Time-frequency and Analog Front-ends Synchronization Approaches of High-speed OFDM Systems for Wireless Control

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Graduate School of Engineering

CHIBA UNIVERSITY

(千葉大学審査学位論文)

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Abstract

The inter-carrier interference (ICI) caused by carrier frequency offset (CFO) and in phase and quadrature (I/Q) imbalance at the tranceiver will bring about serious problems in high-speed orthogonal frequency division multiplexing (OFDM) systems since the orthogonality among subcarriers is destroyed. In addition, an accurate estimation of symbol timing is also crucial since the symbol timing synchronization inaccuracy will bring about phase offset (FO), inter-symbol interference (ISI) and ICI, which have great impact on the system performance. Various techniques have been developed to combat these effects of CFO, the I/Q imbalance in the analog front-end and the symbol timing offset (STO). However, these methods are either too complicated or use a large number of pilot signals. A system less sensitive to STO, CFO and I/Q imbalance is required at the receiver for long term evolution (LTE) or future fifth generation mobile networks (5G) system. In this paper, a novel ICI self-cancellation (ICI SC) scheme is proposed to mitigate the CFO effect in a general implementation framework with I/Q imbalance compensation. ICI and optimum carrier to interference power ratio (CIR) are derived and discussed. Moreover, by utilizing time-frequency interferometry (TFI) pilot signal and energy clipping method, through analysis and simulations, we can determine the predefined weighting coefficients to achieve the best bit error rate (BER) performance. It is demonstrated by the computer simulations that the proposed scheme can make improvement of the BER performance compared to the conventional ICI SC methods. In regarding to the time synchronization problems, we researched on both preamble-aided and non-preamble-aided methods. For preamble-aided method, a novel symbol timing synchronizer is proposed based on the properties of pseudo noise (PN) sequences. By utilizing four identical blocks with inverted signs of the latter two blocks and two different PN sequences with the same period as weighted factors to ensure that the metric function has impulse-like shape at correct timing position without any burst side-peaks. From the simulation results, the proposed scheme achieves accurate timing offset estimation with smaller error mean (EM) and standard deviation (SD) compared with the existing method under flat-fading (FF) channels. When it is subjected to frequency-selective fading (FSF) channels, the performance is relatively better than others with the increases of FFT window size referring to third generation partnership project (3GPP) LTE specifications. For non-preamble-aided, a metric function and a positioning function have been designed by taking advantage of the correlation properties of OFDM signals. Computer simulations demonstrate that the schemes are superior because it achieves accurate timing offset estimation with smaller standard deviation (SD) and relatively lower computational complexity compared with the existing methods based on cyclic prefix (CP) under high signal-noise ratio (SNR) assumption over frequency-selective channels, which can meet the real-time requirements in wireless local area network (WLAN) environments for factory automation (FA).

Chapter 1

General Introduction

1.1 Background and Significance

The fast and convenient information exchange promotes the rapid development of wireless communication systems. With the enlargement of the wireless communication, the field can be divided into microwave communications, radio paging communication, cellular mobile communication, satellite communication, packet radio network communication, and other typical communication systems, among which the mobile communication technology has been widely applied in the world. As one of the important technologies of the twentieth century, wireless communications over the past few decades has changed dramatically. First-generation analog mobile communication systems provide the voice services using frequency division multiple access (FDMA), the second-generation digital mobile communication system based on time division multiple access (TDMA) provides a high-resolution global roaming voice services. The third generation communication system based on code division multiple access (CDMA) offers at least 144kbps (preferable 384 kbps) for highmobility users with wide-area coverage and 2Mbps for low-mobility users with local convergence. The fourth generation communication based on long term evolution (LTE) can achieve the peak bit rate about 100Mbps in the downlink and 50Mbps in the uplink. The fifth generation (5G) technology is expected to provide 100 times the speed than 4G LTE [1]. 5G will not completely replace 4G at the technical level. 4G, WiFi and new techniques will be blended into 5G technology, bringing more rich experience for the users. The system will automatically connected to the network with the best quality, thereby realizing seamless switching. Figure 1.1 shows the development courses of world's mobile communications.



Figure 1.1: Development courses of world's mobile communications.

1.2 Wireless Propagation and Fading

The wireless transmission channel is rather dynamic and unpredictable, which makes an exact estimation of the wireless channel often difficult. In fact, the understanding of wireless channels is crucial to obtain high quality wireless transmission system. There are three basic physical transmission modes in the course of radio propagation in wireless communication: reflection, diffraction, and scattering. Reflection occurs when a radio wave encounters an obstacle surface with much larger dimensions compared to the radio wavelength, it forces the transmit signal power to be reflected back to its origin path. Diffraction occurs when the radio path impinges upon edges of the thin barrier or a surface with sharp irregularities or small openings. Scattering is the physical phenomenon that the radio wave encounters a large number of particles with much smaller dimensions than the transmission wavelength. The propagation path loss of the radio signal emitted from the base station is greatly affected by the ground terrain. The higher the base station is, the farther the signal will be transmitted. Radio wave propagation is extremely complex, easily influenced by multi-path propagation of reflection, diffraction and scattering, which can cause severe signal fading. The propagation of radio waves is also relevant to the frequency, if the frequency is high, thus the path loss is great, the diffraction is week, and the propagation distance is near. Figure 1.2 is the wireless propagation model.

The variation of the amplitude and phase of the non-additive signal over time and frequency is a phenomenon called fading. In the line-of-sight (LOS) environments, there is a direct main path that obeys Rice distribution. The simulation in our work are under non line-of-sight (NLOS) environments with Rayleigh distribution. The fading phenomenon can be divided into two types including large-scale fading and small-scale fading. Large-scale fading contains the shadowing from obstacles that affect the propagation of a transmitted wave and the path loss when a mobile moves through a large distance. On the other hand, small-scale fading mainly refers to multi-path fading and the time variance in a short distances, among which the multipath fading also can be subdivided into flat-fading and frequency-selective fading depending on the signal bandwidth and the channel bandwidth. The transmitted signal is subject to flat fading as long as the bandwidth of the wireless channel is wider than that of the signal bandwidth. On the contrary, the received signal will



Figure 1.2: The radio propagation model.

undergo the frequency-selective fading. Meanwhile, time variation can be classified as either fast fading or slow fading. Slow and fast fading can be distinguished by the changing speed of the signal magnitude and phase. Slow fading arises when the coherence time of the channel is large relative to the delay and the signal is slowly changing while the fast fading occurs when the coherence time of the channel is small relative to the delay and the signal has rapid fluctuations in a short period of time.

Link budget in Figure 1.3 is to show all the gains and losses from the wireless channel to the receiver. Path loss and fading are important factors in link budget [2]. In general, shadowing is modeled by a log-normal distribution. The received signal power must be maintained within 1% - 2% percent in term of the outage rate.



Figure 1.3: The link budget of the fading channel.

1.3 OFDM Strengths and Weaknesses

In conventional multi-carrier transmission system, the whole system band is divided into several narrowband sub-channels, which can be regarded as a kind of FDMA (frequency division multiple access) method. The receiver separates each sub-channel through a filter to obtain the required information. It can avoid interference from different channels but at the expense of frequency utilization costs. When there are a large number of sub-channels, the filter settings for each subchannel signal have become almost impossible. However, OFDM can fully utilize channel bandwidth, avoid high-speed equalization and could confront impulse noise and propagation fading.

OFDM is a parallel multi-carrier transmission system, by extending the transmission symbol period to enhance its ability to resist multipath. Compared with traditional equalizer, it has a simple structure, which can greatly reduce the cost for practical application. OFDM has high spectrum efficiency and a strong antifrequency selective fading ability, it is a very suitable high-speed data transmission technology in wireless environment, which can provide higher quality data services and offer a better solution for 4G/5G wireless network. To solve the diversified problems, Several flexible multi-carrier modulation schemes have been proposed based on the conventional OFDM including generalized frequency division multiplexing (GFDM) and Filter-OFDM (F-OFDM) [3]-[5] to meet the requirements of higher spectrum efficiency and massive connections.



Figure 1.4: The FDM and OFDM transmission schemes.

OFDM main advantages:

(1) OFDM technology can combat the frequency-selective fading and narrowband interference. In a single-carrier system, the fading and interference can cause the entire communication system crashes, but in a multi-carrier system, only a small part of the carriers will be disturbed. These sub-channels can also be corrected by employing error correction code.

(2) High spectral efficiency. This is important in currently scarce spectrum

resources in wireless environments as shown in Figure 1.4. When the number of subcarriers is large, the spectral efficiency of the system tends to 2Baud/Hz.

(3) Simplifies the design of equalizer, or do not need the equalizer, and the data transfer rate is adjustable. OFDM also uses the power control and adaptive modulation coordinated work.

OFDM major deficiencies:

(1) It is sensitive to the frequency offset, timing offset and phase noise. Transmission nonlinearities will cause intermodulation distortion (IMD). At this time, the signal has a high noise level, the signal to noise ratio is generally not too high, and inaccurate synchronization or Doppler shifts will destroy the orthogonality between the sub-channels. Only of 1 percent the offset will result in 30dB decrease of the SNR. So the precise timing and frequency synchronization is particularly important. If the orthogonality can not be guaranteed, it will have a great impact on the system performance.

(2) The high PAPR is one of the most detrimental aspects especially in the uplink since the efficiency of power amplifier is critical due to the limited battery power in a mobile terminal.

(3) The adaptive frequency hopping will increase the complexity of the transmitter and receiver.

1.4 MIMO-OFDM Critical Technologies in 5G

The fifth generation of mobile communication technology provides up to 7.5Gb/sor even higher data rates, supporting from voice to multimedia services, commercial wireless networks, LAN, Bluetooth, TV, satellite communications to achieve seamless connectivity. Data transmission rate can also be dynamically adjusted according to the desired rate. To achieve such a high-speed and large-capacity communication system on the limited spectrum resources, we must improve the bandwidth efficiency. OFDM is efficient technology to use the spectrum resources. MIMO systems can transmit data at a higher speed using antenna diversity techniques without increasing bandwidth and transmission power, it can exponentially increase the channel capacity. MIMO combined with OFDM can overcome frequency selective fading, increase system capacity and spectrum efficiency, it has become one of the key technologies in 4G/5G.

MIMO wireless communications is a major breakthrough in smart antenna, if we have independent channel impulse response, MIMO can create multiple parallel spatial channels. The transmission rate is bound to increase through these parallel spatial channels [6]. Since the signals transmitted from multiple antennas occupy the same frequency band, which can exponentially increase spectral efficiency without increasing the bandwidth as shown in Figure 1.5. With the increases of the antenna numbers, the system capacity and SNR almost linear relationship.



Figure 1.5: The MIMO transmission structure.

critical technologies in MIMO-OFDM system

(1) channel estimation: The purpose of channel estimation is to identify each channel impulse response between transmit and receive antennas. Currently, there are two kinds channel estimation: one is using the training sequence or pilot-based methods, such methods are applicable to time-varying channel, which require periodic training sequences that occupy part of the channel resources, thereby reducing the channel utilization rate. The advantage is that the estimated error is small and fast convergence. The other is blind or semi-blind channel estimation. Blind channel estimation is to use the statistical information to estimate the channel parameters without knowing the pilot or training sequence. These methods have high transmission efficiency for high-speed digital communication, but relatively poor robustness, slow convergence and high computational complexity. Semi-blind treatment is a combination of blind and a small pilot signal or training sequence.

(2) Synchronization: MIMO-OFDM system is still sensitive to timing and frequency offset, so the time-frequency synchronization is particularly important. MIMO-OFDM system synchronization issues including carrier synchronization, symbol synchronization. Carrier frequency offset will destroy the orthogonality between subcarriers, causes amplitude attenuation and signal phase rotation of the output signal, moreover, it will bring more serious ICI, which affect the performance of symbol timing and frame synchronization. Symbol timing is to find the start position of the FFT window, so that the orthogonality is maintained and ISI is eliminated or minimized.

(3) diversity techniques: Wireless communication unreliability is mainly caused by time-varying wireless fading channel and multi-path fading channel. An effective method to overcome above shortcoming are diversity techniques. The fundamental goal of the diversity techniques is to convert an unstable time-varying wireless fading channel into a stable AWGN-like channel without significant instantaneous fading, thereby steepening the BER versus SNR curve. Diversity is currently divided into time diversity, frequency diversity and space diversity and so on. Time diversity is to provide a plurality of signals in the time domain, in order to obtain a good diversity effect, it requires a number of independent time slots that used to transmit signal. Frequency diversity is to provide multiple signal copies on different carrier frequencies, which requires the carrier frequency spacing is greater than the coherence bandwidth, thereby obtaining a better diversity gain. Spatial diversity is the use of multiple antennas to send and receive data, in order to ensure independence among the transmitting or receiving signals, it requires large distance between each antenna. Generally, it is greater than several wavelengths. Each technology has its possible applications, so in the next generation mobile communication system, we must consider the combination of multiple technologies.

OFDM Application and Prospect in the fifth generation mobile communication: 5G is a wireless mobile communication system for the needs of the information society. The goal of 5G is to meet the future development of more than 1000 times the traffic growth or the peak rate of 10Gbit/s and the user experience rate of 100Mbit/s. 5G network can continue to reduce the costs and energy consumption, not only has a big improvement on the speed and throughput, but also has the ability to support business expansion and to improve the reliability of the network, allowing users to get the full experience of science and technology progress. 5G will realize the network integration, network interconnection between the various types of equipment. Furthermore, the network has a stronger self-test, self-healing capabilities, which greatly improve network stability and adaptability. However, despite the advantages of such a property, OFDM / OFDMA also has some shortcomings, the strict time-frequency synchronization constraints have some limitations on the applications in the 5G. In particular, a key issue is the uplink synchronization at different mobile terminal in the cellular network and coordinated use of the base station in the downlink. 5G cellular systems will have some innovative strategies for the existing LTE system, including extensive use of small cells, using millimeter (mm) short-wave communication, a large antenna array that mounted in macro base stations, cloud-based radio access network. All these strategies will influence the physical layer modulation formats. At the same time, 5G mobile network will have more stringent requirements than the LTE on delay, and the use of modulation schemes. In the future 5G, under the premise of preserving OFDM advantage, we must combine with other technologies to make up their own technical difficulties and deficiencies.

