Timing Acquisition in Ultra-wideband Communication Systems

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(Invited Paper)

***Abstract*—The goal of this paper is to highlight the significance of the timing acquisition problem in ultra-wideband (UWB) communication systems and discuss efficient solutions to the problem. We discuss how the distinguishing features of UWB communication systems, such as their wide bandwidth and low transmission power constraints, are responsible for making the acquisition of UWB signals a difficult task. A survey of the current approaches to UWB signal acquisition is also given. In addition, we discuss some of the issues and challenges in UWB signal acquisition which may not have received sufficient attention in existing literature.**

***Index Terms*—Acquisition, impulse radio, spread spectrum, ultra wideband (UWB).**

I. Introduction

A CLASS OF spread spectrum techniques known as ultra-wideband (UWB) communication [1]–[4] has recently received a significant amount of attention from academic researchers as well as from the industry. UWB signaling is being considered for high data rate wireless multimedia applications for the home entertainment and personal computer industry, as well as for low data rate sensor networks involving low power devices. It is also considered a potential candidate for alternate physical layer protocols for the high-rate IEEE 802.15.3 and the low-rate IEEE 802.15.4 wireless personal area network (WPAN) standards [5], [6].

In any communication system, the receiver needs to know the timing information of the received signal to accomplish demodulation. The subsystem of the receiver which performs the task of estimating this timing information is known as the synchronization stage. Synchronization is an especially difficult task in spread spectrum systems which employ spreading codes to distribute the transmitted signal energy over a wide bandwidth. The receiver needs to be precisely synchronized to the spreading code to be able to despread the received signal and proceed with demodulation. In spread spectrum systems, synchronization is typically performed in two stages [7], [8]. The first stage achieves coarse synchronization to within a reasonable amount of accuracy in a short time, and is known as the acquisition stage. The second stage is known as the tracking stage and is responsible for achieving fine synchronization and maintaining synchronization through clock drifts occurring in the transmitter and the receiver. Tracking is typically accomplished using a delay locked loop [7]. Timing acquisition is a particularly acute problem faced by UWB systems, as explained in the following. This paper addresses the significance of the acquisition problem in UWB systems and the ways to efficiently tackle it.

Short pulses and low duty cycle signaling [1] employed in UWB systems place stringent timing requirements at the receiver for demodulation [9], [10]. The wide bandwidth results in a fine resolution of the timing uncertainty region, thereby imposing a large search space for the acquisition system. Typical UWB systems also employ long spreading sequences spanning multiple symbol intervals in order to remove spectral lines resulting from the pulse repetition present in the transmitted signal. In the absence of any side information regarding the timing of the received signal, the receiver needs to search through a large number of phases[1] at the acquisition stage. This results in a large acquisition time if the acquisition system evaluates phases in a serial manner, and results in a prohibitively complex acquisition system if the phases are evaluated in a parallel manner.

Moreover, the relatively low transmission power of UWB systems requires the receiver to process the received signal for long periods of time in order to obtain a reliable estimate of the timing information. In a packet based network, each packet has a dedicated portion known as the acquisition preamble within which the receiver is expected to achieve synchronization. However, for the high data rate applications envisaged for UWB signaling, long acquisition preambles would significantly reduce the throughput of the network.

The transmitted pulse can be distorted through the antennas and the channel, and hence the receiver may not have exact knowledge of the received pulse signal waveform [11]. The short pulses used in UWB systems also result in highly resolvable multipath with a large delay spread at the receiver [12]. The UWB receiver could therefore synchronize to more than one possible arriving multipath component (MPC) and still perform satisfactorily. This means that there could exist multiple phases in the search space which could be considered acceptable and could be exploited to speed up the acquisition process.

These challenges arising from the signal and channel characteristics unique to UWB systems indicate the significance of the acquisition problem in UWB systems and the need to address it efficiently. Addressing some of these issues is the focus

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[1]Traditionally, in direct sequence spread spectrum systems, the chip level timing of the PN sequence is referred to as the phase of the spreading signal. In this paper, we use "phase" and "timing" interchangeably.

of the current paper, which is organized as follows. Section II briefly summarizes the acquisition approaches adopted by traditional spread spectrum systems. Current research on UWB signal acquisition is described in Section III. Some issues in UWB acquisition system design are discussed in Section IV, and the conclusions are presented in Section V.

II. Acquisition Methods in Traditional Spread Spectrum Systems

UWB communication falls in the category of spread spectrum communication systems. In this section, we briefly review the main features of acquisition methods used in traditional spread spectrum systems to put the current approaches to UWB signal acquisition in perspective. There has been extensive research on spreading code acquisition and tracking for spread spectrum systems with direct sequence, frequency hopping, and hybrid modulation formats [13], [8], [7]. We will bring out the main issues by considering the timing acquisition of direct sequence spread spectrum systems.

In a direct sequence spread spectrum system, the receiver attempts to despread the received signal using a locally generated replica of the spreading waveform. Despreading is achieved when the received spreading waveform and the locally generated replica are correctly aligned. If the two spreading waveforms are out of synchronization by even a chip duration, the receiver may not collect sufficient energy for demodulation of the signal. As mentioned before, the synchronization process is typically divided into two stages: acquisition and tracking. In the acquisition stage, the receiver attempts to bring the two spreading waveforms into coarse alignment to within a chip duration. In the tracking stage, the receiver typically employs a code tracking loop which achieves fine synchronization to within a chip duration. If the received and locally generated spreading waveforms go out of synchronization by more than a chip duration, the acquisition stage of the synchronization process is reinvoked. The reason for this two stage structure is that it is difficult to build a tracking loop which can eliminate a synchronization error of more than a fraction of a chip.

A typical acquisition stage attempts to bring the synchronization error down to be within the pull-in range of the tracking loop by searching the timing uncertainty region in increments of a fraction of a chip. A simplified block diagram of an acquisition stage which is optimal in the sense that it achieves coarse synchronization with a given probability in the minimum possible time is the parallel acquisition system [7] shown in Fig. 1. This stage checks all the candidate phases in the uncertainty region simultaneously, and the phase corresponding to the maximum correlation value is declared to be the phase of the received spreading waveform. In an additive white Gaussian noise (AWGN) channel, this acquisition strategy produces the maximum-likelihood estimate (from among the candidate phases) of the phase of the received spreading waveform. However, the hardware complexity of such a scheme may be prohibitive since it requires as many correlators as the number of candidate phases being checked, which may be large depending on the size of the timing uncertainty region. A widely used

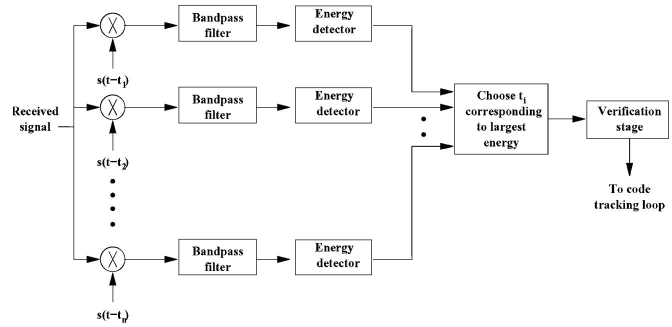

Fig. 1. Block diagram of a parallel acquisition system for direct sequence spread spectrum systems which evaluates the candidate phases $t_1, t_2, \ldots, t_n$. In the ith arm, the decision statistic corresponding to the candidate phase t_i is generated by correlating the received signal with a delayed version of the locally generated spreading waveform $s(t)$.

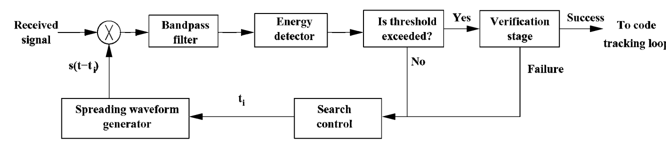

Fig. 2. Block diagram of a serial acquisition system for direct sequence spread spectrum systems which evaluates the candidate phases $t_1, t_2, \ldots, t_n$ serially until the threshold is exceeded. The decision statistic corresponding to the candidate phase t_i is generated by correlating the received signal with a delayed version of the locally generated spreading waveform $s(t)$. If the threshold is not exceeded, the search updates the value of the candidate phase and the process continues.

technique for coarse synchronization, which trades off hardware complexity for an increase in the acquisition time, is the serial search acquisition system shown in Fig. 2. This system has a single correlator which is used to evaluate the candidate phases serially until the true phase of the received spreading waveform is found. Hybrid methods, such as the MAX/TC criterion [14], have also been developed which employ a combination of the parallel and serial search acquisition schemes and reduce the acquisition time at the cost of increased hardware complexity. All the acquisition schemes employ a verification stage [15] which is used to confirm the coarse estimate of the true phase before the control is passed to the tracking loop.

