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Carrier frequency offset estimation method for 2 × 1 MISO TDS-OFDM systems

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Abstract

In time-domain synchronous (TDS)-orthogonal frequency division multiplexing (OFDM) systems, a pseudo noise (PN) sequence is inserted instead of the cyclic prefix. The PN sequence is used not only as a guard interval but also as a training sequence for channel estimation and synchronization in the time domain. Recently, research studies on 2 × 1 multi input-single output (MISO) TDS-OFDM systems have been conducted, and different PN sequences (which are orthogonal to one another or cyclically shifted) are transmitted at each transmit antenna for channel estimation, which are modulated by binary phase shift keying in the same phase angle. However, when the absolute phase difference among the transmitted PN sequences is π, a PN sequence cancellation problem occurs, making the estimation of an accurate carrier frequency offset (CFO) difficult. In this paper, a CFO estimation method with the aid of PN sequences for 2 × 1 MISO TDS-OFDM systems is proposed. In the proposed method, the phase of the PN sequences at each antenna is rotated differently and transmitted to prevent a PN sequence-canceling problem. In addition, a CFO estimation scheme using channel state information is proposed to estimate an accurate CFO in time-varying channels. We show by computer simulations that the mean square error performance of the proposed method over an additive white Gaussian noise environment and time-varying Rayleigh channel is higher than that of the conventional method.

Keywords: TDS-OFDM; MISO; Carrier frequency offset estimation; PN sequence

1 Introduction

Orthogonal frequency division multiplexing (OFDM) systems [1] are popular broadband communication systems and have been well known to be effective against multipath distortion. In general OFDM systems, cyclic prefix (CP) is inserted as a guard interval to efficiently combat multipath channel, and a pilot signal is inserted into the subcarriers for channel estimation and carrier recovery. In time-domain synchronous (TDS)-OFDM systems, a pseudo noise (PN) sequence is employed instead of the CP, which is modulated by binary phase shift keying (BPSK) [2]. The PN sequence is used not only as a guard interval but also as a training sequence for channel estimation and synchronization in the time domain [3]. Thus, it is not needed in the transmission of the pilot signal among subcarriers, and the spectral efficiency of the TDS-OFDM systems is higher than that of the CP-OFDM systems. Currently, digital TV broadcasting services in China employ TDS-OFDM systems [4].

Recently, research on the transmit diversity scheme of the TDS-OFDM systems has been conducted in [5-7], and it mainly focused on the 2 × 1 multi input-single output (MISO) systems to achieve diversity gain. The Alamouti code [8] is a well-known transmission method for 2 × 1 MISO systems and can achieve good spatial diversity gain with minimal decoding complexity. Thus, 2 × 1 MISO TDS-OFDM systems are promising and reliable systems to achieve better performance. In these systems, different PN sequences (which are orthogonal to one another or cyclically shifted) are transmitted at each transmit antenna for channel estimation, which are modulated by BPSK in the same phase angle. However, when the absolute phase difference among the transmitted PN sequences (which are transmitted in the same phase

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angle) is π, a PN sequence cancellation problem occurs and makes the estimation of an accurate carrier frequency offset (CFO) difficult.

In this paper, a CFO estimation method with the aid of PN sequences for 2 × 1 MISO TDS-OFDM systems is proposed. In the proposed method, the phases of the PN sequences at each antenna are rotated differently and transmitted to prevent the PN sequence canceling problem. After the modulations of the received PN sequences are removed, the CFO is estimated using the L&R algorithm [9], which is a type of a maximum likelihood (ML) method with several numbers of auto-correlators. In addition, a CFO estimation scheme using channel state information (CSI) is proposed to estimate an accurate CFO in time-varying channels. Using computer simulations, the mean square error (MSE) performance is measured by employing the proposed transmission method and the L&R algorithm over an additive white Gaussian noise (AWGN) environment and a time-varying Rayleigh channel. The rest of this paper is organized as follows: in Section 2, the PN cancellation problem is analyzed when the transmitted PN sequences have the same phase angle, and the CFO estimation method is introduced even under the presence of the PN cancellation problem; in Section 3, the CFO estimation method and a scheme that uses the CSI are proposed; the computer simulation results of the MSE performance are presented in Section 4; and conclusions are drawn in Section 5.

2 PN cancellation problem and CFO estimation method in conventional 2 × 1 MISO TDS-OFDM systems

2.1 System model

Let us consider a 2 × 1 MISO TDS-OFDM system with two transmit antennas, one receive antenna and N data subcarriers. The frame structure consists of the N_{PN} (long PN sequence as the guard interval) and N subcarriers, as shown in Figure 1 [2,3]. At each transmit antenna, the PN sequences, which are orthogonal to one another or cyclically shifted, are transmitted by BPSK modulation in the same phase angle for MISO channel estimation [5-7]. We assume that the Doppler shifts between all transmit-receive antenna pairs are approximately the same [10].

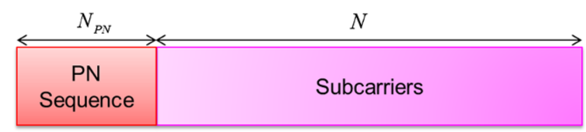

Figure 1 Frame structure of TDS-OFDM systems.

