On Equal-Gain Combining for Acquisition of Time-Hopping Ultra-Wideband Signals

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***Abstract*—The acquisition of ultra-wideband (UWB) signals is a potential bottleneck for system throughput in a packet-based network employing UWB signaling format in the physical layer. The problem is mainly due to the low received signal power and the fine time resolution which forces the acquisition system to process the signal over long periods of time before getting a reliable estimate of the timing of the signal. Hence, there is a need to develop more efficient acquisition schemes by taking into account the signal and channel characteristics. In this paper, we investigate two approaches, the square-and-integrate and the integrate-and-square, which collect the energy in the mulitpaths by performing equal-gain combining (EGC) to improve the acquisition performance. We define the *hit set* as the set of hypothesized phases which can guarantee adequate system performance after acquisition, and also study the effect of the EGC window length on the acquisition performance.**

***Index Terms*—Acquisition, equal-gain combining (EGC), multipath channels, serial search, ultra-wideband (UWB).**

I. Introduction

ULTRA-wideband (UWB) signaling [1]–[4] is under evaluation as a possible modulation scheme for wireless personal area network (PAN) protocols. The features of UWB radio which make it an attractive choice are its multiple-access capabilities [1], [5], lack of significant multipath fading [6]–[8], ability to support high data rates [9], and low transmitter power, resulting in longer battery life for portable devices. The acquisition of the UWB signal is a potential bottleneck for the system throughput in a packet-based network employing UWB signaling in the physical layer. The problem is mainly due to the following two reasons. First, the received signal power is low, which forces the acquisition system to have a large dwell time in order to improve the signal-to-noise ratio (SNR) of the decision statistic. Second, the large system bandwidth results in a very fine time resolution of the ambiguity region, increasing the number of phases in the search space of the acquisition system. Thus, the acquisition system is forced to process the signal over long periods of time before getting a reliable estimate of the timing (phase) of the signal. Hence, there is a need to develop more efficient acquisition schemes by taking into account the signal and channel characteristics.

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The UWB channel is a dense multipath channel without significant fading [7], [10]. In a dense multipath environment, there will be a considerable amount of energy available in the multipath components (MPCs). It seems reasonable to expect that an acquisition scheme which uses the energy in the MPCs would perform better than one which does not. Acquisition schemes which take into account the multipath nature of the channel have been developed for the case of direct-sequence code-division multiple-access (DS-CDMA) signals in [11]–[14]. Many of the previous schemes proposed for UWB acquisition have not considered the UWB channel characteristics for analysis or simulation. For instance, either an additive white Gaussian noise (AWGN) or a flat-fading channel model is assumed for analysis of the schemes discussed in [15]–[19]. Estimation-theoretic approaches to UWB acquisition which exploit the cyclostationarity present in UWB signaling [20], [21], or perform subspace-based spectral estimation [22] have been proposed. The problem of efficient search strategies for UWB acquisition has been investigated in [23]–[25]. The performance of serial search strategies for acquisition of general spread-spectrum signals in dense multipath channels has been investigated in [26]–[28]. A more detailed discussion of proposed UWB acquisition schemes can be found in [29].

Considering that we have no information regarding the channel state, there are essentially two ways in which we can attempt to use this energy in order to develop a more efficient acquisition scheme. In the first approach, the received signal is first squared to eliminate the channel inversion, and then equal-gain combining (EGC) is performed to exploit the rich path diversity present in UWB channels. In the second approach, EGC is performed first, and the integrator output is then squared to generate the decision statistic. In the following, we will refer to the former as square-and-integrate (SAI) and to the latter as integrate-and-square (IAS). It is not exactly clear which approach is more efficient. Also, the choice of the length of the EGC window is not apparent. For instance, in SAI, a small window will not collect enough energy, and thus will result in a low probability of detecting the correct signal phase. A large window may collect a considerable amount of energy, even when the true phase does not match the hypothesized phase, resulting in a high probability of false alarm. In this paper, we derive and compare the performance of both SAI and IAS as a function of the EGC window length.

Finally, in a multipath environment without significant fading, the ambiguity function[1] does not have an impulse-like shape, as it does in the absence of multipaths. In fact, the ambiguity function decays slowly from its maximum value, since a significant amount of energy is collected even if the hypothesized phase is not equal to the true phase but is sufficiently close to it. The fundamental difference between the acquisition problems in a multipath channel and in a channel without multipath is that there are more than one hypothesized phases which can be considered a *hit* or a good estimate of the true signal phase. Thus, there is a need to redefine the set of hypothesized phases which correspond to a hit by taking into account the channel model and the demodulator structure. In this paper, we propose a definition of the set of hypothesized phases which correspond to a good estimate of the true signal phase by considering the demodulation performance subsequent to acquisition. The presence of a high degree of path diversity in the UWB channel motivates the use of a Rake receiver to improve demodulation performance. The three main Rake receiver structures considered for UWB signal demodulation are the all Rake (ARake), the selective Rake (SRake), and the partial Rake (PRake) receivers [8], [30]. The large number of resolvable multipaths in the UWB channel obviates the use of the ARake receiver due to the complexity involved in its implementation. We assume that the receiver uses a partial Rake (PRake) receiver to perform demodulation. Our choice is guided by the fact that the PRake receiver has lower complexity and still achieves comparable bit-error performance relative to the SRake receiver [30].

The paper is organized as follows. In Section II, we describe the system model under consideration. In Section III, we define the set of hypothesized phases which correspond to a hit by considering system performance subsequent to acquisition. We derive expressions for average probabilities of detection and false alarm for SAI and IAS in Sections IV and V, respectively. In Section VI, we give a design criterion for choosing the decision threshold and derive the mean detection time for a serial search strategy as a function of the average probabilities of detection and false alarm. In Section VII, the mean detection time and the probability of a miss are used as performance metrics to compare the two approaches. Section VIII has some concluding remarks.

II. System Model

A. Channel Model

We assume that the propagation channel is modeled by the UWB indoor channel model described in [31]. This model gives a statistical distribution for the path gains based on a UWB propagation experiment, but does not address the issue of characterization of the received waveform shape. Due to the frequency sensitivity of the UWB channel, the pulse shapes received at different excess delays are path-dependent [32]. To enable tractable analysis, we assume that the pulse shapes associated with all the propagation paths are identical. The channel is then a stochastic tapped-delay line model expressed as the impulse response

$$h(t) = \sum_{k=0}^{N_{\mathrm{tap}}-1} p_k h_k f(t - kT_{\mathrm{c}}) \tag{1}$$

where N_{tap} is the number of taps in the channel response, $T_{\mathrm{c}} = 2$ ns is the tap spacing, h_k is the path gain at excess delay kT_{c}, p_k is equally likely to be ± 1 to account for signal inversion due to reflections [33], and $f(t)$ models the combined effect of the transmitting antenna and the propagation channel on the transmitted pulse. The path gains are independent but not identically distributed with Nakagami-m distributions. The average energy gains $\Omega_k = E[h_k^2]$ of the path gains normalized to the total energy received at 1-m distance are given by

$$\Omega_k = \begin{cases} \frac{E_{\mathrm{tot}}}{1+rF(\epsilon)}, & \text{for } k = 0 \\ \frac{E_{\mathrm{tot}}}{1+rF(\epsilon)} r e^{-\left(\frac{(k-1)T_{\mathrm{c}}}{\epsilon}\right)}, & \text{for } k = 1, 2, \ldots, N_{\mathrm{tap}} - 1 \end{cases} \tag{2}$$

where E_{tot} is the total average energy in all the paths normalized to the total energy received at 1-m distance, r is the ratio of the average energy of the second MPC and the average energy of the direct path, ϵ is the decay constant of the power delay profile, and $F(\epsilon) = (1 - \exp[-(N_{\mathrm{tap}}-1)kT_{\mathrm{c}}/\epsilon])/(1 - \exp(-kT_{\mathrm{c}}/\epsilon))$. According to [31], E_{tot}, r, and ϵ are all modeled by lognormal distributions. The Nakagami fading figures $\{m_k\}$ are distributed according to truncated Gaussian distributions whose mean and variance vary linearly with excess delay. In this paper, these long-term statistics are treated as constants over the duration of the acquisition process.

B. Transmitted and Received Signals

The transmitted signal is given by

$$x(t) = \sqrt{P} \sum_{l=-\infty}^{\infty} \psi(t - lT_{\mathrm{f}} - c_l T_{\mathrm{c}}) \tag{3}$$

where $\psi(t)$ is the UWB monocycle waveform, P is the transmitted power, $T_{\mathrm{f}} = N_{\mathrm{f}} T_{\mathrm{c}}$ is the pulse repetition time, $\{c_l\}$ is the pseudorandom time-hopping (TH) sequence with period N_{th} taking integer values between 0 and $N_{\mathrm{h}} - 1$, and T_{c} is the step size of the additional time shift provided by the TH sequence. The pulse repetition time T_{f} is chosen to be not less than $(N_{\mathrm{h}} + N_{\mathrm{tap}})T_{\mathrm{c}}$ to avoid overlap between the multipath responses corresponding to distinct transmitted pulses.

