

**APPARATUS AND METHOD FOR ESTIMATION OF COMMON PHASE
ERROR IN A MIMO-OFDM SYSTEM**

5 The present invention relates to an apparatus and method for estimation of a
common phase error (CPE) component of a wireless communications signal in
a multiple input multiple output orthogonal frequency division multiplexing
(MIMO-OFDM) system. In particular, the invention relates to providing an
estimate of the common phase error component from an initial common phase
10 error estimate.

MIMO-OFDM technology is extensively employed in IEEE802.11n based high-
speed Wireless Local Access Networks (WLAN), mobile systems with
Worldwide Interoperability for Microwave Access (WiMAX), and Digital Video
15 Broadcasting (DVB) systems. It is well known that OFDM is robust in frequency
selective fading channels with a relatively low implementation cost and high
bandwidth efficiency. Moreover, MIMO takes advantage of multipath
propagation to increase throughput, range/coverage, and reliability. By
transmitting multiple signals containing different information streams over the
20 same frequency channel, MIMO provides a means of doubling or tripling
spectral efficiency. The combination of MIMO and OFDM techniques provides
low-cost implementation particularly for high data rate communications.

However OFDM is known to be vulnerable to synchronisation errors, which can
25 cause inter-carrier interference (ICI) and degrade system performance.
Technical publications discussing this issue include, for example, Michael
Speth, Stefan A. Fechtel, Gunnar Fock, and Heinrich Meyr, "Optimum Receiver
Design for Wireless Broadband Systems Using OFDM—Part I", *IEEE Trans. On
Comm.* Vol. 47, No.11, November 1999, pp. 1668-1677, and Hlaing Minn, Vijay
30 K. Bhargava, and Khaled Ben Letaief, "A Robust Timing and Frequency
Synchronisation for OFDM Systems", *IEEE Trans. On Wireless Comm.*, Vol.2,
No.4, July 2003, pp. 822-839. In MIMO-OFDM transmission systems, the

synchronisation tasks include carrier recovery, phase noise suppression and clock timing recovery.

In MIMO-OFDM based WLAN systems, the processing of the MIMO-OFDM wireless communications signal preamble is sufficient for initial synchronisation of the MIMO-OFDM receiver as discussed in the Speth et al and Minn et al publications noted above. However, solely initial synchronisation may not be sufficient due to residual carrier frequency offset (CFO) and timing clock offset (TO) in the system. Though residual CFO/TO can be very small after coarse CFO compensation, accumulated phase rotation and timing drift in the signal can still deteriorate system performance. Consequently, there is a need for an efficient CFO/TO tracking scheme to compensate dynamically for such phase drift. Furthermore, the system will experience phase noise (PN) caused by a combination of mismatches in RF oscillator and a phase-locked loop (PLL); see, for example, P. Robertson and S. Kaiser, "Analysis of the effects of phase noise in orthogonal frequency division multiplexing (OFDM) systems," in *Proc. ICC'95*, vol.3, Seattle, WA, 1995, pp. 1652-1657. PN and residual CFO cause a CPE for all signal carriers and a Gaussian-like ICI term. Initial CPE estimation and compensation may be performed inherently by the channel estimation/equalisation (detection) modules in a receiver apparatus. However CPE changes on a symbol-by-symbol basis, which may mean this initial correction is insufficient for accurate performance. It is, therefore, necessary to track and correct the rotation of the received constellation points caused by the CFO/PN on a symbol-by-symbol basis.

25

One known method for this is to use signal pilot subcarriers, i.e., subcarriers containing known data, which are inserted into the signal data symbols. Rotation of these known pilot symbols, which are observed at the receiver, is a measure for the CPE. Pilot based-CPE estimation methods for OFDM systems can be categorised as:

30

- mean CPE by averaging over multiple pilots with the same or different weighting factors as discussed in, for example, Onizawa, T., Mizoguchi, M., Sakata, T., and Morikura, M., "A new simple adaptive phase tracking scheme employing phase noise estimation for OFDM signals," *IEEE 55th Vehicular Technology Conference, VTC Spring 2002*, Vol. 3, 6-9 May 2002, pp. 1252 -1256; and, Shou-Yin Liu and Jong-Wha Chong, "A Study of Joint Tracking Algorithms of Carrier Frequency Offset and Sampling Clock Offset for OFDM-based WLANs," *IEEE Proc. Of Communications, Circuits and Systems and West Sino Expositions*, Vol.1, June 29 - July 1, 2002, pp. 109-113.
 - the Least Square (LS) CPE estimator as discussed in, for example, Songping Wu and Yeheskel Bar-Ness, "A Phase Noise Suppression Algorithm for OFDM-Based WLANs", *IEEE Communications Letters*, Vol.6, No.12, December 2002, pp. 535-537; and
 - joint CFO/TO weighted LS (WLS) estimator as discussed in, for example, Pei-yun Tsai, Hsin-Yu Kang and Tzi-Dar Chiueh, "Joint Weighted Least Squares Estimation of Frequency and Timing Offset for OFDM Systems over Fading Channels", *IEEE Trans. On Vehicular Technology*, Vol.54, Issue 1, Jan. 2005, pp. 211-223; and Maja Sliskovic, "Carrier and Sampling Frequency Offset Estimation and Correction in Multicarrier Systems", *IEEE Proc. Of GLOBECOM 2001*, Vol.1, pp. 285-289.
- For MIMO-OFDM receivers, some known works deal with CFO/TO tracking as a simple extension of synchronisation tracking from OFDM receivers. Examples of such publications are Seok Ho Won, Deuk-Su Lyu, and Hyeong Jun Park, "Physical Layer Implementation and Evaluation of Multiple Input Multiple Output—Orthogonal Frequency Division Multiplexing (MIMO-OFDM) system", *IEEE Proc. Of International Conference on Communication Technology, ICCT 2003*, Vol. 2, April 9-11, pp. 1348-1352; Chi-Yeh Yu, Zih-Yin Ding, and Tzi-Dar Chiueh, "Design and Simulation of a MIMO OFDM Baseband Transceiver for