In traditional spread spectrum acquisition schemes, the signal-to-noise ratio (SNR) of the decision statistic improves with an increase in the dwell time, which is the integration time of the correlator. Thus the probability of correctly identifying the true phase of the received spreading waveform can be increased by increasing the time taken to evaluate each candidate phase. This tradeoff has been identified and exploited by several researchers for the development of more efficient acquisition schemes, and has led to their classification into fixed dwell time and variable dwell time schemes [8], [7]. The fixed dwell time based schemes are further classified into single and multiple dwell schemes [16]. The decision rule in a single dwell scheme is based on a single fixed time observation of the received signal, whereas a multiple dwell scheme comprises multiple stages with each stage attempting to verify the decision made by a previous stage by observing the received signal over a comparatively longer duration. Variable dwell time methods are based on the principles of sequential detection [17] and are aimed at reducing the mean dwell time. The integration time is allowed to be continuous and incorrect candidate phases are dismissed quickly, which results in a smaller mean dwell time.

Several performance metrics have been used to measure the performance of acquisition systems for spread spectrum systems. The usual measure of performance is the mean acquisition time, which is the average amount of time taken by the receiver to correctly acquire the received signal [7], [18], [8]. The variance of the acquisition time is also a useful performance indicator, but is usually difficult to compute. The mean acquisition time is typically computed using the signal flow graph technique [19]. For parallel acquisition systems, a more appropriate performance measure is the probability of acquisition or alternatively the probability of false lock [20].

In the presence of multipath, there could exist more than one phase which could be considered to be the true phase of the received signal. However, few acquisition schemes for spread spectrum systems [21], [22] have taken this into consideration.

III. Signal Acquisition in UWB Systems

As discussed in Section 1, the distinguishing feature of UWB systems is the wide bandwidth and the relatively low transmission power constraint imposed by regulatory bodies. The wide bandwidth enables fine timing resolution, resulting in a large number of resolvable paths in the UWB channel response. There may be more than one path where a receiver lock could be considered successful acquisition. The stringent power constraint necessitates the use of long spreading sequences which together with fine timing resolution results in a large search space for the acquisition system. So, the main difference between the acquisition problems for UWB systems and traditional spread spectrum systems is the presence of multiple acquisition phases and the relatively large search space in the former.

The large search space obviates the use of a fully parallel acquisition system due to its high hardware complexity. Hence, much of the existing work on UWB signal acquisition has focused on serial and hybrid acquisition systems. Several researchers have tackled the large search space problem by proposing schemes which involve more efficient search techniques. However, the existence of multiple acquisition states has received relatively less attention and has not been sufficiently exploited. Furthermore, a significant portion of the existing work assumes an AWGN channel model for the UWB channel and neglects the effect of multipath in the development and evaluation of the proposed acquisition schemes.

In Section III-A, we describe general models for the propagation channel and the acquisition signal for UWB systems. This model will be used later to describe the main features of some of the proposed schemes for UWB signal acquisition.

A. Signal and Channel Models

The transmitted UWB signal consists of a train of short pulses (monocycles) which may be dithered by a time hopping (TH) sequence to facilitate multiple access and to reduce spectral lines. The polarities of the transmitted pulses may also be randomized using a direct sequence (DS) spreading code to mitigate multiple access interference (MAI). The generalized UWB signal transmitted during the acquisition process for a single user can be expressed as a series of UWB monocycles $\psi(t)$ of width T_{p} each occurring once in every frame of duration T_{f} as

$$x(t) = \sum_{l=-\infty}^{\infty} b_{\lfloor l/N_{\mathrm{b}} \rfloor} a_{[l/N_{\mathrm{ds}}]} \psi\left(t - lT_{\mathrm{f}} - c_{[l/N_{\mathrm{th}}]} T_{\mathrm{c}}\right) \quad (1)$$

where N_b is the number of consecutive monocycles modulated by each data symbol b_l, T_f is the pulse repetition time, T_c is the chip duration which is the unit of additional time shift provided by the TH sequence and $[\cdot]$, $\lfloor \cdot \rfloor$ denote the integer division remainder operation and the floor operation, respectively. The pseudorandom TH sequence $\{c_l\}_{l=0}^{N_{\mathrm{th}}-1}$ has length N_{th} where each c_l takes integer values between 0 and $N_h - 1$ where N_h is less than the number of chips per frame $N_f = T_f/T_c$. The DS sequence $\{a_l\}_{l=0}^{N_{\mathrm{ds}}-1}$ has length N_{ds} with each a_l taking the value $+1$ or -1. Some UWB systems may employ only TH ($a_l = +1$) or only DS ($c_l = 0$) spreading and may not send any data ($b_l = +1$) during the acquisition stage.

The UWB indoor propagation channel can be modeled by a stochastic tapped delay line [23], [12] which can be expressed in the general form in terms of its impulse response as:

$$h(t) = \sum_{k=0}^{N_{\mathrm{tap}}-1} h_k f_k(t - t_k) \quad (2)$$

where N_{tap} is the number of taps in the channel response, and h_k is the path gain at excess delay t_k corresponding to the kth path. Due to the frequency sensitivity of the UWB channel, the pulse shapes received at different excess delays are path-dependent [24]. The functions $f_k(t)$ model the combined effect of the transmit and receive antennas and the propagation channel corresponding to the kth path on the transmitted pulse.

The received signal from a single user can then be expressed as

$$r(t) = \sum_{l=-\infty}^{\infty} b_{\lfloor l/N_b \rfloor} a_{[l/N_{\mathrm{ds}}]} w_r\left(t - lT_f - c_{[l/N_{\mathrm{th}}]} T_c - \tau\right) + n(t) \quad (3)$$

where

$$w_r(t) = \sum_{k=0}^{N_{\mathrm{tap}}-1} h_k \psi_k(t - t_k) \quad (4)$$

is the received waveform corresponding to a single pulse. Here $\psi_k(t) = f_k(t) * \psi(t)$ is the received UWB pulse from the kth path. The duration of the received pulse T_w is assumed to be less than the chip duration T_c. The propagation delay is denoted by τ, and $n(t)$ is a zero mean noise process. Given the received signal, the acquisition system attempts to retrieve the timing offset τ.

B. Current Approaches Toward UWB Signal Acquisition

Acquisition schemes for UWB systems in the literature can be broadly classified into those which follow detection-based

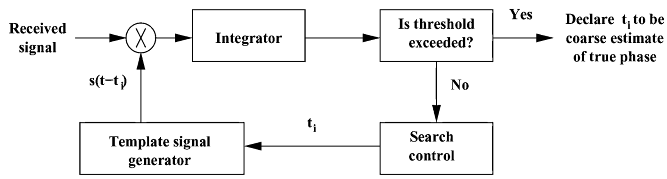

Fig. 3. Block diagram of the acquisition scheme proposed in [25].

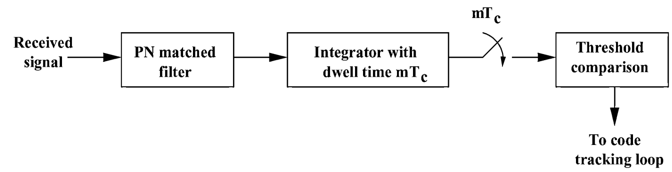

Fig. 4. Block diagram of the acquisition scheme proposed in [26].

approaches, and those which rely on estimation-theoretic strategies. The acquisition methods which employ a detection based approach typically evaluate a candidate phase by first obtaining a measure of correlation between the received signal and a locally generated template signal offset by the candidate phase. This measure of correlation is then compared to a threshold in order to make a decision. These candidate phases could be evaluated in a serial, parallel, or hybrid manner. Among the detection based schemes for UWB acquisition some schemes focus exclusively on the development of efficient search strategies to quickly evaluate the candidate phases in the search space and certain other schemes propose two-stage acquisition methods that achieve a reduction in the search space itself. In the estimation based methods, an estimate of the timing is typically obtained by maximizing a statistic over a set of candidate phases. This statistic is usually obtained from correlation of the received signal with a template signal. These schemes thus do not involve a threshold comparison. Most of the estimation based schemes attempt to exploit the cyclostationarity inherent in UWB signaling due to pulse repetition.

