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Successive Interference Cancellation-Based Weighted Least-Squares Estimation of Carrier and Sampling Frequency Offsets for WLANs

XIAOPING ZHOU[1], KAN ZHENG[1], (Senior Member, IEEE), BIN WU[2], AND HUI ZHAO[1], (Member, IEEE)

[1]School of Information and Communication Engineering, Beijing University of Posts and Telecommunications, Beijing 100876, China
[2]Institute of Microelectronics of the Chinese Academy of Science, Beijing 100029, China

Corresponding author: Xiaoping Zhou (zhouxiaoping@ime.ac.cn)

ABSTRACT A successive interference cancellation (SIC)-based weighted least-squares (WLS) estimation for the carrier frequency offset (CFO) and the sampling frequency offset (SFO) is presented for wireless local area networks (WLANs) based on orthogonal frequency division multiplexing (OFDM). The proposed SIC-based WLS performs the estimation by exploiting the phase rotation in the frequency domain caused by the CFO and the SFO. SIC-based WLS estimates the CFO and the SFO successively instead of by traditional simultaneous estimation. The estimation of CFO based on the Taylor series is performed first, and then the WLS estimation of SFO based on successive cancellation of the CFO is carried out. The simulation results show that the SIC-based WLS can estimate the CFO and the SFO effectively. Compared to the WLS scheme, a performance improvement of more than 0.6 dB is achieved by SIC-based WLS, and nearly 10 percent of the complexity is reduced.

INDEX TERMS Carrier frequency offset, sampling frequency offset, phase tracking and weighted least squares.

I. INTRODUCTION

Orthogonal frequency division multiplexing (OFDM) achieves high spectral efficiency and offers an effective solution in overcoming frequency selectivity [1], [2]; it has been widely used in WLAN systems [3].

Synchronizer error in OFDM systems, such as the carrier frequency offset (CFO) and the sampling frequency offset (SFO), can lead to intolerable performance loss [4]–[6]. A number of CFO and SFO estimation algorithms have been studied throughout the years. These studied algorithms can be classified into two types: blind algorithms that do not use pilot symbols [7]–[9] and data-aided (DA) [10]–[19] algorithms using pilot symbols. Because of their simple form and computational convenience, DA methods have received more attention and are considered in this paper. DA schemes perform estimations based on the phase rotation in the frequency domain, which consists of the common phase error (CPE) and sample timing error (STE) [18], [19] caused by the CFO and the SFO, respectively.

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A commonly used DA estimation method is the maximum likelihood (ML) approach [11]–[14], which can achieve optimal performance. [11] presented an ML algorithm for the CFO and SFO, which required a two-dimensional exhaustive search. To reduce the complexity, [12] derived an estimation of the CFO in closed form with a one-dimensional search. Reference [13] proposed a low-complexity ML method for receivers with a single source. ML estimation requires at least a one-dimensional exhaustive search and is not suitable for hardware implementation. Therefore, many DA suboptimal linear algorithms have been suggested [15]–[19]. A simple joint CFO and SFO estimation was presented in [15], where the value of the CFO was estimated for every symbol, but the SFO was derived from the early estimation of the CFO, resulting in a residual SFO and performance decrease. Dantas [16] presented a pilot frequency index-related SFO estimation method that neglected the different levels of fading of the pilot subcarriers and obtained a limited performance gain. Reference [17] enabled robust estimation using a cyclic delay and pilot pattern to maximize the channel power. Reference [18] proposed a pilot frequency index and fading-level-related weighted least squares (WLS) to achieve good performance. On the basis of WLS, least squares (LS) was to simplify the implementation in [19]. WLS and LS are widely used in real WLAN chips.

In our previous studies, simplified WLS and SFO fitting were proposed in [20] and [21], respectively. The simplified WLS and SFO fitting were based on existing WLS and followed the traditional calculation mechanism, where the CFO and the SFO were calculated simultaneously based on the phase difference between the received signal and the expected pilot. For simplified WLS, a combined pilot scheme was proposed to reduce the complexity of SFO estimation. SFO fitting among consecutive symbols was used to improve the accuracy of SFO estimation, which could work with the proposed method in this paper. Traditionally, simultaneous estimations, such as the widely used WLS and LS, take into account the CFO in calculating the estimation matrix of the SFO and result in a suboptimal weighting coefficient for the SFO.