High Throughput Wireless LAN", *IEEE Proc. Of 2004 Asia-Pacific Conference on Circuit and Systems*, Dec. 6-9, 2004, pp. 205-208; and Allert van Zelst, and Tim C.W.Schenk, "Implementation of a MIMO OFDM-Based Wireless LAN System", *IEEE Trans. On Signal Processing*, Vol. 52, No. 2, February 2004, pp.
 5 483-494.

The invention is defined in the independent claims. Some optional features of the invention are defined in the dependent claims.

10 A known method of obtaining an initial CPE estimate or measure is to perform a cross-correlation of a detected pilot subcarrier and a known pilot subcarrier. The cross-correlation is a CPE estimate disturbed by coloured noise. Some CPE estimates may be seriously affected by coloured noise arriving from MIMO detection (eg, from the MIMO detection matrix $\mathbf{G}(m)$), and the noise power can
 15 vary as a function of channel. Therefore, performance can be greatly enhanced by utilising post-detection signal-to-noise ratio (SNR) characteristics.

A plurality of two-dimensional SNR characteristics each for a selected pilot subcarrier and a selected spatial stream may be used to derive a weighting
 20 factor to weight the initial CPE estimate/measure to achieve minimum variance in estimation error.

The resulting final CPE estimate is actually the exponential function of the rotated phase caused by CFO/PN. (Phase rotation of the known pilot symbol in
 25 the specific spatial stream is a measure of the CPE.) In some apparatuses it is possible to multiply the conjugate of the known pilot symbol with the detected MIMO signal, so that the impairment of CFO and PN is substantially compensated for if the ICI term caused by CFO can be ignored due to a small residual frequency offset. Such an implementation results in less complexity in
 30 CFO/PN tracking than in conventional CPE estimators, which need to extract the rotated phase by use of a trigonometric function.

An advantage arising from weighting of the CPE measure is that measures disturbed by serious coloured noise can be assigned weights which minimise the adverse effects of the coloured noise on estimation accuracy.

- 5 Apparatuses implementing the invention can be used in conjunction with a carrier frequency offset/phase noise-independent timing offset tracking and compensation method for MIMO-OFDM systems as disclosed in commonly-assigned Singapore Patent Application No. 200606234-3, the content of which is hereby incorporated by reference in its entirety.

10

The present invention will now be described, by way of example only, and with reference to the accompanying drawings in which:

- 15 Figure 1 is a block diagram illustrating an apparatus for estimating a CPE component of a MIMO-OFDM wireless communications signal;

Figure 2 is a block diagram illustrating a more detailed view of the CPE estimator of the apparatus of Figure 1 and an optional filter for further processing of the output of the CPE estimator;

- 20 Figure 3 is a process flow diagram illustrating the steps of an algorithm for estimating a CPE component of a MIMO-OFDM wireless communication signal in the apparatus of Figures 1 and 2;

Figure 4 is a graph illustrating the mean square error of the apparatus of Figures 1 and 2 and that of a conventional estimator;

- 25 Figure 5 is a graph illustrating the packet error rate (PER) comparison of the apparatus of Figures 1 and 2 and that of a conventional estimator in channel D; and

Figure 6 is a graph illustrating the bit error rate (BER) comparison of the apparatus of Figures 1 and 2 and that of a conventional estimator in channel D.

- 30 Referring now of Figure 1, a first apparatus 10 for estimating a CPE component of a MIMO-OFDM wireless communications signal will now be described. The apparatus 10 of Figure 1 comprises two receiver antennas (not shown), but the

configuration of two antennas can easily be extended to other MIMO configurations employing higher numbers of antennas. Time-domain signals received at the receiver antennas are first input 11a, 11b to Fast Fourier Transform (FFT) modules 12a, 12b which convert the received time domain signals to the frequency domain. The converted frequency-domain signals comprise frequency-domain high throughput (HT) long preambles which are output by FFT modules 12a, 12b and are received at a channel estimation module 14 which generates a MIMO channel frequency response estimate matrix. The obtained channel frequency response estimate matrix is output both to SNR generator 16 and MIMO detector 18. SNR generator 16 operates to obtain a two-dimensional SNR characteristic – as will be described below – and a first output of SNR generator 16 is input as a first input 20a to CPE estimator 20. CPE estimator 20 obtains a CPE estimate – as will be described in detail below – and the output of CPE estimator 20 is supplied to CPE compensator 22 for compensation of the CPE. The output of CPE compensator 22 is supplied to demapper 24 to obtain the constellation points of the MIMO-OFDM signal with improved accuracy. The post-detection SNR characteristic also aids the third-order timing recovery loop made up of two-dimensional timing drift tracking module 26, phase rotation module 28 and FFT window controller module 30, described in detail in SG200606234-3, as noted above.

