same period T_F. Then, if the periodic frequency $1/T_F$ is set to be at least twice radio signal bandwidth ($= 2B_{rf}$), it is found from the bandpass natural sampling theory [13] that the radio signal can be recovered from only one pulse train in the code sequence. In this way, we can obtain the desired radio signal contaminated by some radio interference signals after the interpolation at the BPF.

3. Theoretical Analysis of Carrier-to-Interference-Plus-Noise Ratio (CINR)

In this section, we theoretically analyze the carrier-to-interference-plus-noise ratio (CINR) of the recovered radio signal at the receiver. We assume that radio signals, $g_j(t)$ $(j=1, 2, \cdots, M)$, are nonmodulated carriers which have the autocorrelations function, $R_g(\tau)$,

$$R_g(\tau)=\frac{1}{2}\cos(2\pi f_{rf}\tau). \tag{7}$$

The autocorrelation function of $i_{S_k}(t)$ expressed as

$$R_S(\tau)=\left(\frac{\alpha P_r}{2}\right)^2 R_g(\tau)(1+R_C(\tau)), \tag{8}$$

where $R_C(\tau)$ is the autocorrelations of the code sequence, $c_k(t)$, given by [14]

$$R_C(\tau)=\frac{1}{L^2}+\frac{L+1}{L^2}\sum_{k=-\infty,k\neq 0}^{\infty}\mathrm{sinc}^2\left(\frac{\pi k T_C}{T_F}\right)\cdot\cos\left(\frac{2\pi k\tau}{T_F}\right). \tag{9}$$

From Eqs. (7)-(9), the carrier power of the recovered radio signal at the output of the BPF is written by

$$C=\frac{1}{2}\left(\frac{\alpha P_r}{2}\right)^2\left(1+\frac{1}{L^2}\right). \tag{10}$$

The autocorrelation function of the interference, $i_I(t)$, is expressed as

$$R_I(\tau)=(M-1)\left(\frac{\alpha P_r}{4}\right)^2 R_g(\tau)(1+2R_C(\tau)+R_\theta(\tau)), \tag{11}$$

where $R_\theta(\tau)$ represents the autocorrelation of a function $\theta_{ik}(t)=c_i(t)c_k(t)$. The $R_\theta(\tau)$ is given by [15]

$$R_\theta(\tau)=\overline{a_0^2}+2\sum_{k=1}^{\infty}\frac{1}{L}\frac{1}{1+(k\pi/L)}\cos\left(\frac{2\pi k\tau}{T_F}\right), \tag{12}$$

where the value of $\overline{a_0^2}$ depends on the kind of used code sequence. For example, for Gold codes and Maximal length codes, $\overline{a_0^2}$ is given by

$$\overline{a_0^2}=\begin{cases}\dfrac{L^2+L-1}{4L^3}, & \text{for Gold code [16]}\\[2ex] \dfrac{1}{L}, & \text{for Maximal length code [15].}\end{cases} \tag{13}$$

From Eqs. (7), (9) and (11)-(12), the interference power at the output of the BPF is obtained as

$$I=\begin{cases}\dfrac{M-1}{2}\left(\dfrac{\alpha P_r}{2}\right)^2\dfrac{L^2+5L-1}{4L^3}, & \text{for Gold code}\\[2ex] \dfrac{M-1}{2}\left(\dfrac{\alpha P_r}{2}\right)^2\dfrac{L+1}{L^2}, & \\ & \text{for Maximal length code.}\end{cases} \tag{14}$$

Hence, the carrier-to-interference power ratio (CIR) is written by

$$CIR=\begin{cases}\dfrac{4L(L^2+1)}{(M-1)(L^2+5L-1)}, & \text{for Gold code}\\[2ex] \dfrac{(L^2+1)}{(M-1)(L+1)}, & \\ & \text{for Maximal length code.}\end{cases} \tag{15}$$

On the other hand, in the case of conventional unipolar type receiving scheme using prime codes, the CIR is given by [5]

$$CIR_{p,uc}\approx\frac{1}{(M-1)}\frac{p^2}{0.29}, \tag{16}$$

where p is a prime number of the prime code, whose length, L, is p^2.