When the symbol timing and frame synchronization are ideal, the received kth PN sequence $y_{PN}(k)$ at the receive antenna in the AWGN channel can be expressed as

$$\begin{aligned} y_{PN}(k) &= [PN_1(k) + PN_2(k)]e^{j(2\pi f_c \cdot iT_s+\theta)} + n(k) \\ &= PN_{Tx}(k)e^{j(2\pi f_c \cdot iT_s+\theta)} + n(k), 0\le k\le N_{PN}-1, \end{aligned} \quad (1)$$

where $PN_1(k)$ and $PN_2(k)$ are the transmitted PN sequences from the first and second transmit antennas, respectively, f_c is the CFO, T_s is the sampling period, θ is an unknown phase offset, and $n(k)$ is the AWGN at the receive antenna.

2.2 Transmitted PN sequence cancellation problem and CFO estimation method

When two PN sequences are transmitted in the same phase angle in the conventional system [5-7], as shown in Figure 2, the received kth PN sequence $PN_{Tx}(k)$ in the ideal channel state can be written as

$$PN_{Tx}(k) = \begin{cases} 2+0j, \quad PN_1(k) = PN_2(k) = 1+0j & \text{(2a)} \\ -2+0j, \quad PN_1(k) = PN_2(k) = -1+0j & \text{(2b)} \\ 0, \quad \text{otherwise when} |\angle PN_1(k) - \angle PN_2(k)| = \pi & \text{(2c)} \end{cases} \quad (2)$$

Whereas $PN_{Tx}(k)$ is constant when $PN_1(k)$ and $PN_2(k)$ have the same phase, as expressed in (2a) and (2b), the PN sequence cancellation problem occurs and $PN_{Tx}(k)$ is equal to zero when the absolute phase difference between $PN_1(k)$ and $PN_2(k)$ is π, as expressed in (2c). To estimate the CFO, the modulation is first removed by multiplying the complex conjugate of the locally generated $PN_{Tx}(k)$ to receive the kth PN sequence $y_{PN,1}(k)$.

$$z(k) = y_{PN}(k)PN^*_{Tx}(k) = \begin{cases} 4e^{j(2\pi f_c kT_s+\theta)} + n'(k), & PN_1(k) = PN_2(k) = 1+0j \text{ or } PN_1(k) = PN_2(k) = -1+0j \quad \text{(3a)} \\ n'(k) & \text{, otherwise when } |\angle PN_1(k) - \angle PN_2(k)| = \pi \quad \text{(3b)} \end{cases}, \quad (3)$$

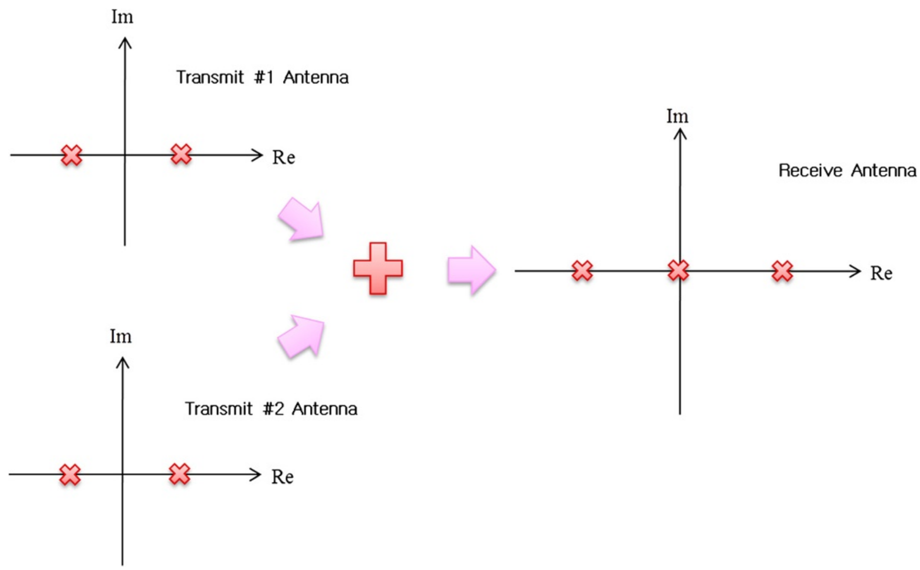

Figure 2 Received PN sequence when transmit PN sequences have the same phase.

where * is the complex conjugate operator and $n'(k) = n(k)PN^*_{Tx}(k)$.

The training-sequence-based method estimates the CFO using the phase difference among successive training sequences. Thus, the phase difference between $z(k)$ and $z(k-1)$ can be obtained as follows:

$$\arg[z(k)z^*(k-1)] = \begin{cases} \arg\left[16\left(e^{j(2\pi f_c kT_s+\theta)} + n'(k)\right)\left(e^{-j(2\pi f_c(k-1)T_s+\theta)} + n'(k-1)\right)\right] \\ \quad = \arg\left[16e^{j2\pi f_c T_s} + n''\right] = 2\pi f_c T_s + n''' & , z(k) = z(k-1) = 4e^{j(2\pi f_c kT_s+\theta)}, \\ n''' & , \text{otherwise} \end{cases} \tag{4}$$

where $n'' = e^{j[2\pi f_c kT_s+\theta]}n'(k-1) + e^{-j[2\pi f_c(k-1)T_s+\theta]}n'(k) + n'(k)n'(k-1)$ and $n''' = \arg[n'']$.

