# Weighted Correlation Receivers for Ultra-wideband Transmitted Reference Systems

Yi-Ling Chao and Robert A. Scholtz

***Abstract*— This paper derives the average likelihood ratio test (ALRT) receiver for ultra-wideband transmitted reference systems in multipath environments with Rayleigh path strength models. A theoretical weighting function is obtained by approximating the ALRT receiver structure, and can be applied to correlation receivers. Compared to a simple correlation receiver [5], results show that theoretical weighted correlation receivers are more robust to different environments, and can reduce bit error probabilities even for IEEE 802.15.3a ultra-wideband channel models with lognormal path strength models.**

## I. Introduction

Ultra-wideband (UWB) impulse radio systems, because of their fine time-resolution capability, require Rake receivers with tens or even hundreds of correlation operations to take full advantage of the available signal energy [1], [2]. Instead of using a Rake reception, Hoctor and Tomlinson proposed an UWB transmitted reference (TR) system with a simple receiver structure to capture all the energy available in a UWB multipath channel [3]. In this TR system, a reference pulse is transmitted before each data-modulated pulse for the purpose of determining the current multipath channel response. The proposed correlation receiver correlates the data signal with the reference to use all the energy of the data signal without requiring additional channel estimation and Rake reception, and only an analog delay line is needed to align the reference and signal pulses. The sampling frequency of the analog-to-digital converter is also reduced. Therefore, a TR system can simplify the implementation not only by eliminating the Rake reception and channel estimation, but also by reducing the digital sampling frequency.

This simple receiver structure has one major drawback, namely the transmitted reference signal used as a correlator template is noisy. Averaging multiple reference pulses to produce a cleaner template can improve the receiver performance [4], [5] but it needs to store and process received waveforms and might not be possible to implement using analog devices. In order to improve bit error probability (BEP) performance and also maintain a simple receiver structure, we start by deriving the average likelihood ratio test (ALRT) receiver in multipath environments with Rayleigh path strength models. A suboptimal weighted correlation receiver, which multiplies the product of the reference and data waveforms by a weighting function before the integration, is developed from the ALRT receiver. This weighted correlation receiver obtained under the Rayleigh environment assumption is evaluated in IEEE 802.15.3a UWB environments with lognormal path strength models [6].

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## II. UWB TR System Modulation

The transmitted signal of a single user UWB TR system with antipodal modulation is

$$s_{\text{tr}}(t) = \sum_{i=-\infty}^{\infty} g_{\text{tr}}(t - iT_{\text{f}}) + b_{\lfloor i/N_{\text{s}}\rfloor} g_{\text{tr}}(t - iT_{\text{f}} - T_{\text{d}}). \quad (1)$$

Here $g_{\text{tr}}(t)$ is a transmitted monocycle waveform that is non-zero only for $t \in (0, T_{\text{w}})$, and $T_{\text{f}}$ is the frame time. Each frame contains two monocycle waveforms. The first is a reference and the second, $T_{\text{d}}$ seconds later, is a data-modulated waveform. The data bits $b_{\lfloor i/N_s \rfloor} \in \{1, -1\}$ are equally likely. The index $\lfloor i/N_{\text{s}} \rfloor$, i.e., the integer part of $i/N_{\text{s}}$, represents the index of the data bit modulating the data waveform in the $i^{\text{th}}$ frame. Hence each bit is transmitted in $N_{\text{s}}$ successive frames to achieve an adequate bit energy in the receiver, and the channel is assumed invariant over one bit time.

In this TR system, $T_{\text{d}}$ is greater than the multipath delay spread $T_{\text{mds}}$ to assure that there is no interference between reference signal and data signal. The frame time is designed to be $T_{\text{f}} \geq 2T_{\text{d}} > 2T_{\text{mds}}$ so that no interframe interference exists. Because the single user case is considered here, the time-hopping or direct sequence modulation which is used to reduce multiuser interference is eliminated for simplicity, but without loss of generality.