1) Detection Based Approaches: Some of the acquisition schemes proposed for UWB signal acquisition involve the straightforward application of traditional spread spectrum acquisition techniques.

In [25], the traditional coarse acquisition scheme where in the search space is searched in increments of a chip fraction is analyzed for the acquisition of TH UWB signals in AWGN noise. Fig. 3 shows a block diagram of the scheme where a particular phase t_i in the search space is checked by correlating the received signal with a locally generated template signal with delay t_i. If the integrator output exceeds the threshold, the phase t_i is declared to be a coarse estimate of the true phase of the received signal. If the threshold is not exceeded, the search control updates the phase to be checked as $t_{i+1} = t_i + \epsilon T_p$, where $\epsilon < 1$ and T_p is the pulse width. This process continues until the threshold is exceeded.

In [26], the output of a matched filter, whose impulse response is a time reversed replica of the spreading code, is integrated over successive time intervals of size mT_c, where $1 < m \leq N_{\rm tap}$ and T_c is the chip duration in an attempt to combine the energy in the multipath. The integrator output is then sampled at multiples of mT_c and compared to a threshold as illustrated in Fig. 4. The performance of this scheme is evaluated in static multipath channels with 2 and 4 paths and is shown to improve mean acquisition time performance.

In [27], the nonconsecutive search proposed in [21] and a simpler version of the MAX/TC scheme [14] called the global MAX/TC are applied to the acquisition of UWB signals in the presence of multipath fading and MAI. In the nonconsecutive search, only one phase in every D consecutive search space phases is tested by correlating the received signal with a template signal with that particular phase. The decimation factor D is chosen to be not larger than the delay spread $N_{\rm tap}$. In the global MAX/TC, a parallel bank of correlators is used to evaluate all the nonconsecutive phases, and the phase corresponding to the correlator output with maximum energy is chosen as the coarse estimate of the true phase.

In [28], a hybrid acquisition scheme called the reduced complexity sequential probability ratio test (RC-SPRT) is presented for UWB signals in AWGN, which is a modification of the multihypothesis sequential probability ratio test (MSPRT) for the hybrid acquisition of spread spectrum signals [29]. In the MSPRT, if the sequential test in one of the parallel correlators identifies the phase being tested as a potential true phase, the control is passed to the verification stage which verifies its decision. In the RC-SPRT, the sequential test in each of the parallel correlators is used only to reject the hypotheses being tested as soon as they become unlikely and replaces them with new hypotheses. The RC-SPRT stops when all the phases except one have been rejected. This scheme has merit at low SNRs where the time required to reject incorrect phases may be much smaller than the time required to identify the true phase.

In [30], [31], the effect of equal gain combining (EGC) on the acquisition of UWB signals with TH spreading is investigated in a multipath environment. The acquisition problem is formulated as a binary composite hypothesis testing problem where the set of phases where a receiver lock results in a nominal uncoded bit error probability constitute the alternate hypothesis. Two schemes based on EGC called the square-and-integrate (SAI) and the integrate-and-square (IAS) are analyzed and compared in [31]. The IAS scheme is similar to the one shown in Fig. 3 with the exception that the template signal is given by

$$s(t) = \sum_{l=0}^{N_{\rm th}-1} v(t - lT_f - c_l T_c) \quad (5)$$

where $v(t) = \sum_{k=0}^{G-1} \psi_r(t - kT_c)$, G is the length of the EGC window and $\psi_r(t)$ is the receiver's estimate of the received pulse shape. Thus in IAS, EGC is done first and then the correlator output is squared to generate the decision statistic. In SAI, the received signal is first squared to eliminate the pulse inversion and then EGC is performed to utilize the energy in the multipath. In this case, the template signal is once again given by (5) with $v(t) = \sum_{k=0}^{G-1} \psi_r^2(t - kT_c)$. It is shown that even though EGC improves the acquisition performance in SAI at low SNRs, the performance of IAS with no EGC is superior to SAI at all SNRs, although the SAI method attempts to collect signal energy from the squared MPCs through EGC.

2) Efficient Search Strategies: A search strategy specifies the order in which the candidate phases in the timing uncertainty region are evaluated by the acquisition system. When there are more than one acquisition phases in the uncertainty region, the serial search which linearly searches the uncertainty region is no longer the optimal search strategy. More efficient nonconsecutive search strategies called the "look-and-jump-by-K-bins" search and bit reversal search are analyzed in the noiseless scenario with mean stopping time as the performance metric in [32]. Suppose that the timing uncertainty region is divided in to bins indexed by $0, 1, \ldots, N_s - 1$. In look-and-jump-by-K-bins search, starting in bin 0, the search continues on to bin K, then to $2K$ and so on. So for $N_s = 9$ and $K = 3$, the look-and-jump-by-K-bins search searches the bins in the following order $\{0, 3, 6, 1, 4, 7, 2, 5, 8\}$. In bit reversal search, the order in which the bins are searched is obtained by reversing the bits in the binary representation of the linear search variable. For instance, when $N_s = 9$, the linear search has the binary representation $\{000, 001, 010, 011, \ldots, 111\}$ and the bit reversal search is obtained by 'bit reversal' by $\{000, 100, 010, 110, \ldots, 111\}$. It then corresponds to the search order $\{0, 4, 2, 6, 1, 5, 3, 7\}$. A generalized flow graph method is presented in [33], [34] to compute the mean acquisition time for different serial and hybrid search strategies. For the case when the acquisition phases are K consecutive phases in the uncertainty region, it has been claimed that the look-and-jump-by-K-bins search is the optimal serial search permutation when K is known, and the bit reversal is the optimal search permutation when K is unknown. Under the assumption that the probability of detection in all the K consecutive acquisition phases is the same and with mean detection time as the performance metric, the optimum permutation search strategy has been found in [35] using techniques in majorization theory. The ith position in the optimal permutation is given by

$$R_i = (i-1)K \pmod{N_s} + \left\lfloor \frac{i-1}{\left(\frac{N_s}{d}\right)} \right\rfloor + 1 \qquad (6)$$

where $i \in \{1, 2, \ldots, N_s\}$ and d is the greatest common divisor (GCD) of N_s and K.

3) Search Space Reduction Techniques: Some acquisition schemes attempt to solve the large search space problem by employing a two-stage acquisition strategy [36]–[40]. The basic principle behind all these schemes is that the first stage performs a coarse search and identifies the true phase of the received signal to be in a smaller subset of the search space. The second stage then proceeds to search in this smaller subset and identifies the true phase. In [36], such a two stage scheme is proposed for the acquisition of time-hopped UWB signals in AWGN noise and multiple acess intereference (MAI). The search space is divided in to Q mutually exclusive groups of M consecutive phases each. In the first stage, each one of the Q groups is checked by correlating the received signal with a sum of M delayed versions of the locally generated replica of the received signal. Once a group is identified as containing the true phase, the phases in the group are searched by correlating with just one replica of the received signal. This is illustrated in Fig. 5 in the absence of noise and MAI. A scheme based on the same principle has been developed independently in [37]. Both of these schemes have been developed under the assumption of an AWGN channel and their performance is likely to suffer in the presence of multipath.

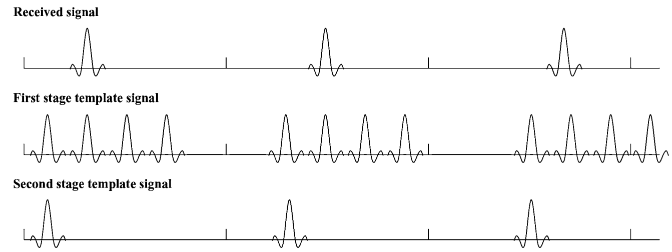

Fig. 5. The template signals used in the two-stage acquisition scheme proposed in [36].

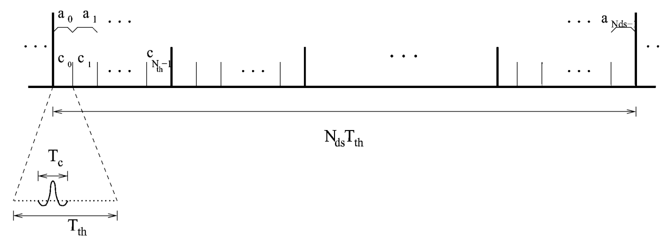

Fig. 6. The hybrid DS-TH signal format used in [39].

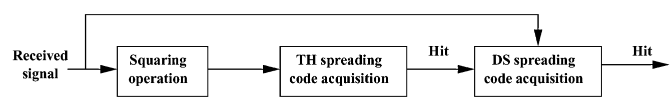

Fig. 7. Conceptual block diagram of the two-stage acquisition scheme proposed in [39].

