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A New Ranging Technique for IEEE 802.16e Uplink

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Abstract

Aiming at the shortcomings of the traditional ranging algorithm, this paper proposes a new ranging technique for 802.16e uplink in the frequency domain. The ranging technique eliminates the effect of the timing offset by computing the frequency difference of adjacent sub-carriers in the ranging channel, which reduces the search space from two dimensions to one dimension. When there is slot, the ranging code can be detected by adopting a threshold. The ranging technique is implemented through one-dimension search in the frequency domain, and consequently the complexity can be reduced and without losing the performance of timing estimation. The simulation results show the good ranging performance of this technique.

Index Terms: 802.16e; Ranging; Uplink

1. Introduction

The multiple access scheme adopted in IEEE 802.16e uplink is Orthogonal Frequency Division Multiple Access (OFDMA), in which different users possess corresponding sets of sub-carriers. When the user attempts to access the network or the handover happens, the process of ranging is required. The ranging process performs two main tasks: detecting all ranging codes in the current ranging slot and estimating the corresponding Return Time Delay (RTD) of these ranging codes. For the OFDMA system using Time Division Duplex (TDD) mode, the ranging process requires the accurate timing synchronization between Base Station (BS) and Mobile Station (MS). The problem is mainly discussed in this paper.

The traditional ranging algorithm is implemented in the frequency domain[1]. In this algorithm the time domain sequences will firstly be transformed into frequency domain through Fast Fourier Transformation (FFT), and then the transformed sequences will correlate with the all ranging codes in the set of ranging codes. Because of the existence of time offset, the reversal of the phase in the frequency domain will happen. The time offsets need be taken into account and the cross-correlation process in the traditional algorithm is a two-dimension and ergodic one. In the IEEE 802.16e there are 256 ranging codes. If the RTD is considered in the

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range of [0,N/2], the computation burden of the two-dimension ergodic search is very high, which enhances the difficulty to implement it. Besides the great complexity, the performance of the traditional algorithm is limited by the frequency selective fading channel[2]. This is because the traditional technique can't eliminate the effect of the channel's phase variation on the correlation function.

A new ranging technique which is implemented in the frequency domain is proposed. The proposed technique computes the difference in frequency domain between the adjacent sub-carriers in the ranging channel and eliminates the effect of the timing offset. Therefore the search space decreases to one-dimension from two-dimensions. At the same time a specific threshold is used to detect the ranging codes in the current ranging slot. The estimation of the corresponding ranging code's RTD is acquired from the phase of the correlation function. Accordingly a less search range is defined based on the estimation and all the possible timing offsets can be searched in this one-dimension space.

2. Description of The Existing Algorithms

The literature [1] divides the whole process of ranging into three steps: Firstly, compute the power in the frequency domain of the received signals and compare this power with the predefined threshold to determine if the ranging code exists in the current FFT window. Secondly, determine the possible timing offset in the FFT window where the ranging code exists in. Finally, eliminate the effect of the timing offset and detect the ranging code through cross-correlation operation in the frequency domain.

The method proposed in [1] works well in the Additional White Gaussian Noise (AWGN) channel and has lower complexity. But this method can't work effectively in the multi-path channel. Because the received power in the frequency domain has large fluctuation which invalidates the method of threshold detection. We can't determine whether the ranging code exists in the current FFT windows through the predefined and invariable threshold.

In order to overcome the shortcomings of the traditional ranging technique (whose performance is limited by the frequency selective fading multi-path channel and the noise), the literature [2] proposed another ranging technique which is based on differential detection combined with multiple length FFT computation. In this method the differential detection is used to counteract the effect of the frequency selective fading in the multi-path channel, and the multiple length FFT computation is used to eliminate the effect of the noise. But the leading modulation sequence in the IEEE 802.16e standard doesn't adopt the scheme of differential code. The method proposed in [2] can't be directly used in the 802.16e standard.