Figure 2 shows components in a more detailed view of the CPE estimator of Figure 1. When compared with Figure 1, like components are denoted by like reference numerals. The apparatus 10 receives frequency-domain subcarrier signals from FFT stages 12a,...,12r for each of N_r receiver antennas. MIMO detector 18 receives the frequency-domain signals from FFT stages 12a,...,12r. Suppose the transmitter has N_t antennas. The output of FFT stages 12 comprises $2J \cdot N_t$ ($-J, \dots, -1, 1, \dots, J$) detected pilot subcarriers which are supplied to a cross-correlation stage 50a,...,50 _{$2J \cdot N_t$} where the detected pilot subcarriers are cross-correlated with a known pilot symbol to obtain initial CPE estimates. The initial CPE estimates output by stage 50 are supplied as a first input to CPE

estimator 20. A second input to CPE estimator is output from the post-detection SNR generator 16.

The output from CPE estimator 20 can either be fed directly to CPE compensation stage 22 or, alternatively can be smoothed in the time domain by filter stage 60 which, in the example of Figure 2 is an infinite impulse response (IIR) filter. The smoothing with filter 60 may provide a yet further improved, second determination of CPE estimation.

The optional filter, in the apparatus of Figure 2, is IIR filter 60 comprising a forgetting factor stage 62, summation stage 64, feedback – weight stage 66 and feedback – delay stage 68. The output of IIR filter stage 60, when utilised, is supplied to the CPE compensation module 22.

The processing of the signals in the apparatuses of Figures 1 and 2 are explained as follows:

Matrix manipulation transformed operators used in the following discussion are given as follows:

- $(\dots)^{-1}$ denotes matrix inverse transform operation
- $(\dots)^T$ denotes matrix transpose transform operation
- $(\dots)^H$ denotes matrix conjugate and transpose transform operation

In a MIMO-OFDM based WLAN system, the transceiver signals may be expressed in the frequency domain using relatively straightforward matrix algebra. Considering a $N_t \times N_r$ MIMO-OFDM system, where N_t and N_r are the numbers of transmitter and receiver antennas respectively, the $N_r \times 1$ received signal vector on the k -th subcarrier may be written as:

$$\mathbf{y}(k) = \mathbf{H}(k)\mathbf{x}(k) + \mathbf{n}(k) \quad (1)$$

Here $\mathbf{y}(k) = (y_1(k), y_2(k), \dots, y_{N_r}(k))^T$, and $\mathbf{x}(k)$ is the $N_t \times 1$ signal vector

transmitted from N_t transmitter antennas on the k -th subcarrier, which may be written as

$$\mathbf{x}(k) = (x_1(k), x_2(k), \dots, x_{N_t}(k))^T.$$

The channel frequency response of the k -th subcarrier from the j -th transmitter antenna to the i -th receiver antenna may be denoted by $H_{i,j}(k)$, and

$$\mathbf{H}(k) = \begin{pmatrix} H_{1,1}(k) & H_{1,2}(k) & \dots & H_{1,N_t}(k) \\ H_{2,1}(k) & H_{2,2}(k) & \dots & H_{2,N_t}(k) \\ \vdots & \vdots & \dots & \vdots \\ H_{N_r,1}(k) & H_{N_r,2}(k) & \dots & H_{N_r,N_t}(k) \end{pmatrix},$$

$$\mathbf{n}(k) = (n_1(k), n_2(k), \dots, n_{N_r}(k))^T.$$

In the above system model, ideal synchronisation in the OFDM receiver has been assumed. However, there is usually a carrier frequency offset and timing clock offset between the receiver and transmitter due to the mismatch between the transmitter and receiver oscillators or the channel Doppler frequency shift. This is true also for the receiver sampling clock. There is also usually a sampling frequency offset between transmitter and receiver side.

In practice, there are multiple radio chains in the MIMO-OFDM transmitter and receiver sides respectively, and these multiple radio chains share the same frequency synthesiser such that multiple radio chains in the transmitter or receiver side are synchronised respectively. Denote the CFO as Δf , and the normalised TO as $\eta = (T' - T)/T$, where T and T' are the nominal sampling period and the actual sampling period at the receiver. Following the system analysis method in Pei-yun Tsai, Hsin-Yu Kang and Tzi-Dar Chiueh, "Joint Weighted Least Squares Estimation of Frequency and Timing Offset for OFDM Systems over Fading Channels", *IEEE Trans. On Vehicular Technology*, Vol.54,

Issue 1, Jan. 2005, pp. 211-223, and “EWC HT PHY Specification”, *Enhanced Wireless Consortium Publication*. V1.27, December 23rd, 2005 (for the synchronisation model for MIMO-OFDM systems), the l -th received OFDM symbol suffering from the synchronisation error can be derived as shown in
 5 Equation (2) below.