If $u(t) = h(t) * x(t)$, the received signal is given by

$$\begin{aligned} r(t) &= u(t) + n(t) \\ &= \sqrt{E_1} \sum_{l=-\infty}^{\infty} w(t - lT_{\mathrm{f}} - c_l T_{\mathrm{c}} - \tau) + n(t) \end{aligned} \tag{4}$$

where

$$w(t) = \sum_{k=0}^{N_{\mathrm{tap}}-1} p_k h_k \psi_{\mathrm{r}}(t - kT_{\mathrm{c}}). \tag{5}$$

Here E_1 is the total received energy at a distance of 1 m from the transmitter, $\psi_{\mathrm{r}}(t) = f(t) * \psi(t)$ is the received UWB pulse of

[1]The ambiguity function is the correlation output of the received signal and the locally generated template signal.

duration $T_{\rm w} < T_{\rm c}$ normalized to have unit energy, τ is the propagation delay, and $n(t)$ is an AWGN process with zero mean and power spectral density (PSD) $N_0/2$.

III. Hit Set Definition

The timing information of the received signal is essential for the performance of a receiver in a wireless communication system. In a multipath channel, the energy corresponding to the true signal phase is spread over several MPCs. The main difference between the acquisition problems in a multipath channel and a channel without multipath is that there are more than one hypothesized phases which can be considered a good estimate of the true signal phase. In a multipath environment, the receiver may lock onto a nonline-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 postacquisition 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. A typical paradigm for transceiver design is the achievement of a certain nominal uncoded bit-error rate (BER) $\lambda_{\rm n}$. Then all those hypothesized phases such that a receiver locked to them achieves an uncoded BER of $\lambda_{\rm n}$ can be considered a good estimate of the true signal phase. We define the *hit set* to be the set of such hypothesized phases.

To simplify the analysis, we assume that the true phase τ is an integer multiple of $T_{\rm c}$. By the periodicity of the transmitted signal, we have $0 \leq \tau \leq (N_{\rm th}N_{\rm f} - 1)T_{\rm c}$. The hypothesized phase $\hat{\tau}$ is also an integer multiple of $T_{\rm c}$ with the same range as τ. Then $\Delta\tau = \hat{\tau} - \tau = \alpha T_{\rm f} + \beta T_{\rm c}$ where α and β are integers such that $-N_{\rm th} + 1 \leq \alpha \leq N_{\rm th} - 1$ and $0 \leq \beta \leq N_{\rm f} - 1$. For a given true phase τ, let $P_{\rm E}(\Delta\tau)$ denote the BER performance of the PRake receiver when it locks to the hypothesized phase $\hat{\tau}$. Let $\Upsilon_{\rm m}$ be the minimum SNR at which the PRake receiver achieves a BER of $\lambda_{\rm n}$ when it locks to the LOS path, that is, $P_{\rm E}(0) \leq \lambda_{\rm n}$ when the SNR is $\Upsilon_{\rm n}$ and $P_{\rm E}(0) > \lambda_{\rm n}$ for all SNRs less than $\Upsilon_{\rm n}$. Then for an SNR $\Upsilon \geq \Upsilon_{\rm n}$ and true phase τ, the hit set is given by

$$S_{\rm h} = \{\hat{\tau} : P_{\rm E}(\Delta\tau) \leq \lambda_{\rm n}\}. \quad (6)$$

To completely characterize the hit set, we need to calculate the error performance of a PRake receiver which is locked to a particular hypothesized phase $\hat{\tau}$. We assume that the modulation format is binary phase-shift keying (BPSK) with $N_{\rm b}$ consecutive UWB monocycles modulated by one bit. The signal received during the demodulation stage is given by

$$r_{\rm b}(t) = \sqrt{E_1}\sum_{l=-\infty}^{\infty} b_{\left\lfloor \frac{l}{N_{\rm b}} \right\rfloor} w(t - lT_{\rm f} - c_l T_{\rm c} - \tau) + n(t) \quad (7)$$

where $b_i \in \{-1, 1\}$ for each i, $\lfloor x \rfloor$ is the largest integer not greater than x

$$w(t) = \sum_{k=0}^{N_{\rm tap}-1} p_k h_k \psi_{\rm r}(t - kT_{\rm c}) \quad (8)$$

and $n(t)$ is a zero-mean AWGN process with PSD $N_0/2$. The PRake receiver is assumed to have $N_{\rm p}$ fingers where $N_{\rm p} \leq N_{\rm tap}$. When the receiver estimates $\hat{\tau}$ to be the true phase in the acquisition stage, the PRake receiver estimates and combines the paths arriving at delays $\hat{\tau} + kT_{\rm c}$ $(k = 0, 1, \ldots, N_{\rm p} - 1)$ to obtain the decision statistic. Since $\hat{\tau} = \tau + (\alpha N_{\rm f} + \beta)T_{\rm c}$, the PRake receiver is estimating the values of $p_{\alpha N_{\rm f}+\beta+i}h_{\alpha N_{\rm f}+\beta+i}$ for $i = 0, 1, \ldots, N_{\rm p} - 1$, where we define $p_k = h_k = 0$ for $k \notin \{0, 1, \ldots, N_{\rm tap} - 1\}$. To make the analysis tractable, we assume that PRake is able to estimate these path gains and inversions perfectly. The decision statistic for the mth bit, Z_m, can be obtained by correlating the received signal with the following template signal:

$$s_{\rm b}(t) = \sum_{l=mN_{\rm b}}^{(m+1)N_{\rm b}-1} v(t - lT_{\rm f} - c_l T_{\rm c} - \hat{\tau}) \quad (9)$$

where

$$v(t) = \sum_{i=0}^{N_{\rm p}-1} p_{\alpha N_{\rm f}+\beta+i} h_{\alpha N_{\rm f}+\beta+i} \psi_{\rm r}(t - iT_{\rm c}). \quad (10)$$

Then we have

$$\begin{aligned} Z_m &= \frac{1}{N_{\rm b}} \int_{\hat{\tau}+mN_{\rm b}T_{\rm f}}^{\hat{\tau}+[(m+1)N_{\rm b}-1]T_{\rm f}} r_{\rm b}(t)s_{\rm b}(t)dt \\ &= b_m\sqrt{E_1}\sum_{i=0}^{N_{\rm p}-1} p^2_{\alpha N_{\rm f}+\beta+i} h^2_{\alpha N_{\rm f}+\beta+i} + n_{\rm b} \\ &= b_m\sqrt{E_1}\sum_{i=0}^{N_{\rm p}-1} h^2_{\alpha N_{\rm f}+\beta+i} + n_{\rm b} \end{aligned} \quad (11)$$

where $n_{\rm b}$ is a zero-mean Gaussian random variable (RV) with variance $\sigma_{\rm b}^2 = (N_0/2N_{\rm b})\sum_{i=0}^{N_{\rm p}-1} h^2_{\alpha N_{\rm f}+\beta+i}$. Then from [34, pp. 268–269], the average probability of error is given by

$$\begin{aligned} P_{\rm E}(\Delta\tau) &= E_h\left[Q\left(\sqrt{\frac{2E_1N_{\rm b}\sum_{i=0}^{N_{\rm p}-1} h^2_{\alpha N_{\rm f}+\beta+i}}{N_0}}\right)\right] \\ &= \frac{1}{\pi}\int_0^{\frac{\pi}{2}} \prod_{i=\alpha N_{\rm f}+\beta}^{\alpha N_{\rm f}+\beta+N_{\rm p}-1} \mathcal{M}_i\left(-\frac{2E_1N_{\rm b}}{N_0\sin^2\theta}\right)d\theta \end{aligned}$$

where $\mathcal{M}_i(\cdot)$, the moment generating function of h_i^2, is given by

$$\mathcal{M}_i(s) = \begin{cases} \left(1 - \frac{\Omega_i s}{m_i}\right)^{-m_i}, & \text{for } i \in \{0, 1, \ldots, N_{\rm tap} - 1\} \\ 1, & \text{otherwise.} \end{cases} \quad (12)$$

Fig. 1 shows the hit set size as a function of the average energy-received-per-pulse-to-noise ratio $E_1E_{\rm tot}/N_0$ for $\lambda_{\rm n} = 10^{-3}$, $N_{\rm b} = 8$ and $N_{\rm p} = 5, 10$. The values of the other system parameters are given in Section VII. This plot confirms our claim in the beginning of this section about the existence of multiple phases where a receiver lock can guarantee adequate demodulation performance.

IV. Performance of SAI

In this section, we derive the performance of an acquisition system which takes the SAI approach.

