Weighted Acquisition of UWB Signals Based on Energy Detection*

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SUMMARY Due to the low complexity and cost characteristics of ultra-wideband (UWB) systems, a weighted acquisition algorithm based on energy detection is proposed in this paper. This method is divided into two steps to acquire the direct path (DP) component. Firstly, weighted energy detection is applied to determine which energy block the DP lies in by generalized likelihood ratio test (GLRT). A sub-optimal weighted vector is obtained, by which the closed form of detection performance is proposed. In the second step, the precise position of DP within the detected energy block is obtained by the statistical characteristics of the channel energy distributions. Key parameters that affect acquisition performance are studied by analytical and numerical methods. Simulations and experiments are carried out for performance and complexity comparison with traditional ones. The results show that weighted acquisition achieves better performance under relative low complexity conditions.

key words: UWB, acquisition, energy detection, weighted

1. Introduction

In recent years, impulse radio ultra-wideband (IR-UWB) communication for personal area networks has attracted significant interest from researchers and engineers in the communication field [1]. UWB is characterized by the huge transmission bandwidth that is obtained through the use of pulses on the order of nanoseconds. It brings out new potentials for low-power, bandwidth-demanding wireless applications, such as low-cost, low-complexity baseband receivers with high user capacity and low-data-rate wireless sensor network applications for positioning and monitoring [2], [3].

Despite these advantages, UWB also introduces unique signal processing challenges, especially for timing synchronization and channel estimation [1]. In an indoor environment, each transmitted pulse generates hundreds of echoes that might be resolved and combined in a Rake receiver, thereby exploiting the rich diversity of the multipath channel. Unfortunately, the implementation of a Rake is complex because of the number of fingers needed to capture a significant part of the signal energy and the computational load involved in the estimation of the gains and delays of the paths [4]. Hence suboptimal receivers that do not require channel estimation are proposed for low complexity and low data rate applications, using either energy detection or transmitted reference signals. According to the Nyquist theory, an extreme high sampling rate is also required in the correlator-based receivers, which increases the cost of the transmitted reference ultra wideband (TR-UWB) system [5]. The energy detection (ED) scheme simply performs “multiply-integration-and-detection” at the receiver, which requires no channel estimation and only frame or symbol rate sampling. These features make ED the simplest and most practical one for implementation [6]. And of course, it debases reasonable performance with a sufficiently low complexity [4].

Timing synchronization is another challenge to be solved. The low duty cycle with very short pulses makes detection-based UWB synchronization algorithms difficult to implement. Even minor misalignments may result in the lack of sufficient energy capture rendering symbol detection nearly impossible [7]. Depending on whether aiding pulses are applied, UWB synchronization is divided into data-aiding (DA) and non-data aiding (NDA). NDA acquisition achieves higher efficiency, while it adds complexity to the design as the cost [8]. Due to the low complexity and cost demands, DA acquisition is applied in most applications. Now most concerned problems in the UWB synchronization field include decreasing the average acquisition time [9], improving the timing accuracy [10], and simplifying the implementation [11].

Aiming at the latter two problems, a timing acquisition method based on energy detection is proposed and evaluated in this paper. This method is divided into two steps. During the first step, a weighted energy detection architecture is applied, which is modeled by the generalized likelihood ratio test (GLRT). A theoretical closed form of detection probability is derived, by which the optimization of integration interval is fulfilled. In the second step, a refined estimation method within the detected interval is proposed based on the energy characteristics of the indoor channel models given by IEEE 802.15 working group [12]. Weighted energy detection is firstly proposed in [13], then both [14] and [15] focus on finding the optimal combing weights. But all the studies are under the assumption that perfect synchronization has been achieved beforehand, which is not practical in most situations. In [16], a similar two-step TOA method...
partly based on energy detection has been proposed in ranging applications. But in the second step, a correlator-based estimation is required, which can not reduce the cost of the UWB transceivers.

In the rest of this paper, firstly the system model is defined, then the acquisition method including both the weighted detection and refined estimation steps is introduced. Further studies on key factors of performance are carried out by analytical and numerical methods. Performance and complexity comparisons between this method and traditional ones are demonstrated in Sect. 5. A conclusion is drawn in the last part.