It is assumed that the information signals are encoded and modulated to form an OFDM symbol $\mathbf{x}_l(k)$ in the frequency domain. After IFFT transformation and pass-band modulation, the time-domain transmitted signals pass through a
 10 MIMO fading channel. The l -th received baseband OFDM symbol after sampling with sampling clock T and removing the guard interval, may be represented by N samples. The demodulation of the received samples via an N -point FFT yields the received OFDM symbols in the frequency domain

$$\mathbf{y}_l(k) = \exp(j[2\pi\Delta f(IN_s + N_g)(1+\eta)T + \phi_l]) \exp\left(j\frac{2\pi k(IN_s + N_g)\eta}{N}\right) \mathbf{H}(k)\mathbf{x}_l(k)s(\pi\phi_{kk}) + \mathbf{ICI}_l(k) + \mathbf{n}_l(k) \quad (2)$$

15

where ϕ_l is phase noise; N is the number of FFT points; N_g and N_s are samples in the guard interval and the whole OFDM symbol including guard interval respectively.

20 Equation (2) shows that the distortion of the k -th subcarrier data caused by synchronisation error includes phase rotation and the magnitude scaling, $s(\pi\phi_{kk})$, respectively. Moreover, the synchronisation error destroys the orthogonality between subcarriers resulting in ICI, which is the function of $s(\pi\phi_{pk})$. As set out in Pei-yun Tsai, Hsin-Yu Kang and Tzi-Dar Chiueh, “Joint
 25 Weighted Least Squares Estimation of Frequency and Timing Offset for OFDM Systems over Fading Channels”, *IEEE Trans. On Vehicular Technology*, Vol.54, Issue 1, Jan. 2005, pp. 211-223, the definition of $s(\pi\phi_{pk})$ may be expressed as:

$$s(\pi\phi_{pk}) = \frac{\sin(\pi\phi_{pk})}{N \sin(\frac{\pi\phi_{pk}}{N})}, \text{ and}$$

$s(\pi\phi_{kk}) = s(\pi\phi_{pk})$ in the case of $p = k$,

where p and k are subcarrier indexes and $\phi_{pk} = (1 + \eta)(\varepsilon + p) - k$, $\varepsilon = \Delta f \cdot T$.

When η and ε are small, $s(\pi\phi_{kk})$ is close to 1 and $s(\pi\phi_{pk})$ is near zero, and the

5 magnitude distortion and the ICI term may be ignored. Thus, a phase rotation occurs when the subcarrier sample $x_l(k)$ is transmitted. The rotated phase is such that

$$\theta_l(k) = 2\pi\Delta f(lN_s + N_g)(1 + \eta)T + \frac{2\pi k}{N}(lN_s + N_g)\eta + \varphi_l \quad (3)$$

10 The first component is the inter-play of carrier and sampling frequency offsets, and the second component is contributed by sampling frequency offset exclusively.

In the case when only CFO/PN exists, a CPE, $\theta_l(k) = 2\pi\Delta f(lN_s + N_g)T + \varphi_l$ is introduced to all carriers in all spatial streams. The CPE changes on a symbol-by-symbol basis. It is, therefore, necessary to estimate and correct the rotation of the received constellation points caused by the CFO/PN on a symbol-by-symbol basis.

As known from a MIMO-OFDM system, the received frequency domain pilot subcarrier before MIMO detection is a combination of all spatial pilot signals through a MIMO frequency-selective multipath fading channel. Thus conventional weighted CPE estimation methods before equalisation may not be applicable. Here a post-MIMO detection CFO/TO tracking algorithm is proposed by utilising two-dimensional post-detection SNR reliability characteristics to derive an optimal linear CPE estimation for MIMO-OFDM systems. The optimal linear CPE estimation can then be utilised to correct the rotation of the received constellation points.

In order to take advantage of its properties as one of the most efficient methods of implementing a spatial multiplexing receiver in practical MIMO systems, linear MIMO detection is employed. This is particularly true for those systems with a large number of transmitter and receiver antennas. Linear receivers
 5 utilise matrix multiplication to separate the data streams transmitted simultaneously, as given by:

$$\begin{aligned}\hat{\mathbf{x}}(k) &= \mathbf{G}(k)\mathbf{y}(k) \\ &= \mathbf{G}(k)\mathbf{H}(k)\mathbf{x}(k) + \mathbf{G}(k)\mathbf{n}(k)\end{aligned}\tag{4}$$

where $\hat{\mathbf{x}}(k)$ is an estimate of the transmitted symbol vector; and detection matrix $\mathbf{G}(k)$ can be computed according to a zero-forcing (ZF) or minimum
 10 mean square error (MMSE) criterion. In this example, the detection is performed assuming that $\mathbf{G}(k)\mathbf{H}(k) = \mathbf{I}$, where $\mathbf{I}$ is an identity matrix. When using MMSE, $\mathbf{G}(k)$ is calculated from, as discussed in Catherine Z.W. Hassell Sweatmean, John S. Thompson, Bernard Mulgrew and Peter M. Grant, "A comparison of detection algorithms including BLAST for wireless
 15 communication using multiple antennas", *Proc. of IEEE PIMRC 2000*, Vol.1. pp. 698-703:

$$\mathbf{G}(k) = (\mathbf{H}^H(k)\mathbf{H}(k) + N_0 \times \mathbf{I})^{-1} \mathbf{H}^H(k)\tag{5}$$

here N_0 is noise spectral density.