Since only limited pilot subcarriers are embedded in the WLAN system, this paper focuses on improving the accuracy of linear CFO and SFO estimation. A successive interference cancellation (SIC)-based WLS estimation method is presented here, which changes the traditional simultaneous estimation mechanism. The proposed SIC-based WLS estimates the CFO and the SFO separately; it first performs the estimation of CFO based on the Taylor series and then carries out the WLS estimation of SFO based on the successive cancellation of CFO. The estimation of CFO based on the Taylor series can decrease the information loss introduced by the arctangent and improve the estimation accuracy of the CFO. After successive cancellation of the CFO, the estimation matrix of the SFO can be optimized, which can enhance the accuracy of the SFO and achieve a better packet error rate (PER). Since the CFO and SFO are estimated separately, the calculation matrix of the SFO can be simplified. Based on the simulation, the scheme is verified in the IEEE 802.11ac system.

The remainder of this paper is formulated as follows: Section II reviews the system model and existing WLS algorithm. The proposed SIC-based WLS scheme is described in section III. The results of the complexity comparison and simulation comparison used to verify the scheme are presented in section IV. Finally, the conclusion is derived in section V.

II. SYSTEM MODEL

A. SYSTEM MODEL

The system model of the CFO and SFO is reviewed in this section. The transmitted baseband signal is written as follows [4]:

$$x(t) = \frac{1}{\sqrt{T_u}} \sum_{l=-\infty}^{+\infty} \sum_{k=-K/2}^{K/2} s_{l,k} \psi_{l,k}(t) \tag{1}$$

where T_u denotes the time of the useful symbol, l is the time index, K is the number of subcarriers used for each OFDM symbol, $s_{l,k}$ denotes the data transmitted with subcarrier k during the lth symbol time and $\psi_{l,k}(t)$ is the subcarrier pulse. Assuming the duration of the cyclic prefix (CP) is T_g, the subcarrier pulse $\psi_{l,k}(t)$ is

$$\psi_{l,k}(t) = e^{(j2\pi k(t-T_g-lT_s)/T_u)} u(t-lT_s) \tag{2}$$

$$u(t) = \begin{cases} 1, & 0 \le t < T_s \\ 0, & else \end{cases} \tag{3}$$

where $e^{(\cdot)}$ represents an e-exponential function, $T_s = T_u + T_g = N_s T$ represents the symbol duration and $u(t)$ is a rectangular pulse of length T_s; $N_s = N + N_g$ is the sample number in each symbol, T is the sampling interval, N is the size of the inverse fast Fourier transform (IFFT) and N_g is the size of the CP.

After the transmitted baseband signal is corrupted by multipath fading, we obtain the received signal

$$y(t) = \sum_i h_i(t) \cdot x(t-\tau_i) + \eta(t) \tag{4}$$

where $h_i(t)$ denotes the ith channel gain with delay τ_i and $\eta(t)$ denotes the additive white Gaussian noise (AWGN).

Assuming the CFO and the SFO are Δf and ζT, the nth sample of the lth received symbol can be represented as

$$\begin{aligned} y_{l,n} &= y(n'T') = e^{(j2\pi\Delta f(n'T'))} \sum_i h_i(n'T') \cdot x(n'T'-\tau_i) \\ &\quad + \eta(n'T') \\ &= e^{(j2\pi\Delta f n'T(1+\zeta))} \\ &\quad \times \left(\sum_i h_i \sum_l \sum_k s_{l,k} \psi_{l,k}(n'T(1+\zeta)-\tau_i) \right) \\ &\quad + \eta(n'T') \end{aligned} \tag{5}$$

with $n' = n + N_g + lN_s$ is the sample index, $T' = (1+\zeta)T$ is the sampling period at the receiver and $\zeta = (T'-T)/T$ is the relative SFO.

Assuming no intersymbol interference (ISI) exists, the signal demodulated through the fast Fourier transform (FFT) in the frequency domain is given by (6):

$$\begin{aligned} z_{l,k} &= e^{(j\pi\phi_k)} e^{(j2\pi((lN_s+N_g)/N)\phi_k)} Sa(\pi\phi_k) s_{l,k} H_k \\ &\quad + \underbrace{\sum_{i,i\neq k} \left(e^{(j\pi\phi_{i,k})} e^{(j2\pi((lN_s+N_g)/N)\phi_i)} \right) Sa(\pi\phi_{i,k}) s_{l,i} H_i}_{ICI} + \eta_{l,k} \end{aligned} \tag{6}$$

where $\phi_{i,k} = (1+\zeta)(\Delta f T_u + i) - k$ denotes the cross-subcarrier, $\phi_k = \phi_{k,k} \approx \Delta f T_u + \zeta k$ denotes the subcarrier offset parameters, $Sa(\pi\phi_k) = \sin(\pi\phi_k)/(\pi\phi_k)$ and $Sa(\pi\phi_{i,k}) = \sin(\pi\phi_{i,k})/(\pi\phi_{i,k})$ denote the magnitude attenuation factors and H_k denotes the frequency channel response of the kth subcarrier.