## III. ALRT Optimal Receiver and Weighted Correlation Receivers

### A. ALRT Optimal Receiver

Assuming the multipath channel is invariant over one bit time, the received TR signal of bit $b_0$ is modelled as

$$r(t) = r_{\text{s}}(t) + n(u, t), \quad (2)$$

$$r_{\text{s}}(t) = \sum_{i=0}^{N_{\text{s}}-1} \sum_{k=0}^{K-1} \big[p_k \alpha_k g_{\text{rx}}(t - iT_{\text{f}} - k\Delta) + p_k \alpha_k b_{\lfloor i/N_{\text{s}}\rfloor} g_{\text{rx}}(t - iT_{\text{f}} - T_{\text{d}} - k\Delta)\big] \quad (3)$$

where $n(u,t)$ represents Gaussian receiver noise with two-sided power spectral density $\frac{N_0}{2}$. This received signal adopts the tapped delay line channel model and assumes the existence of $K$ specular propagation paths. The $k^{\text{th}}$ multipath component's propagation delay is denoted by $k\Delta$, and its amplitude is denoted by $p_k\alpha_k$ in which $p_k \in \{+1,-1\}$ with equal probability, and $\alpha_k \geq 0$ has Rayleigh distribution

$$f(\alpha_k) = \frac{\alpha_k}{\sigma_k^2}\exp\left\{-\frac{\alpha_k^2}{2\sigma_k^2}\right\}, \tag{4}$$

where $2\sigma_k^2$ is the average power. It also assumes $p_k$ and $\alpha_j$ are independent for any $k$ and $j$, and any two paths are spatial uncorrelated. The received monocycle waveform $g_{\text{rx}}(t)$ arriving over a single path can differ in shape from the transmitted waveform [8]. In the design and analysis of the ALRT optimal receiver, we assume that $g_{\text{rx}}(t)$ is known and can be used as a template in a correlator.

We now detect the bit $b_0$ based on the observation $\tilde{r}$ of $r(t)$, $t \in (0, N_{\text{s}}T_{\text{f}})$. Minimizing the bit error probability using ALRT, the decision rule is of the form

$$\frac{p(\tilde{r}|b_0=1)}{p(\tilde{r}|b_0=-1)} \underset{-1}{\overset{1}{\gtrless}} 1. \tag{5}$$

Define $\boldsymbol{\alpha} = [\alpha_0, \alpha_1, ..., \alpha_{K-1}]$, $\boldsymbol{p} = [p_0, p_1, ..., p_{K-1}]$, then

$$p(\tilde{r}|b=1,\boldsymbol{p},\boldsymbol{\alpha}) \equiv \exp\left\{-\frac{1}{N_0}\int_0^{N_{\text{s}}T_{\text{f}}}[r(t)-r_{\text{s}}(t)]^2dt\right\} \tag{6}$$

$$= \prod_{k=0}^{K-1}\exp\{p_k\alpha_kC(k) - \alpha_k^2E\}, \tag{7}$$

where $\equiv$ is an equivalent symbol which ignores irrelevant constants in the right hand side of (6). Equation (7) is simplified from (6) by using the resolvable multipath assumption which is made for analytical reasons, and parameters in it are defined as

$$E = \frac{1}{N_0}\sum_{i=0}^{N_{\text{s}}-1}\int_0^{N_{\text{s}}T_{\text{f}}}[g_{\text{rx}}(t-iT_{\text{f}}-k\Delta) \pm g_{\text{rx}}(t-iT_{\text{f}}-T_{\text{d}}-k\Delta)]^2dt,$$

$$C_{\text{R}}(k) = \sum_{i=0}^{N_{\text{s}}-1}\int_0^{N_{\text{s}}T_{\text{f}}} r(t)g_{\text{rx}}(t-iT_{\text{f}}-k\Delta)dt,$$

$$C_{\text{D}}(k) = \sum_{i=0}^{N_{\text{s}}-1}\int_0^{N_{\text{s}}T_{\text{f}}} r(t)g_{\text{rx}}(t-iT_{\text{f}}-T_{\text{d}}-k\Delta)dt,$$