The two literatures [3] and [4] both present the same idea of difference in the frequency domain as shown in the literature [2]. This idea reduces the number of dimension of the ergodic search process from two to one and can counteract the effect of the frequency selective fading in some extent. The RTD can be simultaneously estimated through the phase of the correlation function. This method can greatly reduce the complexity of the traditional technique, but it will incur some loss in the performance of detection and timing offset estimation. For the detection of the ranging code, the determination of the threshold is very difficult problem. Furthermore the method in [3] and [4] is infeasible when there are several ranging users. This is because the superposition of signals from multiple users will largely impact the performance of timing offset estimation. Moreover the difference computation in frequency domain will reduce the performance of the cross-correlation characteristic of the ranging codes (about 0.8 dB).

3. The New Ranging Technique

3.1. The Principle of The Technique

We denote by y_n the received sequences in the time domain. In the condition of the AWGN channel, we may write

$$y_n = \frac{1}{N}\sum_{k=-N/2}^{N/2} X_k H_k e^{j2\pi n(k+\varepsilon)/N} + w_n \tag{1}$$

where ε is the normalized carrier frequency offset, X_k is the BPSK-modulated ranging code, H_k is the channel response and w_n is the complex Gaussian noise in the time domain.

There are two schemes to implement the initial ranging. The first scheme uses two OFDM symbols to transmit a ranging code, and the second one uses four OFDM symbols to transmit two adjacent ranging codes. Here the first case is discussed and the other case can be processed using the similar manner. The time domain structure of the ranging code defined in the 802.16e standard is shown in Fig. 1.

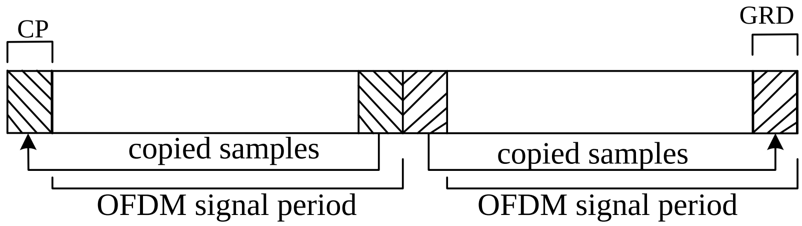

Fig. 1. The time domain structure of the ranging code

As shown in the above Fig. 1, the continuity of the phase is preserved by the two consecutive ranging symbols. Supposing that the RTD is not larger than $N_s + N_g$, then we can obtain a whole ranging symbol from the time domain sequences whose length is equal to the length of two consecutive FFT windows. In this paper we suppose the RTD has the range of $[0, N/2+N_g]$, where N is the length of an OFDM symbol. Supposing the ranging user's RTD is denoted by TO, in the ideal channel the two estimations of the timing offset obtained from the time domain sequences in the two FFT windows are given as

$$\begin{aligned}\delta_1 &= TO \\ \delta_2 &= TO - N_g\end{aligned} \tag{2}$$

After we have obtained the two N-point time domain sequences of the ranging signal, we can transform them into the frequency domain through FFT. For a whole ranging symbol the frequency domain sequences Y_k can be expressed as

$$\begin{aligned}Y_k = X_k H_k \frac{\sin \pi\varepsilon}{N \sin(\pi\varepsilon / N)} \cdot e^{j\pi\varepsilon(N-1)/N} \\ \cdot e^{-j2\pi\Delta nk/N} + I_k + W_k\end{aligned} \tag{3}$$

where Δn is the RTD, I_k is inter-carrier interference (ICI) which is produced due to the carrier frequency offset and the return time delay, and W_k is the term of complex noise. Let I_k be the inter-carrier interference (ICI), then

$$I_k = \sum_{\substack{m=-N/2 \\ m \neq k}}^{N/2} X_m H_m \frac{\sin \pi\varepsilon}{N \sin\left[\pi\left(m-k+\varepsilon\right)/N\right]} \cdot e^{j\pi\varepsilon(N-1)/N} \cdot e^{-j\pi(m-k)/N} \cdot e^{-j2\pi\Delta nk/N} \tag{4}$$