2. System Model

2.1 Signal Model

In a UWB system, every symbol is associated with a “block” made of \( N_f \) repeated pulses (one pulse per frame) in the form of

\[
g_s(t) = \sum_{j=0}^{N_f-1} g(t - jT_f - c_jT_c) \tag{1}
\]

where \( g(t) \) is the Gaussian monocycle, i.e.

\[
g(t) = \left[ 1 - 4\pi \left( \frac{t}{\tau_m} \right)^2 \right] e^{-2\pi(t/\tau_m)^2} \tag{2}
\]

The pulse duration is decided by \( \tau_m \). \( T_f \) is the frame duration that may be a hundred times the pulse duration, \( T_c = T_f/N_c \) is the chip duration to denote \( N_c \) chips per frame, and the sequence \( c_j \) represents the user’s pseudorandom time-hopping (TH) code with \( c_j \in [0, N_c - 1], \forall j \in [0, N - 1] \). The symbol duration spanned by \( g_s(t) \) is thus \( T_s = N_fT_f \). The UWB signals transmitted by the desired user is described by

\[
s(t) = \sum_i b_ig_s(t - iNT_f - a_i\Delta) \tag{3}
\]

where \( a_i \in [0, 1] \) and \( b_i \in \{\pm 1\} \) represent the data symbols that are modeled as binary independent and identically distributed random variables, and \( \Delta \) represents the time shift imposed on all the monocycles of a given block by a unit data symbol. (3) subsumes several widely used modulation schemes: PPM modulation arises when \( b_i = 1, \forall i \), PAM results from \( a_i = 0, \forall i \). Then the received signal \( r(t) \) from RF is

\[
r(t) = s(t) + n(t) \tag{4}
\]

\( n(t) \) is AWGN which is considered to be stationary.

2.2 System Structure and Energy Sequence Model

The weighted receiver structure applied is shown in Fig. 1. After passing through LNA and the square law device, the received signals are divided into two branches corresponding to the two steps of this method. These two branches work simultaneously.

As described in Fig. 1, whether the estimator work or not is dependent on the detection result. If the DP energy block is detected according to the detector, the second step of refined estimation starts.

In both branches, the length of signal block mass handled once is \( T_i \). According to the weighted architecture of the detector in Fig. 2, the signal is divided into \( K \) equal subblocks of size \( T_K = T_i/K \). In \( k \)th sub-block the signal is integrated from \( kT_K \) to \((k+1)T_K \), and multiplied by a weighted factor \( w_k \). The number of energy blocks within one frame is \( M = \lfloor T_i/K \rfloor \). If \( N_f \) pulses are applied, the energy element is

\[
X = [x_1, x_2, ..., x_{N_f}] \tag{5}
\]

where \( x_i = (x_i[0], x_i[1], ..., x_i[M - 1])^T \). And single output of the detector is

\[
x_i[j] = \sum_K \int_{(i-1)T_f+(j-1)T_i}^{(i-1)T_f+(j-1)T_i+kT_k} w_kr^2(t)dt \tag{6}
\]

In the estimation branch, first the received signal of \( T_i \) period is integrated, the energy sequence of the estimator is

\[
Y = [y_1, y_2, ..., y_{N_f}] \tag{7}
\]

where \( y_i = (y_i[0], y_i[1], ..., y_i[M - 1])^T \). It is the same as the traditional energy detectors, i.e.

\[
y_i[j] = \int_{(i-1)T_f+(j-1)T_i}^{(i-1)T_f+(j-1)T_i} r^2(t)dt \tag{8}
\]

\( X \) and \( Y \) are energy block models to be handled in next sections.

2.3 Data Frame Structure

A data aided acquisition is applied in this paper. The data
frame is shown in Fig. 3. Data bits enter the demodulator after acquisition and timing recovery are fulfilled.