$$C(k) = \frac{2}{N_0}[C_{\text{R}}(k)+C_{\text{D}}(k)].$$

Using the assumption that $\boldsymbol{\alpha}$ and $\boldsymbol{p}$ are independent, we can separate the averaging process of nuisance parameters in (7) into two steps. Because of the spatial uncorrelated paths assumption which is often adopted in multipath modelling, the probability density function (pdf) of $\boldsymbol{\alpha}$ and $\boldsymbol{p}$ can be decomposed as

$$f(\boldsymbol{\alpha}) = \prod_{k=0}^{K-1} f(\alpha_k) = \prod_{k=0}^{K-1}\frac{\alpha_k}{\sigma_k^2}\exp\{-\frac{\alpha_k^2}{2\sigma_k^2}\}, \tag{8}$$

$$f(\boldsymbol{p}) = \prod_{k=0}^{K-1} f(p_k) = \prod_{k=0}^{K-1}\left[\frac{1}{2}\delta_{\text{D}}(p_k-1)+\frac{1}{2}\delta_{\text{D}}(p_k+1)\right], \tag{9}$$

where $\delta_{\text{D}}()$ is a dirac delta function. By doing some straight forward integration, the nuisance parameters $\boldsymbol{\alpha}$ can be averaged out from (7). The result is shown in (10) without computation details,

$$p(\tilde{r}|b_0=1,\boldsymbol{p}) = \int_{\infty}^{\infty} p(\tilde{r}|b_0=1,\boldsymbol{p},\boldsymbol{\alpha})f(\boldsymbol{\alpha})d\boldsymbol{\alpha}$$

$$\equiv \prod_{k=0}^{K-1}\left[\frac{w(k)}{\sigma_k^2} + \frac{p_kC(k)\sqrt{2\pi w^3(k)}}{\sigma_k^2}\exp\left(\frac{w(k)C^2(k)}{2}\right) \times Q\left(-p_kC(k)\sqrt{w(k)}\right)\right], \tag{10}$$

and $w(t)$ is defined as

$$w(k) = \frac{\sigma_k^2}{1+2\sigma_k^2E}. \tag{11}$$

It can be seen immediately that $2\sigma_k^2E$ is the average signal-energy-to-noise-power-spectral-density ratio (SNR) in the $k^{\text{th}}$ path, and $w(k)$ is always positive. Then $\boldsymbol{p}$ can be eliminated using (9) and (10), and the result is

$$p(\tilde{r}|b_0=1) \equiv \prod_{k=0}^{K-1}\exp\left(\frac{w(k)C^2(k)}{2}\right)\frac{w(k)}{\sigma_k^2}\times \left\{2\exp\left(-\frac{w(k)C^2(k)}{2}\right) + C(k)\sqrt{2\pi w(k)} \times\left[Q\left(-C(k)\sqrt{w(k)}\right) - Q\left(C(k)\sqrt{w(k)}\right)\right]\right\}. \tag{12}$$

The likelihood function $p(\tilde{r}|b_0=-1)$ can be computed using the same manipulation, and the result is

$$p(\tilde{r}|b_0=-1) \equiv \prod_{k=0}^{K-1}\exp\left(\frac{w(k)D^2(k)}{2}\right)\frac{w(k)}{\sigma_k^2}\times \left\{2\exp\left(-\frac{w(k)D^2(k)}{2}\right) + D(k)\sqrt{2\pi w(k)} \times\left[Q\left(-D(k)\sqrt{w(k)}\right) - Q\left(D(k)\sqrt{w(k)}\right)\right]\right\}, \tag{13}$$

where $D(k) = \frac{2}{N_0}[C_{\text{R}}(k) - C_{\text{D}}(k)]$. Substituting (12) and (13) into (5), taking natural logarithm $\ln()$ on both sides, and eliminating common terms, the decision rule can be simplified to