Assuming Δf and ζ are small enough, $Sa(\pi\phi_k)$ is close to 1 and $Sa(\pi\phi_{i,k})$ is close to 0, the intercarrier interference (ICI) can be ignored. The ICI can be modeled as additional noise $\eta^{\Omega}_{l,k}$, and the signal in (6) is updated to (7).

$$z_{l,k} = e^{(j\pi(1+2N_g/N)\phi_k)} e^{(j2\pi lN_s/N\phi_k)} Sa(\pi\phi_k)\, s_{l,k} H_k + \eta^{\Omega}_{l,k} + \eta_{l,k} \quad (7)$$

As $e^{(j\pi(1+2N_g/N)\phi_k)}$ is time-invariant and is treated as H_k, the phase rotation can be rewritten as (8).

$$\phi_l(k) = 2\pi lN_s/N\phi_k = \underbrace{2\pi lN_s/NT_u\Delta f}_{c_l} + \underbrace{2\pi lN_s/N\zeta k}_{l\delta_l k} = c_l + \delta_l lk \quad (8)$$

The demodulated signal $z_{l,k}$ is as follows:

$$z_{l,k} = e^{(j\phi_l(k))} H_k s_{l,k} + \eta'_{l,k} = e^{(j(c_l+\delta_l lk))} H_k s_{l,k} + \eta'_{l,k} \quad (9)$$

where c_l and δ_l denote the CPE and the STE, respectively, and $\eta'_{l,k} = \eta^{\Omega}_{l,k} + \eta_{l,k}$ denotes the total noise.

B. EXISTING ALGORITHMS

In this section, we take IEEE 802.11ac system to review the existing WLS and LS algorithms. Assuming that V pilots are inserted among N subcarriers, and the frequency indexes of the pilots are $p_v, v = 0, 1, \cdots, V-1$. There are four, six and eight pilots for 20M, 40M and 80M bandwidth respectively. Let $\mathbf{p}_l = \left[p_{l,p_0}\ p_{l,p_1} \cdots p_{l,p_{V-1}}\right]^T$ be the $V \times 1$ pilot vector. $\mathbf{Z}_l = \left[z_{l,p_0}\ z_{l,p_1} \cdots z_{l,p_{V-1}}\right]^T$ and $\mathbf{h}_P = \left[H_{p_0}\ H_{p_1} \cdots H_{p_{V-1}}\right]^T$ denote the $V \times 1$ data vector and channel vector corresponding to the pilots, respectively. Then,

$$\mathbf{Z}_l = \mathbf{P}_l \bar{\mathbf{H}}_P \boldsymbol{\varphi}_l + \boldsymbol{\eta}_l \quad (10)$$

where $(\cdot)^T$ is the transpose of the matrix, $\bar{\mathbf{H}}_P = diag(\mathrm{h}_P)$, $\mathbf{P}_l = diag(\mathrm{p}_l)$, $\boldsymbol{\varphi}_l = \left[e^{(j\phi_l(p_0))}\ e^{(j\phi_l(p_1))} \ldots e^{(j\phi_l(p_{V-1}))}\right]^T$ and $\boldsymbol{\eta}_l$ denotes the noise vector.

To overcome the periodic rotation of the phase, the estimation is based on the phase differences of two consecutive OFDM symbols. Let $\hat{\boldsymbol{\varphi}}'_l$ denote the correlation vector of two consecutive symbols.

$$\hat{\boldsymbol{\varphi}}'_l = \mathbf{A}_l (\mathbf{Z}_{l-1})^* \mathbf{Z}_l = \boldsymbol{\varphi}'_l + \mathbf{e}_l \quad (11)$$

where $()^*$ denotes the conjugate, $\mathbf{A}_l = \left[a_{l,p_0}\ a_{l,p_1} \cdots a_{l,p_{V-1}}\right]^T$, $a_{l,p_V} \in \{-1\ 1\}$ is the differential value of the pilots encoded with the pseudo-noise sequence, and $\mathbf{e}_l$ is the error introduced by $\boldsymbol{\eta}_l$. Additionally, note that

$$\boldsymbol{\varphi}'_l = \begin{bmatrix} e^{(j\phi'_l(p_0))} \\ e^{(j\phi'_l(p_1))} \\ \vdots \\ e^{(j\phi'_l(p_{V-1}))} \end{bmatrix} = \begin{bmatrix} e^{(j(\Delta c_l + p_0\delta_l))} \\ e^{(j(\Delta c_l + p_1\delta_l))} \\ \vdots \\ e^{(j(\Delta c_l + p_{V-1}\delta_l))} \end{bmatrix} \quad (12)$$

with $\Delta c_l = c_l - c_{l-1}$.