3. Acquisition Algorithm

3.1 Acquisition Target and Method

The main difference between the acquisition problems in a multipath channel and a channel without multipath is that there are more than one hypothesized phase which can be considered a good estimate of the true signal phase. From the viewpoint of the postacquisition receiver, if a certain nominal uncoded bit-error rate (BER) is achieved, the signal phase acquired is considered a good estimate of the true signal phase. All these signal phases are called a hit set. Due to the characteristics of UWB transmission, there’re no such definitions like phase or angle. The relative delay within one $T_i$ is defined as the phase of UWB systems [17]. The corresponding hit set is regarded as

$$S_h = \{ \hat{\tau} : P_e(\tau) \leq \lambda_h \}$$  

(9)

In order to clarify the the acquisition target based on the hit set, the evaluation of BER with different $\tau$ is carried out. Taking OOK for example, the decision variable of space and mark under perfect synchronization is

$$H_0 : Z_0 = \int_0^{T_i} [n(t)]^2 dt$$

$$H_1 : Z_1 = \int_0^{T_i} [s(t) + n(t)]^2 dt$$  

(10)

The decision variables of $Z_0$ and $Z_1$ are central and non-central Chi-square distributed respectively, which is difficult to be handled directly. It was shown in [6], [10] by approximating the Chi-square distributions with Gaussian distributions, which becomes more precise for large degrees of freedom (DOF) defined by $2M = 2T_i/W + 1$, $W$ is the signal bandwidth. Then Gaussian approximation of the decision variable is

$$Z_0 \sim N(\mu_0, \sigma_0^2)$$

$$Z_1 \sim N(\mu_1, \sigma_1^2)$$  

(11)

where

$$\mu_0 = MN_0, \sigma_0^2 = MN_0^2$$

$$\mu_1 = MN_0 + E_s, \sigma_1^2 = MN_0^2 + 2N_0E_s$$  

(12)

$E_s$ is defined as the signal energy contained in the integrator, which is also the non-central parameter of non-central Chi-square distribution.

Define $\gamma$ as the decision threshold, likelihood ratio test (LRT) is applied for demodulation. Define $P_{te}$ as the error rate when transmitting bit 'i' ($i=0,1$). BER includes both $P_{te}$ and $P_{te}$, which is defined as

$$P_{te} = \int_0^\infty p_0(z)dz = Q\left(\frac{\gamma - \mu_0}{\sigma_0}\right)$$

$$P_{te} = \int_0^\infty p_1(z)dz = 1 - Q\left(\frac{\gamma - \mu_1}{\sigma_1}\right)$$  

(13)

where $p_0(z)$ and $p_1(z)$ are probability density function (PDF) of $Z_0$ and $Z_1$. In a practical system, the threshold $\gamma$ is customarily set to that $P_{te}$ and $P_{te}$ are equal [18]. Therefore, the BER performance of the non-coherent receiver can be approximately expressed as $P_e = P_{te} = P_{te}$. The threshold is obtained from previous analysis, i.e.

$$\gamma = MN_0 + \frac{E_s}{\sqrt{MN_0^2 + 2E_sN_0}}$$  

(14)

Then

$$\gamma = MN_0 + \frac{E_s}{\sqrt{MN_0^2 + 2E_sN_0}}$$  

(15)

Define $E_s(\tau)$ is the signal energy contained in the demodulation interval when timing error is $\tau$. The closed form of BER and $\tau$ is

$$P_e = Q\left(\frac{E_s(\tau)}{\sqrt{MN_0^2 + 2E_s(\tau)N_0}}\right)$$

$$= Q\left(\frac{1}{\sqrt{MC^2 + 2C}}\right)$$  

(16)

where $C = N_0/E_s(\tau)$ represents the inverse of the contained SNR. Taking IEEE 802.15.4a CM1 channel as an example, a cubic spline function curve fitting process is carried out in Fig. 4. The closed form between BER and relative delay ($\tau$) can be obtained by the fitting coefficients and (16).

Taking a certain BER $(10^{-3})$ as the threshold, it can be seen from Fig. 5 that the demodulation performance is qualified when $\tau$ is between $-4\text{ ns}$ and $2\text{ ns}$ in IEEE 802.15.4a CM1-CM4 models [12]. But the phase of $\tau = 0$ will achieve the best performance which is regarded as the direct path (DP) component in UWB ranging. Similarly, it is suggested to take DP as the acquisition target.