At the output of BPF, we consider the additive noise currents composed of relative intensity noise, the shot noise, the receiver thermal noise, the beat noise among amplified spontaneous emissions (ASE) of an optical amplifier and optical signals, the beat noise among ASE's, and the optical signal beat noise. The total noise power, N, is written by

$$N=N_{RIN}+N_{shot}+N_{th}+N_{s-sp}+N_{sp-sp}+\langle N_{s-s}\rangle \tag{17}$$

and each power is given by

$$N_{RIN}=\zeta_{RIN}\left(\frac{\alpha P_r}{2}\right)^2 B_{rf}, \tag{18}$$

$$N_{shot}=2e\alpha\left(\frac{P_r}{2}+MN_{sp}W\right)B_{rf}, \tag{19}$$

$$N_{th}=\frac{4kT}{R_L}B_{rf}, \tag{20}$$

$$N_{s-sp}=4\alpha^2 MN_{sp}\frac{P_r}{2}B_{rf}, \tag{21}$$

$$N_{sp-sp}=2(\alpha MN_{sp})^2(W-f_{rf}), \tag{22}$$

where, e, ζ_{RIN}, W, k, T and R_L are the electric charge, the power spectral density (PSD) of the relative intensity noise, the bandwidth of optical filter at the CS, Boltzmann constant, the noise temperature and the load resistance, respectively. The PSD of the ASE, N_{sp}, is given by

$$N_{sp}=\frac{\eta_{sp}}{\eta_a}\cdot\frac{G_a-1}{G_a}h\nu, \tag{23}$$

where, G_a, η_{sp}, η_a, and $h\nu$ are the gain, the spontaneous emission factor, the quantum efficiency of the OA and the photon energy, respectively. On the other hand, the optical signal beat noise is caused by an interference among M optical carriers from M RBS's. The optical signal beat noise appears in the radio frequency band after photodetection at the BMPD, but its frequency location and its PSD depend on the frequency difference among M LD's and the spectrum of each optical IM/CDMA signal, respectively. In this analysis, we assume that the optical signal has the single mode gaussian shaped spectrum and also that its center frequency is a random variable with a uniform probability density [2]. Under these assumptions, we derive the PSD of optical signal beat noise and its noise power N_{s-s} in signal band, which should be treated as a random variable. So we investigate the optical signal beat noise power by using its average power$\langle N_{s-s}\rangle$ [12].

From Eqs. (10), (14) and (17), the CINR of the radio signal regenerated at the CS can be represented as

$$CINR=\frac{C}{I+N}. \tag{24}$$

4. Numerical Results of Performance and Discussion

Some numerical results are discussed below with parameters indicated in Table 1. Figure 3 shows the relationship between the code length L and the CIR for the proposed polarity-reversing type photonic receiving scheme and the unipolar type receiving scheme in the case of using bipolar codes for $M=2$. It is seen from Eq. (15) that in the proposed receiving scheme, the CIR is improved as the L increases. On the other hand, in the case of the conventional unipolar type receiving scheme, there is no improvement for both codes. This is due to nonspread component in the interference which the unipolar type receiving scheme cannot remove. It is from Eq. (A·3) that the CIR for both cases using maximal length codes and Gold codes equals to $4/(M-1)$ for large L. This result means that bipolar codes can be applied to radio highway by using the proposed polarity-reversing type photonic receiving scheme. On the other hand, in the case of the conventional unipolar type receiving scheme, only prime codes can be applied. So in the next figure, we compare the number of distinct code sequences which result in the limitation of the number of RBS's connected to the radio highway.

Figure 4 shows the number of distinct code sequences versus the code sequence length. Gold code sequences are generated by combining a pair of preferred Maximal length sequences using modulo-2 addition if the number of preferred maximal length sequences is at least two [16]. On the other hand, in the case of prime codes, the number of distinct code sequences is equal to the prime number p for the code sequence length of p^2 [5]. The number of distinct code sequences for Maximal length codes and Gold codes is larger than that for prime codes. Thus, for example, comparing Maximal length codes and Gold codes of $L=8191$ with prime codes of $p^2=8281$, the number of distinct code sequences for Maximal length codes and Gold codes is 7 times and 90 times larger than that for prime codes, respectively. Therefore, by using the proposed receiving scheme we can assign larger number of distinct code sequences, thereby we can connect larger number of radio base stations to radio highway than the unipolar type receiving scheme using prime codes.