From (11) and (12), we know that the estimated phase differences of the pilot subcarriers between two symbols have the form of (13).

$$\angle\hat{\boldsymbol{\varphi}}'_l = \begin{bmatrix} 1 & p_0 \\ 1 & p_1 \\ \vdots & \vdots \\ 1 & p_{V-1} \end{bmatrix} \begin{bmatrix} \Delta c_l \\ \delta_l \end{bmatrix} + \mathbf{e}^{ang}_l = \mathbf{D} \begin{bmatrix} \Delta c_l \\ \delta_l \end{bmatrix} + \mathbf{e}^{ang}_l \quad (13)$$

where $\mathbf{e}^{ang}_l$ is the error of angle introduced by $\mathbf{e}_l$.

According to [20], the WLS estimation of Δc_l and δ_l can be derived as (14).

$$\begin{bmatrix} \Delta\hat{c}_l \\ \hat{\delta}_l \end{bmatrix} = \left(\mathbf{D}^T\mathbf{W}\mathbf{D}\right)^{-1} \mathbf{D}^T\mathbf{W}\left(\angle\hat{\boldsymbol{\varphi}}'_l\right) \quad (14)$$

where $(\cdot)^{-1}$ is the inverse of the matrix, $\mathbf{W} = diag(w_{p_v})$ is the weighted matrix and w_{p_v} is shown in (15).

$$w_{p_v} = \frac{E_s\left\|H_{p_v}\right\|^2}{\sigma^2_\eta} = \frac{E_s\left(\Re\left(H_{p_v}\right)^2 + \Im\left(H_{p_v}\right)^2\right)}{\sigma^2_\eta} \quad (15)$$

where E_s is the energy of the pilot subcarriers, H_{p_v} denotes the frequency channel response of the $p_v th$ pilot subcarriers, σ^2_η is the variance of the noise, $\Re(\cdot)$ is the real part and $\Im(\cdot)$ is the imaginary part.

Since E_s/σ^2_η is uncharged in a packet transmission for WLANs, w_{p_v} is equivalent to (16).

$$w_{p_v} = \Re\left(H_{p_v}\right)^2 + \Im\left(H_{p_v}\right)^2 \quad (16)$$

In [21], LS is provided to simplify the estimation of Δc_l and δ_l by setting $\mathbf{W} = \mathbf{I}_V$, where $\mathbf{I}_V$ is a $V \times V$ identity matrix.

$$\begin{bmatrix} \Delta\hat{c}_l \\ \hat{\delta}_l \end{bmatrix} = \left(\mathbf{D}^T\mathbf{D}\right)^{-1} \mathbf{D}^T\left(\angle\hat{\boldsymbol{\varphi}}'_l\right) \quad (17)$$

III. PROPOSED ALGORITHM

A. ESTIMATION OF THE CFO BASED ON THE TAYLOR SERIES

From [11], the ML estimation of Δc_l and δ_l is given as (18)

$$\begin{aligned}
&\left(\Delta\hat{c}_l, \hat{\delta}_l\right) \\
&= \arg\min \sum_{v=0}^{V-1} \left\| a_{l,p_v} z_{l,p_v} - \left(e^{(j((c_l - c_{l-1}) + p_v\delta_l))} z_{l-1,p_v}\right)\right\|^2 \\
&= \arg\min \sum_{v=0}^{V-1} \left\| a_{l,p_v} z_{l,p_v} - \left(e^{(j(\Delta c_l + p_v\delta_l))} z_{l-1,p_v}\right)\right\|^2 \\
&= \arg\min \sum_{v=0}^{V-1} \left[\left\|z_{l,p_v}\right\|^2 + \left\|z_{l-1,p_v}\right\|^2 \right. \\
&\quad \left. -2a_{l,p_v}\Re\left(\left(e^{(j(\Delta c_l + p_v\delta_l))} z_{l-1,p_v}\right)^* z_{l,p_v}\right)\right] \quad (18)
\end{aligned}$$

As the value of $\left\|z_{l,p_v}\right\|^2$ and $\left\|z_{l-1,p_v}\right\|^2$ is independent of $\Delta\hat{c}_l$ and $\hat{\delta}_l$, (18) becomes

$$\begin{aligned}&\left(\Delta\hat{c}_l,\hat{\delta}_l\right)\\&=\arg\min\sum_{v=0}^{V-1}\left[-2\Re\left(a_{l,p_v}\left(e^{(j(\Delta c_l+p_v\delta_l))}z_{l-1,p_v}\right)^*z_{l,p_v}\right)\right]\\&=\arg\max\sum_{v=0}^{V-1}\left[2a_{l,p_v}\Re\left(e^{(-j(\Delta c_l+p_v\delta_l))}\left(z_{l-1,p_v}\right)^*z_{l,p_v}\right)\right]\end{aligned}\tag{19}$$

In WLANs, δ_l meets the condition

$$|\delta_l|=|2\pi N_s/N\zeta|\leq 0.000314\tag{20}$$

where the maximum value of N_s/N is 1.25 for a long CP and the absolute value of ζ is no more than 40 ppm.