20 If, say, $2J$ pilots are inserted among N signal subcarriers, the pilot subcarrier indexes are given by $m = -J, \dots, -2, -1, 1, 2, \dots, J$. If implemented with the CFO independent TO estimation and tracking algorithm for MIMO-OFDM systems disclosed in SG200606234-3, only a correction of residual CFO and phase noise must be corrected and, therefore considered. Thus, $s(\pi\phi_{kk})$ is close to 1
 25 with a small ε . In light of these considerations, the detected pilot subcarriers in the l -th OFDM symbol incorporating CFO error and phase noise can be represented as:

$$\hat{\mathbf{x}}_l(m) = \exp(j[2\pi\Delta f(IN_s + N_g)T + \phi_l])\mathbf{x}_l(m) + \mathbf{G}(m)\mathbf{n}_l(m)\tag{6}$$

Equation (6) can be rewritten as $2J \cdot N_t$ independent equations:

$$\begin{cases} \hat{x}_{l,1}(m) = \exp(j[2\pi\Delta f(IN_s + N_g)T + \phi_l])x_{l,1}(m) + \xi_{l,1}(m) \\ \vdots \\ \hat{x}_{l,N_t}(m) = \exp(j[2\pi\Delta f(IN_s + N_g)T + \phi_l])x_{l,N_t}(m) + \xi_{l,N_t}(m) \end{cases} \quad m = -J, \dots, -1, 1, 2, \dots, J. \quad (7)$$

where $\xi_{l,i}(m)$, $i = 1, \dots, N_t$, define coloured noise components due to AWGN noise and MIMO detection, for m pilot subcarriers. This coloured noise component is a function of the MIMO channel. Evidently, the common phase rotation is applied to all pilot subcarriers in all spatial streams. Thus the phase rotation of the known pilot symbol in the specific spatial stream, observed at the MIMO detector output, is a measure for the CPE. There are a total of $2J \cdot N_t$ CPE measures, derived as:

$$\begin{aligned} z_{l,i}(m) &= \hat{x}_{l,i}(m)x_{l,i}^*(m) \\ &= \exp(j2\pi\Delta f(IN_s + N_g)T + \phi_l) |x_{l,i}(m)|^2 + \xi_{l,i}(m)x_{l,i}^*(m) \\ &= CPE_l + e_{l,i}(m) \end{aligned} \quad (8)$$

where $i = 1, \dots, N_t$, and the power of the known pilot symbol is assumed to be unity, i.e., $|x_{l,i}(m)|^2 = 1$. Therefore, a cross-correlation of a detected pilot subcarrier with a known pilot subcarrier symbol yields $2J \cdot N_t$ CPE estimates disturbed by coloured noise. Some of CPE estimates may be seriously affected by noise coloured by the MIMO detection matrix $G(m)$ and, also, noise power can vary as a function of channel. Therefore for linear MIMO detection, post-processing SNR reliability measure is a parameter of interest. This reliability measure can be exploited in a CPE estimator.

Assuming that,

$$\mathbf{Z}_l = [z_{l,1}(-J), \dots, z_{l,N_t}(-J), z_{l,1}(-J+1), \dots, z_{l,N_t}(-J+1), \dots, z_{l,1}(J), \dots, z_{l,N_t}(J)]^T$$

and,

$$\mathbf{e}_l = [e_{l,1}(-J), \dots, e_{l,N_t}(-J), e_{l,1}(-J+1), \dots, e_{l,N_t}(-J+1), \dots, e_{l,1}(J), \dots, e_{l,N_t}(J)]^T$$

then $2J \cdot N_t$ CPE estimates can be expressed as a vector equation,

$$\mathbf{Z}_l = \text{CPE}_l \cdot \mathbf{U} + \mathbf{e}_l \quad (9)$$

where $\mathbf{U} = [1, 1, \dots, 1]^T$ is a $2JN_l \times 1$ unit vector and $\mathbf{e}_l$ is a $2JN_l \times 1$ vector of coloured noise values. Based on the above equation, the optimal linear solution to minimum-variance least squares estimation of CPE can be applied as discussed in Steven M. Kay, "Fundamentals of statistical signal processing: estimation theory", Prentice-Hall, Inc. Upper Saddle River, NJ, USA, 1993, and derived as:

$$\hat{\text{CPE}}_l = (\mathbf{U}^T \mathbf{W} \mathbf{U})^{-1} \mathbf{U}^T \mathbf{W} \mathbf{Z}_l = \frac{\sum_j \left(z_j \sum_i w_{i,j} \right)}{\sum_i \sum_j w_{i,j}} \quad (10)$$

where $w_{i,j}$ is the element in i -th row and j -th column of a $2JN_l \times 2JN_l$ weighting matrix $\mathbf{W}$, and z_j is the j -th element of the vector $\mathbf{Z}_l$. The main ideal of the WLS algorithm is that weighting the specific CPE measure is necessary because the measures disturbed by serious coloured noise should be assigned smaller weights to minimise their adverse effects on estimation accuracy. Namely, the CPE estimate for each pilot subcarrier is weighted by the corresponding reliability measure. The optimal weighting factors are given by weighting matrix $\mathbf{W}$:

$$\begin{aligned} \mathbf{W} &= E\{\mathbf{e}_l \mathbf{e}_l^H\}^{-1} \\ &= E \begin{bmatrix} e_1 e_1^* & e_1 e_2^* & \cdots & e_1 e_{2J \cdot N_l}^* \\ e_2 e_1^* & e_2 e_2^* & \cdots & e_2 e_{2J \cdot N_l}^* \\ \cdots & \cdots & \cdots & \cdots \\ e_{2J \cdot N_l} e_1^* & e_{2J \cdot N_l} e_2^* & \cdots & e_{2J \cdot N_l} e_{2J \cdot N_l}^* \end{bmatrix}^{-1} \end{aligned} \quad (11)$$

where $E\{\cdot\}$ stands for the expected value of the enclosed components, e_i is the i -th element of the vector $\mathbf{e}_l$. The CPE estimator generates a matrix of weighting factors for a plurality of subcarriers and a plurality of spatial streams of the MIMO-OFDM signal and generates elements of the matrix from a correlation of coloured noise values for the plurality of subcarriers and plurality of spatial

streams of the MIMO-OFDM signals. Referring to Equation (6), Equation (7), and Equation (8), we can get,

$$e_i e_j^* = \mathbf{G}_\beta(m) \mathbf{n}_{l,\beta}(m) x_{l,\beta}^*(m) \mathbf{G}_{\beta 1}^*(m) \mathbf{n}_{l,\beta 1}^*(m) x_{l,\beta 1}(m)$$

where $\mathbf{G}_\beta(m)$ and $\mathbf{n}_{l,\beta}(m)$ denote the β -th row of the detection matrix $\mathbf{G}(m)$, and β -th spatial noise vector respectively.

For the case of $i = j$, $e_i e_i^* = E \|\mathbf{G}_\beta(m) \mathbf{n}_{l,\beta}(m)\|^2$.

For the case of $i \neq j$, $e_i e_j^* = 0$, because the noise vector for different subcarrier and different spatial streams is uncorrelated. Thus, the off-diagonal terms of $E\{\mathbf{e}_l \mathbf{e}_l^H\}$ are all zero, and the matrix inverse operation involved in Equation (11) is simplified to an inverse of the corresponding diagonal elements. Therefore, j -th diagonal component of weighting matrix $\mathbf{W}$, is given by:

$$w_{j,j} = \frac{1}{E \|\mathbf{G}_\beta(m) \mathbf{n}_{l,\beta}(m)\|^2} = \frac{1}{N_0 E \|\mathbf{G}_\beta(m)\|^2} \quad (12)$$

The CPE estimator receives an output signal from MIMO detector 14, and generates a weighting factor for the estimation of a CPE component from a value of the MIMO detector detection matrix $\mathbf{G}$.

Assuming that the transmitted signal, $\mathbf{x}(k)$, has been normalised, the derived weighting factor is actually the two-dimensional (spatial and frequency dimensions) SNR of the post MIMO detection for m -th pilot subcarrier and β -th spatial stream. When MMSE MIMO detection is employed, the post-detection SNR reliability in the frequency and spatial domain for j -th sub-stream and m -th pilot subcarrier is given in the following equation as derived from Robert W. Heath Jr. and Arogyaswami Paulraj, "Antenna Selection for Spatial Multiplexing Systems based on Minimum Error Rate", *IEEE International Conference on Communications*, Vol.7, June 2001, pp. 2276-2280:

$$SNR_{m,j} = \frac{1}{N_0 [\mathbf{H}^H(m) \mathbf{H}(m) + N_0 \mathbf{I}]_{j,j}^{-1}} - 1 \quad (13)$$

where, $[\cdot]_{j,j}$ denotes the j -th element on the diagonal of the inverse matrix. Combining the above equations, the final expression of the proposed CPE estimation is obtained,

$$\hat{CPE}_l = \frac{\sum_{i=1}^{N_t} \sum_{m=1}^{2J} SNR_{m,i} (\hat{x}_{l,i}(m) x_{l,i}^*(m))}{\sum_{i=1}^{N_t} \sum_{m=1}^{2J} SNR_{m,i}} \quad (14)$$

5 where $SNR_{m,i}$ is the two-dimensional characteristic from which the weighting factor is derived and $z_{l,i}(m) = \hat{x}_{l,i}(m) x_{l,i}^*(m)$ is the initial CPE estimate for a subcarrier. Therefore, the apparatus of Figures 1 and 2 estimates the CPE component of the MIMO-OFDM wireless communications signal by generating a plurality of two-dimensional post-linear MIMO detection SNR characteristics of the signal, generating a weighting factor from the plurality of said characteristics
10 to apply the weighting factor to an initial CPE estimate thereby to estimate the CPE component.

This can be expressed alternatively as the apparatus of Figures 1 and 2
15 estimates a CPE component of a MIMO-OFDM wireless communications signal, by estimating the CPE from a weighting of an initial CPE estimate, the weighting having a weighting factor generated from a plurality of two-dimensional post-linear MIMO detection SNR characteristics of the MIMO-OFDM signal.