When (3a) exists continuously (meaning that (2a) and (2b2b) exist continuously), the CFO can be normally estimated. These examples are shown in Figure 3.

Only when (2a) and (2b) exist continuously, as shown in Figure 3a,d, can the CFO be estimated normally. However, when (2c) exists, as shown in Figure 3b,c, the phase difference between successive PN symbols cannot be used for the CFO estimation. Therefore, all the PN symbols may not be used for the CFO estimation, and the CFO cannot be estimated accurately using more than one auto-correlators, similar to that in the data-aided (DA)-ML methods [9,11,12].

Finally, the CFO can be estimated when (3a) exists continuously as

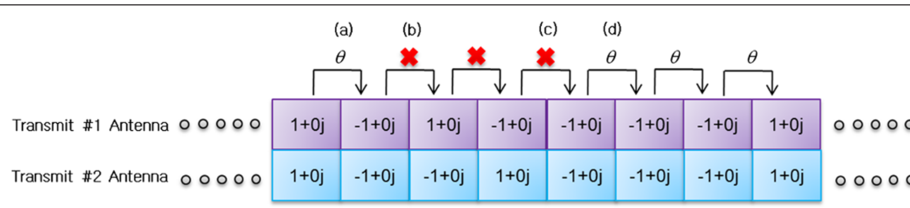

Figure 3 CFO estimation method using transmitted PN sequences that have the same phase. (a,d) Normal CFO estimation condition. **(b, d)** Impossible CFO estimation condition.

$$\nu = \frac{1}{2\pi N_{\arg} T_s} \sum \arg[z(k) z^*(k-1)], \quad \begin{array}{l} PN_1(k-1) = PN_2(k-1) = 1+0j \text{ and } PN_1(k) = PN_2(k) = 1+0j \\ PN_1(k-1) = PN_2(k-1) = 1+0j \text{ and } PN_1(k) = PN_2(k) = -1+0j \\ PN_1(k-1) = PN_2(k-1) = -1+0j \text{ and } PN_1(k) = PN_2(k) = 1+0j \\ PN_1(k-1) = PN_2(k-1) = -1+0j \text{ and } PN_1(k) = PN_2(k) = -1+0j \end{array}, \tag{5}$$

where $N_{\arg}$ is the number of $\arg[z(k)z^*(k-1)]$ calculations.

3 Proposed PN sequence transmission method and CFO estimation method in 2 × 1 MISO TDS-OFDM systems

In this section, the CFO estimation method is proposed, which rotates each phase of the transmitted PN sequences and employs the L& R method in Section 3.1. Furthermore, a CFO estimation scheme that uses the CSI is proposed in Section 3.2 for accurate estimation in the Rayleigh channel.

3.1 Phase rotated PN transmission method and CFO estimation method employing L&R algorithm

In this section, a transmission method that rotates each phase of the transmitted PN sequences to prevent the PN sequence cancellation problem is proposed. In addition, a CFO estimation method employing the L&R algorithm [9] is proposed.

The proposed transmission method is explained as follows: when the number of transmission is two, the phase of $PN_2(k)$ is rotated to make the phase difference between PN_1(A) and $PN_2(k)$ be $\pi/2$ for transmission, as shown in Figure 4. When the proposed method is applied, we can write $PN_{Tx}(k)$ in an ideal channel state as

$$PN_{Tx}(k) = \begin{cases} \frac{1}{\sqrt{2}} + \frac{1}{\sqrt{2}}j, & PN_1(k) = 1+0j, PN_2(k) = 0+1j \\ -\frac{1}{\sqrt{2}} + \frac{1}{\sqrt{2}}j, & PN_1(k) = -1+0j, PN_2(k) = 0+1j \\ \frac{1}{\sqrt{2}} - \frac{1}{\sqrt{2}}j, & PN_1(k) = 1+0j, PN_2(k) = 0-1j \\ -\frac{1}{\sqrt{2}} - \frac{1}{\sqrt{2}}j, & PN_1(k) = -1+0j, PN_2(k) = 0-1j \end{cases} \tag{6}$$

When the phase of the transmitted PN sequences is rotated by the proposed method, the $PN_{Tx}(k)$ sequences for all cases can be used for the CFO estimation, as shown in Figure 5a. In addition, more than one autocorrelators can be used for the CFO estimation, similar to the ML methods [9,11,12], because all consecutive $PN_{Tx}(k)$ sequences can be used, as shown in Figure 5b. Therefore, the accuracy of the estimated CFO can be increased as compared with that of the conventional method.

The frequency offset estimation method is explained as follows: the modulation is removed by multiplying the complex conjugate of the locally generated $PN_{Tx}(k)$ to receive $y_{PN}(k)$.