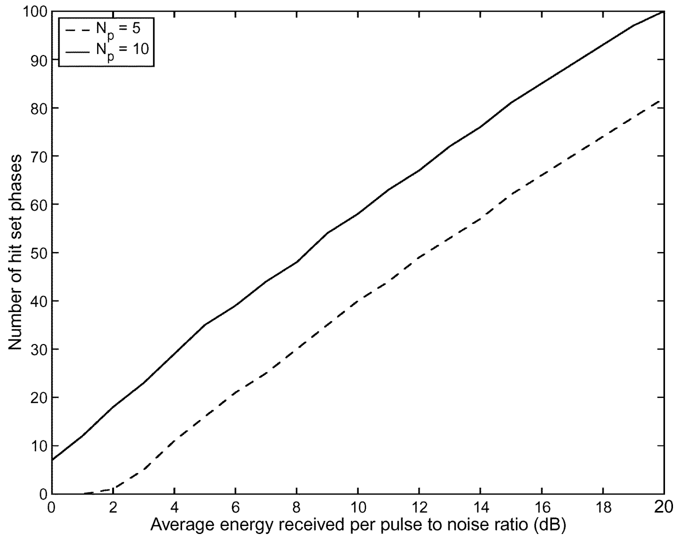

Fig. 1. Hit set size as a function of the average energy-received-per-pulse-to-noise ratio for $N_{\mathrm{P}} = 5, 10$.

A. Derivation of the Decision Statistic

The acquisition system correlates the squared received waveform with a locally generated replica, and compares the correlator output with a threshold to determine whether the hypothesized phase of the replica is correct. If the threshold is exceeded, the hypothesized phase becomes the estimate of the true phase. We assume that the normalized received monocycle waveform $\psi_{\mathrm{r}}(t)$ and the TH sequence $\{c_l\}$ are known to the receiver. We propose to use an equal-gain combiner of window size G. The receiver template signal $w_{\mathrm{r}}(t)$ is given by

$$w_{\mathrm{r}}(t) = \sum_{k=0}^{G-1} \psi_{\mathrm{r}}^2(t - kT_{\mathrm{c}}). \tag{13}$$

The reference TH signal can be obtained by combining the receiver template signal $w_{\mathrm{r}}(t)$ and the known TH sequence as

$$s(t) = \sum_{l=0}^{MN_{\mathrm{th}}-1} w_{\mathrm{r}}(t - lT_{\mathrm{f}} - c_l T_{\mathrm{c}} - \hat{\tau}) \tag{14}$$

where M specifies the number of TH waveform periods in the dwell time and $\hat{\tau}$ is the hypothesized phase. The correlator output is given by

$$\begin{aligned} y &= \frac{1}{MN_{\mathrm{th}}} \int_{\hat{\tau}}^{\hat{\tau}+MN_{\mathrm{th}}T_{\mathrm{f}}} r^2(t)s(t)dt \\ &= \frac{1}{MN_{\mathrm{th}}} \int_{\hat{\tau}}^{\hat{\tau}+MN_{\mathrm{th}}T_{\mathrm{f}}} u^2(t)s(t)dt \\ &\quad + \frac{2}{MN_{\mathrm{th}}} \int_{\hat{\tau}}^{\hat{\tau}+MN_{\mathrm{th}}T_{\mathrm{f}}} u(t)s(t)n(t)dt \\ &\quad + \frac{1}{MN_{\mathrm{th}}} \int_{\hat{\tau}}^{\hat{\tau}+MN_{\mathrm{th}}T_{\mathrm{f}}} n^2(t)s(t)dt. \end{aligned} \tag{15}$$

The first term in (15) can be simplified to

$$R(\Delta\tau; \mathbf{h}) = E_1 R_{\psi_{\mathrm{r}}^2}(0) \sum_{k=0}^{N_{\mathrm{tap}}-1} r_k(\Delta\tau) h_k^2 \tag{16}$$

where $\mathbf{h}$ is an $N_{\mathrm{tap}} \times 1$ vector containing the channel gains, $R_{\psi_{\mathrm{r}}^n}(\nu) = \int_{-\infty}^{\infty} \psi_{\mathrm{r}}^n(t)\psi_{\mathrm{r}}^n(t+\nu)dt$ and $r_k(\Delta\tau)$, the average number of times the energy in the kth MPC is collected by one period of the reference TH signal, is given by

$$r_k(\Delta\tau) = \frac{1}{N_{\mathrm{th}}} \sum_{l=0}^{N_{\mathrm{th}}-1} \sum_{i=0}^{1} \sum_{j=0}^{G-1} \chi(c_l + j + \beta, c_{l+i+\alpha} + k + iN_{\mathrm{f}}) \tag{17}$$

where $\chi(a,b) = 1$ if $a = b$, and 0, otherwise. The value of $r_k(\Delta\tau)$ depends on the particular pseudorandom TH sequence chosen. To simplify the analysis, we assume that the TH sequence is random and that N_{th} is large. Under these assumptions, the mean value of $r_k(\Delta\tau)$ is a reasonable approximation to the actual value. The mean value of $r_k(\Delta\tau)$ is calculated in the Appendix by averaging over the set of random TH sequences.

Conditioned on the random vector $\mathbf{h}$, the second term in (15) is a zero-mean Gaussian RV with variance

$$\sigma_{\mathrm{y}}^2(\Delta\tau; \mathbf{h}) = \frac{2N_0}{M^2 N_{\mathrm{th}}^2} \int_{\hat{\tau}}^{\hat{\tau}+MN_{\mathrm{th}}T_{\mathrm{f}}} u^2(t)s^2(t)dt \tag{18}$$

$$= \frac{2E_1 R_{\psi_{\mathrm{r}}^3}(0) N_0}{MN_{\mathrm{th}}} \sum_{k=0}^{N_{\mathrm{tap}}-1} r_k(\Delta\tau) h_k^2 \tag{19}$$

where the second equality is obtained by exploiting the similarity between the integral in (18) and the first term of (15). We have also used the fact that

$$s^2(t) = \sum_{l=0}^{MN_{\mathrm{th}}-1} \sum_{k=0}^{G-1} \psi_{\mathrm{r}}^4(t - kT_{\mathrm{c}} - lT_{\mathrm{f}} - c_l T_{\mathrm{c}} - \hat{\tau}) \tag{20}$$

which differs from $s(t)$ only in the exponent of the received pulse waveform $\psi_{\mathrm{r}}(t)$.

We approximate the third term in (15) by a Gaussian RV with mean μ_{y} and variance ν_{y}^2, which are given by

$$\begin{aligned} \mu_{\mathrm{y}} &= \frac{1}{MN_{\mathrm{th}}} E\left[\int_{\hat{\tau}}^{\hat{\tau}+MN_{\mathrm{th}}T_{\mathrm{f}}} n^2(t)s(t)dt\right] \\ &= \frac{N_0}{2MN_{\mathrm{th}}} \int_{\hat{\tau}}^{\hat{\tau}+MN_{\mathrm{th}}T_{\mathrm{f}}} s(t)dt \\ &= \frac{GR_{\psi_{\mathrm{r}}}(0)N_0}{2} \\ &= \frac{GN_0}{2} \end{aligned} \tag{21}$$

and

$$\begin{aligned}\nu_{\mathrm{y}}^2 &= \frac{1}{M^2N_{\mathrm{th}}^2}E\left[\left(\int_{\hat{\tau}}^{\hat{\tau}+MN_{\mathrm{th}}T_{\mathrm{f}}} n^2(t)s(t)dt\right)^2\right] - \mu_{\mathrm{y}}^2 \\ &= \frac{E\left[\int_{\hat{\tau}}^{\hat{\tau}+MN_{\mathrm{th}}T_{\mathrm{f}}}\int_{\hat{\tau}}^{\hat{\tau}+MN_{\mathrm{th}}T_{\mathrm{f}}} n^2(t)n^2(u)s(t)s(u)dtdu\right]}{M^2N_{\mathrm{th}}^2} \\ &\quad - \mu_{\mathrm{y}}^2 \\ &= \frac{N_0^2}{2M^2N_{\mathrm{th}}^2}\int_{\hat{\tau}}^{\hat{\tau}+MN_{\mathrm{th}}T_{\mathrm{f}}} s^2(t)dt \\ &\quad + \frac{N_0^2}{4M^2N_{\mathrm{th}}^2}\left(\int_{\hat{\tau}}^{\hat{\tau}+MN_{\mathrm{th}}T_{\mathrm{f}}} s(t)dt\right)^2 - \mu_{\mathrm{y}}^2 \\ &= \frac{N_0^2R_{\psi_{\mathrm{r}}^4}(0)}{2MN_{\mathrm{th}}}\end{aligned} \tag{22}$$

respectively. Note that the expectation in the derivation of μ_{y} and ν_{y}^2 is only with respect to the noise process $n(t)$. This approximation is accurate provided that the product of the integration time $MN_{\mathrm{th}}T_{\mathrm{f}}$ and the bandwidth of the system B is large [35, pp. 240–250], which is the case for the scenarios we consider.

Then the correlator output can be written as

$$y = R(\Delta\tau;\mathbf{h}) + n_{\mathrm{y}} \tag{23}$$

where, conditioned on $\mathbf{h}$, n_{y} is a Gaussian RV with mean μ_{y} and variance $\sigma_{\mathrm{y}}^2(\Delta\tau;\mathbf{h}) + \nu_{\mathrm{y}}^2$.