1.5 Research Focus of the Thesis

OFDM has been considered as a promising technology in wideband digital communication for high data rate applications. Due to higher spectrum efficiency and robustness to multipath fading channels, it has been adopted in various communication standards such as worldwide interoperability for microwave access (WiMAX), wireless local area networks (WLAN), and digital terrestrial video broadcasting (DVB-T). OFDM is sensitive to the distortion associated with the carrier signal including mismatches between I and Q branches in the direct conversion receivers and the CFO due to Doppler frequency or the physically inherent nature of the oscillators. In this case, the orthogonality between subcarriers is destroyed, which implies a subcarrier frequency component is affected by other subcarrier frequency components that leads to ICI. To mitigate the disadvantage, numerous algorithms have been developed in improving the system performance including schemes with improved decoding reliability, windowing function in the time domain, diversitybased ICI cancellation approach, frequency-domain equalization, carrier frequency offset estimation and compensation methods. However, these methods require high implementation complexity and there has been almost no consideration for the possible influence of I/Q imbalance. On the other hand, most of the existing compensation methods for I/Q imbalance are without considering CFO. It is critical to have an efficient and convenient way to solve the CFO problems in the presence of I/Qimbalance. Recently, the ICI self-cancellation (ICI SC) techniques that it is less sensitive to CFO have been extensively investigated. The basic idea of the general ICI SC schemes is that the same data symbol is mapped onto two subcarriers with predefined weighting coefficients to cancel the ICI components. Zhao et al. presented a simple yet effective way for ICI SC that modulate the same data symbol on two adjacently located subcarriers with predefined inversed phase coefficients and the ICI components can be eliminated to some extent either using first-order constant method or first-order linear method. Plural weighted data-conversion algorithm can mitigate phase error and robust to the impact of frequency offset at low CFO. ICI SC of data-conjugate method is studied to reduce ICI effectively. Since the received signal is considered as the circular convolution of transmitted signal, the symmetric conjugate weighting coefficients are used to eliminate the effect of phase noise. This technique can tolerate phase noise of 7 deg with a BER of 10^{-3} and CIR performance is improved. Some methods utilize weighted conjugate transformation with different phase angles. However, it lacks theoretical analysis to determine rotated phases. To analyze the CIR and BER in a generalized frame work, we proposed a real constant and plural weighted data-conjugate scheme in the transmitter, furthermore, the modulation mode of desired signal after I/Q imbalance compensation is discussed in receiver. In the proposed schemes, different data allocation is used to modulate one data symbol onto the next subcarrier with predefined weighting coefficients to mitigate the phase error after I/Q imbalance compensation.

Accurate symbol timing offset (STO) must be implemented since OFDM systems are very sensitive to nonlinear distortion and synchronization errors will result in ISI and ICI. The aim of timing synchronization is to obtain a starting point of a FFT window so that the transmitted datas can be demodulated correctly in the receiver. To avoid these disadvantages, such as ISI and ICI, numerous algorithms have been proposed on the subject of timing synchronization. For non-preambleaided algorithms, the STO can be estimated by minimizing the squared difference between two sliding windows of N_{cp} samples. Although the scheme is simple and effective, it is easily affected by CFO. The algorithm of the maximum correlation (MC) performs better, but still has some problems when CFO exists in the receiving end. The maximum likelihood (ML) scheme is an effective way to find the STO by calculating the maximum value of the likelihood function, which effectively solves the CFO problems. However, it can only make a coarse estimation of STO. These methods without preamble are able to make effective use of spectrum resources, but the timing offset cannot be estimated or be determined accurately. To further enhance the estimation precision, the preamble-based algorithms have been widely

researched [7]-[9]. Some methods utilize more than two OFDM symbols to realize timing synchronization, which result in a waste of frequency resources. An OFDM symbol with a simple form of two identical preambles can be used, the designed repetitive training preamble can be produced by inserting zeros into the even position of the subcarriers in the frequency domain, whereas there exists a plateau in the metric function with the length of N_{cp} samples. To solve the problem, the training symbols are separated from two of the same data blocks into four of the same data blocks and the latter two pieces of blocks are weighted with the minus signs, but the declining curve is not steep enough, which gives rise to timing estimation inaccuracy to some extent. Conjugate symmetry preamble is used to estimate symbol timing offset, which can effectively solve the problem of the steepness of the declining curve. The value of timing metric function is close to the pulse form at the right timing position, nevertheless, there exists side lobe outside the main lobe. Another estimator was proposed by employing a constant envelop preamble with two identical parts weighted by single pseudo-noise (PN) sequence. Although the accuracy of STO is improved, whereas high precision still can not be achieved. In this paper, a novel timing estimator exploits four identical blocks with inverted signs of the latter two blocks and the correlation characteristics of PN sequences to estimate the timing offset. The preamble in the first half can be generated and repeat these samples with sign inversion to form the latter half. The features of the proposed approach is to utilize different PN sequences as weighted factors rather than transmitted signals to ensure that the metric function value reach its maximum at correct timing position. The first and second half of preamble are weighted with PN1 and PN2 respectively. By exploiting the double block correlation which are spaced N/4 samples apart, the metric function appears impulse-shaped form sharply without any large side lobes. When the block windows slide ahead, we introduce the same PN sequences like PN1 and PN2 into the block, which can eliminate the randomness only when the accurate location of the STO is detected. By this method, the difference between the adjacent values of the timing metric function is further enlarged, the performances are

dependent on the structure of the preamble, the property of PN sequences and the design of the metric function. Computer simulations demonstrate that the proposed scheme has a significant probability of getting the accurate estimation of timing offset and relatively smaller error mean and SD compared with the other estimators in flat-fading channels and performs good under frequency-selective fading channels when the FFT size is large in accordance with 3GPP LTE specifications.

Unlike single-carrier systems, OFDM systems demand high synchronization performance to avoid inter-symbol interference (ISI) and inter-channel interference (ICI) which will be generated when the orthogonality between each subcarriers is demolished. Synchronization inaccuracy will lead to some problems like sampling clock offset (SCO), symbol timing offset (STO), and carrier frequency offset (CFO). All of them will lead to a negative impact on the overall system performance especially in the latter two cases. Timing synchronization aims to find a appropriate FFT window so that the transmitted signal can be demodulated precisely in the receiver. In other words, we need to determine the position of the starting point to obtain the exact samples of the sending signal in the OFDM symbol duration. Note that as one kind of sampling offset, the phase offset in the sampling clocks (SPO) is normally small enough which can be considered as a part of STO that will result in frequency domain phase rotation that is proportional to the value of timing offset and subcarrier index. As mentioned above, symbol synchronization is generally divided into two major categories including preamble-aided and non-preamble-aided methods. For preamble-aided approaches, a single training symbol with identically repetitive structure of different periodic length in the time domain can be used. To effectively solve the problem that the declining curve is not steep enough, a circular conjugate symmetry training symbols are employed, so it makes the correct timing position close to the pulse form which can be correctly captured by the synchronizer. Another type of technique is to exploit a synchronization word and CP simultaneously, it provides better performance in the condition that the influence of the multipath is not serious. Using the symbol training sequence can obtain accurate delay estimation, but the introduction of extra redundancy will significantly reduce the bandwidth utilization efficiency. In order to make full use of spectrum resources, recently, blind synchronization (non-preamble-aided algorithms) algorithms have been widely studied [10]-[11]. The STO can be found by seeking the difference between two sliding blocks of N_{cp} samples, although the scheme is concise, it is susceptible to CFO and not stable. The algorithm of maximum likelihood (ML), minimum mean square error (MMSE) and the maximum correlation (MC) can achieve good performance, however, if the signal is subjected to deep frequency selective fading (FSF) channels, estimates point with respect to timing position will shift backwards accordingly due to the ISI. The estimator counteracting multipath fading has been developed, which can be considered as improved algorithm to MMSE and MC. The disadvantage of this approach is that the exact channel length is needed. An article proposed a differential algorithm, measuring the rate of change of the correlation results to determine the timing position. This algorithm is approximately 10 percents heavier than ML algorithm in terms of computational complexity and it still subject to multipath effects, with the increase of the number of channel path, timing performance gradually get worse. Another scheme was present by using 2-D search algorithm that needless to know the channel state information (CSI) which can accurately obtain the original start position of the FFT window, but the computation complexity is increased significantly. This algorithm can be improved by putting the two-dimensional search down to one dimensional search, which greatly reduced the computational complexity. However, it require large amount of OFDM samples, and the width of correlation window is 1, easily affected by noise, which is the deficiency of point correlation algorithm. To mitigate above-mentioned problems, in this paper, a novel timing estimator based on the redundancy of CP is proposed for preambleless OFDM systems over multipath fading channels. The most significant aspect of this approach is to utilize two functions including metric function and positioning function on the premise that we take distinctive correlation characteristics within CP region fully into account. By means of exploiting the block correlation

which are spaced N samples apart in ISI-free sampling region, the correlation value has a platform at the rear-end of CP accordingly. The selection of block length can be decided in our system. Moreover, in order to avoid the deficiencies of fluctuation and instability, a positioning function have been proposed instead of setting a threshold to find the first dip point that below the given value. It is proved that we can find the exact timing location with high accuracy and low-complexity compared with the conventional method. Numerical results show that the proposed approach achieves approximately accurate estimation of STO in high signal to noise ratio and manifests its robustness to multi-path fading channels.

The remainder of the paper is organized as follows. In Chapter 2, a joint I/Q imbalance compensation and ICI SC schemes are proposed in TFI-OFDM system. In Chapter 3, a novel timing synchronization algorithm is proposed based on packet oriented algorithms. In Chapter 4, a novel blind symbol timing synchronization scheme for real-time wireless control network is proposed over multi-path fading channel. Chapter 5 offers the conclusion.



Figure 1.6: The relationship between this dissertation and the previous work.

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Chapter 2

ICI Self-Cancellation Scheme with I/Q Imbalance Compensation in TFI-OFDM Systems

2.1 Introduction

Orthogonal Frequency Division Multiplexing (OFDM) has been considered as a promising technology in wideband digital communication for high data rate applications. Due to higher spectrum efficiency and robustness to multipath fading channels, it has been adopted in various communication standards such as worldwide interoperability for microwave access (WiMAX), wireless local area networks (WLAN), and digital terrestrial video broadcasting (DVB-T). OFDM is sensitive to the distortion associated with the carrier signal [1] including mismatches between I and Q branches in the OFDM direct conversion receivers and the CFO due to Doppler frequency or the physically inherent nature of the oscillators. In this case, the orthogonality between subcarriers is destroyed, which implies a subcarrier frequency component is affected by other subcarrier frequency components that leads to ICI. To mitigate the disadvantage, numerous algorithms have been developed in improving the system performance including schemes with improved decoding reliability [2], windowing function in the time domain [3], diversity-based ICI cancellation approach [4], frequency-domain equalization [5], carrier frequency offset estimation and compensation methods [6]-[9]. However, these methods require high implementation complexity and there has been almost no consideration for the possible influence of I/Q imbalance. On the other hand, most of the existing compensation methods for I/Q imbalance are without considering CFO [17]-[19]. It is critical to have an efficient and convenient way to solve the CFO problems in the presence of I/Q imbalance. Recently, the ICI self-cancellation (ICI SC) techniques that it is less sensitive to CFO have been extensively investigated. The basic idea of the general ICI SC schemes is that the same data symbol is mapped onto two subcarriers with predefined weighting coefficients to cancel the ICI components. Zhao et al. [10] presented a simple yet effective way for ICI SC that modulate the same data symbol on two adjacently located subcarriers with predefined inversed phase coefficients and the ICI components can be eliminated to some extent either using first-order constant method or first-order linear method. Plural weighted data-conversion algorithm in [11] can mitigate phase error compared to [10] and robust to the impact of frequency offset at low CFO. ICI SC of data-conjugate method is studied to reduce ICI effectively in [12]. Since the received signal is considered as the circular convolution of transmitted signal, the symmetric conjugate weighting coefficients are used to eliminate the effect of phase noise in [13]. This technique can tolerate phase noise of 7 deg with a BER of 10^{-3} [16] and CIR performance is improved. The methods in [14]-[15] utilize weighted conjugate transformation with different phase angles. However, it lacks theoretical analysis to determine rotated phases. To analyze the CIR and BER in a generalized frame work, we proposed a real constant and plural weighted data-conjugate scheme in the transmitter, furthermore, the modulation mode of desired signal after I/Q imbalance compensation is discussed in receiver. In the proposed schemes, different data allocation is used to modulate one data symbol onto the next subcarrier with predefined weighting coefficients to mitigate the phase



Figure 2.1: Block diagram of proposed scheme in the TFI-OFDM tranceiver.

error after I/Q imbalance compensation.

2.2 Signal Model and Analysis of ICI

2.2.1 TFI-OFDM Signal Model

In this paper, we use the TFI-OFDM system model [20]-[23] to clarity the benefit of the proposed method. The transmitter block diagram of TFI-OFDM system is illustrated in Figure 2.1. The coded input bit streams are mapped into a series of PSK or QAM symbols, the corresponding transmitted baseband OFDM signal can be described as

$$x(t) = \sum_{i=0}^{N_p + N_d - 1} g(t - iT) \cdot \left\{ \sqrt{\frac{2S}{N_c}} \sum_{k=0}^{N_c - 1} u(k, i) \\ \cdot \exp[j2\pi(t - iT)k/T_s] \right\},$$
(2.1)

where N_d and N_p are the number of data and pilot symbols, T_s is the effective symbol length. After S/P conversion, the duration of transmission time for useful part of each OFDM symbol is extended to N_cT_s . N_c is the total number of subcarriers, S is average transmitting power, g(t) is the transmission pulse, respectively. The frequency separation between adjacent orthogonal subcarriers is $1/T_s$, by using the kth subcarrier of the *i*th modulated symbol X(k,i) with |X(k,i) = 1| for $N_p \leq i \leq$ $N_p + N_d - 1$, the signal after adding scrambling code is given by

$$u(k,i) = c_{PN}(k) \cdot X(k,i), \qquad (2.2)$$

where c_{PN} is a long pseudo-noise (PN) sequence as a scrambling code to reduce the peak average power ratio (PAPR). The PAPR is much lower when we consider random phase. A cyclic prefix T_g is added into the front of each OFDM symbol to reduce the ISI between two successive OFDM symbols. For $0 \le i \le N_p - 1$, the transmitted pilot signal of kth subcarrier is given by

$$X(k,i) = \exp\{-j2\pi k/T_s\} + \exp\{-j4\pi kT_g/T_s\}.$$
(2.3)

In this case, the power of pilot signal is a half of the conventional pilot averaging for $0 \le i \le N_p - 1$. The received baseband signal y(t) can be expressed as

$$y(t) = \int_{-\infty}^{\infty} h(\tau, t) x(t-\tau) d\tau + n(t), \qquad (2.4)$$

where h and τ are the complex channel gain and the time delay, respectively. n(t) is additive Gaussian noise process with a single side-band power spectrum of N_0 . By applying FFT operation, the received signal y(t) is resolved into N subcarriers. The *k*th subcarrier $\overline{\overline{Y}}(k, i)$ is given by

$$\overline{\overline{Y}}(k,i) = \frac{1}{T_s} \int_{iT}^{iT+T_s} y(t) \exp[-j2\pi(t-iT)k/T_s] dt$$

$$= \sqrt{\frac{2S}{N_c}} \sum_{e=0}^{N_c-1} u(e,i) \cdot \frac{1}{T_s} \int_0^{T_s} \exp[j2\pi + (e-k)t/T_s] \left\{ \int_{-\infty}^{\infty} h(\tau,t+iT)g(t-\tau) + \exp(-j2\pi e\tau/T_s)d\tau \right\} dt + \hat{n}(k,i)$$

$$= \sqrt{\frac{2S}{N_c}} H(k,i)u(k,i) + \hat{n}(k,i), \qquad (2.5)$$

where H(f,t) is the Fourier transform of $h(\tau,t)$ named channel transfer function. $\hat{n}(k,i)$ is AWGN noise with zero-mean and a variance of $2N_0/T_s$. After descrambling, the output signal Y(k,i) is given by

$$Y(k,i) = \frac{c_{PN}^*(k)}{|c_{PN}(k)|^2} \overline{\overline{Y}}(k,i) + \hat{n}(k,i)$$

$$= \sqrt{\frac{2S}{N_c}} H(k,i) X(k,i) + \hat{n}(k,i), \qquad (2.6)$$

where $\frac{C_{PN}^{*}(k)}{|C_{PN}(k)|^2}$ is the descrambling operation.