We express the cross-correlation in the frequency domain as $Z_k = Y_k \times \hat{X}_k$, where $\hat{X}_k$ is the candidate in the ranging code set. When the local ranging code is matched with the transmitted code, $\hat{X}_k$ satisfies the following condition of $X_k \times \hat{X}_k = 1$. This is because the ranging code is modulated using BPSK. The differential cross-correlation function of the adjacent sub-carriers is given by

$$\begin{aligned} Z_k Z_{k+1}^* &= \left(Y_k \hat{X}_k\right)\left(Y_{k+1} \hat{X}_{k+1}\right)^* = Y_k Y_{k+1}^* \hat{X}_k \hat{X}_{k+1}^* \\ &= H_k H_{k+1}^* \left[\frac{\sin \pi\varepsilon}{N \sin\left(\pi\varepsilon / N\right)}\right]^2 \cdot e^{-j2\pi\Delta n/N} \\ &\quad + noise\_terms \end{aligned} \tag{5}$$

Ignoring the term of sin[2] and the term of phase, the *noise_terms* in the above equation is written as follows:

$$\begin{aligned} noise\_terms = {} & H_k I_{k+1}^* + H_k W_{k+1}^* + I_k H_{k+1}^* \\ & + I_k I_{k+1}^* + I_k W_{k+1}^* + W_k H_{k+1}^* \\ & + W_k I_{k+1}^* + W_k W_{k+1}^* \end{aligned} \tag{6}$$

We usually suppose that the channel response is independent from the noise and the energy of ICI interference is much less than the power of the useful signal. This hypothesis is reasonable in the actual case. However some correlation is likely to exist between W_k and W_{k+1} in the specific noise environment. But this kind of correlation can't exist in the whole ranging channel.

Superposing the frequency domain cross-correlation of all adjacent sub-carriers in the ranging channel, we can derive the correlation function as follows:

$$\begin{aligned} P = \frac{1}{K} \sum_k Z_k Z_{k+1}^* \cong e^{-j2\pi\Delta n/N} \cdot \\ \frac{1}{K} \left[\frac{\sin \pi\varepsilon}{N \sin\left(\pi\varepsilon / N\right)}\right]^2 \sum_k H_k H_{k+1}^* \end{aligned} \tag{7}$$

where K is the number of the adjacent sub-carriers in the ranging channel. When the ranging channel occupies 6 adjacent sub-carriers, the number of sub-carriers N is 144 and the number of the adjacent sub-carriers K is $6 \times 6 \times 3 = 108$.

We use the magnitude of the correlation function as the metric function and it can be written as

$$M = |P| = \frac{1}{K}\left[\frac{\sin \pi\varepsilon}{N\sin\left(\pi\varepsilon / N\right)}\right]^2 \left|\sum_k H_k H_{k+1}^*\right| \tag{8}$$

Let M_1 denote the metric function obtained from the time domain sequences in the first FTT window and M_2 denote the corresponding one in the second FFT window. According to (2), the estimation of the RTD is given as

$$\begin{aligned} \delta &= round\left\{\frac{N}{2\pi}\arg\left[P_1\right]\right\}, \quad if\ M_1 > M_2 \\ \delta &= round\left\{\frac{N}{2\pi}\arg\left[P_2\right]\right\} + N_g, \quad if\ M_2 > M_1 \end{aligned} \tag{9}$$

If a metric function exceeds the predefined threshold, we can think that the candidate code corresponding with this metric function is the current ranging code which is being transmitted. When we set the threshold we need balance between the probability of losing detection and the probability of false detection. We should emphasize that the larger between M_1 and M_2 is used to detect the ranging code and estimate the timing offset.

3.2. The Range of the Estimated Timing Offset and The Phase Reversal

In the proposed technique the phase of the correlation function is used to estimate the RTD. The range of estimation is limited by the periodic characteristic of the phase. As mentioned in the context the larger between M_1 and M_2 is used to estimate the timing offset. According to (2), the range of estimation of the timing offset δ is given as

$$\delta \in \left[-N/2,\, N/2 + N_g\right] \tag{10}$$

We must notice that the phenomenon of phase reversal may happen when the phase of the correlation function is used to estimate the timing offset. The phase reversal means that the phase of the correlation function turns to π from -π or vice versa. In the non-ideal channel the possible phase reversal will result in two kinds of case. The first is that the estimated timing offset is close to the $-N/2+N$ but not the $-N/2$ when the actual timing offset is close to $-N/2$. The second is that the estimation is likely close to $N/2+N_g-N$, but the actual timing offset is close to $N/2+N_g$. After thorough analysis we can find that phase reversal is likely to happen when the FFT window is lying in the middle of initial ranging signal whose length is equal to the length of the two symbols.