B. Average Probabilities of Detection and False Alarm

For a particular channel realization $\mathbf{h}$ and fixed $\Delta\tau$, the decision statistic y in (23) has a Gaussian distribution with probability density function (pdf)

$$p_{\mathrm{Y}}(y) = \frac{\exp\left[\frac{-(y-R(\Delta\tau;\mathbf{h})-\mu_{\mathrm{y}})^2}{2\left(\sigma_{\mathrm{y}}^2(\Delta\tau;\mathbf{h})+\nu_{\mathrm{y}}^2\right)}\right]}{\sqrt{2\pi\left(\sigma_{\mathrm{y}}^2(\Delta\tau;\mathbf{h})+\nu_{\mathrm{y}}^2\right)}}. \tag{24}$$

The probabilities of false alarm and detection conditioned on the particular channel realization and given the decision threshold γ are given as

$$\begin{aligned}P_{\mathrm{FA}}(\gamma,\Delta\tau|\mathbf{h}) &= \Pr[y>\gamma|\hat{\tau}\notin S_{\mathrm{h}}] \\ &= Q\left(\frac{\gamma - R(\Delta\tau;\mathbf{h})-\mu_{\mathrm{y}}}{\sqrt{\sigma_{\mathrm{y}}^2(\Delta\tau;\mathbf{h})+\nu_{\mathrm{y}}^2}}\right), \quad \hat{\tau}\notin S_{\mathrm{h}}. \\ P_{\mathrm{D}}(\gamma,\Delta\tau|\mathbf{h}) &= \Pr[y>\gamma|\hat{\tau}\in S_{\mathrm{h}}] \\ &= Q\left(\frac{\gamma - R(\Delta\tau;\mathbf{h})-\mu_{\mathrm{y}}}{\sqrt{\sigma_{\mathrm{y}}^2(\Delta\tau;\mathbf{h})+\nu_{\mathrm{y}}^2}}\right), \quad \hat{\tau}\in S_{\mathrm{h}}.\end{aligned}$$

From (16) and (19), one sees that the conditional probabilities of false alarm and detection depend on $\mathbf{h}$ only through $s(\Delta\tau;\mathbf{h}) = \sum_{k=0}^{N_{\mathrm{tap}}-1} r_k(\Delta\tau)h_k^2$. Using (16) and (19), we define

$$I\left(s(\Delta\tau;\mathbf{h})\right) = Q\left(\frac{\gamma - E_1R_{\psi_{\mathrm{r}}^2}(0)s(\Delta\tau;\mathbf{h})-\mu_{\mathrm{y}}}{\sqrt{\frac{2E_1R_{\psi_{\mathrm{r}}^3}(0)N_0}{MN_{\mathrm{th}}}s(\Delta\tau;\mathbf{h})+\nu_{\mathrm{y}}^2}}\right). \tag{25}$$

Since the path gains h_k $(k=0,1,\ldots,N_{\mathrm{tap}}-1)$ are independent, the characteristic function of $s(\Delta\tau;\mathbf{h})$ is given by

$$\Phi_{\mathrm{s}}(\omega;\Delta\tau) = \prod_{k=0}^{N_{\mathrm{tap}}-1}\mathcal{M}_k\left(\mathrm{j}r_k(\Delta\tau)\omega\right) \tag{26}$$

where $\mathcal{M}_k(\cdot)$ is defined in (12). The pdf of $s(\Delta\tau;\mathbf{h})$ is given by $f_{\mathrm{s}}(x;\Delta\tau) = (1/2\pi)\int_{-\infty}^{\infty}\Phi_{\mathrm{s}}(\omega;\Delta\tau)e^{-\mathrm{j}\omega x}d\omega$. Then for $\hat{\tau}\notin S_{\mathrm{h}}$, the probability of false alarm averaged over the channel realizations is given by

$$\begin{aligned}E_{\mathrm{H}}\left[P_{\mathrm{FA}}(\gamma,\Delta\tau|\mathbf{h})\right] &= E_{\mathrm{H}}\left[I\left(s(\Delta\tau;\mathbf{h})\right)\right] \\ &= \int_0^{\infty} I(t)f_{\mathrm{s}}(t;\Delta\tau)dt.\end{aligned} \tag{27}$$

Similarly, for $\hat{\tau}\in S_{\mathrm{h}}$, the average probability of detection is given by

$$E_{\mathrm{H}}\left[P_{\mathrm{D}}(\gamma,\Delta\tau|\mathbf{h})\right] = \int_0^{\infty} I(t)f_{\mathrm{s}}(t;\Delta\tau)dt. \tag{28}$$

The structure of $\mathcal{M}_k(\cdot)$ prevents evaluating $f_{\mathrm{s}}(\cdot)$ in closed form. So we resort to numerical integration to calculate $f_{\mathrm{s}}(\cdot)$ and the average probabilities of false alarm and detection.

V. Performance of IAS

In this section, we derive the performance of an acquisition system which takes the IAS approach. The derivation of the decision statistic in this case is very similar to the decision statistic derivation in the previous section. All the relevant assumptions made in the previous section, to enable tractable analysis, still hold unless stated otherwise. To avoid repetition, we only define those quantities which have not already been defined in the previous section.

A. Derivation of the Decision Statistic

In this approach, the acquisition system correlates the received waveform with a locally generated template signal and squares the integrator output to generate the decision statistic. The receiver template signal $v_{\mathrm{r}}(t)$ is given by

$$v_{\mathrm{r}}(t) = \sum_{k=0}^{G-1}\psi_{\mathrm{r}}(t-kT_{\mathrm{c}}). \tag{29}$$

The reference TH signal is given by

$$q(t) = \sum_{l=0}^{MN_{\text{th}}-1} v_{\text{r}}(t - lT_{\text{f}} - c_l T_{\text{c}} - \hat{\tau}). \tag{30}$$

The decision statistic is given by

$$\begin{aligned} z &= \left[\frac{1}{MN_{\text{th}}} \int_{\hat{\tau}}^{\hat{\tau}+MN_{\text{th}}T_{\text{f}}} r(t)q(t)dt \right]^2 \\ &= \left[\underbrace{\sqrt{E_1} \sum_{k=0}^{N_{\text{tap}}-1} r_k(\Delta\tau) p_k h_k}_{V(\Delta\tau;\mathbf{h})} + n_{\text{z}} \right]^2 \end{aligned} \tag{31}$$

where n_{z} is a zero-mean Gaussian RV with variance $\sigma_{\text{z}}^2 = GN_0/2MN_{\text{th}}$ and $r_k(\Delta\tau)$ is given in (17).

B. Average Probabilities of Detection and False Alarm

For a particular channel realization $\mathbf{h}$ and fixed $\Delta\tau$, the decision statistic z in (31) has a noncentral chi-square distribution with pdf

$$p_{\text{Z}}(z) = \frac{1}{\sqrt{2\pi z}\sigma_{\text{z}}} e^{-\frac{\left(z+V^2(\Delta\tau;\mathbf{h})\right)}{2\sigma_{\text{z}}^2}} \cosh\left(\frac{\sqrt{z}V(\Delta\tau;\mathbf{h})}{\sigma_{\text{z}}^2}\right).$$

The probabilities of false alarm and detection conditioned on the particular channel realization and given the decision threshold $\gamma \geq 0$ are given by

$$\begin{aligned} &P_{\text{FA}}(\gamma, \Delta\tau|\mathbf{h}) \\ &\quad = \Pr[z > \gamma|\hat{\tau} \notin S_{\text{h}}] \\ &\quad = Q\left(\frac{\sqrt{\gamma} - V(\Delta\tau;\mathbf{h})}{\sigma_{\text{z}}}\right) + Q\left(\frac{\sqrt{\gamma} + V(\Delta\tau;\mathbf{h})}{\sigma_{\text{z}}}\right), \\ &\qquad \hat{\tau} \notin S_{\text{h}} \\ &P_{\text{D}}(\gamma, \Delta\tau|\mathbf{h}) \\ &\quad = \Pr[z > \gamma|\hat{\tau} \in S_{\text{h}}] \\ &\quad = Q\left(\frac{\sqrt{\gamma} - V(\Delta\tau;\mathbf{h})}{\sigma_{\text{z}}}\right) + Q\left(\frac{\sqrt{\gamma} + V(\Delta\tau;\mathbf{h})}{\sigma_{\text{z}}}\right), \\ &\qquad \hat{\tau} \in S_{\text{h}}. \end{aligned} \tag{32}$$

Before we derive the average probabilities of detection and false alarm, it is instructive to look at the characteristic function $\Phi_{\text{V}}(\omega;\Delta\tau)$ of $V(\Delta\tau;\mathbf{h})$. Since the polarities p_k and path gains h_k are independent, we have

$$\begin{aligned} &\Phi_{\text{V}}(\omega;\Delta\tau) \\ &\quad = \prod_{k=0}^{N_{\text{tap}}-1} \left[\frac{\phi_k\left(\sqrt{E_1} r_k(\Delta\tau)\omega\right) + \phi_k\left(-\sqrt{E_1} r_k(\Delta\tau)\omega\right)}{2} \right] \end{aligned} \tag{33}$$

where $\phi_k(\cdot)$ is the characteristic function of the Nakagami-m distributed h_k [34]. Since the h_k's are real-valued, the $\phi_k(\cdot)$'s are conjugate symmetric functions, and hence, $\Phi_{\text{V}}(\cdot)$ is a real-valued function.