This weighted acquisition method is divided into two steps, illustrated in Fig. 6. If DP is contained in the $m$th block, and $\delta_{DP}$ is the relative delay, the corresponding TOA to be estimated is

$$\tilde{\tau}_0 = (m_{DP} - 1)T_i + \delta_{DP}$$  

(17)

3.2 DP Block Detection

As described in Sect. 2, the integration interval is $T_i$, the
The detection process is formulated as a hypothesis testing problem. The sub-block in the detector is Chi-square distributed. The energy element of sub-block in the detector is Chi-square distributed. The energy element of sub-block in the detector is Chi-square distributed. The detection process is formulated as a hypothesis testing problem based on Gaussian approximation, i.e.

\[ H_0 : x[n] \sim N(\mu_c, \sigma_c^2) \]
\[ H_1 : x[n] \sim N(\mu_{nc}, \sigma_{nc}^2) \]  

(18)

The parameters in (18) can be obtained as the detector structure in Fig. 2 [13].

\[ \mu_c = w^T M N_0 1, \sigma_c^2 = w^T M N_0^2 w \]
\[ \mu_{nc} = w^T M N_0 1 + w^T E_s \]
\[ \sigma_{nc}^2 = w^T M N_0^2 w + 2 N_0 (w^T H w) \]  

(19)

where \( M \) is DOF of Gaussian approximation (\( 2M = 2T_K W + 1 \)), \( w = [w_0, w_1, ... w_{K-1}]^T \) is the weighted vector, \( E_s = [E_s(0), E_s(1), ... E_s(K-1)]^T \) is the signal energy contained within every sub-block under \( H_1 \), \( H = \text{diag}(E_s) \), \( 1 = [1, ..., 1]^T \), \( N_0 \) is the noise power density.

Then the detection can be modeled as a classical detection problem with unknown parameters in AWGN, for which the GLRT is well motivated. The GLRT rejects \( H_0 \) if

\[ L_G(x) = \frac{p(x[n]; \hat{E}_s, H_1)}{p(x[n]; H_0)} > \gamma \]

(20)

\( \gamma \) is a threshold set up under the desired probability of false alarm (FA), \( \hat{E}_s \) is the estimation of \( E_s \). According to the Neyman-Pearson detection criterion, the false alarm probability \( P_{FA} \) is obtained by

\[ P_{FA} = P(H_1 : H_0) = Q \left( \frac{\gamma - \mu_c}{\sigma_c} \right) \]  

(21)

The threshold is obtained by

\[ \gamma = \sigma_c Q^{-1}(P_{FA}) + \mu_c = \sqrt{w^T M N_0^2 w Q^{-1}(P_{FA}) + w^T M N_0 1} \]  

(22)

And the detection probability is

\[ P_D = P(H_1 : H_1) = Q \left( \frac{\gamma - \mu_{nc}}{\sigma_{nc}} \right) = Q \left( \frac{\sqrt{w^T M N_0^2 w Q^{-1}(P_{FA}) - w^T E_s}}{\sqrt{w^T M N_0^2 w + 2 N_0 (w^T H w)}} \right) \]  

(23)

In order to optimize the detection performance, a proper weighted vector is required. Because the Q-function is monotonic decreasing, the optimal vector is expected that

\[ w_{opt} = \arg \min J(w) \]
\[ = \arg \min \sqrt{w^T M N_0^2 w Q^{-1}(P_{FA}) - w^T E_s} \]
\[ \sqrt{w^T M N_0^2 w + 2 N_0 (w^T H w)} \]  

(24)

Let \( \frac{\partial J(w)}{\partial w} = 0 \), the optimal weighted vector can be obtained. According to the matrix theory and conclusions in [13], the optimal weighted vector is \( w_{opt} = \Lambda^{-1} E_s \), where \( \Lambda = M C^2 T + 2 C H \) is a diagonal positive-definite matrix describing the noise contributions across the \( K \) integration sub-blocks. \( C = N_0 / E_b, T = \text{diag}(t), \) and \( t = [T_0, T_1, ..., T_{K-1}]^T \).

Considering that the sub-blocks are equally divided in this paper for simplification and implementation, \( t = \frac{T}{K}(1, 1, ..., 1)^T \). From (24) and (19), a sub-optimal weighted vector is given.

\[ w_{sub-opt} = [E_s(0), E_s(1), ... E_s(K-1)]^T \]  

(25)

In [10], \( E_s \) is obtained by conditional maximum likelihood estimation, which is not easy to follow. Since the channel
energy characteristics is given by IEEE 802.15.4a working group, a curve fitting method is more convenient and precise.