Figure 2.2: CFO and I/Q imbalance model in a direct conversion receiver.

2.2.2 The Impact Caused by I/Q Imbalance and CFO

Figure 2.2 shows the CFO and I/Q imbalance model in a direct conversion receiver. The received baseband signal with I/Q imbalance and CFO can be represented by

$$\bar{y}(t) = \zeta e^{-j\Delta\omega t} y(t) + \xi e^{j\Delta\omega t} y^*(t), \qquad (2.7)$$

where $\bar{y}(t)$ is the baseband received signal. The signal gain ζ and mirror gain ξ are given as

$$\zeta = \cos(\phi) - j\Delta\sin(\phi) \tag{2.8}$$

$$\xi = \Delta \cos(\phi) + j \sin(\phi), \qquad (2.9)$$

where Δ and ϕ are amplitude and phase mismatches. From Eqs.(2.4) and (2.7), we obtain

$$\bar{y}(t) = \zeta e^{-j\Delta\omega t}(x(t) \otimes h(\tau, t)) + \xi e^{j\Delta\omega t}(x(t) \otimes h(\tau, t))^* + \omega(t), \qquad (2.10)$$

where \otimes represents the convolution operator and $\omega(t)$ is the noise at the output of the receiver front-end, it is given by

$$\omega(t) = \zeta e^{-j\Delta\omega t} (n(t) \otimes h(\tau, t)) + \xi e^{j\Delta\omega t} (n(t) \otimes h(\tau, t))^*.$$
(2.11)

After the FFT operation, Eq.(2.10) becomes

$$\begin{split} \tilde{Y}(k,i) &= \sqrt{\frac{2S}{N}} \Big[\zeta H(k,i) X(k,i) S(0) + \xi H^*(N-1-k,i) \\ &\cdot X^*(N-1-k,i) S(0)^* + \sum_{l \neq k} \zeta H(l,i) X(l,i) S(l-k) \\ &+ \sum_{l \neq N-1-k} \xi H^*(N-1-l,i) \\ &\cdot X^*(N-1-l,i) S(N-1-l-k) \Big] + \omega(k,i), \end{split}$$
(2.12)

where $\tilde{Y}(k,i)$ is received signal in frequency domain and $\omega(k,i)$ is the AWGN in the frequency domain. In the presence of the I/Q imbalance, the subcarriers will be interfered by frequency mirror-image subcarriers. The ICI due to both the CFO and the I/Q imbalance will lead to CIR degradation. Here, S(l-k) is defined as the complex coefficient for the ICI components between l^{th} and k^{th} subcarriers, which can be expressed as

$$S(l-k) = \frac{\sin[\pi(l+\varepsilon-k)]}{N\sin[\pi(l+\varepsilon-k)/N]} \cdot e^{j\pi(l+\varepsilon-k)(N-1)/N}$$
(2.13)

where ε is normalized frequency offset. Observing Eq.(2.12), we can see that the received signals have frequency distortion arising from the mirror signals and ICI. To reduce the effect of IQ imbalance and ICI due to CFO, proper estimation and compensation methods are necessary.

2.2.3 ICI Analysis

For convenience, I abbreviate X(k,i), H(k,i), $\omega(k,i)$, $\tilde{Y}(k,i)$ to X(k), H(k), $\omega(k)$, $\tilde{Y}(k)$ respectively. Due to the susceptibility to small differences in frequency at the OFDM transmitter and receiver, the k^{th} subcarrier contains the ICI from other subcarriers, which implies that the orthogonality among subcarriers is destroyed and not maintained any longer due to the CFO. Here we assume that the accurate CSI (channel state information) is known and $\zeta = 1$, $\xi = 0$ when the two paths are exactly in quadrature. The received frequency-domain symbol stream with CFO can be rewritten as follows

$$Y(k) = X(k)S(0) + \sum_{l=0, l \neq k}^{N-1} X(l)S(l-k) + \omega(k)$$
(2.14)

In order to analyze the influence of ICI from other subcarriers into k^{th} subcarrier, we assume the normalized carrier frequency offset values from 0.2 to 0.5 and N = 16. The complex ICI coefficients |S(l-k)| are plotted for all subcarriers in Figure 2.3 and Figure 2.4, respectively.

The two figures indicate that the weight of the desired signal component S(0) decreases whereas the undesired weights of the ICI components increase for a larger ε . Note that the adjacent carrier has the maximum contribution to the ICI. The CIR is an indicator that used as a measure of the carrier component power to the interference components power. Here we assume that the modulated data symbol X(k) is statistically independent and identically distributed, and the CIR of standard OFDM transmission is given by

$$CIR = \frac{E[|C(k)|]}{E[|ICI(k)|]} = \frac{E[|X(k)|^{2}] \cdot E[|S(0)|^{2}]}{E[|X(l)|^{2}]\sum_{l=0, l \neq k}^{N-1} |S(l-k)|^{2}}$$



Figure 2.3: ICI coefficients for the subcarriers from k=0 to N-1. (N = 16, l = 0).

$$=\frac{|S(0)|^2}{\sum_{l=1}^{N-1}|S(l)|^2}.$$
(2.15)

Equations (2.13), (2.15) show that CIR depends upon subcarrier number N and carrier frequency offset ε . Furthermore, the CIR has a maximum change of 0.068 dB when $N \geq 8$ [10]. Therefore, the CIR of normal OFDM systems can be considered as relation of the normalized frequency offset approximately.



Figure 2.4: ICI coefficients for the subcarriers from (-k)=0 to N-1. (N = 16, l = 0).

2.2.4 Conventional ICI Cancellation Algorithms

2.2.4.1 Data-conversion Algorithm

Data-conversion scheme in [10] is a simple and effective technique, the adjacent subcarrier signals are remapped in the form of

$$X'(k) = X(k), X'(k+1) = -X(k), \ k \in \{0, 2, ..., N-2\}.$$

In the receiver, the desired signal demodulation is performed as

$$Y''(k) = \frac{Y'(k) - Y'(k+1)}{2}$$

= $\frac{1}{2} \sum_{l=0,l=even}^{N-2} X(l) [-S(l-k-1)+2S(l-k)-S(l-k+1)]$
+ $\omega''(k).$ (2.16)
where

$$\begin{cases} Y'(k) = \sum_{l=0, l=even}^{N-2} X(l) [S(l-k) - S(l-k+1)] + \omega(k) \\ \omega''(k) = \omega(k) - \omega(k+1) \end{cases}$$
(2.17)

The theoretical CIR of this technique is given by

$$CIR = \frac{|2S(0) - S(1) - S(-1)|^2}{\sum_{l=2, l=even}^{N-2} |2S(l) + S(l+1) - S(l-1)|^2}$$
(2.18)

2.2.4.2 plural weighted data-conversion algorithm

This scheme [11] is based on the data allocation on adjacent subcarriers of

$$X'(k) = X(k), X'(k+1) = e^{-j\pi/2}X(k),$$

$$k \in \{0, 2, ..., N-2\}.$$

The demodulation scheme and the CIR have the form as follows

$$Y''(k) = \frac{Y'(k) - e^{-j\pi/2}Y'(k+1)}{2},$$
(2.19)

$$CIR = \frac{\left|2S(0) + e^{-j\pi/2}[S(1) - S(-1)]\right|^2}{\sum_{l=2, l=even}^{N-2} \left|2S(l) + e^{-j\pi/2}[S(l+1) - S(l-1)]\right|^2}.$$
(2.20)

2.2.4.3 Data-conjugate Algorithm

In date-conjugate ICI cancellation scheme [12], subcarrier signal is mapped in the form of

$$X'(k) = X(k), X'(k+1) = -X^*(k), \ k \in \{0, 2, ..., N-2\}.$$

The demodulation scheme and the CIR have the form as follows

$$Y''(k) = \frac{Y'(k) - Y'^{*}(k+1)}{2},$$
(2.21)

$$CIR = \frac{|S(0) + S^*(0)|^2 + |S(1) - S^*(-1)|^2}{\sum_{l=2, l=even}^{N-2} |S(l) + S^*(l)|^2 + |S(l+1) - S^*(l-1)|^2}.$$
(2.22)

2.2.4.4 Symmetric data-conjugate scheme

The symmetric data-conjugate scheme [13] is based on the data allocation of

$$\begin{split} X(N-1) &= -X^*(0), X(N-2) = -X^*(1), \\ X(N-1-k) &= -X^*(k), k \in \{0, 2, ..., N/2 - 1\} \end{split}$$

The demodulation scheme and the CIR have the form as follows

$$Y''(k) = \frac{Y'(k) - Y'^*(N - 1 - k)}{2},$$
(2.23)

$$CIR = \frac{|S(0) + S^*(0)|^2}{\sum_{l=0}^{N/2-1} |S(N-1-l) + S^*(l-N+1)|^2 + \sum_{l=1}^{N/2-1} |S(l) + S^*(-l)|^2}.$$
(2.24)

2.2.4.5 Fixed plural weighted data-conjugate algorithm A

This scheme [14] is based on the data allocation of

$$X'(k) = X(k), X'(k+1) = e^{j\pi/2} X^*(k), \ k \in \{0, 2, ..., N-2\}.$$

The demodulation scheme and the CIR have the form as follows

$$Y''(k) = \frac{Y'(k) - e^{-j\pi/2}Y'^*(k+1)}{2},$$
(2.25)

$$CIR = \frac{|S(0)+S^*(0)|^2 + \left|e^{j\pi/2}S(1)-e^{-j\pi/2}S^*(-1)\right|^2}{\sum_{l=2,l=even}^{N-2} |S(l)+S^*(l)|^2 + \left|e^{j\pi/2}S(l+1)-e^{-j\pi/2}S^*(l-1)\right|^2}.$$
(2.26)

2.2.4.6 Fixed plural weighted data-conjugate algorithm B

This scheme [15] is based on the data allocation of

$$\begin{split} X^{'}(k) &= X(k), X^{'}(k+1) = e^{-j\pi/2} X^{*}(k) \,, \\ &\quad k \in \{0,2,...,N-2\}. \end{split}$$

The demodulation scheme and the CIR have the form as follows

$$Y''(k) = \frac{Y'(k) - e^{j\pi/2}Y'^*(k+1)}{2},$$
(2.27)

$$CIR = \frac{|S(0)+S^*(0)|^2 + \left|e^{-j\pi/2}S(1) - e^{j\pi/2}S^*(-1)\right|^2}{\sum_{l=2, l=even}^{N-2} |S(l)+S^*(l)|^2 + \left|e^{-j\pi/2}S(l+1) - e^{j\pi/2}S^*(l-1)\right|^2}.$$
(2.28)

2.3 Proposed Joint I/Q Imbalance Compensation and ICI SC Schemes

2.3.1 I/Q imbalance estimation and compensation

In this paper, we use the I/Q Imbalance estimation method based on TFI-OFDM in [23] to achieve an accurate compensation with small number of pilot symbols. The pilot signal of the kth subcarrier can be expressed as

$$\begin{split} \tilde{Y}(k,i) &= \sqrt{\frac{2P}{N}} \Big[\zeta H(k) X(k) S(0) + \xi H^* (N-1-k) \\ &\cdot X^* (N-1-k) S(0)^* + \sum_{l \neq k} \zeta H(l) X(l) S(l-k) \\ &+ \sum_{l \neq N-1-k} \xi H^* (N-1-l) \\ &\cdot X^* (N-1-l) S(N-1-l-k) \Big] + \omega(k), \end{split}$$
(2.29)

where P is the power of pilot signals. Here, we define V(k) as

$$V(k) = \begin{cases} \frac{\tilde{Y}^*(N-1-k)}{\tilde{Y}(k)} & k \text{ for odd} \\ \frac{\tilde{Y}^*(k)}{\tilde{Y}(N-1-k)} & k \text{ for even.} \end{cases}$$
(2.30)

The transmitted pilot signal X(k) in TFI-OFDM is given by Eq.(2.3). Due to the characteristics of X(k), the average power of pilot signal is 1/2. We can obtain $\sum_{k=0}^{N_c-1} X(k) = \{1, 0, \dots, 1, 0\}$ as the pilot signal from Eq.(2.3) and separate the desired and mirror signals. By using Eq.(2.3), Eq.(2.30) can be rewritten as

$$V(k) = \frac{\xi^* \sqrt{\frac{2P}{N}} H(k) X(k) + \omega^*(k)}{\zeta \sqrt{\frac{2P}{N}} H(k) X(k) + \omega(k)}.$$
(2.31)

For simplicity, assuming that the noise power is small and can be ignored, Eq.(2.31) can be further rewritten as

$$V(k) = \frac{\xi^*}{\zeta} = \frac{\Delta \cos(\phi) - j\sin(\phi)}{\cos(\phi) - j\Delta\sin(\phi)}.$$
(2.32)

Observing Eq.(2.32), if ϕ is small, we can approximate $\Delta \sin \phi \approx 0$. Eq.(2.32) can be written as

$$V(k) = \frac{\xi^*}{\zeta} \approx \Delta - j \tan \phi.$$
(2.33)

As a result, we can obtain Δ and ϕ as

$$\Delta = \operatorname{Re}\{V(k)\} \tag{2.34}$$

$$\phi = \arctan\left[-\operatorname{Im}\{V(k)\}\right]. \tag{2.35}$$

We can select N_s subcarriers of the largest power and calculate V[k] by Eq.(2.30). After that we can reduce the influence of the noise power by averaging V[k] and then I/Q imbalances can be estimated by using Eqs.(2.34) and (2.35).

I/Q imbalance compensation is easy to implement in time domain [24] by using

$$\hat{y}_{com}(t) = \frac{\zeta^* \bar{y}(t) - \xi \bar{y}^*(t)}{|\zeta|^2 - |\xi|^2},\tag{2.36}$$

where $\hat{y}_{com}(t)$ is the compensated received signal in time domain.