$$q(\tilde{r}|b_0=1) \underset{-1}{\overset{1}{\gtrless}} q(\tilde{r}|b_0=-1),$$

where

$$q(\tilde{r}|b_0=1) = \sum_{k=0}^{K-1}\left[\frac{w(k)C^2(k)}{2} + \ln\left\{\exp\left(-\frac{w(k)C^2(k)}{2}\right) + \sqrt{\frac{\pi w(k)C^2(k)}{2}}\left(1-2Q\left(\sqrt{w(k)C^2(k)}\right)\right)\right\}\right], \tag{14}$$

$$q(\tilde{r}|b_0=-1) = \sum_{k=0}^{K-1}\left[\frac{w(k)D^2(k)}{2} + \ln\left\{\exp\left(-\frac{w(k)D^2(k)}{2}\right) + \sqrt{\frac{\pi w(k)D^2(k)}{2}}\times\left(1-2Q\left(\sqrt{w(k)D^2(k)}\right)\right)\right\}\right].$$

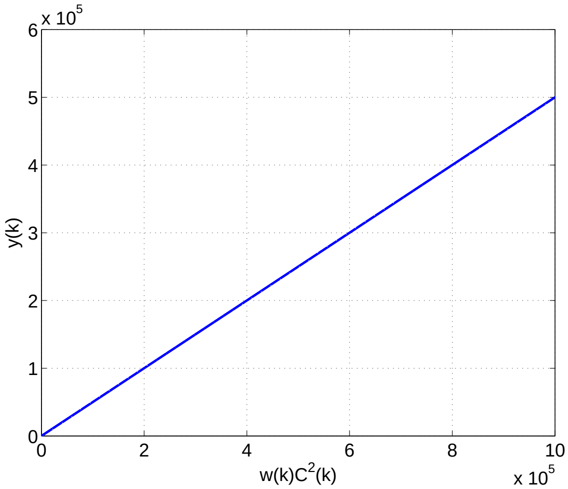


Fig. 1. $w(k)c^2(k)$ versus $y(k)$ for $k = 0, 1, \ldots, K-1$.

The decision statistic, $q(\tilde{r}|b_0 = 1)$, is a function of $w(k)C^2(k)$. Without receiver noise, $C(k)$ is equal to zero when $b_0 = -1$. Therefore, with moderate or high SNR, $C(k)$ and also $w(k)C^2(k)$ should be close to zero if $b_0 = -1$, and should have significant positive values if $b_0 = 1$. Define $y(k)$ as the term in the brackets in (14). Figure 1 shows that $y(k)$ is close to $w(k)c^2(k)/2$. This indicates the effect of the logarithm term in $q(\tilde{r}|b_0 = 1)$ is small. The same explanation can apply to $q(\tilde{r}|b_0 = -1)$. Therefore, abandoning the logarithm term is a reasonable way to simplify the receiver structure, and the sub-optimal receiver is

$$\sum_{k=0}^{K-1} w(k)C^2(k) \underset{-1}{\overset{1}{\gtrless}} \sum_{k=0}^{K-1} w(k)D^2(k). \tag{15}$$

Expanding $C^2(k)$ and $D^2(k)$, and eliminating common terms on both sides, the decision rule in (15) is equivalent to

$$\sum_{k=0}^{K-1} w(k)C_{\mathrm{R}}(k)C_{\mathrm{D}}(k) \underset{-1}{\overset{1}{\gtrless}} 0. \tag{16}$$