In [38], an acquisition scheme for UWB signals with TH spreading called n-scaled search is presented, where the search space is divided into groups of $M = N_f/2^n$ where $n \geq 1$. The TH sequence used to generate the replica of the received signal is also modified by neglecting the n least significant bits of each additional shift c_l. Although the actual scheme involves chip rate sampling of a matched filter output, it is equivalent to correlating the received signal with M delayed versions of the modified replica of the received signal. In this sense, it is similar in spirit to the schemes described above.

A two-stage scheme which achieves search space reduction by employing a hybrid DS-TH spreading signal format (shown in Fig. 6) is described in [41], [39]. In the first stage, the DS spreading is removed by squaring the received signal and the timing of the TH spreading code, which has a relatively small length, is acquired. Once this is done, the acquisition of the DS spreading code is performed by searching the search space in increments equal to the length of the TH code. Fig. 7 shows a conceptual block diagram of this system.

Another two stage acquisition scheme for UWB signals with DS spreading which employs a special signal format is presented in [40]. The signal transmitted during the acquisition process is a sum of two signals, a periodic pulse train and a pulse train with DS spreading, as shown in Fig. 8. In the first stage, the timing of the periodic pulse train is acquired by correlating the received

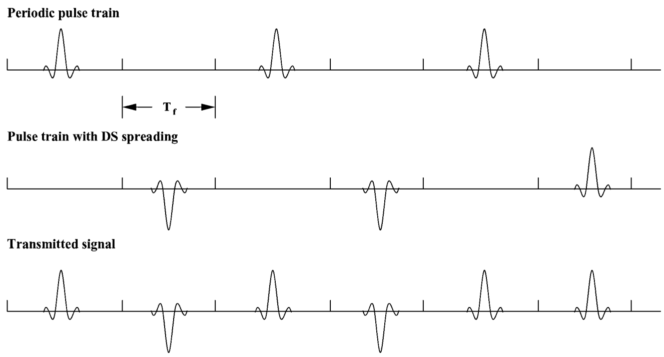

Fig. 8. The transmitted signal along with its component signals used in [40].

signal with a replica of the periodic pulse train. This is an easy task considering that the uncertainty region is just twice the pulse repetition time T_f. Once this is done, the chip boundaries of the DS spreading sequence are known and the second stage needs to only search in increments of $2T_f$ to acquire the timing of the DS spreading sequence.

4) Estimation-Based Schemes: Certain approaches toward acquisition in UWB systems have employed estimation-theoretic methods to obtain timing information of the received signal. The nondata aided timing estimation approaches in [42], [43] exploit cyclostationarity, inherent in UWB signaling due to pulse repetition, to estimate timing information of the received signal. These schemes require frame rate sampling in the acquisition stage and pulse rate sampling during the tracking stage. The signal model assumes only TH spreading and no polarity randomization of the pulses; i.e., $a_l = 1$. It is also assumed that the received pulses from all paths $\psi_k(t) = \psi(t)$, for $k = 0, 1, \ldots, N_{\rm tap} - 1$, and the period of the TH sequence is equal to a symbol duration; i.e., $N_b = N_{\rm th}$. The timing offset is assumed to be confined to a symbol duration and is expressed as $\tau = N_\epsilon T_f + \epsilon$, where $N_\epsilon \in [0, N_{\rm th} - 1]$ and $\epsilon \in [0, T_f)$ represents the pulse level offset. The acquisition system estimates the frame level timing offset by estimating N_ϵ. To do this, a sliding correlator correlates the received signal with the template $\psi(t)$ and frame rate samples $z(n) = \int_{nT_f}^{(n+1)T_f} \psi(t - nT_f) r(t)$ are obtained. Under certain conditions, it is observed that the autocorrelation $R_z(n;\nu) = E\{z(n)z(n+\nu)\}$ of $z(n)$ is periodic in n with period $N_{\rm th}$ and hence $z(n)$ is a cyclostationary process. Estimates $\hat{R}_z(n;\nu)$ of $R_z(n;\nu)$ are obtained by sample averaging and the frame-level timing estimate is obtained by picking the peak of the periodically time varying correlation of the sampled correlator output [43] and is given by

$$\hat{N}_\epsilon = \text{round}\{[\text{argmax}_{\nu} \hat{R}_z(n,\nu) + n]_{N_{\rm th}}\} \tag{7}$$

where $\text{round}\{\cdot\}$ denotes the rounding operation. A slightly more robust approach in [42], [43] estimates the Fourier coefficients $\hat{\mathcal{R}}_z(n,\nu)$ of the periodic sequence $R_z(n;\nu)$ via sample averaging which are then used to estimate the frame level timing as

$$\hat{N}_\epsilon = \text{round}\left\{\left[\frac{1}{2}\left(\nu - \hat{\theta}(n;\nu)\frac{N_{\rm th}}{n\pi}\right)\right]_{N_{\rm th}}\right\} \tag{8}$$

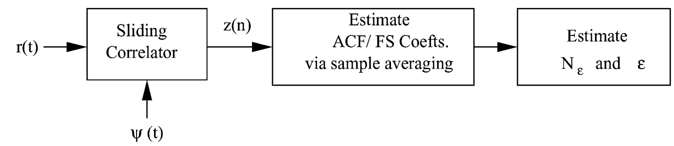

Fig. 9. Autocorrelation function (ACF) of correlator outputs $z[n]$ or its Fourier series (FS) coefficients estimated via sample averaging and used to estimate timing offset.

where $\theta(n;\nu) = \angle\hat{\mathcal{R}}_z(n,\nu)$. The estimation of the pulse level timing offset ϵ, is done using a similar method but requires pulse rate sampling of the correlator output. These schemes are conceptually illustrated in Fig. 9.

In [44], a maximum likelihood (ML) timing estimation scheme is presented for data aided and nondata aided methods and a tradeoff between acquisition accuracy and complexity is discussed. A data aided timing estimation scheme employing a Rake-like structure with L_c fixed tap delays is analyzed in [44], assuming the timing offset to be less than a symbol duration and identical received pulses from all paths $\psi_k(t) = \psi(t)$, for $k = 0, 1, \ldots, N_{\rm tap} - 1$. The signal component of the received signal is modeled as

$$r_s(t) = \sum_{l=0}^{L_c-1} \gamma_l x(t - N_\epsilon T_f - lT_c) \tag{9}$$

where γ_l are the gains corresponding to each bin. The frame-level timing offset N_ϵ is then estimated from the observation of M symbols through a grid search and is given by

$$\hat{N}_\epsilon = \arg\max_{N_\epsilon \in [0,\ldots,N_f-1]} \sum_{l=0}^{L_c-1} J^2(N_\epsilon; lT_c) \tag{10}$$

where $J(N_\epsilon; lT_c) = \sum_{k=0}^{M-1}\int_{-\infty}^{\infty} r(t) b_k \psi(t - kN_bT_f - N_\epsilon T_f - lT_c)dt$ denotes the sum of M pulse rate samples from the lth correlator.

Another data aided timing estimation scheme is developed in [45] where the timing estimation problem is translated to an ML amplitude estimation problem and a generalized likelihood ratio test to detect the presence or absence of a UWB signal is developed, which makes use of the ML timing estimates. The symbol level timing offset is obtained through a line search and a closed form solution is derived for the frame level timing offset, utilizing symbol rate samples. In [46], a method is presented for optimizing allocation of pulses in training and information symbols used for acquisition, channel estimation, and symbol detection.

Least squares estimates of the timing and the channel impulse response, using Nyquist rate samples of the received signal, are obtained in [47] under the restrictive assumption that $\tau < T_f$; therefore, this method is not practical for timing acquisition. A nondata aided timing estimation method called timing with dirty templates (TDT) is presented in [48] which, in the absence of intersymbol interference (ISI), makes use of cross correlations between adjacent symbols to estimate timing information of the received signal. In this scheme, a symbol length segment of the received waveform is used as a template and correlated with the subsequent symbol length segment, and the symbol rate correlator output samples are summed over K pairs of symbols to estimate the timing information τ, assumed to be within a symbol duration, as

$$\hat{\tau} = \arg \max_{\tau \in [0, N_b T_f)} \sum_{k=1}^{K} \left(\int_{(2k-1)N_b T_f}^{2k N_b T_f} r(t) r(t - N_b T_f) dt \right)^2 . \tag{11}$$

A training sequence design method for a similar data aided scheme is presented in [49].