2.3.2 Proposed Transceiver Model of ICI SC scheme

After I/Q imbalance compensation, a novel real constant and plural weighted data-conjugate ICI-Cancellation scheme is proposed in this paper. The data symbols are remapped as the form of

$$X'(k) = \alpha X(k), X'(k+1) = \beta e^{-j\varphi} X^*(k),$$

$$k \in \{0, 2, ..., N-2\}.$$

All data symbol is mapped onto a pair of two adjacent subcarriers in frequency domain to achieve frequency diversity. Unlike the conventional methods, it can be proved that the proposed method is robust to the artificial phase rotation φ at any angel in the evaluation of CIR. Furthermore, the optimal values of α and β are achieved not only considering the CIR but also the BER performance, which is more important. Hence, the received data signal at the k^{th} and $(k + 1)^{th}$ subcarriers are

$$Y'(k) = \sum_{l=0, l=even}^{N-2} \alpha X(l) S(l-k) + \beta e^{-j\varphi} X(l)^* S(l+1-k) + \omega(k)$$
(2.37)

$$Y'(k+1) = \sum_{l=0, l=even}^{N-2} \alpha X(l) S(l-k-1) + \beta e^{-j\varphi} X(l)^* S(l-k) + \omega(k+1)$$
(2.38)

After combining the real constant and plural weighted coefficients, the received signal is determined as

$$Y''(k) = \frac{1}{\alpha - \beta} [Y'(k) - e^{-j\varphi} Y'^{*}(k+1)]$$

$$= \frac{1}{\alpha - \beta} \sum_{l=0, l=even}^{N-2} \{X(l)[\alpha S(l-k) - \beta S^{*}(l-k)]]$$

$$+ X^{*}(l)[\beta e^{-j\varphi} S(l+1-k) - \alpha e^{-j\varphi} S^{*}(l-k-1)]\} + \omega'(k)$$

$$= \frac{1}{\alpha - \beta} \{\underbrace{X(k)[\alpha S(0) - \beta S^{*}(0)] + X^{*}(k)[\beta e^{-j\varphi} S(1) - \alpha e^{-j\varphi} S^{*}(-1)]}_{C(k)}$$

$$+ \sum_{l=0, l \neq k, l=even}^{N-2} \{X(l)[\alpha S(l-k) - \beta S^{*}(l-k)] + X^{*}(l)[\beta e^{-j\varphi} S(l+1-k) - \alpha e^{-j\varphi} S^{*}(l-k-1)]\}$$

$$ICI(k)$$

$$+ \omega'(k). \qquad (2.39)$$

Thus, CIR of proposed ICI SC scheme is given by

$$CIR_{Pro} = \frac{m(\alpha, \beta, \varphi)}{n(\alpha, \beta, \varphi)}$$

$$= \frac{|\alpha S(0) - \beta S^*(0)|^2 + |\beta e^{-j\varphi} S(1) - \alpha e^{-j\varphi} S^*(-1)|^2}{\sum_{l=2, l=even}^{N-2} |\alpha S(l) - \beta S^*(l)|^2 + |\beta e^{-j\varphi} S(l+1) - \alpha e^{-j\varphi} S^*(l-1)|^2}.$$
(2.40)

In order to get the optimum value of CIR at the given values of ε and N, we have to analyse the proper numerical coefficients α , β and the phase rotation factor φ . This is equivalent to finding the maximum value of the ternary function. Derivation process is as follows

$$m(\alpha, \beta, \varphi) = [\alpha S(0) - \beta S^{*}(0)][\alpha S^{*}(0) - \beta S(0)] + [\beta e^{-j\varphi} S(1) - \alpha e^{-j\varphi} S^{*}(-1)[\beta e^{j\varphi} S^{*}(1) - \alpha e^{j\varphi} S(-1)] = (\alpha^{2} + \beta^{2})|S(0)|^{2} - \alpha\beta(S^{2}(0) + S^{*2}(0) + S(1)S(-1) + S^{*}(1)S^{*}(-1)) + \alpha^{2}|S(-1)|^{2} + \beta^{2}|S(1)|^{2}$$
(2.41)

Similarly

$$n(\alpha, \beta, \varphi) = \sum_{l=2, l=even}^{N-2} (\alpha^2 + \beta^2) |S(l)|^2 - \alpha\beta (S^2(l) + S^{*2}(l) + S(l+1)S(l-1) + S^*(l+1)S^*(l-1)) + \alpha^2 |S(l-1)|^2 + \beta^2 |S(l+1)|^2$$

$$(2.42)$$

take the partial derivative with respect to φ , let

$$\frac{\partial CIR}{\partial \varphi} = \frac{\frac{\partial m}{\partial \varphi} \cdot n - \frac{\partial n}{\partial \varphi} \cdot m}{n^2} = 0$$
(2.43)

where $\frac{\partial m}{\partial \varphi} = 0$, $\frac{\partial n}{\partial \varphi} = 0$. The specially designed modulation approach allows the phase rotation at any angel since there is no variable in Eq.(2.41) and Eq.(2.42). Next, take the partial derivative with respect to α , let

$$\frac{\partial CIR}{\partial \alpha} = \frac{\frac{\partial m}{\partial \alpha} \cdot n - \frac{\partial n}{\partial \alpha} \cdot m}{n^2} = 0$$
(2.44)

$$\frac{\partial m}{\partial \alpha} = 2\alpha |S(0)|^2 - \beta (S^2(0) + S^{*2}(0) + S(1)S(-1)) + S^*(1)S^*(-1)) + 2\alpha |S(-1)|^2, \qquad (2.45)$$

$$\frac{\partial n}{\partial \alpha} = \sum_{l=2, l=even}^{N-2} 2\alpha |S(l)|^2 - \beta (S^2(l) + S^{*2}(l) + S(l+1)S(l-1) + S^*(l+1)S^*(l-1)) + 2\alpha |S(l-1)|^2.$$
(2.46)

After substituting the values from Eq.(2.45) and Eq.(2.46) into Eq.(2.44), we have

$$C_0 \alpha^3 + C_1 \alpha^2 + C_2 \alpha + C_3 = 0 \tag{2.47}$$

where

$$\begin{cases} C_{0} = 0 \\ C_{1} = (S^{2}(0) + S^{*2}(0) + S(1)S(-1) + S^{*}(1)S^{*}(-1))\sum_{l=2,l=even}^{N-2} (|S(l)|^{2} + |S(l-1)|^{2}) \\ -(|S(0)|^{2} + |S(-1)|^{2})\sum_{l=2,l=even}^{N-2} S^{2}(l) + S^{*2}(l) + S(l+1)S(l-1) + S^{*}(l+1)S^{*}(l-1) \\ C_{2} = 2\beta \{ (|S(0)|^{2} + |S(-1)|^{2})\sum_{l=2,l=even}^{N-2} |S(l+1)|^{2} + (|S(-1)|^{2} - |S(1)|^{2})\sum_{l=2,l=even}^{N-2} |S(l)|^{2} \\ -(|S(0)|^{2} + |S(1)|^{2})\sum_{l=2,l=even}^{N-2} |S(l-1)|^{2} \} \\ C_{3} = \beta^{2} \{ (|S(0)|^{2} + |S(1)|^{2})\sum_{l=2,l=even}^{N-2} |S(l)| + S^{*2}(l) + S(l+1)S(l-1) + S^{*}(l+1)S^{*}(l-1) \\ -(S^{2}(0) + S^{*2}(0) + S(1)S(-1) + S^{*}(1)S^{*}(-1))\sum_{l=2,l=even}^{N-2} |S(l)|^{2} + |S(l+1)|^{2} \} \end{cases}$$

In the same way, take the partial derivative with respect to β

$$\frac{\partial CIR}{\partial \beta} = \frac{\frac{\partial m}{\partial \beta} \cdot n - \frac{\partial n}{\partial \beta} \cdot m}{n^2} = 0$$
(2.49)

$$\frac{\partial m}{\partial \beta} = 2\beta |S(0)|^2 - \alpha (S^2(0) + S^{*2}(0) + S(1)S(-1)) + S^*(1)S^*(-1)) + 2\beta |S(1)|^2, \qquad (2.50)$$

$$\frac{\partial n}{\partial \beta} = \sum_{l=2,l=even}^{N-2} 2\beta |S(l)|^2 - \alpha (S^2(l) + S^{*2}(l) + S(l+1)S(l-1) + S^*(l+1)S^*(l-1)) + 2\beta |S(l+1)|^2.$$
(2.51)

Substituting Eq.(2.50) and Eq.(2.51) into Eq.(2.49), it can be formulated as

$$C_{0}'\beta^{3} + C_{1}'\beta^{2} + C_{2}'\beta + C_{3}' = 0$$
(2.52)

$$\begin{cases} C_{0}^{\prime} = 0 \\ C_{1}^{\prime} = (S^{2}(0) + S^{*2}(0) + S(1)S(-1) + S^{*}(1)S^{*}(-1)) \sum_{l=2,l=even}^{N-2} (|S(l)|^{2} + |S(l+1)|^{2}) \\ - (|S(0)|^{2} + |S(1)|^{2}) \sum_{l=2,l=even}^{N-2} S^{2}(l) + S^{*2}(l) + S(l+1)S(l-1) + S^{*}(l+1)S^{*}(l-1) \\ C_{2}^{\prime} = 2\alpha \{ (|S(0)|^{2} + |S(1)|^{2}) \sum_{l=2,l=even}^{N-2} |S(l-1)|^{2} + (|S(1)|^{2} - |S(-1)|^{2}) \sum_{l=2,l=even}^{N-2} |S(l)|^{2} \\ - (|S(0)|^{2} + |S(-1)|^{2}) \sum_{l=2,l=even}^{N-2} |S(l+1)|^{2} \} \\ C_{3}^{\prime} = \alpha^{2} \{ (|S(0)|^{2} + |S(-1)|^{2}) \sum_{l=2,l=even}^{N-2} |S(l)| + S^{*2}(l) + S(l+1)S(l-1) + S^{*}(l+1)S^{*}(l-1) \\ - (S^{2}(0) + S^{*2}(0) + S(1)S(-1) + S^{*}(1)S^{*}(-1)) \sum_{l=2,l=even}^{N-2} |S(l)|^{2} + |S(l-1)|^{2} \} \end{cases}$$

Through the analysis of equation Eq.(2.47) and Eq.(2.52), we can find the optimal solutions of α and β from the above equations theoretically at a given ε under $\alpha \neq 0$ and $\beta \neq 0$. We can see the relative variation regularity from Figure 2.5 and Figure 2.6. The optimum CIR are taken with the approximately inverted signs of α and β for normalized $CFO \in [0, 0.15]$.

2.3.3 Energy Clipping Scheme in The Transmitter to Achieve Better BER

In order to get the best performance after I/Q imbalance compensation, based on the point of opposite in sign between α and β as discussed above, we have found that the energy clipping in the transmitted signal can contribute a lot to the BER performance by using the proposed demodulator. That is, in this way, the transmitted weighted coefficients α and β are given as $\alpha=0.5$, $\beta=-0.5$, respectively. We can get the conclusion by simulations that the reduction of signal energy in the transmitter are much more effective than the manipulation between adjacent subcarriers in the receiver.

2.4 Simulation Results

In this section, Monte Carlo simulations have been conducted to evaluate the performance of proposed scheme compared to the existing methods. On the transmitter, the pilot signals are assigned by using Eq.(2.3). In this case, the TFI-OFDM system can multiplex the same impulse responses in twice on the time domain without overlapping each other. Here, the convolution codes (rate R = 1/2, constraint length $\kappa = 7$) with bit interleaving are used. The packet consists of $N_p = 1$ pilot symbol and $N_d = 20$ data symbols. The subcarriers of 256 samples (N = 256) and CP length of 16 samples ($N_{cp} = 16$) are considered. Modulation type is QPSK and unless stated otherwise, the simulation is run by 1000 trials. The system per-

Data Modulation	QPSK
FEC	Convolutional code
	$(R=1/2,\kappa=7)$
Frame size	21 symbol
	$(N_p = 1, N_d = 20)$
FFT size	256
Number of subcarrier	256
Guard interval	16 sample times
Applicable channel model	AWGN, flat-fading
The gain mismatch Δ	0.2
phase mismatch ϕ	10°
Weighted coefficients $\alpha \beta$	0.5 and -0.5
CFO ϵ	0.001 0.05 0.1 0.15

Table 2.1: Simulation parameters.

formance is evaluated in terms of BER besides CIR. As mentioned above, $\alpha=0.5$, $\beta=-0.5$ are chosen respectively. The conventional algorithms were also implemented for the purpose of comparison. In this project, the simulations were performed in an AWGN channel model and it can be easily adapted to a flat-fading channel with channel estimation. The self cancellation schemes have widely been used because it does not require very complex hardware or software for implementation. Table 2.1 shows the simulation parameters.

The theoretical CIR curve of conventional and the proposed ICI SC schemes are shown in Figure 2.7. Algorithms in [10]-[11] were labeled as data-conversion, plural weighted data-conversion, algorithms in [12]-[13] were labeled as data-conjugate, symmetric data-conjugate, and algorithms in [14]-[15] were labeled as fixed plural weighted data-conjugate algorithm A and B, respectively. The CIR performance of ICI SC scheme is found to be better than that of the standard-OFDM system. Furthermore, it can be seen that the CIR performance of proposed scheme is inferior than the methods in [10], [11], but higher than the normal OFDM and has nearly identical performance as the methods in [13], [14], and [15]. From Fig. 2.7, we know, in general, the data conversion methods have better CIR performance than conjugate methods. Note that CIR is one indicator of system performance, we also have to evaluate the BER performance. There are some relationship between CIR and BER from the analysis of previous methods but no data to indicate that they are directly related to the each other.

Figure 2.8-Figure 2.11 show the BER performance of proposed method compared to the conventional method after I/Q imbalance compensation. As seen in these figures, the proposed algorithm performs better than the others for different values of CFO. It depends on the design of the data allocation in the transmitter and the demodulation scheme in the receiver. When CFO is very small as shown in Fig. 2.8, except for proposed scheme and plural weighted data-conversion method, the dataconversion method outperforms the others at a low SNR below 6dB while inferior than others when SNR is greater than it. From Figure 2.9-Figure 2.10, we know that both the data-conversion method and plural weighted data-conversion method are inferior than the other approaches. Among the comparatively good schemes, fixed plural weighted data-conjugate algorithm A exhibits better performance than B when SNR is less than 10dB in Figure 2.9 whereas B is superior than A when SNR is less than 10dB in Figure 2.10. As described in [15], data-conjugate algorithm B gives better BER than A as ε increases. Overall, symmetric data-conjugate scheme and data-conjugate scheme have almost the same performance.

To make clear what are the effectiveness of proposed method when considering the actual OFDM, the system of 2GHz 3GPP and 2.3GHz mobile wimax are considered here. The subcarrier spacing are 15kHz and 9.769kHz respectively. The proposed algorithm can support the moving velocity up to 409km/h and 229km/h if the 10^{-3} BER is required when SNR equals 10dB. Analytical and simulation results show that the algorithm achieves better performance than other algorithms due to its higher utilizability and more robust property to normalized CFO.