Based on the Taylor series expansion of the exponential e^x, (19) can be linearized.

$$\begin{aligned}&\left(\Delta\hat{c}_l,\hat{\delta}_l\right)\\&=\arg\max\sum_{v=0}^{V-1}\left[2a_{l,p_v}\Re\left(e^{(-j\Delta c_l)}e^{(-jp_v\delta_l)}\left(z_{l-1,p_v}\right)^*z_{l,p_v}\right)\right]\\&=\arg\max\sum_{v=0}^{V-1}\left[2a_{l,p_v}\Re\left(e^{(-j\Delta c_l)}\left(1-p_v\delta_l\right)\left(z_{l-1,p_v}\right)^*z_{l,p_v}\right)\right]\\&\approx\arg\max\sum_{v=0}^{V-1}\left[2e^{(-j\Delta c_l)}a_{l,p_v}\Re\left(\left(z_{l-1,p_v}\right)^*z_{l,p_v}\right)\right]\end{aligned}\tag{21}$$

From (21), we can obtain the estimation of Δc_l.

$$\Delta\hat{c}_l=\angle\sum_{v=0}^{V-1}a_{l,p_v}\left(z_{l-1,p_v}\right)^*z_{l,p_v}\tag{22}$$

B. WLS ESTIMATION OF THE SFO BASED ON SUCCESSIVE CANCELLATION OF THE CFO

Equation (11) can be expanded as (23).

$$\hat{\varphi}'_{l,p_v}=a_{l,p_v}\left(z_{l-1,p_v}\right)^*z_{l,p_v}\tag{23}$$

From (12) and (23), we have

$$\angle\hat{\varphi}'_{l,p_v}=\angle\left(a_{l,p_v}\left(z_{l-1,p_v}\right)^*z_{l,p_v}\right)=\Delta c_l+p_v\delta_l+e^{ang}_{l,p_v}\tag{24}$$

where e^{ang}_{l,p_v} is the angle error for pilot p_v.

Once the estimation of Δc_l is obtained, we can remove it from $\angle\hat{\varphi}'_{l,p_v}$ and obtain $\hat{\phi}^{\delta}_{l,p_v}$.

$$\hat{\phi}^{\delta}_{l,p_v}=\angle\hat{\varphi}_{l,p_v}-\Delta\hat{c}_l=p_v\delta_l+e^{ang}_{l,p_v}+e^{\mathrm{CFO}}_{l,p_v}=p_v\delta_l+e'_{l,p_v}\tag{25}$$

where e^{CFO}_{l,p_v} is the estimation error of Δc_l and $e'_{l,p_v}=e^{ang}_{l,p_v}+e^{\mathrm{CFO}}_{l,p_v}$ is the total error.

By stacking (25) for $v=0,1,\cdots,V-1$ and expressing the equations in vector form, we obtain $\angle\hat{\boldsymbol{\varphi}}^{\delta}_l$:

$$\angle\hat{\boldsymbol{\varphi}}^{\delta}_l=\begin{bmatrix}\hat{\phi}^{\delta}_{l,p_0}\\\hat{\phi}^{\delta}_{l,p_1}\\\vdots\\\hat{\phi}^{\delta}_{l,p_{V-1}}\end{bmatrix}=\begin{bmatrix}p_0\\p_1\\\vdots\\p_{V-1}\end{bmatrix}\delta_l+\begin{bmatrix}e'_{l,p_0}\\e'_{l,p_1}\\\vdots\\e'_{l,p_{V-1}}\end{bmatrix}=\mathbf{D}_f\delta_l+\mathbf{e}'_{l,p}\tag{26}$$

where $\mathbf{D}_f=\left[p_0\ p_1\ \cdots\ p_{V-1}\right]^T$ and $\mathbf{e}'_{l,p}=[e'_{l,p_0}\ e'_{l,p_1}\ \cdots\ e'_{l,p_{V-1}}]^T$.

Then, the WLS estimation of $\hat{\delta}_l$ based on successive cancellation of the CFO can be calculated as

$$\hat{\delta}_l=\left(\mathbf{D}_f^T\mathbf{W}\mathbf{D}_f\right)^{-1}\mathbf{D}_f^T\mathbf{W}\left(\angle\hat{\boldsymbol{\varphi}}^{\delta}_l\right)\tag{27}$$

where $\mathbf{W}=diag\left(w_{p_v}\right)$ is the combining coefficient, which is derived from the related channel information and is calculated as (15).

The proposed SIC-based WLS estimation of CFO and SFO is shown in Table 1.

TABLE 1. Proposed algorithm of SIC-based WLS.