Transform-domain methods, which obtain estimates of channel parameters employing sub-Nyquist sampling rates, are presented in [50]–[52] where the joint channel and timing estimation problem is translated into a harmonic retrieval problem. These methods obtain samples $F_r[n]$ of the Fourier transform, $F_r(\omega)$ of the received signal and use them to estimate the excess delays t_k employing standard spectral estimation techniques. However, these schemes can estimate t_ks only after the timing offset τ is known, and hence cannot be used for timing acquisition.

In [53], the Cramer-Rao lower bounds (CRLBs) for the time delay estimation problem are derived for UWB signals in AWGN and multipath channels. It is shown that a larger number of multipath results in higher CRLBs and a potentially inferior performance for unbiased estimators.

5) Miscellaneous Approaches: An acquisition strategy for impulse radio which makes use of relative timing between pulses in specially chosen TH sequences is presented in [54] in the absence of multipath. This scheme may not be applicable in the presence of multipath, which is usually the case with UWB systems. An acquisition scheme implemented on UWB-based positioning devices which use a coded beacon sequence in conjunction with a bank of correlators is presented in [55] and assumes absence of multipath. A distributed synchronization algorithm for a network of UWB nodes, motivated by results from synchronization of pulse coupled oscillators in biological systems such as synchronized flashing among a swarm of fireflies and synchronous spiking of neurons, is presented in [56].

IV. Issues and Challenges in the Design of UWB Acquisition Systems

In this section, we discuss some of the issues and challenges in UWB signal acquisition which may not have received sufficient attention in the existing literature.

A. Hit Set

In a multipath channel, the energy corresponding to the true signal phase is spread over several multipath components (MPCs). The primary difference between the acquisition problems in a multipath channel and a channel without multipath is that there is more than one hypothesized phase which can be considered a good estimate of the true signal phase. In a multipath environment, the receiver may lock onto a non-line-of-sight (non-LOS) path and still be able to perform adequately as long as it is able to collect enough energy. From the viewpoint of post-acquisition receiver performance, a receiver lock to any one of such paths can be considered successful acquisition. Thus we require a precise definition of what can be considered a *good* estimate of the true signal phase.

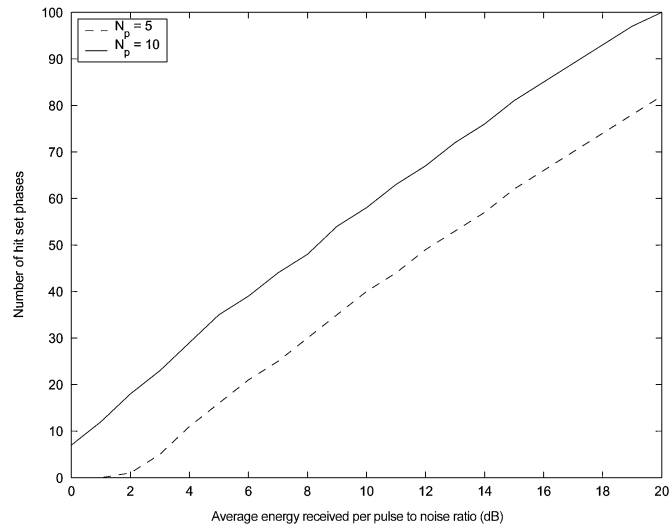

Fig. 10. The hit set size as a function of the average energy received per pulse to noise ratio for $N_p = 5$ and 10.

A typical paradigm for transceiver design is the achievement of a certain nominal uncoded bit error rate (BER) λ_n. Then, all those hypothesized phases such that a receiver locked to them achieves an uncoded BER of λ_n can be considered a good estimate of the true signal phase. We define the *hit set* to be the set of such hypothesized phases. For a given true phase τ, let $P_E(\Delta\tau)$ denote the BER performance of the receiver when it locks to the hypothesized phase $\hat{\tau}$ where $\Delta\tau = \hat{\tau} - \tau$. Let Υ_m be the minimum SNR at which the receiver achieves a BER of λ_n when it locks to the LOS path, that is, $P_E(0) \leq \lambda_n$ when the SNR is Υ_n and $P_E(0) > \lambda_n$ for all SNRs less than Υ_n. Then for an SNR $\Upsilon \geq \Upsilon_n$ and true phase τ, the hit set is given by

$$\mathcal{H} = \{\hat{\tau} : P_E(\Delta\tau) \leq \lambda_n\}. \tag{12}$$

The hit set when a partial Rake (PRake) receiver [57] is employed for demodulation has been derived in [39], [31]. Fig. 10 shows a plot of the number of phases in the hit set as a function of the SNR when $\lambda_n = 10^{-3}$ and the PRake receiver has $N_p = 5$ and 10 fingers. It is observed that the cardinality of the hit set could be significantly large depending upon the operating SNR.

A design for an acquisition system which does not take the hit set into account can result in a significant performance degradation. For instance, in serial acquisition schemes, such as the one shown in Fig. 3, the decision threshold is usually set such that the average probability of false alarm is constrained by a small positive constant $\delta \ll 1$, i.e.,

$$\gamma_d = \underset{\gamma}{\operatorname{argmin}} \max_{\hat{\tau} \notin \mathcal{H}} E_h[P_{\mathrm{FA}}(\gamma, \Delta\tau)] \leq \delta. \tag{13}$$

Fig. 11 shows two receiver operating characteristics (ROCs) for an acquisition scheme where the received signal is correlated

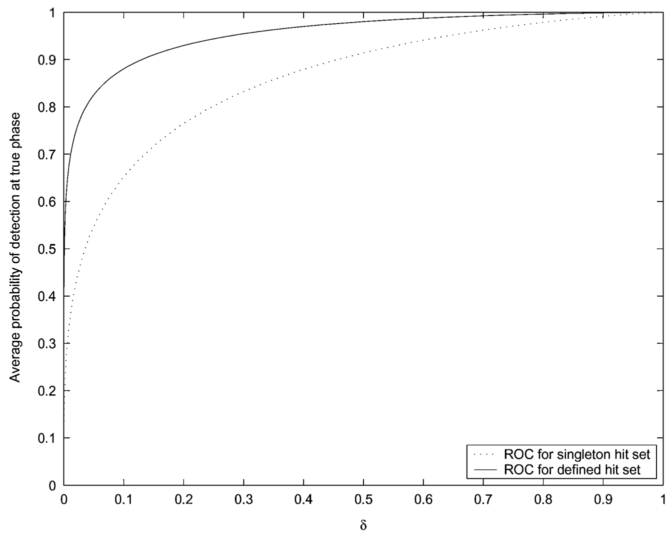

Fig. 11. The ROCs when the threshold is set for a singleton hit set containing only the true phase and for a hit set defined in (12) with $\lambda_n = 10^{-3}$.

with a template signal and the correlator output is squared and compared to a threshold. The detailed derivation of the performance analysis can be found in [31]. For one of the ROCs, the threshold was set assuming that the hit set consists of only the true phase τ, and for the other the hit set definition in (12) was used assuming a PRake receiver with $N_p = 5$ fingers with the nominal BER requirement $\lambda_n = 10^{-3}$ and the average energy received per pulse to noise ratio equal to 5 dB. When the hit set contains only the true phase τ, the threshold needs to be set much higher in order to prevent the decision statistics for the other phases in the multipath profile, which have significant energy, from exceeding it. This causes the degradation in the probability of detection when $\hat{\tau} = \tau$.

B. Asymptotic Acquisition Performance of Threshold Based Schemes

A typical threshold based timing acquisition system consists of a verification stage in which a threshold crossing at a candidate phase is checked to see if it was a false alarm or a true detection event. The usual procedure for implementing the verification stage is to have a large dwell time for the correlator [7]. The large dwell time increases the effective SNR of the decision statistic and in the absence of channel fading, this results in accurate verification; i.e., the probabilities of a false alarm and a miss can be made arbitrarily small. However, for threshold based acquisition schemes in multipath fading channels, it was shown [58] that no matter how large the SNR is or how we choose the threshold, it may not be possible to make the probabilities of detection and false alarm arbitrarily small. In particular, the asymptotic performance of two typical threshold based acquisition schemes for TH UWB signals was calculated in [59]. It was shown that if the threshold is such that the average probability of false alarm is less than a given tolerance, then there is a nontrivial lower bound on the asymptotic average probability of miss. This lower bound translates to an upper bound on the asymptotic average probability of detection. These results suggest that it may not be possible to build a good verification stage for UWB signal acquisition systems by just increasing the dwell time. They also suggest that the principles underlying the design of efficient UWB signal acquisition schemes may be very different from the traditional spread spectrum acquisition schemes.