2.5 Conclusion

In this paper, a novel ICI SC scheme has been presented in the presence of I/Q imbalance to improve the system performance in TFI-OFDM systems. The proposed ICI cancellation scheme repeats the data symbol with phase rotation and constant weighted data-conjugate over a pair of subcarriers. Unlike conventional methods, the designed transmission model is robust to phase rotation and it achieves relatively better CIR and the best BER performance for lower and higher values of CFOs. Due to the repetition symbols on adjacent subcarriers, it achieves frequency diversity by reducing bandwidth efficiency. However, the proposed scheme is effective for combating the impact of ICI and easy for hardware implementation since no further channel equalization is needed and the system complexity is not increased. The proposed scheme shows good property compared with conventional methods for practical application.



Figure 2.5: CIR versus β . ($N = 256, \alpha = 1$).



Figure 2.6: CIR versus α . $(N = 256, \beta = -1)$.



Figure 2.7: CIR for the conventional and proposed algorithm versus Normalized frequency offset.



Figure 2.8: BER for the conventional and proposed algorithm versus SNR with I/Q imbalance compensation ($N = 256, \varepsilon = 0.001$).



Figure 2.9: BER for the conventional and proposed algorithm versus SNR with I/Q imbalance compensation ($N = 256, \varepsilon = 0.05$).



Figure 2.10: BER for the conventional and proposed algorithm versus SNR with I/Q imbalance compensation ($N = 256, \varepsilon = 0.1$).



Figure 2.11: BER for the conventional and proposed algorithm versus SNR with I/Q imbalance compensation ($N = 256, \varepsilon = 0.15$).

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Chapter 3

Time Synchronization Approach for Coded OFDM Systems Based on The Preamble Embedded by Moving Pseudo-Noise Sequences

3.1 Introduction

Accurate symbol timing offset (STO) must be implemented since OFDM systems are very sensitive to nonlinear distortion [1] and synchronization errors will result in ISI and ICI. The aim of timing synchronization is to obtain a starting point of a FFT window so that the transmitted data can be demodulated correctly in the receiver. To avoid these disadvantages, such as ISI and ICI, numerous algorithms have been proposed on the subject of timing synchronization. Figure 3.1 shows the four cases of OFDM symbol starting point. For non-preamble-aided algorithms, the STO can be estimated in [2] by minimizing the squared difference between two sliding windows of N_{cp} samples. Although the scheme is simple and effective, it is easily affected by CFO. The algorithm of the maximum correlation (MC) [3] performs better than [2], but still has some problems when CFO exists in the receiving end. The maximum likelihood (ML) scheme in [4] is an effective way to find the STO by calculating the maximum value of the likelihood function, which effectively solves the CFO problems. However, it can only make a coarse estimation of STO. These methods without preamble are able to make effective use of spectrum resources, but the timing offset cannot be estimated or be determined accurately. To further enhance the estimation precision, the preamble-based algorithms have been widely researched. The methods in [5, 6] utilize more than two OFDM symbols to realize timing synchronization, which result in a waste of frequency resources. An OFDM symbol with a simple form of two identical preambles is used in [7], the designed repetitive training preamble can be produced by inserting zeros into the even position of the subcarriers in the frequency domain, whereas there exists a plateau in the metric function with the length of N_{cp} samples. To solve the problem, the training symbols in [8] are separated from two of the same data blocks into four of the same data blocks and the latter two pieces of blocks are weighted with the minus signs, but the declining curve is not steep enough, which gives rise to timing estimation inaccuracy to some extent. Conjugate symmetry preamble is used in [9] to estimate symbol timing offset, which can effectively solve the problem of the steepness of the declining curve. The value of timing metric function is close to the pulse form at the right timing position, nevertheless, there exists side lobe outside the main lobe. Another estimator was proposed in [10] by employing a constant envelop preamble with two identical parts just like [7] weighted by single pseudo-noise (PN) sequence. Although the accuracy of STO is improved, whereas high precision still can not be achieved. In this paper, a novel timing estimator exploits four identical blocks with inverted signs of the latter two blocks and the correlation characteristics of PN sequences to estimate the timing offset. The preamble in the first half can be generated in the same way like [7] and repeat these samples with sign inversion to form the latter half. The features of the proposed approach is to utilize different PN sequences as weighted factors rather than transmitted signals to ensure that the



Figure 3.1: Four cases of OFDM symbol starting point.

metric function value reach its maximum at correct timing position. The first and second half of preamble are weighted with PN1 and PN2 respectively. By exploiting the double block correlation which are spaced N/4 samples apart, the metric function appears impulse-shaped form sharply without any large side lobes. When the block windows slide ahead, we introduce the same PN sequences like PN1 and PN2 into the block, which can eliminate the randomness only when the accurate location of the STO is detected. By this method, the difference between the adjacent values of the timing metric function is further enlarged, the performances are dependent on the structure of the preamble, the property of PN sequences and the design of the metric function. Computer simulations demonstrate that the proposed scheme has a significant probability of getting the accurate estimation of timing offset and relatively smaller error mean and SD compared with the other estimators in flatfading channels and performs good under frequency-selective fading channels when the FFT size is large in accordance with 3GPP LTE specifications.

3.2 OFDM Signal Model and Preamble-Based Time Synchronization Method

3.2.1 Signal Model

The input bit streams are mapped into a series of QPSK or QAM symbols which are then converted into N parallel signals in the OFDM transmitter. These parallel signals are modulated onto N different orthogonal subcarriers which are realized by inverse fast fourier transform (IFFT). Hence, the corresponding discrete-time OFDM symbol can be described as

$$\mathbf{x}(n) = \frac{1}{\sqrt{N}} \sum_{k=0}^{N-1} X(k) e^{j2\pi k n/N}$$

$$n \in \{0, 1, \dots, N-1\}$$
(3.1)

where X(k) denotes the data transmitted over the kth subcarrier with a length of T_s . After S/P conversion, the duration of transmission time for useful part of each OFDM symbol (*Nsamples*) is extended to $T = NT_s$. N is the IFFT window size. A cyclic prefix with the length of N_{cp} is inserted into the front of each OFDM symbol to reduce the impact of ISI between two consecutive OFDM symbols. In the receiver, here we assume there is no effect of sampling clock offset, and then the received baseband discrete-time signal on the nth sample can be represented as

$$r(n) = \sum_{l=0}^{L-1} h(l)x(n-l)$$

$$n \in \{0, 1, \dots, N + N_{cp} - 1\}$$
(3.2)

where h(l) is the channel impulse response. Since the symbol timing offset θ and the normalized carrier frequency offset ε are modeled as a delayed signal in the receiver and a phase offset in the time domain respectively. Considering these, the received signal can be expressed as

$$y(n) = e^{j2\pi n\varepsilon/N} r(n-\theta) + \omega(n)$$

$$n \in \{0, 1, \dots, N + N_{cp} - 1\}$$
(3.3)

where $\omega(n)$ is a complex Gaussian noise process with zero-mean and the variance of σ_n^2 .

3.2.2 Preamble-Based Representative Time Synchronization Method

In order to obtain accurate timing synchronization, the form and structure of the preamble and the design of the metric function is very important. Let us briefly introduce existing representative methods.

3.2.2.1 Schmidl and Cox Algorithm

Schmidl and Cox's method in [7] uses only one training symbol to implement timing synchronization, the training symbol in the time domain is divided into two blocks of data, the content of the two data block is exactly the same, the preamble structure in time domain is as follows:

$$Preamble_{sch\&cox} = \begin{bmatrix} A_{N/2} & A_{N/2} \end{bmatrix}$$

where $A_{N/2}$ is half a symbol (excluding cyclic prefix) with N/2 samples. Schmidl and Cox define a time measurement, when the value of time measurement exceeds a certain threshold, it can be considered the arrival of OFDM frame, when the value reach the maximum, we can find the starting point of the FFT window. The timing metric is given by

$$M_{\rm sch\&cox}(d) = \frac{\left|P_{\rm sch\&cox}(d)\right|^2}{\left(R_{\rm sch\&cox}(d)\right)^2} \tag{3.4}$$

$$P_{\text{sch}\&cox}(d) = \sum_{i=0}^{N/2-1} y^*(d+i)y(d+i+N/2)$$
(3.5)

$$R_{\rm sch\&cox}(d) = \sum_{i=0}^{N/2-1} |y(d+i+N/2)|^2.$$
(3.6)

The timing metric has a flat interval over the length of CP, which leads to some uncertainty to locate the STO.

3.2.2.2 Minn's Algorithm

An improved algorithm in [8] eliminates the flat area, make regular measurement curve form a peak on the best timing point, which is more conducive to determine the starting position of OFDM symbol. Minn's preamble has the form as follows:

$$Preamble_{Minn} = [A_{N/4} \ A_{N/4} \ -A_{N/4} \ -A_{N/4}]$$

the timing metric is expressed as

$$M_{Minn}(d) = \frac{|P_{Minn}(d)|^2}{(R_{Minn}(d))^2}$$
(3.7)

where

$$P_{Minn}(d) = \sum_{k=0}^{1} \sum_{m=0}^{N/4-1} y^* (d+2\frac{N}{4}k+m) \cdot y(d+2\frac{N}{4}k+m+\frac{N}{4})$$
(3.8)

$$R_{Minn}(d) = \sum_{k=0}^{1} \sum_{m=0}^{N/4-1} \left| y(d+2\frac{N}{4}k+m+\frac{N}{4}) \right|^2.$$
(3.9)

Despite the reduction of plateau over the length of CP, the values of the timing metric around the correct timing position are nearly identical that lead to erroneous estimation.

3.2.2.3 Park's Algorithm

The method in [9] enlarges the difference between two adjacent values of the timing metric. Park's preamble has the following form

$$Preamble_{Park} = \begin{bmatrix} A_{N/4} & B_{N/4} & A^*_{N/4} & B^*_{N/4} \end{bmatrix}$$

where $B_{N/4}$ represents N/4 length samples generated by IFFT of PN sequence. $B_{N/4}$ is constructed to be symmetric with $A_{N/4}$. The whole preamble presents characteristics of circular conjugate even symmetry and the timing metric is given by

$$M_{Park}(d) = \frac{|P_{Park}(d)|^2}{(R_{Park}(d))^2}$$
(3.10)

where

$$P_{Park}(d) = \sum_{i=0}^{N/2} y(d-i+N/2)y(d+i+N/2)$$
(3.11)

$$R_{Park}(d) = \sum_{i=0}^{N/2} |y(d+i+N/2)|^2.$$
(3.12)

This method has an impulse-shaped timing metric, allowing it to achieve a more accurate timing offset estimation.

3.3 Proposed Symbol Synchronization Algorithm

3.3.1 Feasibility of Adding PN to Training Preamble

PN sequences have some similar statistical characteristics with random noise, but unlike a real random signal, it can be produced and processed repeatedly which are widely used in spreading spectrum communication systems and secret communication systems. The autocorrelation of PN sequence is a two-valued function with speciality of a sharp autocorrelation peak when there is no shift. Although the cross correlation properties of PN sequence have not been described perfectly yet, the cross correlation properties of the same order PN sequences can be considered as a correlation function between a PN sequence and its sampling sequence [11, 12], and the calculation of the correlation function is only related to the selection of sampling factor. Here we consider two same normalized complex QPSK signals P1 and P2with the length of 16 samples. To verify whether the additive PN can significantly reduce the correlation value as weighted factors, the QPSK signals P1 and P2 are



Figure 3.2: Correlation value between the former part P1 and the latter part P2.

assembled head-to-tail in sequence to form P with the length of 32 samples. The PN sequences $s_k, k \in \{0, 1, ..., N-1\}$ are weighted on P. The performance are shown in Figure 3.2, the correlation value between P1 and P2 after adding PN is significantly reduced especially there is no difference offset as indicated in the middle point in the picture. We can also know that whether P1 and P2 are uncorrelated or part correlated, the correlation value after adding PN can be reduced to some extent.

3.3.2 Proposed Timing Synchronization Scheme

To reduce the error probability occurring in the above-mentioned methods and to avoid burst wide range deviation and further enlarge the difference of the timing metric value around the correct starting point, we design a new preamble by introducing two different PN sequences with the same period. The double PN weighted factors including PN1 and PN2 can be defined as

$$S_i, \quad i \in \{0, 1, \dots, N/2 - 1\}$$

 $Q_i, \quad i \in \{0, 1, \dots, N/2 - 1\}$

The sequences of S_i and Q_i are mapped into +1 or -1 and the original discretetime training sequences have the structure that satisfies

$$X_{i} = X_{i+N/4} = -X_{i+N/2} = -X_{i+3N/4}$$

$$i \in \{0, 1, \dots, N/4 - 1\}$$
(3.13)

The new preamble can be defined as

$$X' = \begin{cases} X_i S_i, & i \in \{0, 1, \dots, N/2 - 1\} \\ X_i Q_{i-N/2}, & i \in \{N/2, N/2 + 1, \dots, N - 1\} \end{cases}$$
(3.14)

so the proposed preamble are designed to be the form

$$Preamble_{proposed} = \begin{bmatrix} A_{[N/4]} S_{[N/4]} & A_{[N/4]} S_{[N/4]} \\ & - A_{[N/4]} Q_{[N/4]} & - A_{[N/4]} Q_{[N/4]} \end{bmatrix}$$

where $A_{[N/4]}$ represents samples of N/4 length produced by IFFT of a PN sequence. $S_{[N/4]}$ and $Q_{[N/4]}$ are weighted PN sequences with N/4 samples. Note that the PN in the frequency-domain and the time-domain have completely different effects. The former is used to avoid high peak-to-average power ratio (PAPR) and the latter is employed to diminish the two adjacent values of the timing metric because it is shown that they share the same sum of the pairs of product, with the exception of only two product terms in the previous research [7, 8]. Therefore, to enlarge the difference between the two adjacent values of the timing metric and avoid sudden timing shift in [9] and non-robust under low SNR in [10], it is necessary to maximize the different pairs of product between them. Then the timing metric is given by

$$M_{proposed}(d) = \frac{|P_{proposed}(d)|^2}{\left(R_{proposed}(d)\right)^2}$$
(3.15)

where

$$P_{proposed}(d) = \sum_{i=0}^{N/4-1} S_i S_{i+N/4} y^*(d+i) y(d+i+N/4) + \sum_{i=0}^{N/4-1} Q_i Q_{i+N/4} y^*(d+i+N/2) y(d+i+3N/4)$$
(3.16)

$$R_{proposed}(d) = \frac{1}{2} \sum_{i=0}^{N-1} |y(d+i)|^2.$$
(3.17)

We can see from (3.16) that the introduction of PN weighted factors ensures the proposed timing metric gets its maximum at the accurate timing point which is taken as the starting position of the useful part of the training symbol, whereas the remaining values at the other sample points are fairly low relative to the correct location. Therefore, the impacts are small which leads to a much smaller error rate of symbol timing estimation. All samples over one symbol period are utilized in calculation of half symbol energy as shown in (17). $P_{proposed}(d)$, and $R_{proposed}(d)$ can be calculated iteratively. The goal to minimize the correlation value between



Figure 3.3: Proposed preamble-based synchronization scheme.

two partly related blocks and maximize different pairs of product between two adjacent values are achieved by the proposed method. Note that the randomness of PN sequences can be eliminated only at perfect timing position. When the FFT window size gets bigger, the length of PN weighted factors gets longer that closer to the random signal and then the autocorrelation properties of m-sequence becomes very sharp towards impulse form, the system performance will be further improved. Figure 3.3 showns the proposed preamble-based synchronization scheme

3.4 Simulation Results

Computer simulations have been conducted to evaluate the performance of proposed timing offset estimator compared to the existing methods. An OFDM system with 128 subcarriers (N = 128) and CP length of 16 samples ($N_{cp} = 16$) is considered. We assume that the symbol timing offset and normalized carrier frequency offset are $\theta = 3$ and $\varepsilon = 0.2$, respectively. Modulation type is QPSK and unless stated otherwise, the simulations are run for 2000 trials. Two channel models including flat-fading channels and frequency-selective channels are considered here in the simulations. The system performance is evaluated in terms of symbol timing error rate, error mean and standard deviation. All of them are the functions of SNR. In order to get the most accurate timing offset, for convenience of comprehension, we take the absolute value of difference between estimated $\hat{\theta}$ and standard θ as reference value $|\hat{\theta} - \theta|$, if $|\hat{\theta} - \theta| = 0$, it means the estimated timing offset is precise without any mistake. In other words, the symbol timing error rate is zero at this time and $\hat{\theta}$ is the unbiased estimation of θ which can be perceived as the right time to detect the starting point of each OFDM symbol (with CP excluded) to obtain the exact samples. The error mean is the arithmetic average value of the error $|\hat{\theta} - \theta|$ and the standard deviation is the square root of the variance of $\left|\hat{\theta} - \theta\right|$ that used to measure the amount of variation or dispersion from the mean value. The algorithms in [7]-[10] were also implemented for the purpose of comparison. They were labeled as Sch and Cox, Minn, Ren and Park successively.