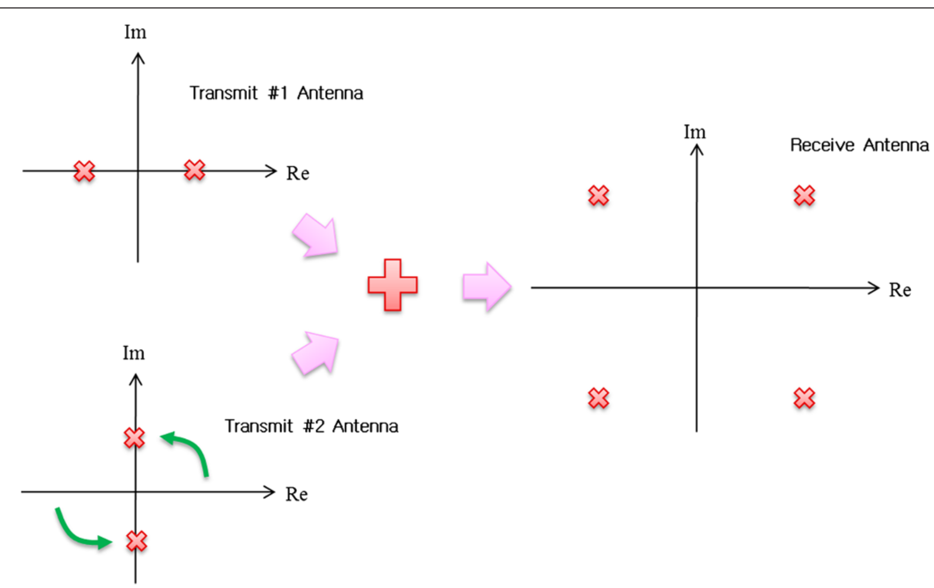

Figure 4 Proposed PN sequence transmission method using phase rotation.

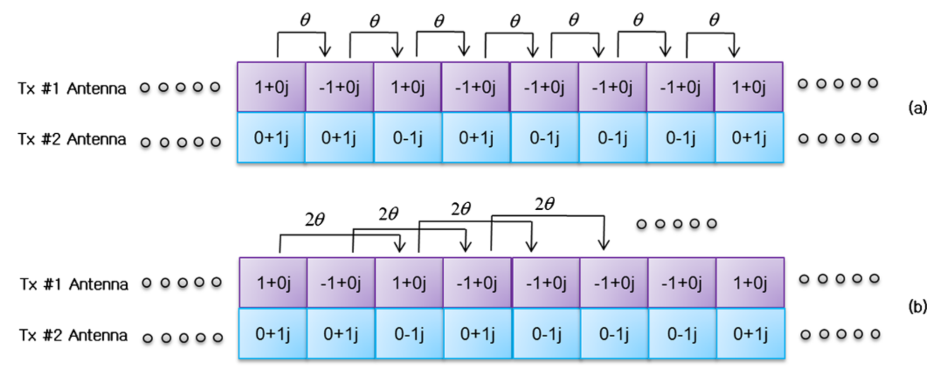

Figure 5 Frequency offset estimation with PN sequences employing proposed phase rotation technique. (a) CFO estimation with a single auto-correlator. **(b)** CFO estimation with multiple auto-correlators.

$$z_{PN}(k) = y_{PN}(k)PN^{*}_{Tx}(k) = e^{j(2\pi f_c kT_s+\theta)} + n^{'}(k), \quad (7)$$

where $n^{'}(k) = n(k)PN^{*}_{Tx}(k)$.

After removing the modulation of the received $y_{PN}(k)$, the phase difference is calculated from the argument of the multiplication between $z_{\mathrm{PN}}(k)$ and $Z^{*}_{PN}(K-1)$.

$$\begin{aligned}&\arg\left[z_{PN}(k)z^{*}_{PN}(k-1)\right]\\ &\quad=\arg\left[\left(e^{j(2\pi f_c kT_s+\theta)}+n^{'}(k)\right)\left(e^{-j(2\pi f_c(k-1)T_s+\theta)}+n^{'}(k-1)\right)\right]\\ &\quad=\arg\left[e^{j2\pi f_c T_s}+n^{''}(k)\right]=2\pi f_c T_s+n^{'''}(k),\end{aligned} \quad (8)$$

where $n^{''} = e^{j[2\pi f_e kT+\theta]}n^{'}(k-1) + e^{-j[2\pi f_e(k-1)T_2+\theta]}n^{'}(k) + n^{'}(k)n(k-1)$ and $n^{'''} = \arg[n^{''}]$.

In Equation 8, the CFO term $2\pi f_c T_s$ is estimated in the presence of a noise term $n^{'''}$. To reduce the effect of such noise, multiple auto-correlators are employed for CFO estimation in the L&R algorithm. $R(m)$, which is the auto-correlation function of $z_{PN}(k)$, can be written as

$$R(m) = \frac{1}{N_p - m}\sum_{k=m}^{N_p-1} z_{PN}(k)z^{*}_{PN}(k-m), \quad (9)$$

where N_p is the number of PN sequences. Finally, the CFO can be estimated as

$$\hat{\nu} = \frac{1}{\pi(N_R+1)T_s}\arg\left[\sum_{m=1}^{N_R} R(m)\right], \quad (10)$$

where N_R is the number of auto-correlators. The accuracy of the estimated CFO increases while the CFO estimation range decreases as N_R increases [9]. In addition, as N_R increases, the number of $R(m)$ in Figure 6 increases, and the hardware complexity increases.

Figure 7 shows the normalized CFO estimation range of the proposed method, and the N_R value of the L&R method is equal to eight. When N_R is set to eight, the normalized CFO estimation range is from −0.11 to +0.11, as shown in Figure 7. This estimation range is sufficient in general terrestrial digital TV broadcasting environment [13]. If a wider estimation range is required, N_R can be decreased or the M&M method [12], which shows an almost full estimation range, can be employed.