Taking DP as the starting point, the normalized signal energy within this block is represented by $e(t)$ when the integration interval is $t$. During the channel energy curve fitting process, the environment conditions applied are IEEE 802.15.4a CM1 and CM2, for typical LOS/NLOS scenarios. 1000 channel realizations are taken to average the energy results. The normalized energy curve can be fitted as exponential distribution in Fig. 7, i.e.

$$e(t) = 1 - e^{-\alpha t} \quad (26)$$

where $\alpha$ is dependent on the given channel model, listed in Table 1.

In appendix A, $E_s$ is obtained, i.e.

$$E_s(k) = \frac{E_b}{K} \left[ k + 1 + \frac{1}{\alpha T_K} (e^{-\alpha(k+1)T_K} - 1) \right]$$

(27)

where $k = 0, 1, ... K - 1$. $E_b$ is the energy of one transmitted pulse. According to appendix B, $E_b$ can be obtained by

$$E_b = \frac{1}{N_f} \sum_{n=0}^{N-1} \int_0^{T_f} r^n(t)dt - \int_0^{T_f} n^n(t)dt \quad (28)$$

From (25) and (27), the sub-optimal weighted vector is

$$\mathbf{w}_{\text{sub–opt}} = \left[ E_s(0), E_s(1), ... E_s(K - 1) \right]^T$$

$$= \left\{ \frac{E_b}{K} \left[ 1 + \frac{e^{-\alpha T_K}}{\alpha T_K} - 1, ... K + \frac{e^{-\alpha K T_K}}{\alpha T_K} - 1 \right] \right\}^T \quad (29)$$

The closed form of detection probability is

$$P_D = Q \left( \frac{\mathbf{w}^T M N_0^2 \mathbf{w} Q^{-1}(P_{FA}) - \mathbf{w}^T \mathbf{E}_s}{\sqrt{\mathbf{w}^T M N_0^2 \mathbf{w} + 2 N_0 \mathbf{w}^T H \mathbf{w}}} \right)$$

$$= Q \left( \sqrt{\frac{M N_0^2 \sum_{k=0}^{K-1} E_s^2(k) - \sum_{k=0}^{K-1} E_s^2(k)}{M N_0^2 \sum_{k=0}^{K-1} E_s^2(k) + 2 N_0 \sum_{k=0}^{K-1} E_s^2(k)}} \right)$$

(30)

where $E_s$ is demonstrated in (27).

3.3 Refined Estimation of DP

According to the receiver structure described in Fig. 1, if the threshold is exceeded in the detection step, the estimation module starts to acquire the precise position of DP within this detected energy block. In traditional TOA measurements, the middle point of the energy block is regarded as the DP position by approximation [20], [21]. In [16], a two-step TOA estimation is introduced as shown in Fig. 8, where matched filtering (MF) is adopted in the second step. It is restricted in lots of applications for its extreme high sampling rate demands. In our method, the precise position of DP is obtained by applying multiple acquisition-aided pulses and the energy characteristics of UWB channel models, which requires low complexity and computational conditions.

In Sect. 2, the energy sequence model is introduced in both branches of the receiver. Similar with $M$ depicted in the previous section, $M'$ is the DOF of $y_f[m]$, and $2M' = 2T_f W + 1$. When the aiding pulse number for acquisition is $N_f$, the energy element is

$$y_f[m_{DP}] = \frac{1}{N_f} \sum_{i=0}^{N_f-1} y_f[m_{DP}]$$

(31)

The signal energy contained in the DP block is
\[ y_i[mDP] \approx \int_{(mDP - 1)T_i}^{mDP T_i} r_i^2(t) dt - \int_{(mDP - 1)T_i}^{mDP T_i} n_i^2(t) dt \]  

by approximation. The noise energy \( E_n = \int T_i n_i^2(t) dt \) can be obtained from the energy block before DP arrives. Then \( y(mDP) \) is obtained by averaging the multiple transmitted pulses. According to the energy statistics of UWB channels, the relative delay of \( \delta_{DP} \) can be calculated.

\[
y[mDP] = \frac{1}{N_f} \sum_{i=0}^{N_f-1} y_i[mDP] = E_b[1 - e^{-\alpha(T_i - \delta_{DP})}] \quad (33)
\]

By which the relative delay of DP is

\[
\hat{\delta}_{DP} = \frac{1}{\alpha} \ln \left( 1 - \frac{1}{N_f E_b} \sum_{i=0}^{N_f-1} y_i[mDP] \right) - T_i \quad (34)
\]

Since the right of (33) is a statistical value, considering the random error occurred in one single transmission, a simple way to solve this problem is to take multiple pulses to average the unavoidable fluctuations. A detailed discussion is carried out later. After acquiring \( m_{DP} \) and \( \hat{\delta}_{DP} \), the signal TOA is obtained by (17).