Figure 3.4: Symbol timing error rate for the conventional and proposed algorithms versus SNR in flat-fading channels.


Figure 3.5: Error mean of $|\hat{\theta} - \theta|$ for the conventional and proposed algorithms versus SNR in flat-fading channels.



Figure 3.6: SD of $|\hat{\theta} - \theta|$ for the conventional and proposed algorithms versus SNR in flat-fading channels.



Figure 3.7: Symbol timing error rate for the conventional and proposed algorithms versus SNR in frequency-selective fading channels (L = 5, N = 128).



Figure 3.8: Error mean of $|\hat{\theta} - \theta|$ for the conventional and proposed algorithms versus SNR in frequency-selective fading channels (L = 5, N = 128).



Figure 3.9: SD of $|\hat{\theta} - \theta|$ for the conventional and proposed algorithms versus SNR in frequency-selective fading channels (L = 5, N = 128).

Figure 3.4 shows the symbol timing error rate in flat-fading channels. Because the phase and amplitude have the same changes on the transmitted signal under flat-fading channels, it can be modeled as transmitted signal multiplied by the factor (rand(1) + sqrt(-1) * rand(1))/sqrt(2). As seen in Figure 3.4, the proposed algorithm performs better than the others especially at a low SNR because the further reduction of the influence of the side-peaks besides the correct position. The EM and SD performance are shown in Figure 3.5 and Figure 3.6 respectively. It indicates that the proposed method is almost the same as Park and Ren when the SNR is greater than 10 dB but greatly outperforms the others when SNR less than it. In [7], the mean value is a plot of the expected value of timing metric versus SNR at both the best timing instant and a point outside the training symbol, it has a different meaning compared to the EM in this paper. Although the method in [7] is the foundation of all the improved methods, it simply find the maximum of the timing metric that leads to some uncertainty as to the start of the frame. Therefore, it has bad performance in the evaluation of EM and SD. We do not use logarithmic coordinates in Figure 3.6 for the reason that the proposed scheme achieves a zero SD except a few points in the beginning. The system performance is further tested when under frequency-selective fading channels as shown in Figure 3.7, Figure 3.8, and Figure 3.9. All the channels contain 5 Rayleigh-fading taps (L=5) with the exponential energy decay of 1dB between adjacent paths. Channel responses were produced by using of complex Gaussion random variable with zero mean and unit variance which satisfies the condition that the power summation of each normalized path amplitude is equivalent to $\sum_{l} |h(l)|^2 = 1$. We can see that the proposed method still exhibits good performance on the error probability, but do not give satisfactory results in the light of EM and SD because once we have wrong timing offset estimation, a rather big error will occur, here we call it the burst large errors. The properties of Park's and Ren's methods also degraded severely than the conventional Sch and Cox's and Minn's method under frequency-selective fading channels. Here we consider another system profile resembles 3GPP LTE criterion [13] with FFT



Figure 3.10: Symbol timing error rate for the conventional and proposed algorithms versus SNR in frequency-selective fading channels (L = 5, N = 1024).

size of 1024 samples and 12.5 percent cyclic prefix of 128 samples. As illustrated in Figure 3.10 the symbol timing error rate of proposed algorithm and Ren's method performs better than the others, and the proposed scheme has similar performance with Minn's method which is superior to the rest methods in the respect of EM in Figure 3.11, since park's method have not any improvement with the increase of FFT window size, on the contrary, it appears large error mean with the average around 600 samples, therefore it is not displayed in the graphic. In terms of SD, although the proposed algorithm is slightly worse than Minn's method, it provides relatively good performance compared with the remaining approaches as shown in Figure 3.12.

3.5 Conclusion

A novel preamble-based symbol timing synchronizer has been presented for OFDM systems to improve the timing precision. By taking advantage of the special properties of PN sequences, we use two different PN sequences with the same period as weighted factors to ensure that the metric function has impulse-like shape at correct timing position without any burst side-peaks. The computer simulations show that the proposed algorithm achieves precise timing offset estimation with smaller EM and SD compared with the existing methods under flat-fading channels. In terms of the performance under frequency-selective fading channels, in spite of some unsatisfactory results when the FFT window size is small, the performance is relatively better than others with the increases of FFT size in accordance with 3GPP LTE specifications. Moreover, the system is robust to CFO and shows good properties especially under the low SNR assumption.



Figure 3.11: Error mean of $|\hat{\theta} - \theta|$ for the conventional and proposed algorithms versus SNR in frequency-selective fading channels (L = 5, N = 1024).



Figure 3.12: SD of $|\hat{\theta} - \theta|$ for the conventional and proposed algorithms versus SNR in frequency-selective fading channels (L = 5, N = 1024).

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Chapter 4

Blind Symbol Timing Synchronization Scheme for Real-time Industrial Wireless Control Network using High-Speed Preambleless OFDM Systems over Fading Channels

4.1 Introduction

Unlike single-carrier systems, OFDM systems demand high synchronization performance to avoid inter-symbol interference (ISI) and inter-channel interference (ICI) which will be generated when the orthogonality between each subcarriers is demolished. Synchronization inaccuracy will lead to some problems like sampling clock offset (SCO), symbol timing offset (STO), and carrier frequency offset (CFO). All of them will lead to a negative impact on the overall system performance especially in the latter two cases. Timing synchronization aims to find a appropriate FFT window so that the transmitted signal can be demodulated precisely in the receiver. In other words, we need to determine the position of the starting point to obtain the exact samples of the sending signal in the OFDM symbol duration. Note that as one kind of sampling offset, the phase offset in the sampling clocks (SPO) is normally small enough which can be considered as a part of STO that will result in frequency domain phase rotation that is proportional to the value of timing offset and subcarrier index. Symbol synchronization is generally divided into two major categories, preamble-aided and non-preamble-aided. For preamble-aided approaches, a single training symbol with identically repetitive structure of different periodic length in the time domain can be used in [2, 3]. To effectively solve the problem that the declining curve is not steep enough, a circular conjugate symmetry training symbols are employed, so it makes the correct timing position close to the pulse form which can be correctly captured by the synchronizer [4]. Another type of technique is to exploit a synchronization word and CP simultaneously [5], it provides better performance than the one mentioned in [8] in the condition that the influence of the multipath is not serious. Using the symbol training sequence can obtain accurate delay estimation, but the introduction of extra redundancy will significantly reduce the bandwidth utilization efficiency. In order to make full use of spectrum resources, recently, blind synchronization algorithms have been widely studied. The STO can be found in [6, 7] by seeking the difference between two sliding blocks of N_{cp} samples, although the scheme is concise, it is susceptible to CFO and not stable. The algorithm of maximum likelihood (ML) [8], minimum mean square error (MMSE) [9] and the maximum correlation (MC) [10] can achieve good performance, however, if the signal is subjected to deep frequency selective fading (FSF) channels, estimates point with respect to timing position will shift backwards accordingly due to the ISI. The estimator counteracting multipath fading has been developed in [11] which can be considered as improved algorithm to MMSE and MC. The disadvantage of this approach is that the exact channel length is needed. The article in [12]

proposed a differential algorithm, measuring the rate of change of the correlation results to determine the timing position. This algorithm is approximately 10 percents heavier than ML algorithm in terms of computational complexity and it still subject to multipath effects, with the increase of the number of channel path, timing performance gradually get worse. In [13], another scheme was presented by using 2-D search algorithm that needless to know the channel state information (CSI) which can accurately obtain the original start position of the FFT window, but the computation complexity is increased significantly. This algorithm is improved by putting the two-dimensional search down to one dimensional search in [14], which greatly reduced the computational complexity. However, it requires large amount of OFDM samples, and the width of correlation window is 1, easily affected by noise, which is the deficiency of point correlation algorithm. To mitigate above-mentioned problems, in this paper, a novel timing estimator based on the redundancy of CP is proposed for preambleless OFDM systems over multipath fading channels. The most significant aspect of this approach is to utilize two functions including metric function and positioning function on the premise that we take distinctive correlation characteristics within CP region fully into account. By means of exploiting the block correlation which are spaced N samples apart in ISI-free sampling region, the correlation value has a platform at the rear-end of CP accordingly. The selection of block length can be decided in our system. Moreover, in order to avoid the deficiencies of fluctuation and instability, a positioning function has been proposed instead of setting a threshold to find the first dip point that below the given value. It is proved that we can find the exact timing location with high accuracy and low-complexity compared with the conventional method. Numerical results show that the proposed approach achieves approximately accurate estimation of STO in high signal to noise ratio and manifests its robustness to multi-path fading channels.

4.2 OFDM System Model and Correlation Properties

4.2.1 System Model

Consider a discrete time signal containing N subcarriers in OFDM system. First of all, the transmitted data is mapped into complex symbols in frequency domain and then the complex signals $\{X(k)\}_{k=0}^{N-1}$ are modulated onto the N orthogonal subcarriers, which are implemented by inverse fast fourier transform(IFFT). It can be described as

$$\mathbf{x}(n) = \frac{1}{\sqrt{N}} \sum_{k=0}^{N-1} X(k) e^{j2\pi kn/N}$$

$$n \in \{0, 1, ..., N-1\}$$
(4.1)

where X(k) denotes the data transmitted in the OFDM symbol over the kth subcarrier with a length of T_s . Hence, the duration of transmission time for a single OFDM symbol is equivalent to $T = NT_s$. A CP with the length of N_{cp} is inserted into the front of each OFDM symbol so as to form a guard interval between two consecutive OFDM symbols to avoid ISI. After parallel-to-serial (P/S) conversation, the signal that will be subsequently transmitted turns out to be x(n), $n \in \{0, 1, \ldots, N + N_{cp} - 1\}$, consists of $N + N_{cp}$ samples. Let it pass through frequency selective fading channel with a lagged channel response of L - 1 samples. Notably, channel length is assumed to be less than the length of the CP. Here we ignore the influence of frequency offset in sampling clocks and then the baseband discrete-time signal on the *n*th sample can be represented as

$$r(n) = \sum_{l=0}^{L-1} h(l)x(n-l)$$

$$n \in \{0, 1, \dots, N + N_{cp} - 1\}$$
(4.2)

where h(l) is the channel impulse response. Since the symbol timing offset η and the normalized carrier frequency offset ϵ (ratio of the CFO to subcarrier spacing Δf , shown as $\epsilon = f_{offset}/\Delta f$) are modeled as a delayed signal in the receiver and a phase offset in time domain respectively. Considering these, the received signal can be expressed as

$$y(n) = e^{j2\pi n\varepsilon/N} r(n-\eta) + \omega(n)$$

$$n \in \{0, 1, \dots, N + N_{cp} - 1\}$$
(4.3)

where $\omega(n)$ denotes the complex white Gaussian noise process with zero mean and the variance of σ_n^2 .

4.2.2 Correlation Properties

Consider the correlation properties of the received OFDM signals y(n) and y(n+N). Here, the intervals are divided into three parts namely $A_1 = \{\eta, \eta+1, ..., \eta+L-2\}$, $A_2 = \{\eta+L-1, \eta+L, ..., \eta+N_{cp}-1\}$ and $A_3 = \{\eta+N_{cp}, \eta+N_{cp}+1, ..., \eta+N_{cp}+L-2\}$. Take (4.2) and (4.3) into account, the correlation between y(n) and y(n+N) thus satisfies

$$E\{y(n)y^*(n+N)\}$$

$$= e^{-j2\pi\varepsilon} \sum_{l=0}^{L-1} \left| h(l)^2 \right| E\{x(n-l-\eta)x^*(n+N-l-\eta)\}$$

$$= \begin{cases} Q_1(n)e^{-j2\pi\varepsilon}, & \text{for} & n \in A_1 \\ \sigma_m^2 e^{-j2\pi\varepsilon}, & \text{for} & n \in A_2 \\ Q_2(n)e^{-j2\pi\varepsilon}, & \text{for} & n \in A_3 \\ 0, & \text{otherwise} \end{cases}$$

$$(4.4)$$

where

$$Q_1(n) = \sum_{l=0}^{n-\eta} |h(l)|^2 \sigma_x^2 < \sigma_m^2,$$
(4.5)

$$\sigma_m^2 = E\{|y(n)|^2\} - \sigma_n^2 = \sum_{l=0}^{L-1} |h(l)|^2 \sigma_x^2, \qquad (4.6)$$

$$Q_2(n) = \sum_{l=n-\eta-N_{cp}+1}^{L-1} |h(l)|^2 \sigma_x^2 < \sigma_m^2.$$
(4.7)



Figure 4.1: block diagram of baseband preambleless OFDM and proposed scheme for WCN.