In traditional spread spectrum acquisition systems, the decision threshold is chosen such that the probability of false alarm in each of the nonhit set phases is small. The verification stage helps the acquisition system recover from false alarm events when they occur. Considering that the construction of a verification stage in some UWB signal acquisition systems may be difficult, a more appropriate choice of decision threshold is one which restricts the probability that the acquisition process encounters a false alarm to be small. So if $P_F(\gamma)$ is the average probability that the acquisition process ends in a false alarm, then the decision threshold γ_d is chosen such that $P_F(\gamma)$ is constrained by a small positive constant $\delta \ll 1$,

$$\gamma_d = \operatorname*{argmin}_{\gamma} P_F(\gamma) \leq \delta. \tag{14}$$

The performance of spread spectrum acquisition systems has typically been characterized by the calculation of mean acquisition time [7], [19]. In mean acquisition time calculations, a false alarm penalty time is assumed which is the dwell time of the verification stage; i.e., the time required by the acquisition system to recover from a false alarm event. Thus mean acquisition time calculations implicitly assume the existence of a verification stage. For UWB signal acquisition systems, if the threshold is set according to (14), the mean detection time is a reasonable metric for system performance. The mean detection time is defined as the average amount of time taken by the acquisition system to end in a detection, conditioned on the nonoccurrence of a false alarm event. The calculation of the mean detection time thus does not require any assumption on the verification stage.

Finally, several detection based schemes for UWB signal acquisition have proposed using some form of EGC to improve the acquisition performance by combining the energy in the multipath [36]–[38]. The asymptotic performance of threshold based UWB signal acquisition schemes using EGC has been calculated in [59]. It has been shown that EGC may lead to a significant performance degradation. Fig. 12 shows the asymptotic receiver operating characteristic (AROC) for different values of the EGC window size G when the threshold is set according to (13). The AROC is very good for $G = 1$, and degrades significantly as G increases. This is because as G increases, when the candidate phase is the true phase, the EGC window collects multiple paths which may have opposing polarities resulting in cancellations, and hence a decrease in the probability of detection.

C. The Search Space in UWB Signal Acquisition

The large search space in UWB signal acquisition poses significant challenges in the design and implementation of practical systems. Although detection based schemes which evaluate the phases in the search space one at a time have a simple hardware implementation, they may suffer from a large mean detection

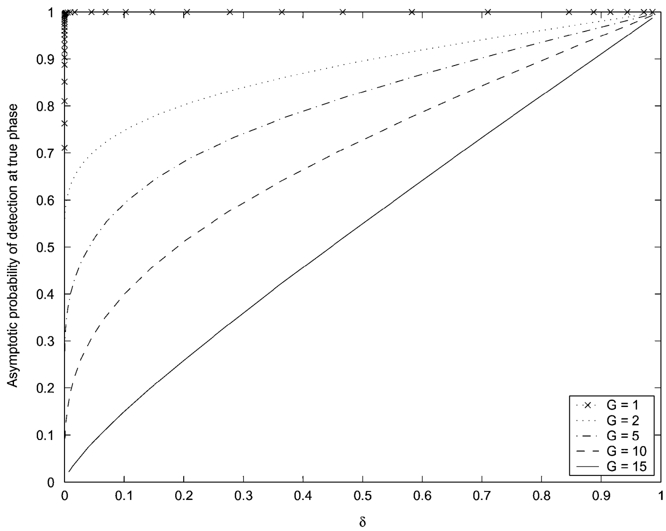

Fig. 12. The AROC for a detection based acquisition system using EGC for different values of EGC window size G.

time which makes them unsuitable for high data rate applications. Furthermore, it was shown in [31] that the time spent by the acquisition system in evaluating and rejecting the nonhit set phases was the dominant part of the mean detection time causing it to decrease only marginally with increase in SNR. Thus acquisition techniques capable of reducing the search space are crucial in the design of efficient acquisition schemes. For example, the two stage scheme described in [39] achieves a significant reduction in the mean detection time.

Another approach to solve the search space problem is by designing the higher layers in the network architecture carefully. A multiple access protocol which employs continuous physical layer links in the network in order to avoid repeated acquisition is presented in [60]. The timing uncertainty region may be reduced significantly if a beacon-enabled network is employed, where the medium access is coordinated by a central node which periodically transmits beacons to which other nodes synchronize and follow a slotted medium access approach.

D. Generalized Likelihood Ratio Test for UWB Signal Acquisition

There has not been much effort in the direction of finding the optimal detectors for the timing acquisition problem in UWB systems. Most detection based schemes for UWB signal acquisition have been ad hoc schemes based on the principles of traditional spread spectrum acquisition systems. In the context of the hit set and the dense multipath in UWB systems, a reasonably systematic approach to detector design is the generalized likelihood ratio test (GLRT). It is instructive to examine the structure of the GLRT detector used by a serial acquisition system which tries to find the true phase by evaluating the phases in the search space one at a time. Although the GLRT is not an optimal test, it has been known to work quite well in general [61]. The GLRT has been shown to be asymptotically uniformly most powerful among the class of invariant tests [62].

The received signal is observed over a duration of M periods of the DS sequence, which is assumed without loss of generality to be longer than the TH sequence, and this observation is denoted by $\bar{r}$. The acquisition system is to determine whether a hypothesized phase $\hat{\tau}$ can be considered the true phase of the received signal. It is assumed that the hypothesized phase is a multiple of the chip duration T_c. To enable tractable analysis, it assumed that the true phase is also a multiple of T_c. The number of phases in the search space is thus $N_{\mathrm{ds}}N_f$. From the previously mentioned definition of the hit set, it is clear that there exist many ways in which $\hat{\tau}$ can be considered to be the true phase. Without loss of generality, suppose that the hit set is $\{\tau - \Delta_b T_c, \tau - (\Delta_b - 1)T_c, \ldots, \tau + \Delta_f T_c\}$ where Δ_b and Δ_f are integers between 0 and $N_{\mathrm{ds}}N_f/2$. Also suppose that an all-ones data training sequence is sent in the acquisition preamble. This results in a composite hypothesis testing problem whose hypotheses can be formulated as follows:

$H_0 : \hat{\tau}$ is not an acceptable phase, i.e., $\tau \notin \mathcal{S}(\hat{\tau})$

$H_1 : \hat{\tau}$ is an acceptable phase, i.e., $\tau \in \mathcal{S}(\hat{\tau})$,

where $\mathcal{S}(\hat{\tau}) = \{\hat{\tau} - \Delta_f T_c, \hat{\tau} - (\Delta_f - 1)T_c, \ldots, \hat{\tau} + \Delta_b T_c\}$. The GLRT is given by

$$\Lambda(\bar{r}) = \frac{\max_{\{\mathbf{h},\nu \in \mathcal{S}(\hat{\tau})\}} p(\bar{r} \mid \mathbf{h}, \nu)}{\max_{\{\mathbf{h},\nu \notin \mathcal{S}(\hat{\tau})\}} p(\bar{r} \mid \mathbf{h}, \nu)} \underset{H_0}{\overset{H_1}{\gtrless}} \gamma \tag{15}$$

where the column vector $\mathbf{h}$, of channel gains $\{h_k\}$, is assumed to be deterministic but unknown and γ is the decision threshold. It can easily be shown using techniques similar to those used in [63], [64] that when $n(t)$ is an AWGN process with power spectral density $N_0/2$, the choices of ν and $\mathbf{h}$ which maximize $p(\bar{r} \mid \mathbf{h}, \nu)$ in the numerator in (15) are given by

$$\begin{aligned} \tau_1 &= \arg\, \max_{\nu \in \mathcal{S}(\hat{\tau})} \mathbf{C}^{\mathbf{T}}(\nu)\mathbf{C}(\nu) \\ \mathbf{h_1} &= \frac{1}{MN_{\mathrm{ds}}R_{\psi_r}}\mathbf{C}(\tau_1) \end{aligned} \tag{16}$$

and similarly for the denominator in (15)

$$\begin{aligned} \tau_0 &= \arg\, \max_{\nu \notin \mathcal{S}(\hat{\tau})} \mathbf{C}^{\mathbf{T}}(\nu)\mathbf{C}(\nu) \\ \mathbf{h_0} &= \frac{1}{MN_{\mathrm{ds}}R_{\psi_r}}\mathbf{C}(\tau_0) \end{aligned} \tag{17}$$

where $R_{\psi_r} = \int_0^{T_w} \psi_r^2(t)$ and $\mathbf{C}(\nu) = [C_0(\nu), C_1(\nu), \ldots, C_{N_{\mathrm{tap}-1}}(\nu)]^T$ with

$$C_k(\nu) = \int_0^{MN_{\mathrm{ds}}T_f} r(t)s_k(t-\nu) \tag{18}$$

where

$$s_k(t-\nu) = \sum_{l=0}^{MN_{\mathrm{ds}}-1} a_{[l/N_{\mathrm{ds}}]}\psi_r(t - kT_c - lT_f - c_lT_c - \nu)dt. \tag{19}$$