Figure 5 shows the relationship between the peak transmitted optical power P_s and the CINR for Maximal length codes and Gold codes of $L=8191$ and prime codes of $p^2=8281$. From Eqs. (15) and (16), large L or p^2 is needed to obtain large CINR for both receiving schemes because the CIR is the upper limit of the CINR. In the conventional unipolar type receiving

Table 1 Parameters used in calculations.

ζ_{RIN}	-152dB/Hz	f_{rf}	1.9GHz
α	0.8A/W	R_L	50 Ω
η_{sp}	2.0	T	300K
η_a	0.5	G_a	6.5dB
B_{rf}	300KHz	W	1THz

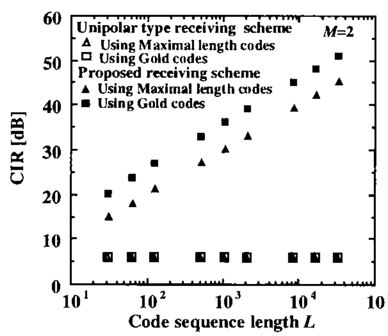

Fig. 3 Relationship between code sequence length L and CIR.

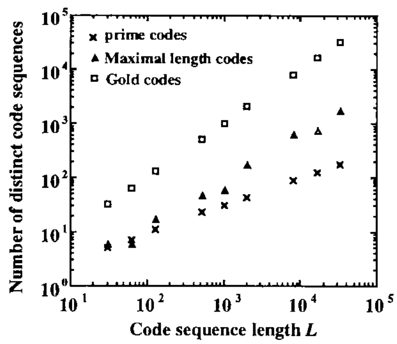

Fig. 4 Number of distinct code sequences versus code sequence length L.

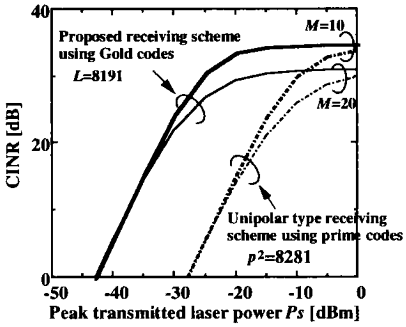

Fig. 5 Relationship between peak transmitted optical power P_S and CINR.

scheme using prime codes, the carrier-to-noise power ratio (CNR) awfully deteriorates the CINR as P_S decreases because the signal power is in proportion to P_S^2/p^2. However, in the proposed receiving scheme using Gold codes, the CINR is not so abruptly deteriorated with the CNR as P_S decreases because the signal power is in proportion to $(P_S/2)^2$ as given in Eq. (10). From this result, the proposed polarity-reversing type photonic receiving scheme using Gold codes can reduce the peak laser power at the RBS in the case that the peak transmitted optical power is limited.

5. Conclusion

In this paper, we have newly proposed the polarity-reversing type photonic receiving scheme for optical CDMA signal in radio highway. The carrier-to-interference-plus-noise ratio of the recovered radio signal at the control station has been theoretically analyzed. Analysis results have clarified that bipolar codes can be applied to optical CDMA for radio highway. It is also found that by using the proposed receiving scheme we can more improve the limitation of the number of radio base stations connected to radio highway and a little better carrier-to-interference ratio can be obtained by using Gold codes than the unipolar type receiving scheme using prime codes. It is also found that the proposed receiving scheme can reduce the peak laser power at radio base station. In this analysis, we have not considered the dependence on both the polarization and the optical frequency at the OPRC. Further study for them will be required.

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Appendix

In the case of the unipolar type receiving scheme, from Eq. (2), the interference current, $i_I(t)$, is given by

$$i_I(t)=\sum_{j=1,j\neq k}^{M}\frac{aP_r}{4}g_j(t)(1+c_j(t))(1+c_k(t)). \tag{A·1}$$

Autocorrelation function of $i_I(t)$ is given by

$$R_I(\tau)=(M-1)\left(\frac{aP_r}{4}\right)^2 R_g(\tau)(1+2R_c(\tau)+R_\theta(\tau)). \tag{A·2}$$

Then, from Eqs. (7), (9), (10), (12) and (A·2), the CIR in the unipolar type receiving scheme can be obtained as

$$CIR_{G,uc}=\frac{16L(L^2+1)}{(M-1)(4L^3+L^2+9L-1)}, \quad \text{for Gold code}$$

$$CIR_{M,uc}=\frac{4(L^2+1)}{(M-1)(L^2+L+2)}, \quad \text{for Maximal length code.} \tag{A·3}$$