3.2 Frequency offset estimation scheme using CSI

The received kth PN sequence $y_{PN}(k)$ at the receive antenna in the time-varying Rayleigh channel can be written as

$$\begin{aligned}y_{PN}(k) &= [h_{11}PN_1(k) + h_{12}PN_2(k)]e^{j(2\pi f_c \cdot iT_s+\theta)}\\ &\quad + n(k)\\ &= PN_{Tx}(k)e^{j(2\pi f_c \cdot iT_s+\theta)} + n(k),\end{aligned} \quad (11)$$

where h_{11} is the channel impulse response (CIR) from the first transmit antenna to the receive antenna and h_{12} is the CIR from the second transmit antenna to the receive antenna. In the Rayleigh channel, h_{11} and h_{12} are

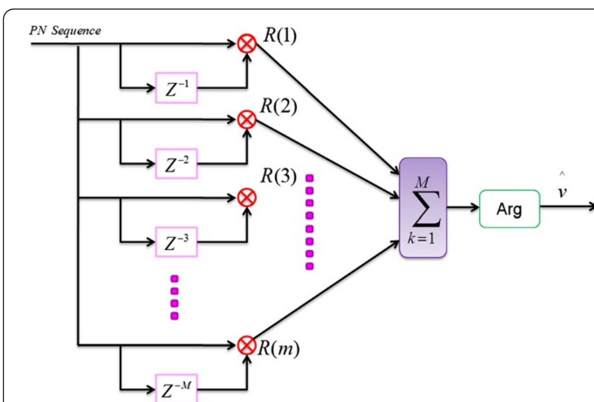

Figure 6 L&R CFO estimation algorithm.

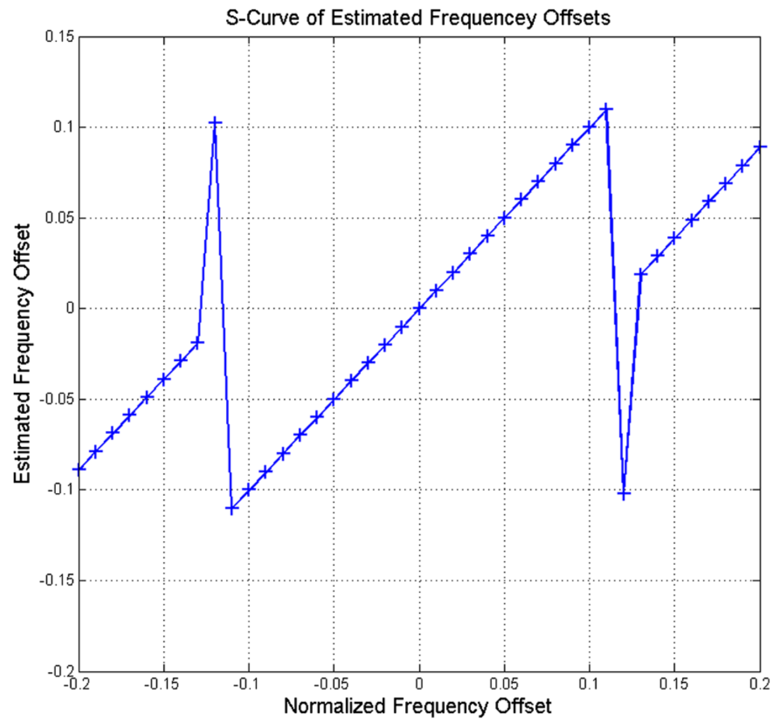

Figure 7 **Normalized frequency offset estimation range of proposed method.**

time varying, and the variation rate depends on the Doppler frequency.

The proposed PN sequence transmission method rotates the phase of the PN sequences to prevent the PN sequence cancellation problem. However, when the absolute phase difference between the time-varying h_{11} and h_{12} is $\pi/2$, the PN sequence cancellation problem occurs again, and the accuracy of the estimated CFO decreases in contrast to that in the AWGN channel. In other words, as the absolute phase difference between h_{11} and h_{12} tends closer to $\pi/2$, the accuracy of the estimated CFO decreases; when it becomes smaller than $\pi/2$, the accuracy increases. Thus, the frequency offset estimation scheme using CSI is proposed, as shown in Figure 8.

The proposed scheme consists of the coarse carrier frequency recovery (CFR), which is a feedback structure, fine CFR, which is a feedforward structure, and phase