The meaning of (16) can be explained as follows. The transmitted bit is embedded in the phase difference of the reference waveform and data waveform. Therefore, without receiver noise, the polarity of $C_{\mathrm{R}}(k)C_{\mathrm{D}}(k)$ for all $k$ should be the same and is the transmitted bit. It is possible that we have detection errors by using $C_{\mathrm{R}}(k)C_{\mathrm{D}}(k)$ as a decision statistic because of the receiver noise, but the error probability decreases as the SNR increases. Therefore, we should give $C_{\mathrm{R}}(k)C_{\mathrm{D}}(k)$ different weights for different $k$ according to both the a priori information and received information, i.e., the average SNR and sample SNR in the $k^{\text{th}}$ path. For the suboptimal detection in (16), we discard the information of sample SNR, i.e., logarithm terms in $q(\tilde{r}|b_0 = 1)$ and $q(\tilde{r}|b_0 = -1)$, in order to reduce the receiver complexity. We only generate weights for $\{C_{\mathrm{R}}(k)C_{\mathrm{D}}(k)\}_{k=0}^{K-1}$ according to the a priori knowledge, and the weights are $\{w(k)\}_{k=0}^{K-1}$. For a specified value of $E$, $w(k)$ is an increasing function of $\sigma_k^2$, and $C_{\mathrm{R}}(k)C_{\mathrm{D}}(k)$ gets a large weight if the average power of the $k^{\text{th}}$ multipath component is large.

### B. *Weighted Correlation Receivers*

If we ignore the weights in (16), the decision rule becomes

$$\sum_{k=0}^{K-1} C_{\mathrm{R}}(k)C_{\mathrm{D}}(k) \underset{-1}{\overset{1}{\gtrless}} 0, \tag{17}$$

which is recognized as the generalized likelihood ratio test for UWB TR systems [5]. Comparing (17) and the decision rule of a simple correlation receiver [5] which is rewritten in (18)

$$\sum_{j=0}^{N_{\mathrm{s}}-1} \int_{jT_{\mathrm{f}}+T_{\mathrm{d}}}^{jT_{\mathrm{f}}+T_{\mathrm{d}}+T_{\mathrm{mds}}} r(t-T_{\mathrm{d}})r(t)dt \underset{-1}{\overset{1}{\gtrless}} 0 \tag{18}$$

in which $r(t-T_{\mathrm{d}})$ and $r(t)$ represent the received reference and data waveforms, some similarities can be seen between them. The integration, $r(t-T_{\mathrm{d}})$, and $r(t)$ in (18) correspond to the summation, $C_{\mathrm{R}}(k)$, and $C_{\mathrm{D}}(k)$ in (17), and the summation in (18) is imbedded in $C_{\mathrm{R}}(k)$ and $C_{\mathrm{D}}(k)$. Therefore we can also multiply a weighting function $w(t)$ , which is a continuous function interpolated from $\{w(k)\}_{k=0}^{K-1}$, to $r(t-T_{\mathrm{d}})r(t)$ in (18) before the integration to provide the a priori information in the decision. Then the decision statistic of the weighted correlation receiver becomes

$$D_{\mathrm{s}}(u) = \sum_{j=0}^{N_{\mathrm{s}}-1} \int_{jT_{\mathrm{f}}+T_{\mathrm{d}}}^{jT_{\mathrm{f}}+T_{\mathrm{d}}+T_{\mathrm{mds}}} r(t-T_{\mathrm{d}})r(t)w(t-jT_{\mathrm{f}}-T_{\mathrm{d}})dt, \tag{19}$$

and we say the transmitted bit is 1 if $D_{\mathrm{s}}(u) > 0$, otherwise $-1$. This is called a theoretical weighted correlation receiver, and $w(t)$ is a theoretical weighting function (WF). The derivation of the BEP is eliminated because of the space limitation. Given a channel impulse response, the BEP is

$$P_{\text{bit}} \cong Q\left(\frac{\sqrt{N_{\mathrm{s}}}\int_0^{T_{\mathrm{mds}}} w(t)g^2(t)dt}{\sqrt{N_0\int_0^{T_{\mathrm{mds}}} w^2(t)g^2(t)dt + \frac{BN_0^2}{2}\int_0^{T_{\mathrm{mds}}} w^2(t)dt}}\right) \tag{20}$$

where $g(t)$ is the convolution of the transmitted pulse and the channel impulse response, and $B$ is the one-sided receiver bandwidth.