20

The apparatus can estimate the CPE of the MIMO-OFDM signal by implementing the steps above across a plurality of sub-carriers (and a plurality of spatial streams) of the MIMO-OFDM signal. That is, the CPE estimator generates a plurality of weighting factors for a plurality of subcarriers of the
25 MIMO-OFDM wireless communications signal and applies the plurality of weighting factors to a respective plurality of initial CPE estimates thereby to estimate the CPE of the MIMO-OFDM signal. Thus, the CPE of the signal can be monitored and derived dynamically. The apparatus of Figure 2, when

implementing the optional IIR filter provides a further (second) improved CPE estimate.

The signal-to-noise ratio generator can generate characteristics as weighting factors where the weighting factors are different for different subcarriers of different spatial streams. That is, the CPE estimate for each subcarrier is weighted by a corresponding reliability measure, as noted above. To implement this feature, the CPE estimator generates first and second weighting factors, the first weighting factor being different from the second weighting factor. In some apparatuses, all weighting factors generated by the CPE estimator are different. All, or at least the first and second, different weighting factors are then applied to respective different subcarriers.

When the residual frequency error is small, i.e., $\varepsilon N \ll 1$, the CPE shift over one OFDM symbol is small, which means that the CPE estimates among continuous OFDM symbols are correlated. Moreover, the phase noise and thermal noise contribute more to the overall CPE in most applications, which is a slow random process and is invariant within continuous OFDM symbols. In this case, the IIR filter is used to improve the estimation performance further to suppress the variation in phase noise and thermal noise. The improved CPE estimate $\hat{CPE}_t$ is given by:

$$\hat{CPE}_t = \gamma \hat{CPE}_t + (1 - \gamma) \hat{CPE}_{t-1} \quad (19)$$

where $\hat{CPE}_t$ is the output of the IIR filter, and γ is the forgetting factor, $0 < \gamma < 1$. The applied forgetting factor's value depends on the power spectrum density of the phase noise and thermal noise to be suppressed as discussed in P. Robertson and S. Kaiser, "Analysis of the effects of phase noise in orthogonal frequency division multiplexing (OFDM) systems," in *Proc. ICC'95*, vol.3, Seattle, WA, 1995, pp. 1652-1657.

As discussed in Songping Wu and Yeheskel Bar-Ness, "A Phase Noise Suppression Algorithm for OFDM-Based WLANs", IEEE Communications

Letters, Vol.6, No.12, December 2002, pp. 535-537, when utilising one method to improve the estimation accuracy, equation (19) can be expressed as:

$$\hat{I}_m = \gamma \tilde{I}_m + (1 - \gamma) \tilde{I}'_m \quad (20)$$

At first glance, Equations (19) and (20) look similar. However, they are different
5 in three aspects.

1. Equation (19) is an IIR filter, the new output at l -th time instant, is weighted sum of this feedback at the previous time instant $l-1$ -th, and the new input at l -th time instant.
- 10 2. Equation (20) in reference is not an IIR filter, because there is no feedback. Equation (20) weights and combines two new inputs to get the output.
3. In a word, IIR filters only need one input, while Equation (20) is not an IIR filter (no feedback). Their mechanisms are totally different.
- 15 The processing of the signals as explained above may be better understood with reference to Figure 3. The process starts at step 100. At step 102, FFT stage 12 performs a Fast Fourier Transform on the received time-domain signals. MIMO detection module 18 performs detection on the MIMO-OFDM signal at step 104. The resulting signal after detection at step 106 are
20 frequency domain post-detection signals with residual carrier frequency offset and phase noise. The subcarriers are represented by Equations (6) and (7) above.

The $2J \cdot N_t$ pilot subcarriers are output by MIMO detector 18 to cross-correlation
25 stage 50. The output of cross-correlation stage 50 is the vector result defined by Equation (9) as shown at step 108. At step 114, post detection signal-to-noise generator 16 derives the two-dimensional SNR characteristic given by Equation (13) above.

30 At step 116, common phase estimator 20 derives the improved CPE estimate in accordance with Equation (14) given above. Optionally, the algorithm can end

at step 118 or, alternatively, the improved CPE estimate is passed through the filter 60 of Figure 2 at step 117 before the algorithm ends at step 118.

Referring again to Figures 1 and 2, the second CPE estimate is actually the exponential function of the rotated phase caused by CFO/PN. Thus multiplying the conjugate of this value with the detected MIMO signal, the impairment of CFO and PN is substantially compensated for if the ICI term caused by CFO can be ignored due to small residual frequency offset. This compensation is implemented in CPE compensator 22. The apparatus of Figures 1 and 2 may introduce much lesser complexity to the CFO/PN tracking than conventional CPE estimators, which extract the phase from a trigonometric function.

System performance may be considered with respect to an example employing a 2×2 MIMO-OFDM based WLAN system platform. Two spatial streams are transmitted over a MIMO fading channel with 20MHz frequency band by utilising spatial division multiplexing (SDM) techniques. Convolutional codes with different puncturing rates are adopted. One suitable detailed system specification is referred to the WLAN 802.11n standard draft as specified in "EWC HT PHY Specification", *Enhanced Wireless Consortium Publication*. V1.27, December 23rd, 2005. A type D MIMO fading channel model as specified in "IEEE P802.11 Wireless LANs TGn Channel Models", *doc.: IEEE 802.11-03/940r4*, May 2004 is employed in the system performance assessment. The antenna configuration is of the linear array type and distance between adjacent two antennas is a half carrier wavelength.