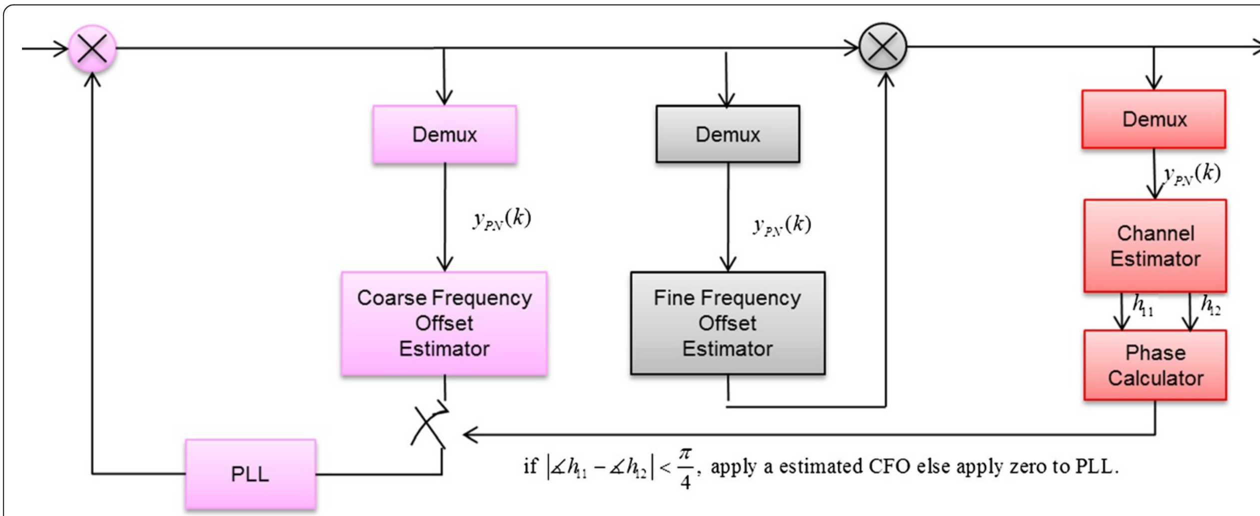

Figure 8 **Proposed carrier frequency offset estimation scheme using CSI.**

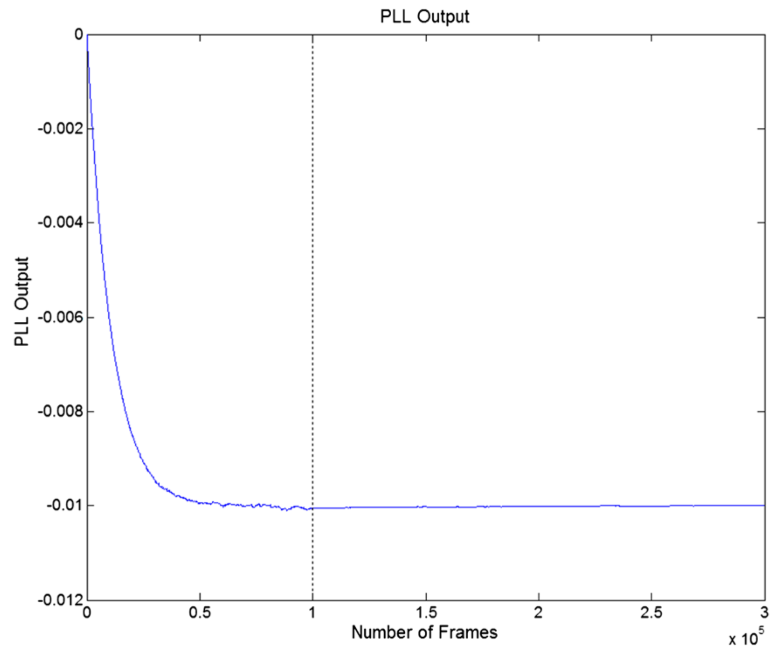

Figure 9 PLL output of coarse CFR using the CSI.

calculator that uses CSI. For the estimation scheme, the proposed PN sequence rotation and the CFO estimation method are employed for the coarse CFR, and the pilot block correlation method in [14] can be employed for the fine CFO estimation algorithm. In the proposed scheme, the fine CFR estimates the CFO and applies it to the compensator using the PN sequence at every frame. On the other hand, the coarse CFR first estimates the CFO at every frame and applies it to the compensator using the CSI. The phase estimator estimates the CSI using PN sequences over one PN block (guard interval) after the CFR and applies the estimated CFO to the phase locked loop (PLL) only when the absolute phase difference between h_{11} and h_{12} is smaller than $\pi/4$. If the absolute phase difference between h_{11} and h_{12} is greater than $\pi/4$, zero is applied to the PLL.

However, the CSI may not be acquired until CFR is roughly achieved. Thus, the proposed scheme controls the PLL using the CSI only when a rough CFR is achieved.

Table 1 Computer simulation parameters

Parameter	Value
Symbol rate f_s	10.76 MHz
Center frequency f_c	476 MHz
Size of PN sequence	201
Size of data symbol	3,780
Carrier frequency offset	1% of symbol rate
Maximum Doppler frequency	17.64 Hz (40 Km/h), 35.28 Hz (80 Km/h), 52.93 Hz (120 Km/h)
Number of auto-correlators	8

Figure 9 shows the PLL output of a coarse CFR. CSI is not used for the CFR when the variance in the PLL output is smaller than 10^{-4} or 100,000 frames are used for the CFR (dotted line). After 100,000 frames are used, the CSI is used for the phase difference calculation. In Figure 9, the PLL converges roughly without using the CSI, and rough CFR can be achieved without using the CSI. In addition, a fine CFO can be recovered after a coarse CFR; we can therefore decide whether to apply an estimated CFO to the PLL with the aid of the CSI after the PLL has roughly converged.

The PN cancellation problem in the conventional method can also be solved by employing the proposed frequency estimation scheme. For the conventional system, if the absolute phase difference between h_{11} and h_{12} over one block is greater than $\pi/4$, the CFO is estimated by employing several auto-correlators and applying them to the PLL. However, the proposed PN sequence transmission method that uses the proposed scheme is suitable for both the AWGN and time-varying Rayleigh channels, whereas the conventional method that uses the proposed scheme only works well in the time-varying channel.