The Gil–Pelaez lemma [36] gives an alternative form of the Q function as

$$Q(x) = \frac{1}{2} - \frac{1}{\pi} \int_0^{\infty} \frac{1}{t} e^{-\frac{t^2}{2}} \sin(tx) dt. \tag{34}$$

Substituting this form of the Q function in (32), the probability of false alarm averaged over the channel realizations, for $\hat{\tau} \notin S_{\text{h}}$, is given by

$$\begin{aligned} &E_{\text{H}}\left[P_{\text{FA}}(\gamma,\Delta\tau|\mathbf{h})\right] \\ &\quad = 1 - \frac{1}{\pi} \int_0^{\infty} \frac{1}{t} e^{-\frac{t^2}{2}} E_{\text{H}} \\ &\qquad \times \left[\sin\frac{t\left(\sqrt{\gamma} - V(\Delta\tau;\mathbf{h})\right)}{\sigma_{\text{z}}} + \sin\frac{t\left(\sqrt{\gamma} + V(\Delta\tau;\mathbf{h})\right)}{\sigma_{\text{z}}} \right] dt \\ &\quad = 1 - \frac{2}{\pi} \int_0^{\infty} \frac{1}{t} e^{-\frac{t^2}{2}} \sin\left(\frac{\sqrt{\gamma}t}{\sigma_{\text{z}}}\right) E_{\text{H}}\left[\sin\frac{V(\Delta\tau;\mathbf{h})t}{\sigma_{\text{z}}}\right] dt \\ &\quad = 1 - \frac{2}{\pi} \int_0^{\infty} \frac{1}{t} e^{-\frac{t^2}{2}} \sin\left(\frac{\sqrt{\gamma}t}{\sigma_{\text{z}}}\right) \operatorname{Im}\left\{E_{\text{H}}\left[e^{\mathrm{j}\frac{V(\Delta\tau;\mathbf{h})t}{\sigma_{\text{z}}}}\right]\right\} dt \\ &\quad = 1 - \frac{2}{\pi} \int_0^{\infty} \frac{1}{t} e^{-\frac{t^2}{2}} \sin\left(\frac{\sqrt{\gamma}t}{\sigma_{\text{z}}}\right) \Phi_{\text{V}}\left(\frac{t}{\sigma_{\text{z}}};\Delta\tau\right) dt \end{aligned} \tag{35}$$

where the last equality follows from our observation that $\Phi_{\text{V}}(\cdot)$ is real-valued. Similarly, for $\hat{\tau} \in S_{\text{h}}$, the average probability of detection is given by

$$\begin{aligned} &E_{\text{H}}\left[P_{\text{D}}(\gamma,\Delta\tau|\mathbf{h})\right] \\ &\quad = 1 - \frac{2}{\pi} \int_0^{\infty} \frac{1}{t} e^{-\frac{t^2}{2}} \sin\left(\frac{\sqrt{\gamma}t}{\sigma_{\text{z}}}\right) \Phi_{\text{V}}\left(\frac{t}{\sigma_{\text{z}}};\Delta\tau\right) dt. \end{aligned} \tag{36}$$

VI. Mean Detection Time Analysis of Serial Search

We define a *hit* or detection event as the event when the decision threshold is exceeded for some $\hat{\tau} \in S_{\text{h}}$. We define a *miss* as the event when the decision threshold is not exceeded for all $\hat{\tau} \in S_{\text{h}}$. Although the average probability of a miss is a potential indicator of acquisition system performance, the mean acquisition time is usually the metric used to evaluate the performance of acquisition systems [35]. The calculation of the mean acquisition time enumerates all false alarms which occur before a detection event and associates a false-alarm penalty time T_{fa} to each one of them. The false-alarm penalty time is equal to the dwell time of a verification stage in the acquisition system which aids in the confirmation of detection events and rejection of false-alarm events with high probability. In other words, a good verification stage simultaneously achieves low probabilities of miss and false alarm. The choice of the mean acquisition time as the performance metric implicitly assumes that one can construct such a verification stage. However, for threshold-based UWB acquisition systems, it is shown [37] that the average probabilities of false alarm and miss cannot be made arbitrarily small, even in the asymptotic scenario of the SNR approaching infinity. Thus, it is not apparent how one would build a good verification stage for such systems. We propose to deal with this problem by choosing the decision threshold γ such that the average probability that the acquisition process will end in a false alarm, $P_{\rm F}(\gamma)$, is small. The justification for this design is that a false alarm is a more serious problem in the absence of a verification stage. Then if $P_{\rm F}(\gamma)$ is small enough, we can use the mean detection time as the performance metric. The mean detection time is defined as the average time it takes for the acquisition process to end in a detection event in the absence of false alarms.

In the Appendix, we calculate $P_{\rm F}(\gamma)$ as a function of the average probabilities of detection in the hit set and the average probabilities of false alarm at the phases not in the hit set. We choose the decision threshold $\gamma_{\rm d}$ to be the minimum threshold, such that $P_{\rm F}(\gamma)$ is not greater than a given positive constant $\delta \leq 1$

$$\gamma_{\rm d} = \inf\left\{\gamma | P_{\rm F}(\gamma) \leq \delta\right\}. \tag{37}$$

If the correlator outputs for different phase evaluations are assumed to be independent, then the average probability of a hit for a particular $\hat{\tau}$ is $E_h[P_{\rm D}(\gamma_{\rm d}, \Delta\tau|\mathbf{h})]$, and the average probability of a miss is given by

$$P_{\rm M} = \prod_{\hat{\tau}\in S_{\rm h}} \left(1 - E_h\left[P_{\rm D}(\gamma_{\rm d}, \Delta\tau|\mathbf{h})\right]\right). \tag{38}$$

The mean detection time depends on the search strategy employed, and to illustrate the effect of the choice of G, we assume that a serial search strategy is used. Owing to our definition, the hit set $S_{\rm h}$ consists of a contiguous set of H hypothesized phases within the search space. The search space is the set $S_p = \{nT_{\rm c} : n \in Z \text{ and } 0 \leq n \leq N_{\rm s} - 1\}$, where $N_{\rm s} = N_{\rm th}N_{\rm f}$. Let the first phase of the hit set be at position A in the search space $S_{\rm p}$. Then the hit set consists of the phases $\{(A-1)T_{\rm c}, AT_{\rm c}, \ldots, (A+H-2)T_{\rm c}\}$. The initial value of the hypothesized phase which corresponds to the starting point of the search is chosen at random from the set S_p. Thus, there is no loss of generality in assuming that $A = 1$.

We need to consider all possible sequences of events leading to a hit or detection event. The mean detection time can then be calculated as the average time taken for each of the detection events. A detection event is defined by a particular position n of the initial value of the hypothesized phase in S_p, the position i of the hypothesized phase in $S_{\rm h}$ where we have a hit and a particular number of misses j of $S_{\rm h}$. Let $T_{\rm det}(n)$ be the mean detection time conditioned on the event that the serial search starts at the nth position in S_p, i.e., the initial value of the hypothesized phase is $(n-1)T_{\rm c}$. Then the mean detection time is

$$\overline{T}_{\rm det} = \frac{1}{N_{\rm s}} \sum_{n=1}^{N_{\rm s}} T_{\rm det}(n). \tag{39}$$

First, suppose that the initial value of the hypothesized phase lies to the right of the hit set, i.e., $n \in \{H+1, H+2, \ldots, N_{\rm s}\}$. The total detection time for a particular detection event defined by (n, j, i) is then

$$\begin{aligned} T(n,j,i) &= (N_{\rm s} - n + 1)T + jN_{\rm s}T + iT \\ &= (N_{\rm s} - n + 1 + jN_{\rm s} + i)T \end{aligned} \tag{40}$$

where T is the dwell time for the evaluation of one hypothesized phase. Let $P_{\rm d}(i)$ denote the average probability of detection of the ith phase of the hit set. The average probability of the serial search missing the hit set is $P_{\rm M} = \prod_{i=1}^{H}[1 - P_{\rm d}(i)]$. Then the probability of j misses of $S_{\rm h}$ followed by a hit at the phase in $S_{\rm h}$ which is at the ith position of the hit set is $P_{\rm M}^j P_{\rm h}(i)$ where $P_{\rm h}(i) = P_{\rm d}(i)\prod_{r=1}^{i-1}[1 - P_{\rm d}(r)]$. The mean detection time conditioned on the starting point of the serial search is given by

$$\begin{aligned} T_{\rm det}(n) &= \sum_{i=1}^{H}\sum_{j=0}^{\infty} T(n,j,i)P_{\rm M}^j P_{\rm h}(i) \\ &= \sum_{i=1}^{H}\left[\frac{N_{\rm s} - n + i + 1}{1 - P_{\rm M}} + \frac{N_{\rm s}P_{\rm M}}{(1-P_{\rm M})^2}\right] TP_{\rm h}(i) \\ &= (N_{\rm s} - n + 1)T + \frac{N_{\rm s}P_{\rm M}T}{1 - P_{\rm M}} \\ &\quad + \frac{\sum_{i=1}^{H} iTP_{\rm h}(i)}{1 - P_{\rm M}} \end{aligned} \tag{41}$$

where we have used the identities $\sum_{i=1}^{H} P_{\rm h}(i) = 1 - P_{\rm M}$ in obtaining the third equality.