Algorithm: SIC-based WLS
Input:
pilot vector $\mathbf{p}_l=\left[p_{l,p_0}\quad p_{l,p_1}\quad\cdots\quad p_{l,p_{V-1}}\right]^T$
received pilot data vector $\mathbf{Z}_{l-1}=\left[z_{l-1,p_0}\quad z_{l-1,p_1}\quad\cdots\quad z_{l-1,p_{V-1}}\right]^T$
received pilot data vector $\mathbf{Z}_l=\left[z_{l,p_0}\quad z_{l,p_1}\quad\cdots\quad z_{l,p_{V-1}}\right]^T$
channel response vector $\mathbf{h}_p=\left[H_{p_0}\quad H_{p_1}\quad\cdots\quad H_{p_{V-1}}\right]^T$
Estimation:
$\Delta\hat{c}_l$: use (22) to perform the estimation
$\hat{\delta}_l$: use (25) and (26) to obtain $\angle\hat{\boldsymbol{\varphi}}^{\delta}_l$, and use (27) to perform the estimation
Output:
$\Delta\hat{c}_l$ and $\hat{\delta}_l$

IV. SIMULATION RESULTS AND ANALYSIS

A. COMPLEXITY COMPARISON

We present a comparison of the complexity of different algorithms, which is shown in Table 2.

To demonstrate the complexity effect of the estimation of CFO based on the Taylor series and the WLS estimation of SFO based on successive cancellation of the CFO, WLS and LS with the estimation of CFO based on Taylor series only are provided for comparison, and they are labeled "WLS_CFO" and "LS_CFO", respectively. The proposed SIC-based WLS is labeled "SIC_WLS".

As illustrated in Table 2, ML is the most complex method, and it requires hundreds of searches. Compared to ML, LS and WLS are simple. The simplest method is LS, which assumes that the pilot subcarriers suffer from the same

TABLE 2. Complexity comparison for different related methods.

complexity	ML	WLS	WLS_ CFO	LS	LS_ CFO	SIC_ WLS
addition	$(6V$-$1)s_{ml}$	$10V$-4	$10V$-5	$4V$-2	$5V$ -3	$7V$-4
multiplication	$(8V$+$1)s_{ml}$	$10V$+8	$10V$+3	$4.5V$	$4.5V$	$9V$+1
arctangent	0	V	V +1	V	V+1	V +1
exponent	Vs_{ml}	0	0	0	0	0

* s_{ml} is the search time of ML

TABLE 3. PER simulation parameters.

Parameters	Value
Center frequency	5.8 GHz
Bandwidth	20 MHz
MCS	MCS 2, MCS 4 and MCS 7
Packet length	10000 [byte]
Normalized CFO	0.02
Normalized SFO	20 ppm

fading level. WLS requires matrix operations to calculate the weights and is a complex method.

For LS, the estimation of CFO based on Taylor series requires an extra V-1 additions and one arctangent. Compared to WLS, estimation of CFO based on Taylor series uses one arctangent to replace the matrix operations needed for CFO estimation and has removed one addition and five multiplications but added one arctangent. Since a coordination rotation digital computer (CORDIC) [22] is used to finish arctangent calculation, the arctangent calculation is simpler than multiplication. Therefore, WLS with the estimation of CFO based on a Taylor series only has a lower complexity than WLS.

Compared to WLS_CFO, nearly 30 percent of the addition and 10 percent of the multiplication is reduced by SIC_WLS. SIC_WLS has lower complexity than WLS_CFO. Compared to WLS and WLS_CFO, SIC_WLS reduces complexity by more than 10 percent.

B. PERFORMANCE COMPARISON

The root mean square (RMS) and PER performance for SIC-based WLS for the CFO and SFO are presented and compared to ML, WLS, WLS with the estimation of CFO based on Taylor series only, LS and LS with the estimation of CFO based on Taylor series only. 10000 bytes per burst with different modulation and coding scheme (MCS): MCS2, MCS4 and MCS7 are transmitted based on the IEEE 802.11ac standard, where the performance under a frequency-selective fading channel TGac-B [23] is simulated. According to [5], the normalized CFO is within 0.02, and the normalized SFO is 20 ppm in this simulation.

The RMS results for the CFO and SFO obtained by SIC-based WLS ("SIC_WLS"), ML, WLS and LS are shown in Figures 1 and 2, respectively.

As shown in Figure 1, ML achieves the minimum RMS for the CFO. The estimation of the CFO based on the Taylor series has a better RMS than WLS and LS, which benefits from the reduced information loss introduced by arctangent operations.