In channel modelling, multipath channels are often assumed having an exponential average power profile, i.e., $2\sigma_k^2 = \Omega\exp(\frac{-k\Delta}{\Gamma})$ where $\Omega$ is the average power in the first path and $\Gamma$ is the time constant. This assumption is reasonable even for channel models with clusters like IEEE 802.15.3a which can be observed from the figures in [6]. Use this exponential average power assumption,

$$\begin{aligned} w(k) &= \frac{\sigma_k^2}{1+2\left(\sum_{k=0}^{K-1}\sigma_k^2 E\right)\times\frac{\sigma_k^2}{\sum_{k=0}^{K-1}\sigma_k^2}} \\ &= \frac{\frac{\Omega}{2}\exp(\frac{-k\Delta}{\Gamma})}{1+(\text{A-SNR})\times\frac{\exp(\frac{-k\Delta}{\Gamma})}{\sum_{k=0}^{K-1}\exp(\frac{-k\Delta}{\Gamma})}}, \end{aligned} \tag{21}$$

where A-SNR denotes the average signal-energy-to-noise-power-spectral-density ratio of the received reference/data

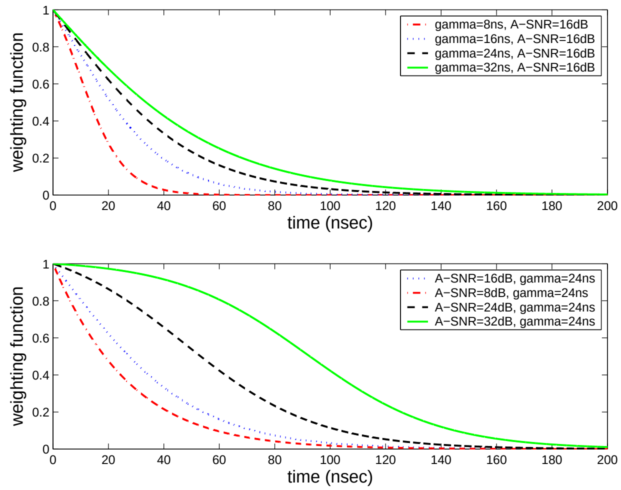


Fig. 2. Weighting functions with different values of A-SNR and Γ.

waveform. The theoretical WF $w(t)$ for the correlation receiver can be obtained from (21) by interpolation and is expressed as

$$w(t) = \frac{\exp(\frac{-t}{\Gamma})}{1 + (\text{A-SNR}) \times \frac{\exp(\frac{-t}{\Gamma})}{\sum_{k=0}^{K-1}\exp(\frac{-k\Delta}{\Gamma})}}. \quad (22)$$

It is worth noting that $w(t)$ can be multiplied by any constant without affecting the BEP. Figure 2 shows the shape of $w(t)$ with different values of A-SNR and Γ. We can see that increasing Γ lengthens the tail of $w(t)$, while increasing A-SNR increases the height.

The theoretical weighting function $w(t)$ is not the only choice for a weighting function. If we inspect (18) in details, the length of the integration interval can also affect the BEP performance. The integration time, which is defined as $T_{\text{corr}}$ here, should follow the constraint that $T_{\text{corr}} \leq T_{\text{mds}}$. Under the constraint, increasing $T_{\text{corr}}$ can increase the incoming signal energy, but we also get more noise power through the noise cross noise term. Since the tail response of the channel is usually small, it might not be good to use $T_{\text{corr}} = T_{\text{mds}}$ like in (18). We should adjust $T_{\text{corr}}$ according to the channel response to minimize the BEP, and this adjustment is equivalent to applying a square weighting function

$$w_{\text{s}}(t) = \begin{cases} 1, & 0 \leq t \leq T_{\text{corr}} \\ 0, & \text{otherwise.} \end{cases}$$