Now suppose that the initial value of the hypothesized phase falls in the hit set, i.e., $n \in \{1, 2, \ldots, H\}$. Let m be the total number of phases evaluated for a particular detection event. We can partition the set of detection events into two sets, one containing those events for which $m \leq H - n + 1$ and the other containing those events for which $m > H - n + 1$. The mean detection time for events in the first set is just mT, and for events in the second set, it is $T_{\rm det}(H+1) + (H - n + 1)T$, where $T_{\rm det}(H+1)$ is obtained from (41). Averaging over the total number of phases evaluated, we get

$$\begin{aligned} T_{\rm det}(n) &= \sum_{i=n}^{H}(i - n + 1)TP_{\rm d}(i)\prod_{j=n}^{i-1}(1 - P_{\rm d}(j)) \\ &\quad + (T_{\rm det}(H+1) + (H - n + 1)T)\prod_{j=n}^{H}(1 - P_{\rm d}(j)). \end{aligned} \tag{42}$$

From (41) and (42), we obtain the conditional mean detection times $T_{\rm det}(n)$ for all values of $n \in \{1, 2, \ldots, N_{\rm s}\}$. The mean detection time is obtained by substituting these values in (39).

VII. Numerical Results

In this section, we compare the performance of SAI and IAS in terms of the average probability of a miss $P_{\rm M}$ and the mean detection time $\overline{T}_{\rm det}$. We also investigate the effect of increasing the EGC window length on these performance metrics for both schemes. We choose the following values for the system parameters: the TH sequence period $N_{\rm th} = 256$, $N_{\rm h} = 16$, the length of the channel response $N_{\rm tap} = 100$, $N_{\rm f} = 116$, and the number of monocycles modulated by one bit $N_{\rm b} = 8$. The nominal uncoded BER requirement is set to be $\lambda_{\rm n} = 10^{-3}$. The

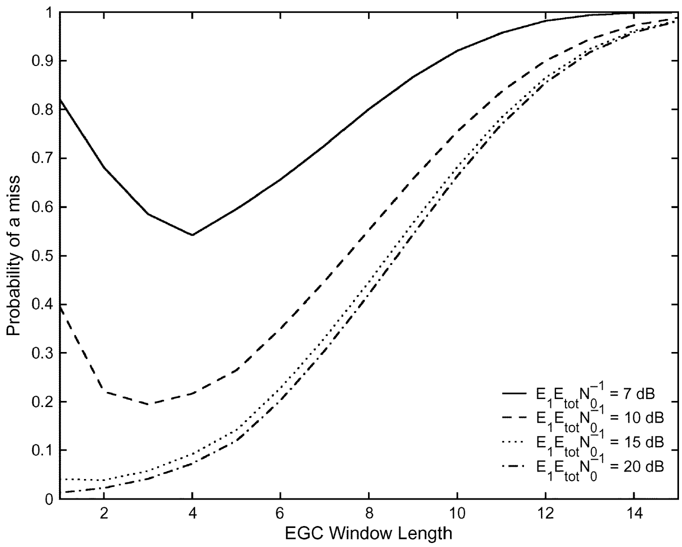

Fig. 2. Effect of EGC window length on the probability of a miss for SAI when $N_{\rm p} = 5$.

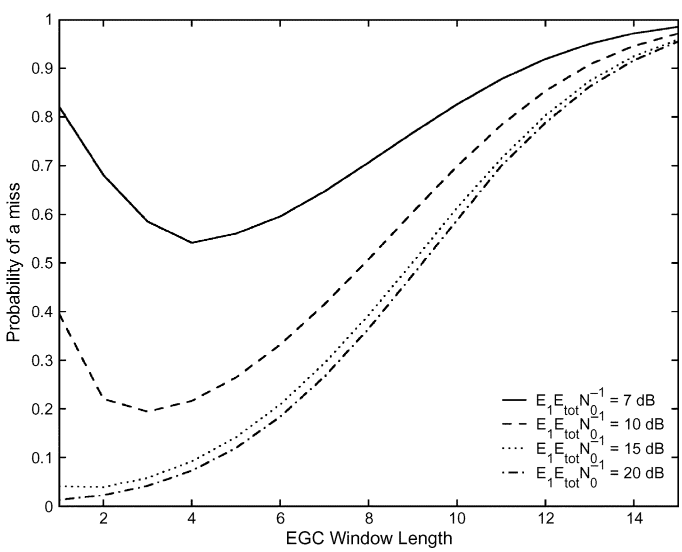

Fig. 3. Effect of EGC window length on the probability of a miss for SAI when $N_{\rm p} = 10$.

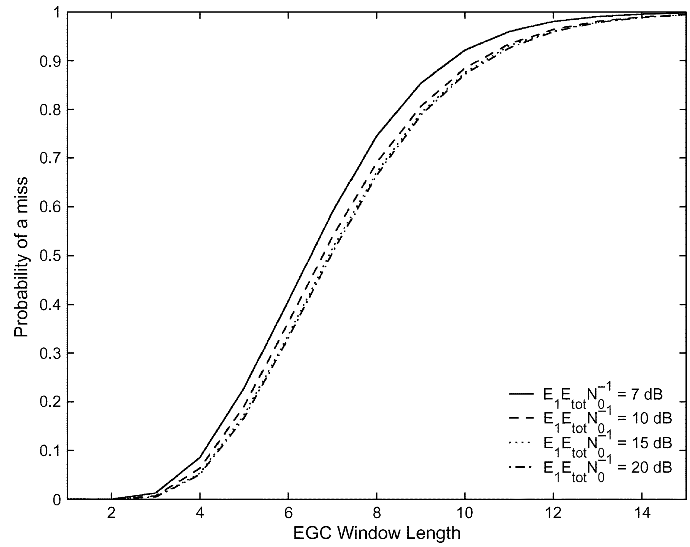

Fig. 4. Effect of EGC window length on the probability of a miss for IAS when $N_{\rm p} = 5$.

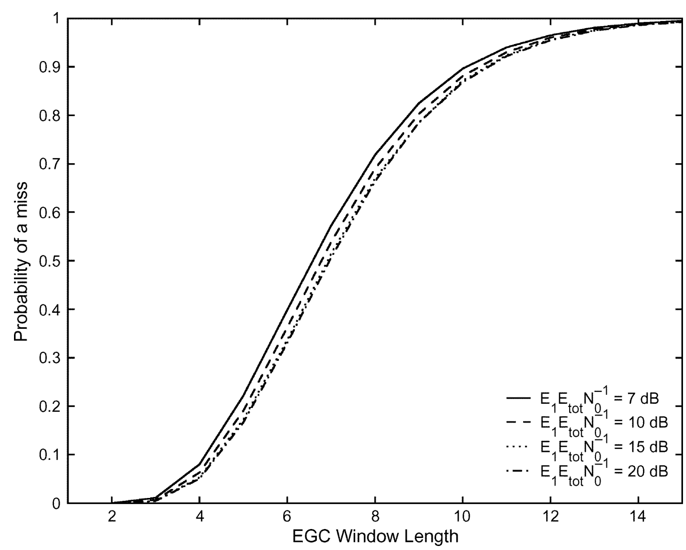

Fig. 5. Effect of EGC window length on the probability of a miss for IAS when $N_{\rm p} = 10$.

decision threshold $\gamma_{\rm d}$ is chosen to be the minimum threshold such that the bound on $P_{\rm F}(\gamma_{\rm d})$ is $\delta = 0.05$. We assume that $E_{\rm tot} = -20.4$ dB, which is its mean value when the transmitter–receiver (T–R) separation is 10 m [31]. We choose the power ratio $r = -4$ dB, decay constant $\epsilon = 16.1$ dB, and fading figures $m_k = 3.5 - (kT_{\rm c}/73)$, $0 \le k \le N_{\rm tap} - 1$, which are their mean values given in [31].