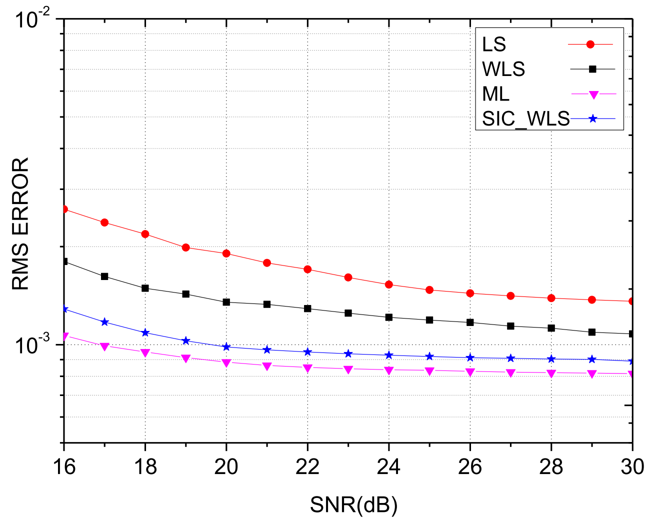

FIGURE 1. CFO RMS errors of different algorithms.

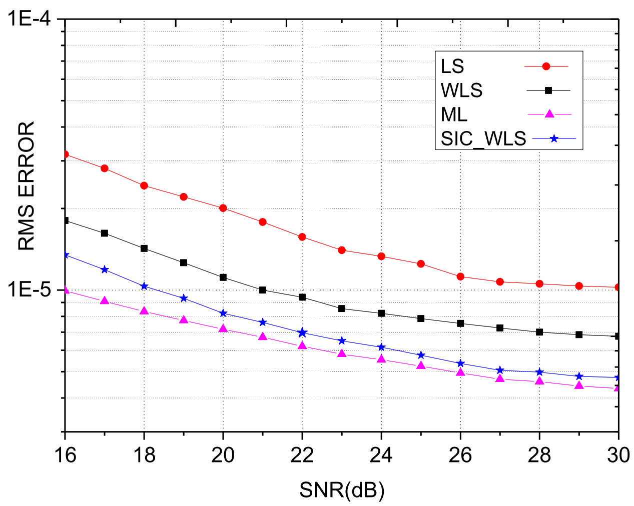

FIGURE 2. SFO RMS errors of different algorithms.

In Figure 2, ML achieves the minimum RMS for the SFO. Compared to WLS and LS, SIC-based WLS improves the estimation accuracy of the SFO, mainly due to the optimized estimation matrix of the SFO. As the accuracy of the CFO improved in high signal to noise ratio (SNR) cases, the performance difference between ML and SIC-based WLS decreased.

The performance comparison of SIC-based WLS, ML, WLS, WLS with the estimation of CFO based on Taylor series only ("WLS_CFO"), LS and LS with the estimation of CFO based on Taylor series only ("LS_CFO") for MCS2, MCS4 and MCS7 are shown in Figures 3, 4 and 5, respectively.

In the IEEE 802.11ac system, the PER needs to be lower than 10^{-1}, and the corresponding SNR can be used as a performance indicator.

As shown in Figures 3, 4 and 5, when the PER is 10^{-1}, ML has the best performance in all cases. ML achieves a nearly 0.25 dB better gain than SIC-based WLS, which is a close approach to the ML method over a wide range of SNR values.

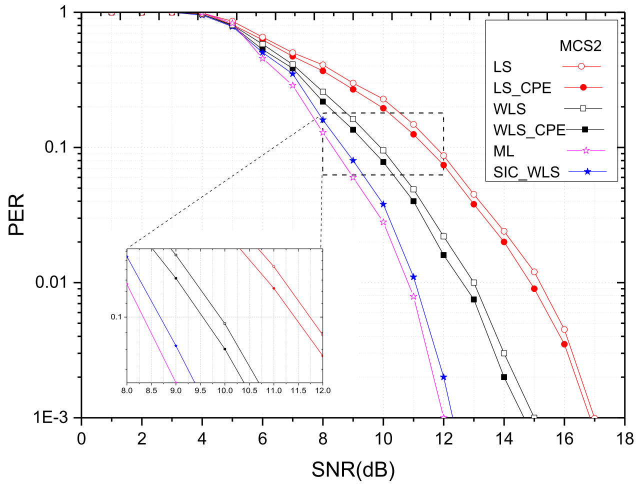

FIGURE 3. PER of MCS2 for different algorithms in TGac-B.

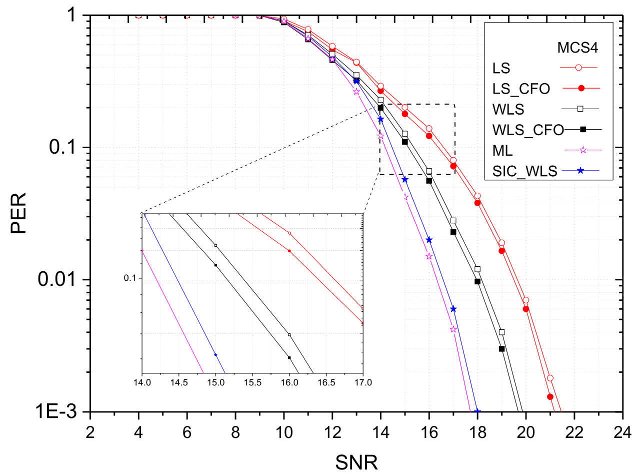

FIGURE 4. PER of MCS4 for different algorithms in TGac-B.