Also, it can be easily shown that the test in (15) can be written as

$$\Lambda(\bar{r}) = \left[\mathbf{h_1^T C}(\tau_1) - \mathbf{h_0^T C}(\tau_0)\right] - \frac{MN_{\rm ds}R_{\psi_r}}{2}\left[\mathbf{h_1^T h_1} - \mathbf{h_0^T h_0}\right] \underset{H_0}{\overset{H_1}{\gtrless}} \frac{N_0}{2}\gamma \tag{20}$$

Using (16) and (17), the GLRT in (20) reduces to

$$\Lambda(\bar{r}) = \max_{\nu\in\mathcal{S}(\hat{\tau})}\sum_{k=0}^{N_{\rm tap}-1} C_k^2(\nu) - \max_{\nu\notin\mathcal{S}(\hat{\tau})}\sum_{k=0}^{N_{\rm tap}-1} C_k^2(\nu) \underset{H_0}{\overset{H_1}{\gtrless}} \frac{N_0 MN_{\rm ds}R_{\psi_r}}{2}\gamma \triangleq \gamma'. \tag{21}$$

The threshold γ' can be set such that the probability of false alarm $P_{\rm FA} = \Pr\{\Lambda(\bar{r}) > \gamma' \mid H_0\} < \delta$, where δ is a specified false alarm tolerance. It can be observed from (21) that the test statistic given by the GLRT amounts to correlating the received signal with $N_{\rm tap}$ different templates, each corresponding to a different MPC, summing the squared outputs of each of these correlators, maximizing this sum for two disjoint sets of phases, and comparing the difference to a threshold as illustrated in Fig. 13. This test statistic thus attempts to collect the energy from all the MPCs through a form of equal gain combining. However, it is immediately clear that such an implementation is prohibitively complex to realize. Thus other suboptimal strategies need to be explored which would collect energy from the MPCs in an alternative way.

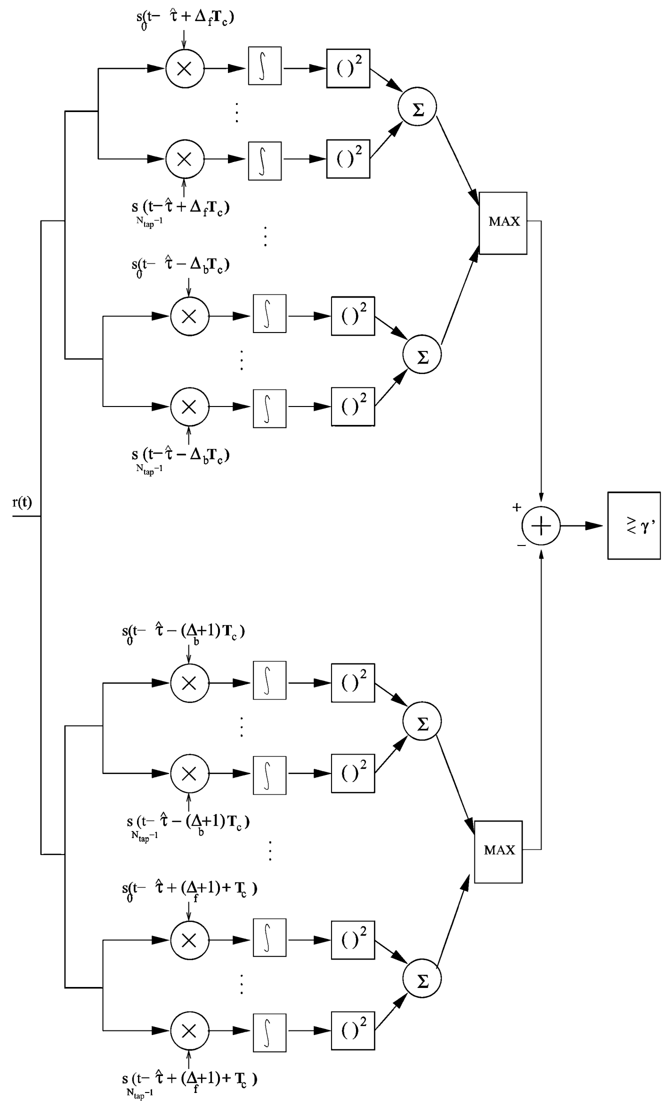

Fig. 13. Generalized likelihood ratio test for evaluation of phase $\hat{\tau}$. The upper and lower MAX operations evaluate the maximum of $\sum_{k=0}^{N_{\rm tap}-1} C_k^2(\nu)$ over $\nu \in \mathcal{S}(\hat{\tau})$ and $\nu \notin \mathcal{S}(\hat{\tau})$, respectively.

E. Transmitted Reference UWB Systems

There has been renewed interest in the so-called transmitted reference (TR) systems [65] for impulse radio [66]–[68] due to their simple receiver structure. In typical TR signaling for UWB systems, each modulated pulse is preceded by an unmodulated (reference) pulse in the transmitted signal. At the receiver, the noisy reference pulse is used to demodulate the signal using a simple *delay-and-correlate* receiver [67]–[69]. However, the BER performance of a TR receiver is usually worse compared to a Rake receiver due to energy wasted on the reference pulse and due to the noise-noise cross terms resulting in the delay-and-correlate receiver [70]. However, the TR-UWB receiver does not require a channel estimator and need not estimate the shapes of the received pulses. The acquisition problem in TR-UWB systems is also less acute due to a relaxed timing requirement by the TR demodulator. This relaxation in the timing requirement can be illustrated by an example. Consider a TR-UWB system employing antipodal data modulation and DS signaling, with the transmitted signal given by

$$x(t) = \sum_{l=-\infty}^{\infty} d_{[l/M_{\rm ds}]}\left[\psi(t-lT_f) + b_{\lfloor l/N_b\rfloor}a_{[l/N_{\rm ds}]}\psi(t-lT_f-T_d)\right] \tag{22}$$

where T_d is the delay between the reference and data pulses and $b_l \in \{-1,+1\}$. The *outer* DS sequence $\{d_l\}$ and the *inner* DS sequence $\{a_l\}$ take values in $\{-1,1\}$ and have lengths $M_{\rm ds}$ and $N_{\rm ds}$, respectively. The number of frames modulated by a bit is given by N_b. The delay $T_d = N_dT_c$ between the reference and modulated pulses is chosen to be larger than the multipath spread $T_m = N_{\rm tap}T_c$ to avoid interference between the multipath responses of the reference and data modulated pulses. Here we assume that the duration of the received pulse $\psi_k(t) = f_k(t) * \psi(t)$ corresponding to the kth path is less than T_c. The frame duration T_f is chosen to be equal to $2T_d$ to avoid interframe interference. The received signal with a timing offset of τ can then be expressed as the sum of signal and noise components as

$$r(t) = \sum_{l=-\infty}^{\infty} d_{[l/M_{\rm ds}]}\left[w_r(t-lT_f-\tau) + b_{\lfloor l/N_b\rfloor}a_{[l/N_{\rm ds}]}w_r(t-lT_f-T_d-\tau)\right] + n(t) \tag{23}$$

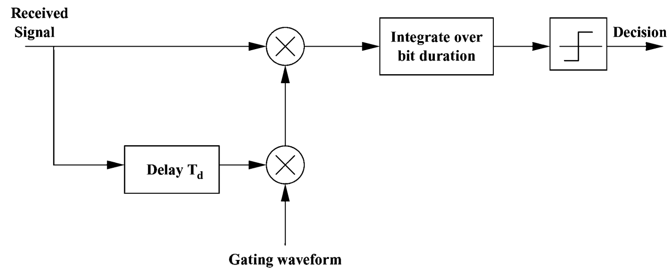

Fig. 14. Block diagram of the autocorrelation receiver.