Note that since CP is to extend the OFDM symbol length by copying the last N_{cp} samples into its front, for $n, n + n_1 \in \{0, 1, ..., N_{cp} - 1\}$, the correlation relationship between x(n) and $x(n + n_1)$ can be shown as

$$E[x(n)x^*(n+n_1)] = \begin{cases} \sigma_x^2, n_1 = 0 \text{or} N\\ 0, \text{otherwise} \end{cases}$$
(4.8)

where $\sigma_x^2 = E[|x(n)|^2]$. As shown above, the correlation characteristics are decided by the channel condition. It can be seen from (4.4) and (4.6), when the signal y(n)is located in A_2 region, known as ISI-free region, the correlation value will reach the peak simply because it contains the information of all the channel taps. Otherwise, it will be reduced correspondingly on account of ISI under frequency-selective fading channels.

4.3 Proposed Symbol Synchronization Algorithm

Figure 4.1 shows the block diagram of the discrete-time baseband preambleless OFDM and the proposed scheme. It is by two steps to complete the synchronization process including coarse synchronization and fine synchronization, respectively.

4.3.1 Metric Function for Coarse Synchronization

A metric function is proposed by means of utilizing sampling data that is not influenced by ISI. It is derived on the basis of the algorithm in [8]. Since the exact channel length is not clearly known, in order to avoid the influence of multipath fading, we use the block length $b(b < N_{cp})$ instead of using all the samples in the CP. We perform the correlation calculation between two data blocks at a span of N samples. The block length b must be set appropriately so that the probability that the overall block falls into ISI-free region is high. The purpose of designing the metric function is to find a point in the interval A_2 irrespective which point it is on the premise that we do not know the exact channel length L. The metric function can be expressed as

$$\phi(\mathbf{n}) = \left| \sum_{k=0}^{b-1} E\{y(n+k)y^*(n+N+k)\} \right| \\ -\frac{\rho}{2} \sum_{k=0}^{b-1} [E\{|y(n+k)|^2\} + E\{|y(n+N+k)|^2\}] \\ n \in \{0, 1, ..., N+N_{cp}-1\},$$
(4.9)

where

$$\rho = \{E[|y(n)|^2] - E[|\omega(n)|^2]\} / E[|y(n)|^2]$$

= $\sigma_{\rm m}^2 / (\sigma_{\rm m}^2 + \sigma_{\rm n}^2) = SNR / (SNR + 1).$ (4.10)

From the discussion above, the function in (10) becomes

$$\phi(\mathbf{n}) = \left| \sum_{k=0}^{b-1} E\{y(n+k)y^*(n+N+k)\} \right| - b\sigma_m^2.$$
(4.11)

we make an analysis as follows:

Case i: for $n + b - 1 < \eta + L - 1$,

$$\phi(n) < b\sigma_m^2 - b\sigma_m^2 = 0. \tag{4.12}$$

Case ii: for $(n \leq \eta + L - 2) \cap (n + b - 1 \geq \eta + L - 1)$, here we assume the length of the block that falls within the interval A_1 is b1, it yields

$$\phi(n) = b_1 Q_1(n) + (b - b_1) \sigma_m^2 - b \sigma_m^2$$

< $b \sigma_m^2 - b \sigma_m^2 = 0.$ (4.13)

Case iii: for $(n \ge \eta + L - 1) \cap (n + b - 1 < \eta + N_{cp})$,

$$\phi(n) = b\sigma_m^2 - b\sigma_m^2 = 0. \tag{4.14}$$

Case iv: for $(n + b - 1 \ge \eta + N_{cp}) \cap (n \le \eta + N_{cp} - 1)$, here we assume the length of the block that falls within the interval A_2 is b2, $\phi(n)$ is obtained as

$$\phi(n) = b_2 \sigma_m^2 + (b - b_2) Q_2(n) - b \sigma_m^2$$

$$< b \sigma_m^2 - b \sigma_m^2 = 0.$$
(4.15)

Case v : for $n > \eta + N_{cp} - 1$,

$$\phi\left(n\right) < b \cdot 0 = 0,\tag{4.16}$$

$$n_{max} = \arg\max_{n} \{\phi(n)\}.$$
(4.17)

In regard to deciding b, Fig. 4.2 shows the lock-in probability outside the interval A_2 as a function of block length in FSF channel at the SNR of 22dB. To verify whether the value of b = 4 can be commonly used, we consider the other two cases when the system parameters such as the N_{cp} and L are changed to show its generality. In high speed WCN, the duration of transmission time for one OFDM symbol $(N + N_{cp} = 160)$ is set to 10μ s with N_{cp} is about 2μ s in our system. In general, short-range communication of WCN for FA has a small delay spread. To show the feasibility of the model, we choose the maximum delay about 0.5μ s in [16], which is enough to satisfy conditions of real environment. Based on this point, we select L to be 5 and 10 with the minimum and maximum respectively to determine the block length. The simulation results show that with the increase in block length, the probability

will rise correspondingly except for the first point. Hence, the block length is set to be 4 samples (b = 4) to achieve the best performance. Furthermore, the lockin probability in A_2 is also improved with the increase of CP length at a given channel length, however it is reduced with the increase of channel length at a given CP length. Note that the metric function is used for coarse timing synchronization, that means we do not require strict control of b and a small deviation of b has almost no effect on the whole system. In the design of actual OFDM system, the GI length must be taken by longer than the possible maximum channel length L but its GI length should be taken as close to L as possible because the longer GI length leads to the degradation of transmission efficiency in real-time wireless control networks, therefore the realistic parameters L = 10 and $N_{cp} = 32$ are chosen in our system.

Known from above, under the frequency selective fading channels, due to the influence of the multipath, new timing metric function will have a period of platform with zero value and the length of p at the rear-end of cyclic prefix. The plateau will appear in A_2 region. Furthermore, after the platform there is a falling edge, as a result, channel length can be roughly estimated, $L = N_{cp} - b - p + 2$. Although the formula of calculating the channel length is given in the paper, it is not the key point that we focuses on. In fact the detailed procedures for deciding p can be figured out by finding the first dip where $\phi(n)$ is below the given threshold which is zero theoretically from (4.14). However, it is difficult and not necessary to get accurate w because the signal is easily affected by noise. In other words, if the length of the platform p in the interval A_2 were precisely to be known, the symbol timing synchronization could be completed and the estimated starting point of OFDM symbol should locate in the first dip point plus (b-1) samples and thus we do not need to use positioning function any more. Since the estimation of channel length is outside the scope of this paper, it will not be further discussed here. After that we search to obtain the value n_{max} in (4.17) that correspond to the maximum point in the platform regardless of which point it is for the reason that it has little impact on the whole system performance and we have the positioning function for further fine timing synchronization. Fig. 4.3 show the estimation accuracy achieved by the first step of metric function with different GI length and channel length. Obviously, with the increase of SNR, the lock-in probability in A_2 is improved.

4.3.2 Positioning Function for Fine Synchronization

By combining n_{max} , we put forward a novel positioning function, which is of great importance for ultimately achieving synchronization process. The positioning function should have some particularity at the right timing instance (at the end of CP) by utilizing the correlation characteristics, so we want to have a monotonic increasing function when the independent variable is within CP and begin to fall when the independent variable is outside it. Here let the coefficients in the first term and the second term in positioning function to be X and Y respectively. It can be represented as

$$\psi(d) = X \left| \sum_{m=0}^{d} E[y(n_{\max} + m)y^{*}(n_{\max} + m + N)] \right|$$

-Y $\sum_{m=0}^{d} \{E[|y(n_{\max} + m)|^{2}]$
+E[|y(n_{\max} + m + N)|^{2}] \}
 $d \in \{0, 1..., N_{cp} - 1\}.$ (4.18)

Similarly, from (4.6), we can derive that the function is equivalent to

$$\psi(d) = X \left| \sum_{m=0}^{d} E[y(n_{\max} + m)y^{*}(n_{\max} + m + N)] \right| -(d+1) \cdot 2Y(\sigma_{m}^{2} + \sigma_{n}^{2}), \qquad (4.19)$$

when $(\eta + L - 1 \le n_{max} + d \le \eta + N_{cp} - 1) \cap (0 \le d \le N_{cp} - 1)$, combined with (4.4), it yields

$$\psi(d) = X(d+1)\sigma_m^2 -(d+1) \cdot 2Y(\sigma_m^2 + \sigma_n^2)$$

= $(d+1) \cdot [X\sigma_m^2 - 2Y(\sigma_m^2 + \sigma_n^2)]$
 $d \in \{0, 1..., \eta + N_{cp} - n_{max} - 1\}.$ (4.20)

here we take $X\sigma_m^2 - 2Y(\sigma_m^2 + \sigma_n^2)$ to be z, z > 0 is needed to meet the condition of monotonically increasing of the function in specified interval as mentioned above. When $\eta + N_{cp} - n_{max} \le d \le N_{cp} - 1$, it can be further written as

$$\psi(d) = (\eta + N_{cp} - n_{\max}) \cdot z$$

+(d - \eta - N_{cp} + n_{\max} + 1)
\cdot [XQ_2(n) - 2Y(\sigma_m^2 + \sigma_n^2)]
< (\eta + N_{cp} - n_{\max}) \cdot z

$$+(d - \eta - N_{cp} + n_{\max} + 1)$$

$$\cdot [X\sigma_m^2 - 2Y(\sigma_m^2 + \sigma_n^2)]$$

$$= (\eta + N_{cp} - n_{\max}) \cdot z$$

$$+(d - \eta - N_{cp} + n_{\max} + 1) \cdot z.$$
(4.21)

From above, we know, to ensure that the function value get its maximum at the right timing position, the term of last line in (4.21) must equal zero. When $\sigma_m^2 >> \sigma_n^2$, the term $\sigma_m^4/(\sigma_m^2 + \sigma_n^2)$ can be approximated as σ_m^2 , we know if z satisfy the following conditions:

$$z = X\sigma_m^2 - 2Y(\sigma_m^2 + \sigma_n^2)$$

= $\sigma_m^2 - \sigma_m^4 / (\sigma_m^2 + \sigma_n^2)$
= $\Delta(\sigma_m^2, \sigma_n^2)$ (4.22)

where $\Delta(\sigma_m^2, \sigma_n^2) = \sigma_m^2 \sigma_n^2 / (\sigma_m^2 + \sigma_n^2) > 0$, the variables X and Y are the functions of ρ . At this time, we can find such a positioning function that meet the above requirements at high SNRs. After solving the binary indeterminate equation, an optimum solution of the equation which can apply to WLAN environments for FA can be obtained, the solution can be express as

$$\begin{cases} X = 1 + \rho \\ Y = \rho^2 \end{cases}$$
(4.23)

Therefore, after calculation, we have

$$\psi(d) < (\eta + N_{cp} - n_{\max})\Delta(\sigma_m^2, \sigma_n^2).$$

$$(4.24)$$

It is improved that the maximum of function $\psi(d)$ is located in the point $\eta + N_{cp} - n_{max} - 1$. when $d > \eta + N_{cp} - n_{max}$, all the function values are less than $(\eta + N_{cp} - n_{max})\Delta(\sigma_m^2, \sigma_n^2)$. Hence the fine timing synchronization will be completed by seeking the maximum of $\psi(d)$,

$$d_{\max} = \arg\max_{d} \{\psi(\mathbf{d})\}.$$
(4.25)

Based on (4.17) and (4.25), the estimation of timing offset can be developed as

$$\hat{\eta} = n_{max} + d_{max} - N_{cp} + 1. \tag{4.26}$$

As you see, through the use of two functions, with the combination of n_{max} and d_{max} , the final timing offset estimation can be achieved from (4.26).

4.4 Simulation Results

4.4.1 Simulation Parameters

Monte Carlo simulations have been conducted to evaluate the performance of proposed symbol synchronization estimator. We consider an OFDM system with 128 subcarriers (N = 128) and CP length of 32 $(N_{cp} = 32)$. Modulation mode is QPSK. The symbol timing offset and normalized carrier frequency offset are set to be $\eta = 3$ and $\varepsilon = 0.2$, respectively. In the simulations, in order to make it easier to understand, $\hat{\eta} - \eta$ will be taken as reference for the reason that we strive to find the perfect timing point in the OFDM receiver. However, in general, ICI and ISI-free demodulation of OFDM signals does not require accurate symbol timing synchronization. Therefore, depending on the location of the estimated starting point of OFDM symbol, it is necessary to consider the early and late situations to the exact point. In the case of early synchronization when the starting point of the OFDM symbol is prior to the end of the lagged channel response to the previous symbol corresponding to the interval A_1 , orthogonality among subcarriers is destroyed and furthermore, ICI occurs. Next, when the detected symbol timing is faster than the perfect point, yet after the maximum lagged channel response corresponding to the interval A_2 , the current symbol is not overlapped with the previous symbol, which leads only phase offset but no ISI and ICI. The phase offset can be compensated by a frequency-domain equalizer. In the case of late synchronization in which the estimated starting point is later than the exact point corresponding to the interval A_3 , the received signal involves the information from the next symbol, which have a significant impact on the performance due to the occurrence of ISI and ICI.

At each simulation loop, the parameter ρ can be calculated by solving the real root of a cubic equation as mentioned in [15], here we assume that ρ was known as shown in (10) and SNR varies from 0dB to 30dB. Block length have been fixed at 4 samples (b = 4) unless otherwise specified. Frequency selective fading channels with 10 Rayleigh-fading taps (L = 10) were considered in this paper. Channel responses were produced by taking advance of complex Gaussion random variable with zero mean and unit variance which satisfies that the power summation of every normalized path amplitude is equal to $\sum_{l} |h(l)|^2 = 1$. The system performance is measured in terms of BER performance subject to STO, symbol timing error rate, error mean and standard deviation (SD), which are the functions of SNR. Error mean is the arithmetic mean of the error value $\hat{\eta} - \eta$, standard deviation is the square root of the variance of $\hat{\eta} - \eta$ that used to measures the amount of variation or dispersion from the average. All of them were averaged by 2000 trials. The algorithms in [6]-[10] were also implemented for comparison purpose. Algorithms [6, 7] were labeled as DF, DF1, schemes in [8]-[10] were labeled as ML, MC, MMSE and the methods in [11] was labeled as MC1, MMSE1. Since the channel length must precisely be known in MC1 and MMSE1 methods, which do not belong to the category of blind synchronization, they are not shown here.

4.4.2 BER Performance by Proposed Estimator

Fig. 4.4 and Fig. 4.5 show the BER performance of proposed symbol synchronization method compared with the existing methods. Note that in order to see the effect of STO, the signal after taking FFT has not been compensated in any form with the exception of channel compensation in Fig. 4.4 while the BER performance in Fig. 4.5 is evaluated by assuming the ideal frequency domain equalization of the phase offset due to STO in the interval A_2 with ICI and ISI-free demodulation. That means if the estimated starting point of OFDM symbol $\hat{\eta}$ is smaller than η but larger than the maximum lagged channel response, we do not consider it as a mistake. If $\hat{\eta} - \eta$ is zero, it means $\hat{\eta}$ is the accurate estimation which can be deemed as the perfect starting point of FFT window. The CFOs in Fig. 4.4 and Fig. 4.5 are perfectly compensated in order to demonstrate the effectiveness of the proposed symbol timing synchronization method. Comparing the two figures, it can be seen that the potential STO problems will result in ISI and ICI, which have great impact on the signal constellation in the receiver particularly on the conventional methods. The performance of our proposed estimator performs better than the remaining estimators when SNR is greater than 10dB in both cases.