3.3. The Optimization of The Timing Offset Estimation

1) The technique eliminating the effect of the phase reversal

In the above section we have depicted the phenomenon of phase reversal and the condition when it maybe happens. Now we will research the technique to eliminate the effect of the phase reversal. When one metric function exceeds the threshold, the processing method is described as follows:

If M_2 is greater than M_1 and the estimation of timing offset is negative, we think that the phase reversal has happened and add N to the estimated value of the timing offset. If M_1 is greater than M_2 and the estimation of timing offset is positive, we think that the phase reversal has happened and subtract N from the estimated value of the timing offset.

But there are some hidden troubles in the above method. For example, when the timing offset is greater than zero but less than N_g, M_1 is likely to be greater than M_2. At this time if we only adopt the latter processing means, we will improperly process the estimated value of the timing offset.

When the actual timing offset is close to the $-N/2$, the phase reversal brings a $+N$ offset to the estimation and as a result the final estimation is close to $N/2$. But when the actual timing offset is really close to $N/2$, it is impossible that M_1 is greater than M_2. Based on the above fact we can set a better threshold than zero. In this case the threshold can be set as $(N_g + N/2)/2$.

When the actual timing offset is close to the $N/2+N_g$, the phase reversal brings a $-N$ offset to the estimation and as a result the final estimation is close to $-N/2+N_g$. Similarly when the actual timing offset is really close to $-N/2+N_g$, it is impossible that M_2 is greater than M_1. In this case the threshold can be set as $[-N_g+(-N/2+N_g)]/2$.

In the IEEE 802.16e standard the maximal N_g can be equal to $N/4$, therefore the equation $-N/2+N_g$ is fitly equal to $-N_g$. At this condition M_2 is likely to be greater than M_1. But two facts prevent the appearance of this case. The first fact is that the WiMAX forum selects $N/8$ as the length of cyclic prefix (CP) of mobile WiMAX. The second fact is that the actual timing offset is mainly concentrated on the positive section and the negative estimation with small absolute value only exists in the case of periodic ranging.

After using (9) to estimate the RTD, we can use the following technique to eliminate the effect of phase reversal:

$$\begin{aligned} \delta' &= \delta - N, if\ (M_1 > M_2)\, and \left(\delta > \left(\mathrm{N_g}+\mathrm{N}/2\right)/2\right) \\ \delta' &= \delta + N, if\ (M_2 > M_1)\, and \left(\delta < -\mathrm{N}/4\right) \end{aligned} \tag{11}$$

2) The technique improving the performance of timing offset estimation

Except the effect of the noise, the method of estimating the RTD in (9) will be influenced by the phase change of the channel response of adjacent sub-carriers. In the case of multi-users ranging these users' data will overlap on the same sub-carrier set in frequency domain. This will influence the estimation of phase and consequently will influence the estimation of RTD. The interference among the multi-users has large impact on the performance of estimating the timing offset. So we can conclude that the technique using the phase to estimate the timing offset is infeasible in the case of multiple users ranging.

The traditional ranging technique estimating the timing offset is mainly influenced by the multi-path channel. The peak values proportional with the channel coefficients will appear at the locations of each path [5]. The primary simulation result shows that the performance of the traditional ranging technique is much better than the performance of the technique using the phase to estimate the timing offset in the case of multiple users ranging. We will combine these two techniques to improve the performance of the timing offset estimation. The complexity due to this combination is low and acceptable. The process of the algorithm is given as follows:

a) Use (7) to (9) to detect the ranging code in the current ranging slot and estimate the corresponding timing offset;

b) Use (11) to eliminate the effect of phase reversal;

c) Set a search range at the both sides of the current timing offset estimation and compensate the phase for the timing offset candidates in this search range; then compute the correlation function in the frequency domain;

d) Select the timing offset candidate corresponding to the peak value of correlation function as the new timing offset estimation.