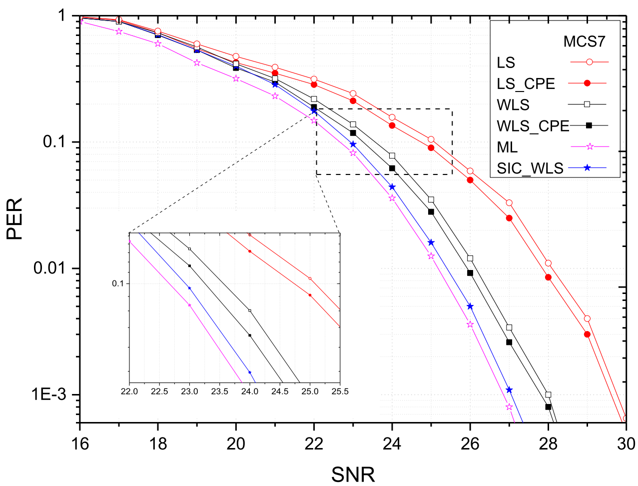

FIGURE 5. PER of MCS7 for different algorithms in TGac-B.

Compared to WLS, ML achieves a minimum performance improvement of 0.9 dB and maximum improvement of 1.6 dB for MCS7 and MCS2, respectively. For LS, the performance gain can reach 2.5 dB and even 3.5 dB for MCS7 and MCS2, respectively.

When the PER is 10^{-1}, compared to WLS and LS, WLS with the estimation of CFO based on Taylor series only and LS with the estimation of CFO based on Taylor series only could obtain a nearly 0.2 dB performance improvement for all cases.

SIC-based WLS achieves a better performance gain than WLS, WLS with the estimation of CFO based on Taylor series only, LS and LS with the estimation of CFO based on Taylor series only in all cases. As shown in Figures 4 and 5, SIC-based WLS can achieve a nearly 0.6 dB performance gain over WLS for MCS4 and MCS7. The performance gain obtained between SIC-based WLS and WLS can reach 1.25 dB for MCS2. When the MCS number decreases enough, the performance gain achieved by SIC-based WLS increases, mainly because the number of OFDM symbols is much larger and it is more sensitive to the influence of the SFO.

V. CONCLUSION

A SIC-based WLS estimation method for the CFO and SFO is proposed in this paper, in which the CFO and SFO are estimated separately, rather than simultaneously as in traditional methods. The method of obtaining the CFO based on the Taylor series is to estimate the CFO first, after which the WLS estimation of the SFO-based successive cancellation of the CFO is carried out, which can effectively improve the accuracy of the SFO estimation. Based on simulations, this scheme is verified in the IEEE 802.11ac system. The SIC-based WLS is a close approach to ML but uses a simple computation method. Compared to WLS, SIC-based WLS achieved a minimum 0.6 dB performance improvement, while 10 percent of the complexity was reduced. In the future, we will focus on the estimation of the CFO and SFO in MIMO systems.

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XIAOPING ZHOU graduated from the Beijing University of Posts and Telecommunications (BUPT), China, in 2011. She is currently pursuing the Ph.D. degree with BUPT. Her research interests include 5G wireless communication systems and VLSI design.

KAN ZHENG (Senior Member, IEEE) received the B.S., M.S., and Ph.D. degrees from the Beijing University of Posts and Telecommunications (BUPT), China, in 1996, 2000, and 2005, respectively. He is currently a Full Professor with BUPT. He has rich experiences on the research and standardization of new emerging technologies. He has authored over 200 journal articles and conference papers in the field of wireless networks, the Internet-of-Things, and so on.

BIN WU received the M.Sc. degree in electronics from Chongqing University, Chongqing, China, in 2002, and the Ph.D. degree in electronics from the Institute of Microelectronics of Chinese Academy of Science (IMECAS), China, in 2011. He is currently a Professor with IMECAS. His current research interests include wireless communication systems and VLSI design.

HUI ZHAO (Member, IEEE) received the M.S. degree from Tianjin University, in 2003, and the Ph.D. degree from the Beijing University of Posts and Telecommunications (BUPT), China, in 2006. She is currently an Associate Professor with BUPT. She has accumulated rich experience in the field of LTE-advanced, ultra-high throughput WLAN, and 5G.