### 4. Performance Analysis

In this section, the performance related parameters are analyzed. Since the detection probability is presented with a closed form, it’s more convenient to study than the second step, during which the analysis is based on the numerical results.

#### 4.1 Key Factors of Detection Step

The analysis of detection step is based on (30) which is verified in Fig. 9 under conditions of \( K = 2 \), \( E_b/N_0 = 10 \) dB, and \( 1 \leq M \leq 50 \).

The simulation results differ from the model when DOF is small due to the error introduced by Gaussian approximation. When a relative large integration interval is applied, the closed form of DP detection is accurate enough.

During the detection step, the integration interval \( T_i \) and sub-block number \( K \) are analyzed respectively based on (30). The optimization of integration interval is investigated by the hypothesis testing problem formulated in Sect. 3, and DOF is taken as the bridge between the detection performance and the interval.

From (30), the influence of \( M \) on detection performance is illustrated. Considering \( 2M = 2T_K W + 1 \) and \( T_i = K T_K \), it can be seen that under the conditions of Fig. 10 (CM1, \( E_b/N_0 = 10 \) dB), the optimal sub-interval \( T_K \) varies with \( K \). But the whole interval \( (T_i) \) preferred in Fig. 10 remains about 20 ns, for \( T_i = K T_K \).

In Fig. 11, the theoretical detection performance with different sub-block number is given. It is clear that the larger value of \( K \) is, the better results are achieved, which is also demonstrated in Fig. 10. The comparison between the weighted and non-weighted detection method is also shown. The non-weighted method achieves a comparative performance with the \( K = 1 \) case. When \( K \) exceeds 2, the weighted detection will have an obvious performance advantage while it requires an increasing complexity of the receiver as the cost.

#### 4.2 Key Factors in Estimation Step

During the estimation step, the related parameters include...
$T_i$ and the acquisition-aided pulse number $N_f$. Because no closed form is given, Monte-Carlo simulation for performance evaluation in CM1 is carried out. From Fig. 12, it can be seen that when the interval increases, the mean absolute error (MAE) scale increases correspondingly, because large $T_i$ implies a large uncertainty region for estimation.

Since the estimation is based on the channel statistical characteristics, the results seem to be meaningless when $N_f = 1$. In order to average the unpredictable fluctuation of one single transmission, it’s evident of utilizing multiple pulses to achieve better performance. From Fig. 12, if the detecting pulses increase, the estimation error decreases. And when detecting pulses exceeds 50, the performance promotion is not obvious. So it’s suggested to use no more than 50 pulses, in view of the system efficiency.

5. Simulations and Experiments

5.1 Simulation and Result Analysis

Based on the discussion in Sect. 4, the performance of the weighted energy acquisition method is evaluated by simulation. CM1 and CM2 of IEEE 802.15.4a are employed as the channel model. 1000 different realizations for each channel model are generated, and the TOA of DP is uniformly distributed within $(0, T_f)$. Gaussian monocycle pulse with 1 ns duration is considered throughout the simulations and $W$ is regarded as nearly 1 GHz. The transmission data rate is 12.5 Mbps, and $T_f = 80$ ns. The false alarm probability applied in N-P criterion is $P_{FA} = 10^{-3}$. The integration interval applied is 20 ns in CM1, and 30 ns in CM2. The sub-block number tested is $K = 2, 10$. Pulse number applied for acquisition is $N_f = 5, 50$. In Fig. 13, this weighted acquisition results are compared with two well investigated TOA algorithms: Threshold Crossing (TC) and Maximum Energy Selection-Search Back (MES-SB). The related parameters including the normalized threshold $\xi_{norm}$, the search back window length $W_{sb}$ and the energy block length $T_b$ are set as [21].