Supposing the search length in single side is L , the search rang R can be determined according to the estimation δ' in (11) as follows:

$$R = \{ n \mid \delta' - L \le n \le \delta' + L \} \tag{12}$$

If the above search range exceeds the bound given in (10), the search range will be truncated. Compensating the phase for every candidate of the timing offset and computing the correlation function in the frequency domain, we have

$$\begin{aligned} P_t(n) &= \sum_k Y_k X_k^* e^{j2\pi kn/N}, if\ \ M_1 > M_2 \\ P_t(n) &= \sum_k Y_k X_k^* e^{j2\pi k(n-N_g)/N}, if\ \ M_2 > M_1 \\ M_t(n) &= |P_t(n)| \end{aligned} \tag{13}$$

Then we select the timing offset candidate corresponding to the maximal metric function as the new timing offset estimation which is expressed as

$$\hat{\delta} = \arg\left\{ \max_n \left[M_t(n) \right] \right\} \tag{14}$$

3.4. The Setup of The Threshold

Based on the metric functions given in (8) of all possible ranging codes at the location of first and second FFT windows, we can respectively compute the rms value. Setting the threshold used to detect the ranging code at the value which is *m* dB larger than rms, we have

$$T_s = 10^{m/10} \sqrt{\sum_{r \in I} M_{s,r}^2}, s = 1, 2 \tag{15}$$

where I is the set of the ranging codes and the value of m is determined through simulation. The ranging code candidate corresponding to the metric function exceeding the threshold is detected as the ranging code in the current ranging slot.

4. Simulation Result and Performance Analysis

In order to validate the performance of the ranging technique proposed in this paper, we simulate the proposed technique and compare the result with others in literature [3] and [4]. Fig. 2 shows the performance of the technique proposed in this paper compared with which were given in literature [3] and [4].

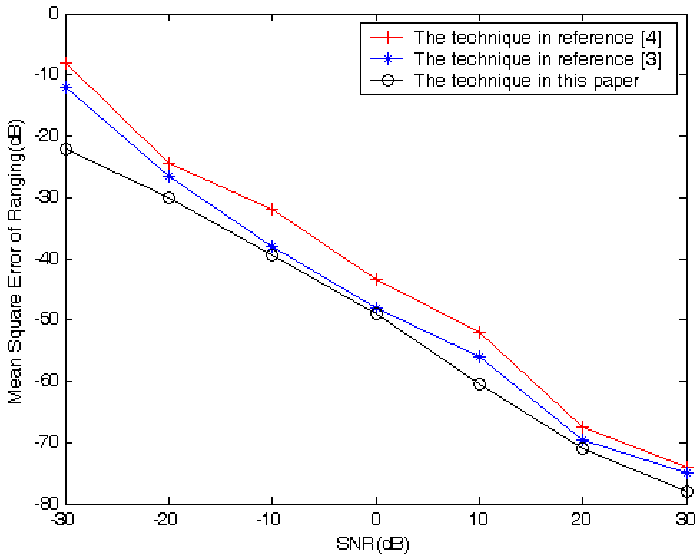

Fig. 2. The mean square error of ranging technique

The simulation condition is the uplink PUSC. The size of FFT is 1024. The adopted channel is the Vehicle A test channel with 100 Hz Doppler frequency offset. The modulation scheme is QPSK. The result in Fig. 2 shows that the performance of the technique proposed in this paper is close to the techniques in reference [3] and [4] and is a little better than them.

5. Conclusion

Aiming at the shortcomings of the traditional ranging algorithm, this paper proposes a new ranging technique for 802.16e uplink in the frequency domain. The ranging technique eliminates the effect of the timing offset by computing the frequency difference of adjacent sub-carriers in the ranging channel, which reduces the search space from two dimensions to one dimension. And the threshold is used to detect the ranging code existing in the current ranging slot. The simulation result shows the good ranging performance of this technique.

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