In order to demodulate the bit b_k, the receiver in Fig. 14 correlates the received signal with a product of the delayed received signal and the return-to-zero (RZ) gating waveform $c_k(t-\hat{\tau})$ given by

$$c_k(t-\hat{\tau}) = \sum_{l=kN_b}^{(k+1)N_b-1} a_{[l/N_{\mathrm{ds}}]} p_m(t - lT_f - T_d - \hat{\tau}) \quad (24)$$

where $\hat{\tau}$ is the receiver's estimate of the timing offset and $p_m(t) = 1$ for $t \in [0, T_m)$ and zero otherwise. The RZ gating waveform has unit amplitude and the same polarity as the inner code over those periods where the product signal does not depend on the outer code. It has zero amplitude over those periods where the product signal depends on the outer code. The product signal is multiplied with the RZ gating waveform and integrated over a bit duration to get the decision statistic for a particular bit. The advantage of using the two level DS signaling format and the gating waveform is made clear in the following discussion.

Fig. 15 illustrates the polarities due to DS signaling and modulation of the received signal, the delayed version of the received signal and the product signal in the absence of noise. Note that the polarity of every alternate received pulse in the product signal depends only on the modulating bit and the inner code $\{a_l\}$, and is independent of the outer code $\{d_l\}$. Thus, if the phase of the inner code is known, the bit can be demodulated without knowledge of the phase of the outer code $\{d_l\}$ by using the RZ gating waveform shown in the figure. The purpose of the RZ gating waveform is two-fold: to despread the inner code $\{a_l\}$, and to restrict the period of integration to those times where the product signal does not depend on the outer code. Thus, the gating waveform accomplishes a function similar to the multiple integration windows employed in [68].

This method of achieving demodulation without knowledge of the outer code $\{d_l\}$ is advantageous because the burden of eliminating the spectral lines can be placed on the outer code by choosing its length M_{ds} to be large. This does not increase the size of the search space, which is proportional to the length of the inner code N_{ds}. The inner code $\{a_l\}$ results in a peaky ambiguity function as in traditional spread spectrum systems and hence is essential for good acquisition performance. The inner code can be chosen to be of relatively smaller length compared to the outer code.

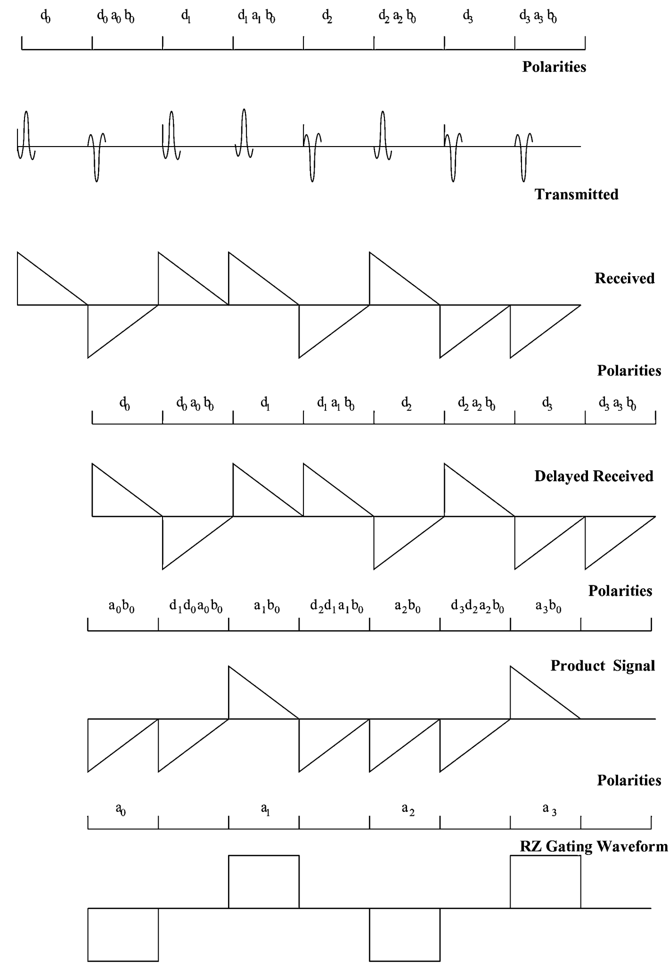

Fig. 15. Illustration of the delay and multiply operation on the noiseless received signal.

Clearly the purpose of the acquisition system is to align (at least approximately) the RZ gating waveform with the useful part of the product signal. Although the size of the search space is now proportional to the length of the relatively short inner code, it can still be large due to the fine timing resolution of the UWB system. So, we propose a two-stage acquisition system which solves the problem of the RZ gating waveform alignment in two steps. In the first stage, the acquisition system attempts to find the phase of the inner code modulating the product signal by correlating the product signal with a locally generated replica of the nonreturn-to-zero (NRZ) inner code DS waveform with chip duration T_f. As illustrated in Fig. 16, the phase of the locally generated NRZ inner code DS waveform can suffer a large margin of error and still be successful in despreading the inner code corresponding to the useful part of the product signal. This suggests that the phases in the search space can be evaluated, not necessarily serially but in increments proportional to the allowed margin of error, which is precisely the search strategy used in the first stage. Once the approximate phase of the inner code is found in the first stage, the second stage of the acquisition system proceeds to align the RZ gating waveform with the useful part of the product signal by searching serially through the

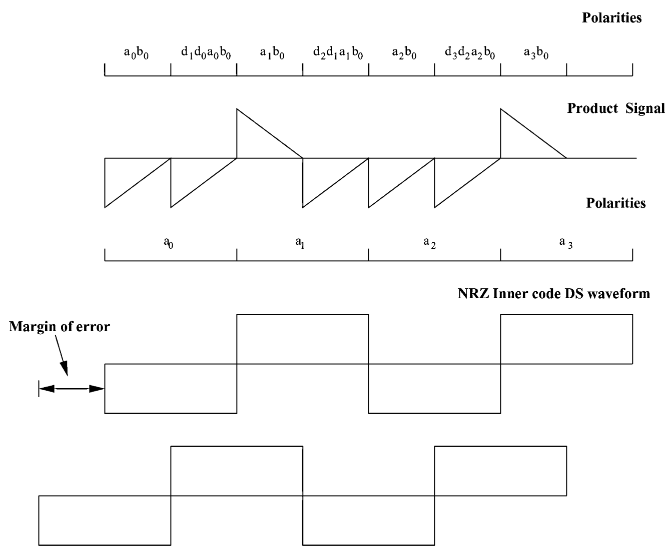

Fig. 16. Illustration of the margin of error tolerable in the despreading of the inner code in the first stage.

phases around the estimated phase. A more precise description of the two-stage acquisition system for TR DS-UWB can be found in [71].

F. Challenges for Estimation Based UWB Timing Acquisition

Most estimation-based timing acquisition schemes for UWB systems are based on the ML principle, and hence involve the simultaneous calculation of the likelihood function corresponding to each one of the phases in the search space followed by a maximum operation. When the search space is large, a fully parallel implementation of this scheme is not feasible, and one may have to resort to a serial or hybrid implementation where the system calculates the likelihood functions for small groups of phases in the search space sequentially. The likelihood functions calculated at each intermediate step need to be stored until all the phases are evaluated. The likelihood functions calculated at each step correspond to different noise realizations, and so a simple maximum operation may not be a good method to find the true phase especially at low SNRs. A more robust approach might be repeated calculation of the likelihood function at each phase followed by averaging to reduce the variations due to noise. This effectively amounts to trading off hardware complexity for an increase in the acquisition time to achieve similar acquisition performance. However, the performance of such reduced complexity estimation-based acquisition schemes in terms of estimation accuracy and acquisition time is still an open research direction.

V. Conclusion

The significance, issues, and challenges of the timing acquisition problem in UWB systems have been presented in this paper. Currently proposed UWB acquisition schemes, classified into detection based and estimation based approaches, are briefly surveyed. We identify and discuss several issues inadequately addressed in existing literature. Some of these considerations, such as the hit set concept, the search space reduction techniques, and the use of TR signaling for acquisition, may lead to more efficient acquisition schemes and hence merit further investigation.

Finally, we note that only impulse radio UWB systems have been considered in this paper. Another form of UWB signaling known as multiband orthogonal frequency division multiplexing (OFDM) is also being considered in WPAN standards [72]. Due to the lack of significant research on timing acquisition for multiband OFDM systems, the acquisition problem for such systems has not been addressed in this paper.

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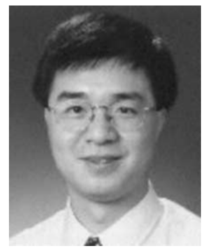

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