4 Computer simulation results

The MSE performance is measured by applying the proposed PN transmission method and the L&R algorithm over the AWGN and Rayleigh channels. Ideal symbol timing recovery and frame detection situations are assumed, and the simulation parameters are listed in Table 1. The symbol rate and the center frequency of

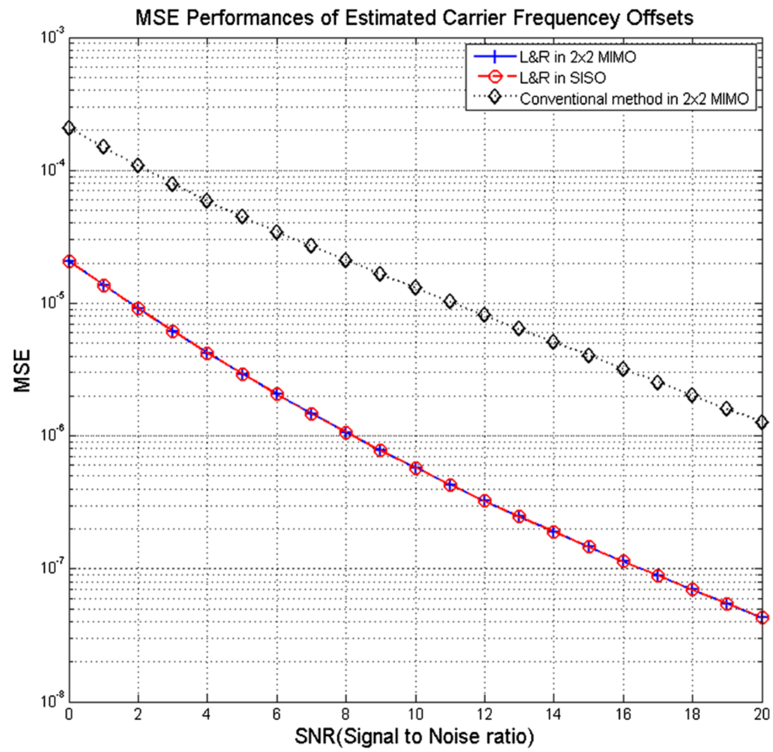

Figure 10 MSE performance in AWGN.

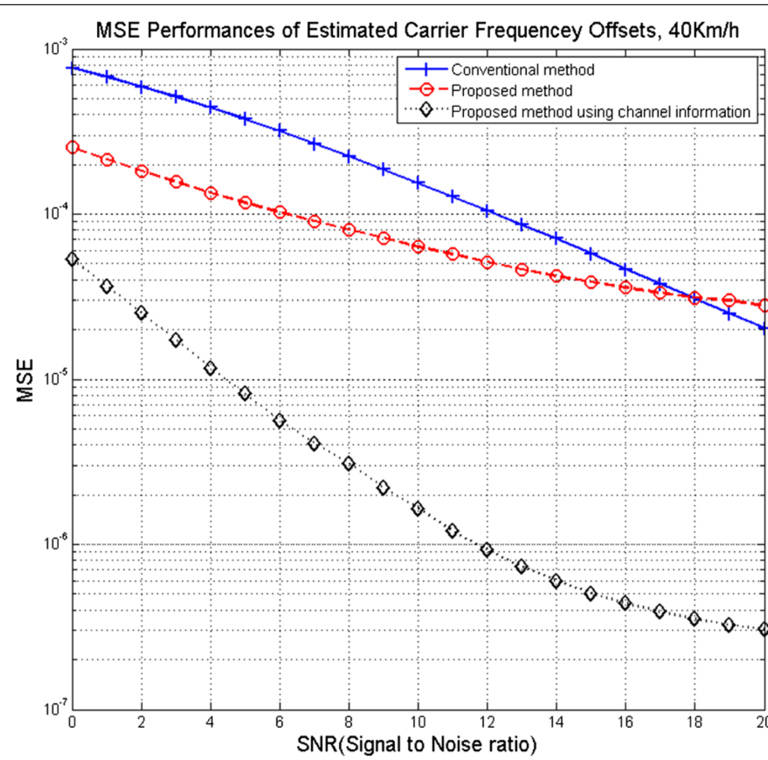

Figure 11 MSE performance over Rayleigh channel when the receiver's velocity is 40 km/h.