The total timing error of this weighted acquisition is consisted with detection error and estimation error, i.e.

$$e_{TOTAL} = e_{DET} + e_{EST} \quad (35)$$

When the SNR is low, the error is mainly the detection error $e_{DET}$. The detection error includes the missing detection error $e_M$ and the false detection error $e_{FA}$. $e_M$ is generated when all the energy blocks can not be detected over the threshold. According to [22], the maximum energy block is chosen as the DP energy block, which will introduce an unpredictable error. The effective ways of improving the performance is to increase the sub-block number ($K$), choosing a preferred integration interval and adding the synchronization-aided pulses. $e_{EST}$ is the main part of $e_{TOTAL}$ under high SNR conditions. In Fig. 13, if the SNR exceeds 16 dB in CM1, the weighted method will have a gain of 2–4 dB when 50 pulses are applied. In CM2, the gain under high SNR conditions is even higher.

From Fig. 13, it’s shown that when SNR is high enough, there exists a lower bound on the estimation accuracy. Under high SNR conditions, as discussed above, $e_M \approx 0$, $e_{DET}$ is mainly introduced by $e_{FA}$. The FA block is uniformly distributed between the first block and DP block, i.e. $n_{FA} \sim U(n_1, n_{DP})$. Then the detection error is

$$e_{LB-DET}(n, n_{DP}, T_i) = P_{FA} \cdot \frac{1}{n_{DP} - n_1} |n - n_{DP}| T_i \quad (36)$$

For the DP block is uniformly lied in between $n_1$ and $n_M$, the
where \( n_M = T_f / T_i \), then

\[
e_{LB-DET}(T_i) = E(e_{LB-DET}(n, n_{DP}, T_i)) = P_{FA} \cdot \frac{T_i}{2} \tag{37}
\]

It can be seen that the detection part of error lower bound is mainly affected by the preset \( P_{FA} \) and \( T_i \). Taking \( P_{FA} = 10^{-3} \), \( T_i = 20\)ns for example, the detection error lower bound is 0.2 ns, which is a minor part of the total error.

As to the estimation error \( e_{EST} \), it’s affected by more factors than detection model. In this section, a numerical method is carried out with different \( T_i \) under CM1, where SNR and \( N_f \) are ideal. The estimation results is shown in Fig. 14, which can be fitted with a linear function

\[
e_{LB-EST}(T_i) = 0.0163T_i + 1.357 \tag{38}
\]

The lower bound from analysis above is

\[
e_{LB-TOTAL} = e_{LB-DET} + e_{LB-EST} = \left( P_{FA} / 2 + 0.0163 \right) T_i + 1.357 \tag{39}
\]

After performance comparison with typical DP detection algorithms, it can be seen that during high SNR conditions, the performance of this weighted acquisition is close to the traditional ones; meanwhile the DP refined estimation in the second step also benefits the whole performance. Under low SNR conditions, this weighted method has an obvious advantage over the traditional ones for its high detection probability with the weighted mechanism, which can promote the detection performance evidently, on the cost of an increasing complexity of the receiver. But this method reserves the low sampling-rate demands of energy detection receivers, which makes it practical in the most applications.

5.2 Experiments and Complexity Comparison

In order to validate the practicability of the weighted acquisition method, both the weighted and non-weighted algorithms that have the same second step are realized in a digital UWB receiver board (Fig. 15). A simple case of weighted acquisition \( (K = 2) \) is implemented. And a same differential demodulation mechanism is applied after weighted/non-weighted acquisition [19]. The transmitted UWB signal is near-Gaussian pulse. The frequency bandwidth of the signaling is between 0.4G and 1.4 GHz according to the transceiver antennas. On the receiver board, a piece of stratiﬁ II FPGA (EP2S60F1024C) is utilized, and the complexity of different algorithms are evaluated by the compiler of Quartus II 7.0, in Table 2

BER test experiment is carried out by transmitting a 128K bytes text file at a data rate of 12.5 Mbps, demonstrated in Fig. 16. The synchronization-aided pulses applied in 50-bit-length. Due to some environment influences such as the narrow band interference and clock drift between transceivers, the data frame is 4096-bit-length, which shares the same structure as Fig. 3. The measurement distance varies from 3 m to 6 m, and the average SNR is calculated from the collected waveforms. Different \( \alpha \) from 0.01 to 0.1 at a step of 0.01 is tested since typical values are given in Table 1. The corresponding BER results are shown in Fig. 17. According to the definition of \( \alpha \), an ideal method should be based on channel estimation. When the amplitude and relative delay of each multipath component is obtained, the
precise $\alpha$ of a certain environment can be realized. But as introduced before, the channel estimation module will add large amount of complexity to the receiver. From Fig. 17, it can be seen that BER changes not so fast with different $\alpha$ due to the characteristics of energy detection. Then a preferred method is to preset $\alpha$ by typical channel parameters to avoid channel estimation, which could reach a good tradeoff between performance and complexity.