Figs. 2 and 3 show the effect of increasing G on the average probability of a miss $P_{\rm M}$ for SAI when the number of PRake fingers are $N_{\rm p} = 5, 10$, respectively. For each value of $N_{\rm p}$, we plot $P_{\rm M}$ for the average energy-received-per-pulse-to-noise ratio $E_1E_{\rm tot}/N_0 = 7, 10, 15, 20$ dB. When $E_1E_{\rm tot}/N_0$ is low, $P_{\rm M}$ decreases at first as G increases, and then begins to increase. When $E_1E_{\rm tot}/N_0$ is low, increasing G helps combat the effect of the noise by collecting more energy when the hypothesized phase belongs to the hit set. This is the reason for the initial decrease in $P_{\rm M}$. At the same time, the energy collected in the nonhit set phases begins to increase, resulting in a much higher threshold being chosen to ensure that $P_{\rm F}(\gamma_{\rm d})$ does not exceed δ. Consequently, the probabilities of detection decrease, causing $P_{\rm M}$ to increase. When $E_1E_{\rm tot}/N_0$ is high, the detrimental effect of the noise is not very significant, and hence, increasing G does not improve the probabilities of detection significantly. But the probabilities of detection suffer from the stringent choice of threshold required to keep $P_{\rm F}(\gamma_{\rm d})$ small. This is the reason for the increase in $P_{\rm M}$ with G for high values of $E_1E_{\rm tot}/N_0$. The values of $P_{\rm M}$ show a slight decrease for large G in the case when $N_{\rm p} = 10$, compared with the case when $N_{\rm p} = 5$. This is because the hit set is larger when $N_{\rm p} = 10$, but the probabilities of detection in the additional phases becomes significant only for large G.

Figs. 4 and 5 show the effect of increasing G on the average probability of a miss $P_{\rm M}$ for IAS when the number of PRake fingers are $N_{\rm p} = 5, 10$, respectively. For all values of $E_1E_{\rm tot}/N_0$ we considered, $P_{\rm M}$ increases with G from a value close to zero to a value close to one. Thus, performing EGC

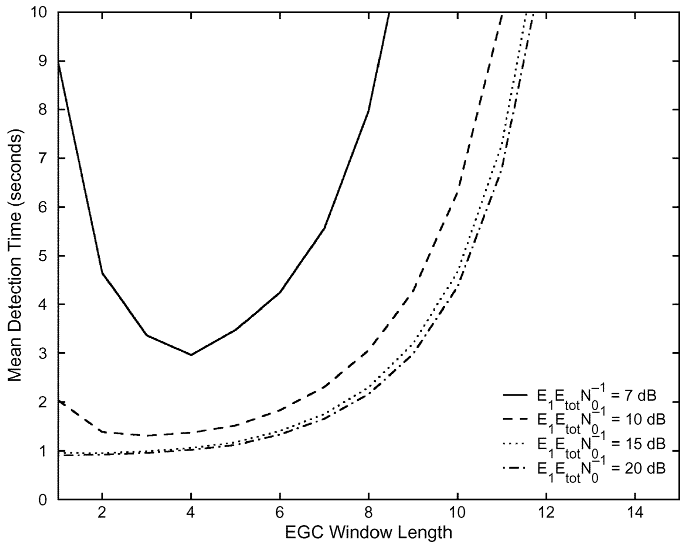

Fig. 6. Effect of EGC window length on the mean detection time for SAI when $N_\mathrm{p} = 5$.

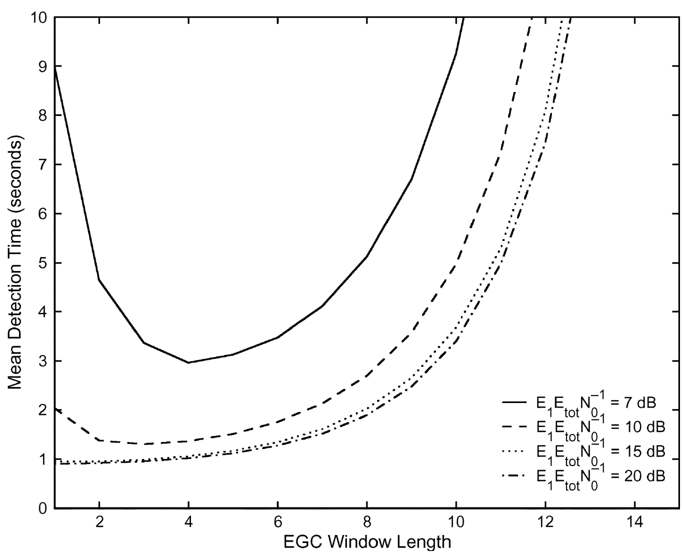

Fig. 7. Effect of EGC window length on the mean detection time for SAI when $N_\mathrm{p} = 10$.

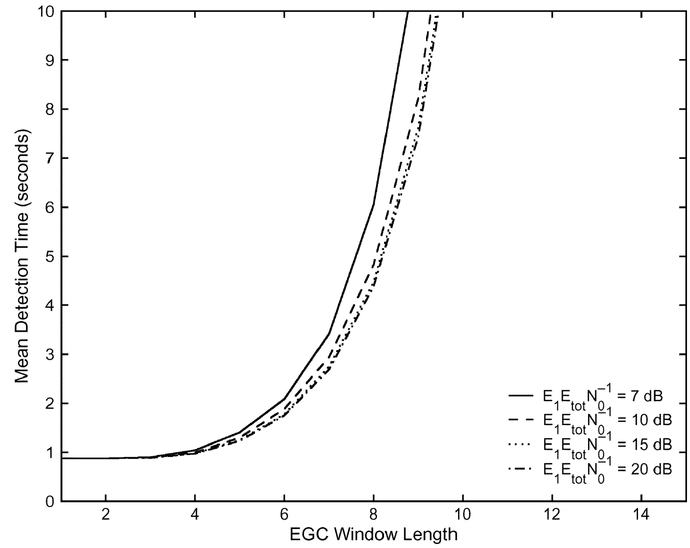

Fig. 8. Effect of EGC window length on the mean detection time for IAS when $N_\mathrm{p} = 5$.

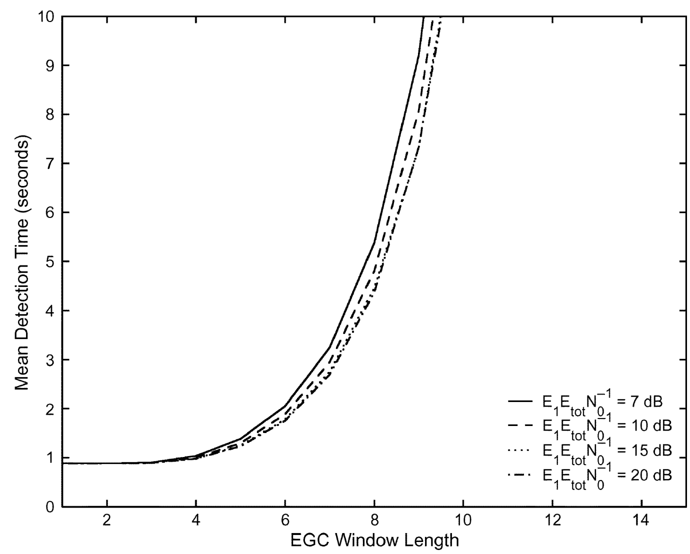

Fig. 9. Effect of EGC window length on the mean detection time for IAS when $N_\mathrm{p} = 10$.

in the IAS approach actually results in a degradation in performance. As G increases, the EGC window collects multiple paths which may have opposing polarities resulting in cancellations, and hence, a decrease in the probabilities of detection in the hit set phases. This is the reason for the increase in P_M with G. This cancellation effect is also present in the nonhit set phases, resulting in a less stringent choice for the threshold needed to keep $P_\mathrm{F}(\gamma_\mathrm{d})$ small. This effect is absent in the case of SAI, because the squaring operation eliminates the path polarities. The small values of P_M (for small G) suggest that IAS does a better job of averaging out the effect of the noise than SAI. Squaring the received signal before integrating seems to be preventing this averaging effect in SAI, resulting in higher values of P_M when E_1E_tot/N_0 is low. When E_1E_tot/N_0 is high, the less stringent threshold results in smaller values of P_M for IAS in comparison to SAI. Once again, the presence of a larger hit set for $N_\mathrm{p} = 10$ results in smaller values of P_M for IAS, in comparison with the case of $N_\mathrm{p} = 5$.

To compare the performance of the two schemes in terms of the mean detection time $\overline{T}_\mathrm{det}$, we assume that the dwell time is equal to one period of the TH sequence, i.e., $M = 1$ and $T = N_\mathrm{th}N_\mathrm{f}T_\mathrm{c}$. Figs. 6 and 7 show the mean detection time in seconds (at different values of E_1E_tot/N_0) for SAI as a function of G for $N_\mathrm{p} = 5, 10$, respectively. Figs. 8 and 9 show the corresponding plots for IAS. For both schemes, the effect of increasing G on the mean detection time mirrors its effect on P_M for the same reasons mentioned in the previous paragraph. Once again, performing EGC is beneficial in the SAI approach, and causes performance degradation in the IAS approach. For SAI, we observe that the minimum mean detection time is achieved for some value of G larger than one. This value of G changes with E_1E_tot/N_0, but is the same even when the number of PRake fingers N_p is increased from five to ten. Increasing N_p keeping the E_1E_tot/N_0 fixed increases the size of the hit set. But the probabilities of detection in the additional phases becomes significant only for large values of G where the probabilities of detection have already been lowered by the stringent choice of threshold. On the other hand, the IAS approach achieves mean detection times which are significantly lower than the corresponding values for SAI when $E_1 E_{\text{tot}}/N_0$ is low, and hence, is a more efficient scheme. Even though the probabilities of detection get better as $E_1 E_{\text{tot}}/N_0$ increases, the mean detection time does not change significantly, since it is dominated by the time the acquisition system spends in the nonhit set phases. The minimum mean detection time is seen to be of the order of a second, which is too high from a practical system viewpoint. This is due to the large search space and the fact that the serial search has to evaluate a considerable number of phases on the average before it encounters the hit set. This issue can be alleviated by a parallel search strategy.