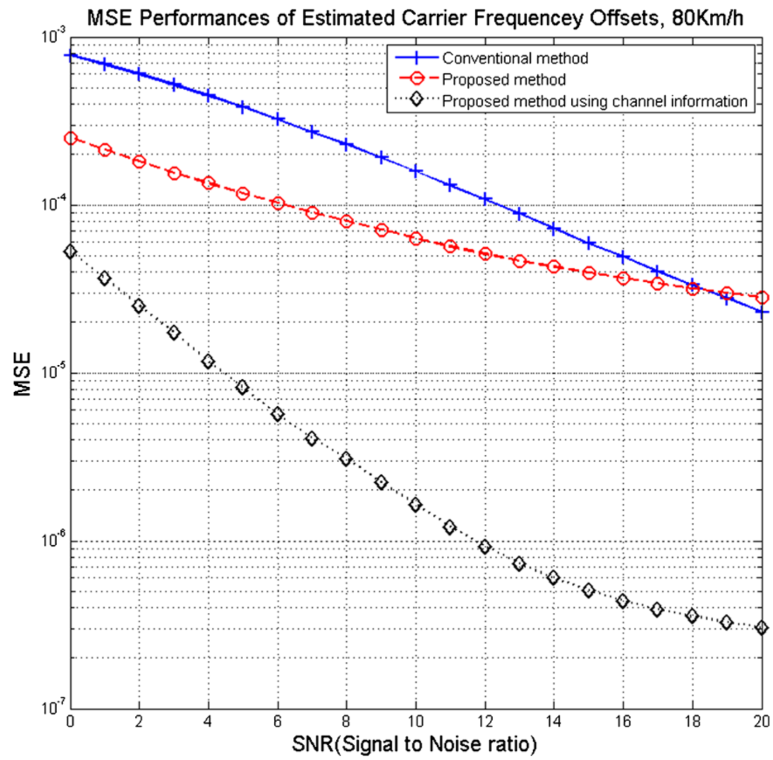

Figure 12 MSE performance over Rayleigh channel when the receiver's velocity is 80 km/h.

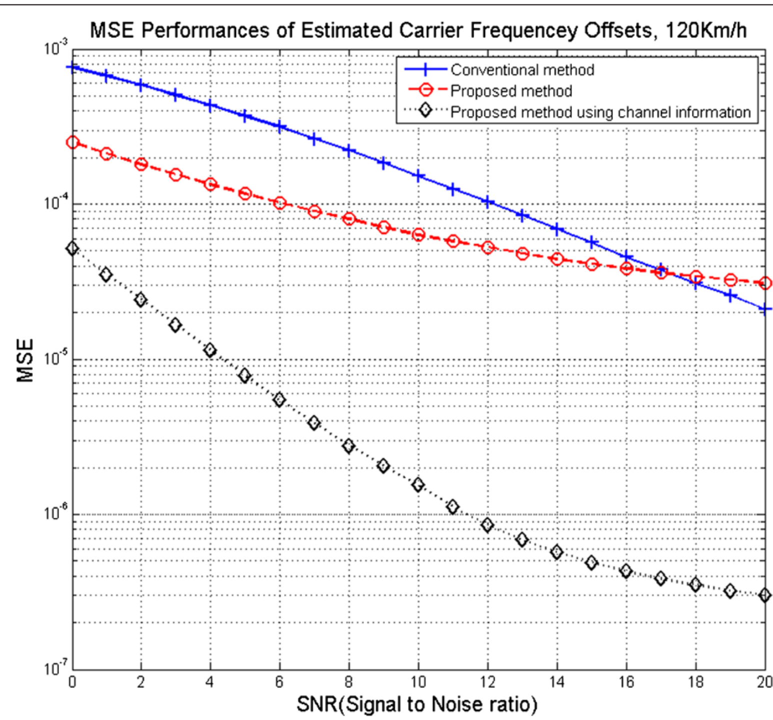

Figure 13 MSE performance over Rayleigh channel when the receiver's velocity is 120 km/h.

the Korean DTV environment are used, and the CFO is set to 1% of the symbol rate [13]. We assume that the distance between the transmitter and the receiver is sufficiently long, and the reception times from each transmit antenna to the receive antenna are the same [10].

Figure 10 shows the MSE performance over the AWGN environment; the MSE performance of the proposed method is higher than that of the conventional method.

The MSE performance over the Rayleigh channel by varying the receiver's velocity from 40 to 120 km/h is shown in Figures 11, 12 and 13. We assume that the CSI is perfectly estimated. In contrast to the performance over the AWGN environment, the more the signal-to -noise ratio becomes higher, the more the MSE of the conventional method becomes smaller as compared with that of the proposed method. Because some of the CFOs are estimated by employing the proposed PN transmission method when the phase difference between h_{11} and h_{12} is closer to $\pi/2$, the accuracy of the estimated CFOs is lower than that of the conventional method. However, the MSE performance is much higher than that of the conventional method when the proposed transmission method and scheme using CSI are employed. In addition, the performance, which depends on the receiver's velocity, is almost the same and shows only a slight difference. Because the CFO is estimated using the phase difference of the consecutive PN sequences for short duration or seven delayed PN sequences for maximum duration, the channel variation over the PN sequences is not very high.

5 Conclusions

In this paper, we have proposed a PN sequence phase rotation transmission method and a frequency estimation scheme using CSI. The proposed PN transmission method rotated the transmitted PN sequences differently from one another to prevent the PN sequence cancellation problem and made possible the use of all consecutive PN sequences for the CFO estimation. In addition, the CFO was accurately estimated by the L&R algorithm, which is a type of DA-ML algorithm with multiple auto-correlators. For the time-varying Rayleigh channel, the frequency estimation scheme using the CSI has also been proposed. The simulation results show that the MSE performance of the proposed transmission method and estimation scheme over the AWGN and Rayleigh channels is higher than that of the conventional method.

Competing interests

The authors declare that they have no competing interests.

Authors' contributions

The work presented here was carried out in collaboration with all the authors. JGO, JH and JTK defined the research theme. JGO designed methods and experiments, carried out the laboratory experiments, analyzed the data, interpreted the results and wrote the paper. JH and JTK co-designed the dispersal and colonization experiments and co-worked on associated data collection and their interpretation. JGO and JTK co-designed experiments, discussed analyses, interpretation and presentation. All authors read and approved the final manuscript.

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