4.4.3 Positioning Precision by Proposed Estimator versus SNR

The symbol timing error rate as illustrated in Fig. 4.6 and Fig. 4.7 is to evaluate the timing accuracy, if $\hat{\eta} - \eta \neq 0$, it will be counted as one mistake. In a similar way, Fig. 4.7 is evaluated with the ICI and ISI-free demodulation interval of A_2 excluded. Furthermore, CFO ($\varepsilon = 0.2$) is added here. It can be seen that the performance of proposed algorithm in Fig. 4.6 is significantly improved than the others when SNR equals from 10dB to 30dB. It is mainly because under a high SNR assumption, the term $\sigma_m^2 - \sigma_m^4/(\sigma_m^2 + \sigma_n^2)$ is approximately equivalent to zero so that the peak value of the function $\psi(d)$ necessarily points to the correct position $\eta + N_{cp} - n_{max} - 1$ whereas the other algorithms have unnecessary summation lengths for assumed channel mode which result in high error rate. By comparision, we know that the positioning in the interval A_2 is the biggest influencing reason for the symbol timing error rate of the proposed scheme whereas the positioning in the ICI and ISI occurrence interval A_1 and A_3 affects the system most by using the rest methods.

4.4.4 EM and SD by Proposed Estimator versus SNR

In the evaluation of symbol timing synchronization performance, the EM and SD are the most important indicators since they can reflect synchronizer performance directly. We should take $\hat{\eta} - \eta$ as reference rather than $|\hat{\eta} - \eta|$ because the early and late symbol timing synchronization have different level of effects on the system.

CFO ($\varepsilon = 0.2$) is also added here. The error mean and SD performances are shown in Fig.4.9, Fig.4.10 respectively. It shows that due to inherent characteristic of positioning function, the proposed scheme greatly outperforms the others after SNR equals around 13dB. The mean of the timing offset is able to achieve an accurate estimation with $\hat{\eta} - \eta$ approximately equals to zero and the SD performance is much smaller than the aforementioned algorithms.

Several methods can be used to achieve higher symbol timing estimation accuracy for OFDM system even under lower SNR such as preamble-aided schemes with various structures of training sequences. But the purpose of proposed preambleless OFDM synchronization scheme is to realize efficient industrial wireless control network (WCN) which are characterized by strict real-time requirement. In real-time WCN such as fieldbus motion control in FA, there are a master node and hundreds of slave nodes which are connected to a controller and actuators respectively. The data communication between them must be completed within $250\mu s$. To meet these needs, first, preambleless OFDM system has been employed in the uplink to reduce the transmission time. Second, high data-rate communication system such as 802.11 WLAN is necessary in order to further shorten the transmission time of the payload. According to the shannon formula, in the case of limited bandwidth resources, the only way to gain satisfactory transmission rate is to increase SNR. As a result, in general, the SNR value is usually set to be equal or great than 20dB and the signal field strength in coverage area is no less than -75dBm in 802.11 WLAN planning and design. The performance of proposed approach is superior than the conventional methods when SNR is more than around 15dB, which can completely satisfy the real-time requirements in FA WLAN environments. Fig. 4.8 is shown as wireless control network structure.

4.4.5 Performance by Proposed Estimator versus Channel Length

Finally, the system performance is further tested when the channel length L varied from 2 to 16 (SNR=22dB). The results are shown in Fig. 4.11, Fig. 4.12. Approximately accurate estimation of STO have been achieved by the proposed method regardless of channel length. However, the error mean of the STO in the rest methods gradually increases with the number of the channel length. As to the MSE, although the proposed algorithm has slight fluctuation, it is superior to the others under a given channel length.

4.4.6 Computational Complexity

The computational complexity is calculated in terms of complex multiplications. Since difference algorithms in [6, 7] are needless to do complex multiplications, it has the advantages of simplicity, but at the same time, only low symbol timing precision is obtained. In our system, the mathematical OFDM symbol numbers are assumed to be S. That means S consecutive preambleless OFDM symbols are sent and should be timing synchronized at a time. The computational complexity that required in MC, MC1, MMSE, MMSE1, ML algorithm are $SN_{cp}(N + N_{cp})$, $S(N_{cp} + L - 1)(N + N_{cp})$ N_{cp}), $3SN_{cp}(N + N_{cp})$, $3S(N_{cp} + L - 1)(N + N_{cp})$, $3SN_{cp}(N + N_{cp})$ respectively. Note that the MC1 and MMSE1 methods are used under the condition that the summation lengths have been increased (from N_{cp} to $N_{cp}+L-1$) to take into account the channel length. Therefore, when calculating the complex multiplications, the results come out to be $S(N_{cp} + L - 1)(N + N_{cp})$ and $3S(N_{cp} + L - 1)(N + N_{cp})$ respectively. The 2-D search algorithm in [13] can achieve high STO estimation accuracy, however, the complexity is $3SN_{cp}(N_{cp}+1)(N+N_{cp})/2$ which is far larger than the others. Under the given system parameters, we perform the metric function with varying b to decide the appropriate block length in the first OFDM symbol and use the decided b in the remaining OFDM symbols. As a result, the proposed algorithm has the complexity of $3(N + N_{cp})(2 + 20) \cdot 19/2 + 3N_{cp} + (S - 1)[3b(N + N_{cp})(2 + 20) \cdot 19/2 + 3N_{cp} + (S - 1)[3b(N + N_{cp})(2 + 20) \cdot 19/2 + 3N_{cp} + (S - 1)[3b(N + N_{cp})(2 + 20) \cdot 19/2 + 3N_{cp} + (S - 1)[3b(N + N_{cp})(2 + 20) \cdot 19/2 + 3N_{cp} + (S - 1)[3b(N + N_{cp})(2 + 20) \cdot 19/2 + 3N_{cp} + (S - 1)[3b(N + N_{cp})(2 + 20) \cdot 19/2 + 3N_{cp} + (S - 1)[3b(N + N_{cp})(2 + 20) \cdot 19/2 + 3N_{cp} + (S - 1)[3b(N + N_{cp})(2 + 20) \cdot 19/2 + 3N_{cp} + (S - 1)[3b(N + N_{cp})(2 + 20) \cdot 19/2 + 3N_{cp} + (S - 1)[3b(N + N_{cp})(2 + 20) \cdot 19/2 + 3N_{cp} + (S - 1)[3b(N + N_{cp})(2 + 20) \cdot 19/2 + 3N_{cp} + (S - 1)[3b(N + N_{cp})(2 + 20) \cdot 19/2 + 3N_{cp} + (S - 1)[3b(N + N_{cp})(2 + 20) \cdot 19/2 + 3N_{cp} + (S - 1)[3b(N + N_{cp})(2 + 20) \cdot 19/2 + 3N_{cp} + (S - 1)[3b(N + N_{cp})(2 + 20) \cdot 19/2 + 3N_{cp} + (S - 1)[3b(N + N_{cp})(2 + 20) \cdot 19/2 + 3N_{cp} + (S - 1)[3b(N + N_{cp})(2 + 20) \cdot 19/2 + 3N_{cp} + (S - 1)[3b(N + N_{cp})(2 + 20) \cdot 19/2 + 3N_{cp} + (S - 1)[3b(N + N_{cp})(2 + 20) \cdot 19/2 + 3N_{cp} + 3$ N_{cp} + $3N_{cp}$] with the complexity of metric function is included. The selected block length satisfies b = 4, moreover, N = 128, $N_{cp} = 32$, and L = 10 to L = 16are considered respectively. After calculation, we know if the worst case scenario is considered (only one OFDM symbol), we know the complexity relations among these algorithms are as follows:

$$MC = 5120 < MC1 = 6560$$

$$< ML = MMSE = 15360 < MMSE1 = 19680$$

$$< proposed = 100416 < [13] = 253440$$
(4.27)

Since a large amount of preambleless OFDM symbols are sent successively in the actual uplink WCN systems and each symbol needs to be synchronized. An actual situation is discussed here, if the OFDM symbol numbers are larger than 33 (including 33) samples, we have the following property of complexity relations:

$$Proposed < MC < MC1 < ML = MMSE < MMSE1 < [13]$$
(4.28)

in fact, when S=33, we have

$$proposed = 164928 < MC = 168960$$

$$< MC1 = 216480 < ML = MMSE = 506880$$

$$< MMSE1 = 649440 < [13] = 8363520$$
(4.29)

Obviously, the proposed algorithm has lower computational complexity for practical application.

4.5 Conclusion

A novel blind symbol timing synchronizer has been presented for OFDM systems over frequency-selective fading channels. By utilizing a metric function featuring block correlation which are spaced N samples apart to find a sampling point in ISI-free region. And then we further designed a novel positioning function to complete the final symbol timing synchronization. It has been proved by the computer simulations that the proposed algorithm achieves better BER performance subject to STO, accurate timing offset estimation with smaller SD and relatively lower computational complexity compared with the existing method based on CP. In addition, the system performance is robust to CFO and irrespective of channel length when it is subjected to frequency-selective fading channels with high SNR assumption, which can be applied to WCN systems for FA according to the 802.11 WLAN Specification.



Figure 4.2: The probability of incorrect positioning in A_2 by metric function versus block length in frequency-selective fading channels.(N = 128, SNR=22dB).



Figure 4.3: The probability of incorrect positioning in A_2 by metric function versus SNR in frequency-selective fading channels. (N = 128, b = 4).


Figure 4.4: The BER performance subject to STO for the conventional and proposed algorithms versus SNR in frequency-selective fading channels. (L = 10, b = 4).



Figure 4.5: The BER performance subject to STO with perfect equalization for the conventional and proposed algorithms versus SNR in frequency-selective fading channels.(L = 10, b = 4).



Figure 4.6: Symbol timing error rate for the conventional and proposed algorithm versus SNR in frequency-selective fading channels.(L = 10)



Figure 4.7: Symbol timing error rate with A_2 excluded for the conventional and proposed algorithm versus SNR in frequency-selective fading channels.(L = 10)



Figure 4.8: Wireless control network structure.



Figure 4.9: Error mean of $\hat{\eta} - \eta$ for the conventional and proposed algorithm versus SNR in frequency-selective fading channels.(L = 10)



Figure 4.10: SD of $\hat{\eta} - \eta$ for the conventional and proposed algorithm versus SNR in frequency-selective fading channels.(L = 10)



Figure 4.11: Error mean of $\hat{\eta} - \eta$ for the conventional and proposed algorithm versus channel length. (SNR=22dB)



Figure 4.12: SD of $\hat{\eta} - \eta$ for the conventional and proposed algorithm versus channel length. (SNR=22dB)

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Chapter 5

Conclusion

Asynchronous problems, such as the I/Q imbalance at the transceiver, the ISI and ICI caused by either the STO or CFO need to be solved urgently in OFDM system in order to adapt to the future development of mobile communication. There are some previous literatures on solving these problems. However, these methods have some disadvantages like low precision or high complexity. In this paper, we aim to enhance the performance of the previous synchronization algorithms.

In Chapter 2, we investigates a joint synchronization algorithm for dealing with the ICI problem in the presence of I/Q imbalance in TFI-OFDM systems. To reduce the sensitivity of the ICI caused by the CFO, a novel ICI SC scheme is proposed in a general implementation framework. Unlike conventional methods, the designed transmission model has fewer pilot signals, which achieves relatively better CIR and the best BER performance for lower and higher values of CFOs. Due to the repetition symbols on adjacent subcarriers, it achieves frequency diversity by reducing bandwidth efficiency. However, the proposed scheme is effective for combating the impact of ICI and easy for hardware or software implementation since no further channel equalization is needed and the system complexity is not increased. The proposed scheme shows good property compared with conventional methods for practical application. The system model is applicable to AWGN and a flat-fading channel. Further work can be done by investigating the performance of our schemes in more complex channel situations without accurate CSI.

In Chapter 3, we researched on the preamble-aided methods to solve the time synchronization problems. A novel symbol timing synchronizer is proposed based on the properties of PN sequences. By utilizing four identical blocks with inverted signs of the latter two blocks and two different PN sequences with the same period as weighted factors, the proposed metric function has impulse-like shape at correct timing position without any burst side-peaks. The first and second half of preamble are weighted with PN1 and PN2 respectively. We use the double block correlation which are spaced N/4 samples apart, and when the block windows slide ahead, we introduce the same PN sequences including PN1 and PN2 into the block, which can eliminate the randomness only when the accurate location of the STO is detected. By this means, the difference between the adjacent values of the timing metric function is further enlarged. The simulations reveal that the proposed scheme has a significant probability of getting the accurate STO and has relatively smaller EM and SD compared with the other estimators in flat-fading channels. When considering the frequency-selective fading channels, the proposed method shows favorable performance when a large number of samples is exploited within the FFT interval in accordance with 3GPP LTE specifications.

In Chapter 4, we research on the non-preamble-aided timing synchronization method to meet the development of bandwidth-efficient and high-speed requirement for 5G, blind timing synchronization aims to find the appropriate FFT window so that the transmitted signal can be demodulated precisely in the receiver. It is usually performed in two phases including coarse and fine synchronization. Coarse timing synchronization attempts to generate the possible starting of an OFDM symbol within the CP. Nevertheless, fine synchronization improves the precision to a difference of few samples from the perfect point. In this chapter, a metric function and a positioning function have been proposed for coarse and fine synchronization respectively by taking advantage of the correlation characteristics of OFDM signals. Through derivation and numerical analysis, the proposed methods result in accurate timing offset estimation with smaller SD and lower computational complexity compared with the existing methods. Furthermore, the system performance is robust to CFO and irrespective of channel length with realistic SNR assumption, which can be applied to WCN systems for FA according to the 802.11 WLAN Specification.

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List of Related Papers

Journal Paper

- Wenjian Wang, Ye Tian, and Jun-Ahn Chang, "Performance analysis of an improved ICI-Cancellation scheme with I/Q imbalance compensation in TFI-OFDM systems," Journal of Selected Areas in Telecommunications, vol. 6, no. 2, pp.1-8, 2016.
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International Conference Paper

- Naotoshi Yoda, <u>Wenjian Wang</u>, and Chang-Jun Ahn, "Single Symbol Decodable QO-STBC with Full Diversity," Proc. of IEEE International Conference on Signal Processing and Communication Systems (ICSPCS2013), pp.265-269, Gold Coast, Australia, December 2013.
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- Wenjian Wang, Chang-Jun Ahn, Tatsuya Omori, and Ken-ya Hashimoto, "A Novel Symbol Timing Synchronization Scheme for OFDM systems based on variational pseudo-noise preamble," IEICE Technical Report, CS2014-106, vol. 114, no. 463, pp. 81-86, February 2015.
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