VIII. Conclusions

We have analyzed two approaches, SAI and IAS, for the acquisition of UWB signals which perform EGC to use the energy in the multipaths. By considering system performance subsequent to acquisition, the set of phases which can be considered a hit was obtained. In the SAI approach, performing EGC improves acquisition performance at low SNRs, while it causes performance degradation in the IAS approach. With mean detection time as the metric for system performance, we observe that the IAS approach outperforms the SAI significantly. Thus, EGC may not be a good method to use the energy available in the multipaths to improve acquisition performance. Finally, the far-from-practical values of the mean detection time obtained motivate the need for a parallel search strategy and the development of acquisition schemes capable of reducing the search space.

Appendix

A. Average Number of MPCs Collected

The calculation of the expected value of $r_k(\Delta\tau)$ is made easy by the observation that it is a sum of Bernoulli distributed RVs

$$r_k(\alpha T_{\text{th}} + \beta T_c) = \frac{1}{N_{\text{th}}} \sum_{l=0}^{N_{\text{th}}-1} \sum_{i=0}^{1} \underbrace{\sum_{j=0}^{G-1} \chi(c_l + j + \beta, c_{l+i+\alpha} + k + iN_{\text{f}})}_{\text{Bernoulli RV}}.$$

Hence, the expected value is just the sum of the probabilities of the events of each Bernoulli RV taking the value one

$$\begin{aligned}
\Pr&\left[\sum_{j=0}^{G-1} \chi(c_l + j + \beta, c_{l+i+\alpha} + k + iN_{\text{f}}) = 1\right] \\
&= \Pr\left[\bigcup_{j=0}^{G-1} (c_{l+i+\alpha} + k + iN_{\text{f}} = c_l + j + \beta)\right] \\
&= \sum_{j=0}^{G-1} \Pr[c_{l+i+\alpha} + k + iN_{\text{f}} = c_l + j + \beta] \\
&= \sum_{j=0}^{G-1} \sum_{m=j+\beta}^{j+\beta+N_{\text{h}}-1} \times \Pr[c_{l+i+\alpha} + k + iN_{\text{f}} = m | c_l + j + \beta = m] \\
&\quad \times \Pr[c_l + j + \beta = m] \\
&= \sum_{j=0}^{G-1} \sum_{m=j+\beta}^{j+\beta+N_{\text{h}}-1} \frac{1}{N_{\text{h}}} \\
&\quad \times \Pr[c_{l+i+\alpha} + k + iN_{\text{f}} = m | c_l + j + \beta = m] \\
&= \sum_{j=0}^{G-1} \sum_{m=j+\beta}^{j+\beta+N_{\text{h}}-1} \frac{1}{N_{\text{h}}} \Pr[c_{l+i+\alpha} + k + iN_{\text{f}} = m] \\
&= \sum_{j=0}^{G-1} \sum_{m=\max\{j+\beta, k+iN_{\text{f}}\}}^{\min\{j+\beta+N_{\text{h}}-1, k+iN_{\text{f}}+N_{\text{h}}-1\}} \frac{1}{N_{\text{h}}^2} \\
&= \sum_{j=0}^{G-1} \sum_{m=\max\{j+\beta, k+iN_{\text{f}}\}}^{\min\{j+\beta, k+iN_{\text{f}}\}+N_{\text{h}}-1} \frac{1}{N_{\text{h}}^2}.
\end{aligned} \tag{43}$$

The sixth equality in the above calculation is due to the random sequence assumption which does not hold if $i + \alpha = 0(\text{mod} N_{\text{th}})$. If there is an $i_1 \in \{0, 1\}$ such that $i_1 + \alpha = 0(\text{mod} N_{\text{th}})$, we have

$$\Pr\left[\sum_{j=0}^{G-1} \chi(c_l + j + \beta, c_{l+i_1+\alpha} + k + i_1 N_{\text{f}}) = 1\right] = U(\beta + G - 1, k + i_1 N_{\text{f}}) U(k + i_1 N_{\text{f}}, \beta) \tag{44}$$

where $U(a, b) = 1$ if $a \geq b$ and 0, otherwise. In general, we have

$$E\left[r_k(\alpha T_{\text{f}} + \beta T_{\text{c}})\right] = U(\beta + G - 1, k + i_1 N_{\text{f}}) U(k + i_1 N_{\text{f}}, \beta) + \sum_{\substack{i=0 \\ i \neq i_1}}^{1} \sum_{j=0}^{G-1} \sum_{m=\max\{j+\beta, k+iN_{\text{f}}\}}^{\min\{j+\beta, k+iN_{\text{f}}\}+N_{\text{h}}-1} \frac{1}{N_{\text{h}}^2} \tag{45}$$

where the expectation is over the set of random TH sequences.

B. Average Probability That the Acquisition Process Will End in a False Alarm

Most of the notation used in the following has been defined in Section VI during the calculation of the mean detection time. As in the calculation of the mean detection time, we assume that the hit set consists of the phases $\{0, T_{\text{c}}, 2T_{\text{c}}, \ldots, (H-1)T_{\text{c}}\}$ in the search space $S_{\text{p}} = \{nT_{\text{c}} : n \in Z \text{ and } 0 \leq n \leq N_{\text{s}} - 1\}$. Let $P_{\text{F}}(\gamma, n)$ be the average probability that the acquisition process will end in a false alarm conditioned on the event that the serial search starts at the nth position in S_{p}. Then we have

$$P_{\text{F}}(\gamma) = \frac{1}{N_{\text{s}}} \sum_{n=1}^{N_{\text{s}}} P_{\text{F}}(\gamma, n). \tag{46}$$

First, suppose that the initial value of the hypothesized phase lies to the right of the hit set, i.e., $n \in \{H+1, H+2, \ldots, N_{\text{s}}\}$.

Let $P_{\rm f}(i)$ denote the average probability of false alarm at the ith position of the search space when $H+1 \le i \le N_{\rm s}$. Then

$$\begin{aligned}
&P_{\rm F}(\gamma,n)\\
&=1-\prod_{i=n}^{N_{\rm s}}(1-P_{\rm f}(i))+\left[\prod_{i=n}^{N_{\rm s}}(1-P_{\rm f}(i))\right]\\
&\quad\times\left[1-\prod_{i=H+1}^{N_{\rm s}}(1-P_{\rm f}(i))\right]P_{\rm M}+\left[\prod_{i=n}^{N_{\rm s}}(1-P_{\rm f}(i))\right]\\
&\quad\times\left[\prod_{i=H+1}^{N_{\rm s}}(1-P_{\rm f}(i))\right]\times\left[1-\prod_{i=H+1}^{N_{\rm s}}(1-P_{\rm f}(i))\right]P_{\rm M}^2\\
&\quad+\left[\prod_{i=n}^{N_{\rm s}}(1-P_{\rm f}(i))\right]\left[\prod_{i=H+1}^{N_{\rm s}}(1-P_{\rm f}(i))\right]^2\\
&\quad\times\left[1-\prod_{i=H+1}^{N_{\rm s}}(1-P_{\rm f}(i))\right]P_{\rm M}^3+\cdots\\
&=1-\prod_{i=n}^{N_{\rm s}}(1-P_{\rm f}(i))\\
&\quad+\frac{\left[\prod_{i=n}^{N_{\rm s}}(1-P_{\rm f}(i))\right]\left[1-\prod_{i=H+1}^{N_{\rm s}}(1-P_{\rm f}(i))\right]P_{\rm M}}{1-P_{\rm M}\prod_{i=H+1}^{N_{\rm s}}(1-P_{\rm f}(i))}\\
&=1-\left[\prod_{i=n}^{N_{\rm s}}(1-P_{\rm f}(i))\right]\left[\frac{1-P_{\rm M}}{1-P_{\rm M}\prod_{i=H+1}^{N_{\rm s}}(1-P_{\rm f}(i))}\right].
\end{aligned} \tag{47}$$

Now suppose that the initial value of the hypothesized phase falls in the hit set, i.e., $n \in \{1,2,\ldots,H\}$. Then

$$P_{\rm F}(\gamma,n)=P_{\rm F}(\gamma,H+1)\prod_{i=n}^{H}(1-P_{\rm d}(i)). \tag{48}$$

The average probability that the acquisition process will end in a false alarm is now obtained by substituting (47) and (48) in (46).

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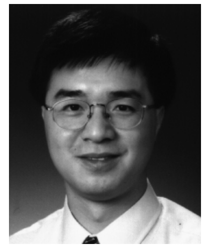

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