25

In the considered system, a residual CFO of 1000Hz and optimal TO tracking are assumed. The least square MIMO channel estimation without smoothing and MMSE MIMO detection are adopted in the receiver. Moreover, Viterbi decoding with trace-back length of 128 is employed. With the above established example, the MSE (mean square error) of the invented post-detection CPE estimator for Type D MIMO channel is obtained and compared with the conventional average CPE estimator, as illustrated. Figure 4 shows

30

that the MSE of the estimator implementing the techniques described above is much less than that of the conventional mean CPE estimator over a wide SNR range. Moreover, the PER (packet error rate) and BER (bit error rate) comparisons between the described post-detection CPE estimator and the
5 conventional average CPE estimator over Type D MIMO channel are shown in Figure 5 and Figure 6 respectively. The comparison involves various modulation schemes, namely, BPSK, 16QAM and 64QAM with their respective coding rates $\frac{1}{2}$, $\frac{1}{2}$ and $\frac{5}{6}$. The results in Figure 5 and Figure 6 clearly show that the described post-detection CPE estimator out-performs the conventional
10 average CPE estimator.

It will be appreciated that an invention has been described by way of example only and various modifications to the invention may be made without departing from the spirit and scope of the appended claims.

CLAIMS

1. Apparatus for estimating a common phase error component of a MIMO-OFDM wireless communications signal, the apparatus comprising:
 - 5 a signal-to-noise ratio generator configured to generate a plurality of two-dimensional post-linear MIMO detection signal-to-noise ratio characteristics of the MIMO-OFDM wireless communications signal; and
 - a common phase error estimator configured to generate a first weighting factor and a second weighting factor from the plurality of two-dimensional post-linear MIMO detection signal-to-noise ratio characteristics and to apply the first and second weighting factors to an initial common phase error estimate thereby to estimate the common phase error component, wherein the first weighting factor is different from the second weighting factor.
- 15 2. Apparatus according to claim 1, wherein the common phase error estimator is configured to generate a plurality of weighting factors for a plurality of subcarriers of the MIMO-OFDM wireless communications signal and to apply the plurality of weighting factors to a respective plurality of initial common phase error estimates thereby to estimate the common phase error.
- 20 3. Apparatus according to claim 1 or 2, wherein the signal-to-noise generator is configured to receive an output signal from a MIMO detector, the common phase error estimator being configured to generate a weighting factor from a value of a detection matrix of the MIMO detector.
- 25 4. Apparatus according to any preceding claim further comprising a filter configured to receive the common phase error estimate and to output a second common phase error estimate.
- 30 5. Apparatus according to claim 4, wherein the apparatus is configured to convert the common phase error estimate to a time-domain signal and wherein the filter is an IIR filter configured to receive the time-domain signal.

6. Apparatus for estimating a common phase error component of a MIMO-OFDM wireless communications signal, the apparatus being configured to estimate the common phase error component from a weighting of an initial
5 common phase error estimate, the weighting having a first weighting factor and a second weighting factor generated from a plurality of two-dimensional post-linear MIMO detection signal-to-noise ratio characteristics of the MIMO-OFDM signal, wherein the first weighting factor is different from the second weighting factor.
- 10
7. A method of estimating a common phase error component in an apparatus for receiving a MIMO-OFDM wireless communications signal, the method comprising:
- controlling a signal-to-noise ratio generator to generate a plurality of two-
15 dimensional post-linear MIMO detection signal-to-noise ratio characteristics of the MIMO-OFDM wireless communications signal; and
- controlling a common phase error estimator to generate a first weighting factor and a second weighting factor from the plurality of two-dimensional post-linear MIMO detection signal-to-noise ratio characteristics and to apply the
20 weighting factor to an initial common phase error estimate thereby to estimate the common phase error component, wherein the first weighting factor is different from the second weighting factor.
- 25
8. The method of claim 7, wherein controlling the common phase error estimator comprises generating a plurality of weighting factors for a plurality of subcarriers of the MIMO-OFDM wireless communications signal and applying the plurality of weighting factors to a respective plurality of initial common phase error estimates thereby to estimate the common phase error
- 30
9. The method of claim 7 or 8, comprising receiving an output signal from a MIMO detector at the signal-to-noise generator and controlling the common

phase error estimator to generate a weighting factor from a value of a detection matrix of the MIMO detector.

10. The method of any of claims 7 to 9, further comprising receiving, at a
5 filter of the apparatus, the common phase error estimate and outputting, from the filter, a second common phase error estimate.

11. The method of claim 10, further comprising converting the common
phase error estimate to a time-domain signal and wherein the filter is an IIR
10 filter configured to receive the time-domain signal.

12. A method of estimating a common phase error component of a MIMO-
OFDM wireless communications signal, the method comprising estimating the
common phase error component from a weighting of an initial common phase
error estimate, the weighting having a first weighting factor and a second
15 weighting factor generated from a plurality of two-dimensional post-linear MIMO
detection signal-to-noise ratio characteristics of the MIMO-OFDM signal,
wherein the first weighting factor is different from the second weighting factor.