Although the energy demodulator is not so sensitive to the timing error as correlator-based demodulator, it can be seen that weighted acquisition achieves a performance advantage over non-weighted ones from Table 3. And the additional complexity is acceptable according to Table 2. Furthermore, it does not require extreme high sampling rate, which reserves the low cost merit of energy detection.

6. Conclusion

In this paper, a weighted acquisition algorithm for UWB signals based on energy detection is proposed to acquire the DP component in typical multipath environments. A mathematical model is set up for this method, and key parameters analyses are carried out. Performance and complexity comparison between this method and traditional ones shows that weighted acquisition has an obvious performance advantage, especially under low SNR conditions. The additional complexity is acceptable according to our practical implementation, which makes it feasible in most applications.

References

Appendix A: Closed Form of $E_s$

When DP lies in the first energy sub-block, the signal energy contained within the $k + 1$th sub-block is defined as $h_k$ for assistance. As depicted in Fig. A.1 Considering the appearance of DP is uniformly distributed within the first energy sub-block, i.e. $p_{DP}(t) = 1/T_K$, the expectation of $h_k$ is obtained by

$$h_0 = E_b \int_0^{T_{K}} e(t)p_{DP}(t) dt$$

$$= \frac{E_b}{T_K} \int_0^{T_{K}} (1 - e^{-\alpha t}) dt$$

$$= E_b \left[ 1 + \frac{1}{\alpha T_K} (e^{-\alpha T_K} - 1) \right]$$

$$...$$

$$h_{K-1} = E_b \left[ 1 + \frac{1}{\alpha T_K} (e^{-\alpha K T_K} - 1) \right]$$

(A-1)

Actually, the position of DP is also uniformly distributed among all the sub-blocks, i.e. $P_{DP} = 1/K$. Then $E_s$ is achieved by

$$E_s(k) = \frac{E_b}{K} \sum_{i=0}^{k} h_i$$

$$= \frac{E_b}{K} \left( k + 1 + \frac{1}{\alpha T_K} (e^{-\alpha (k+1) T_K} - 1) \right)$$

(A-2)

where $k = 0, 1, ... K - 1$. And $E_s$ is shown in (27).

Appendix B: Numerical Method of $E_b$

According to the definition of $E_b$, it means the signal energy of one transmitted pulse, i.e.

$$E_b = \int_0^{T_{f}} s^2(t) dt$$

(A-3)

In (4), the received signal is consisted with signal and noise. Then (A-3) can be rewritten as

$$E_b = \int_0^{T_{f}} r^2(t) dt$$

For $r(t) = s(t) + n(t)$, (A-4) is

$$E_b = \int_0^{T_{f}} r^2(t) dt + \int_0^{T_{f}} n^2(t) dt - 2 \int_0^{T_{f}} r(t)n(t) dt$$

(A-4)

Because the transmitted signal and noise are non-coherent, i.e.

$$\int_0^{T_{f}} s(t)n(t) dt = 0$$

(A-6)

(A-5) can be rewritten as

$$E_b = \int_0^{T_{f}} r^2(t) dt - \int_0^{T_{f}} n^2(t) dt$$

(A-7)

Based on (A-7), $E_b$ can be obtained from the received signal $r(t)$ and noise $n(t)$ by following steps:

1. Initialize the parameters, such as $N_f$, $T_f$, etc.
2. Obtain the noise energy $E_n$ by

$$E_n = \frac{1}{N_f} \sum_{n=0}^{N_f-1} \int_0^{T_{f}} n^2(t) dt$$

(A-8)

3. Turn on the transmitter to begin acquisition.
4. Obtain the received signal $E_r$ by

$$E_r = \frac{1}{N_f} \sum_{n=0}^{N_f-1} \int_0^{T_{f}} r^2(t) dt$$

(A-9)

5. $E_b$ is obtained by $E_b = E_r - E_n$. 

![Fig. A.1 Definition of $h_k$.](image)
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