The decision rule in (18) is modified to

$$\sum_{j=0}^{N_{\text{s}}-1} \int_{jT_{\text{f}}+T_{\text{d}}}^{jT_{\text{f}}+T_{\text{d}}+T_{\text{mds}}} r(t-T_{\text{d}})r(t)w_{\text{s}}(t-jT_{\text{f}}-T_{\text{d}})dt \underset{-1}{\overset{1}{\gtrless}} 0, \quad (23)$$

and the BEP given a channel response is

$$P_{\text{bit}} \cong Q\left(\left[\frac{2}{N_{\text{s}}}\left(\frac{N_0}{\eta E_{\text{f}}}\right) + \frac{2BT_{\text{corr}}}{N_{\text{s}}}\left(\frac{N_0}{\eta E_{\text{f}}}\right)^2\right]^{-\frac{1}{2}}\right) \quad (24)$$

where $E_{\text{f}} = 2\int_0^{T_{\text{mds}}} g^2(t)dt$ is the received signal energy in one frame, and $\eta = \int_0^{T_{\text{corr}}} g^2(t)dt / \int_0^{T_{\text{mds}}} g^2(t)dt$ is the efficiency factor.

## IV. Numerical Results

Two parameters decide the shape of a theoretical weighting function $w(t)$. One is A-SNR, and the other one is Γ which is the time constant of the exponential average power profile of channels. Theoretical weighted correlation receiver is tested on the IEEE 802.15.3a UWB channel model cm1, cm2, cm3, and cm4 [6], which have lognormal path strength models and poisson arrival clusters and rays, representing four different environments. The single received pulse $g_{\text{rx}}(t)$ used in simulations is a second derivative Gaussian pulse with duration 0.7 nsec, the one-sided receiver bandwidth $B = 4$GHz, $\Delta = 0.7$ nsec, and $N_{\text{s}} = 1$. In this section, (20), (24), and 1000 equal power channel realizations for each channel model in IEEE 802.15.3a are used to obtain the average BEPs. Although paths arrive in clusters in IEEE 802.15.3a, equivalent channels models without clusters and with exponential average power profiles exist for analytical reasons [7]. The parameter A-SNR can be predicted from the link budget computation. Given A-SNRs, equivalent Γs for four environments are acquired by minimized average BEPs (which are 1e-4) at those A-SNRs, and listed in Table I. For $w_{\text{s}}(t)$, we can obtain the optimal $T_{\text{corr}}$s which make the correlation receiver achieves the average BEP=1e-4 with minimum $E_{\text{b}}/N_0$, and list those values in Table I. Using those optimal parameter values in Table I for $w(t)$ and $w_{\text{s}}(t)$, BEP curves in four environments are plotted in Figure 3. Curves show BEPs depend on application environments, and using theoretical weighting function outperforms using square weighting function. This also means applying a theoretical weighting function to a simple correlation receiver can improve the BEP performance.

| | CM1 | CM2 | CM3 | CM4 |
|---|---|---|---|---|
| theoretical WF, $w(t)$ | | | | |
| A-SNR (dB) | 17.80 | 18.45 | 19.35 | 20.40 |
| Γ (nsec) | 7.5 | 10.0 | 20.5 | 39.5 |
| square WF, $w_{\text{s}}(t)$ | | | | |
| Tcorr (nsec) | 20 | 25 | 50 | 75 |

TABLE I

Optimal parameter values of theoretical and square weighting functions at BEP=1e-4.

Figure 4 shows the performance degradation of a theoretical weighted correlation receiver in four environments if the value of A-SNR or Γ is not optimized. The x-axis is A-SNR (upper figure) or Γ (lower figure), and the y-axis is the required $E_{\text{b}}/N_0$ to achieve average BEP=1e-4. The circle on each line marks the position of the optimal value of A-SNR or Γ. Results show the receiver is not sensitive to the A-SNR deviation. As for Γ, underestimation degrades the performance more than overestimation. Since Γ is related to application environments, we should consider large values of Γ if the parameter is not adjustable in the receiver.

Figure 5 compares the performance degradation in cm1 (upper figure) and cm4 (lower figure) environments using theoretical and square weighting functions while parameters are not optimized. Because $w(t)$ and $w_{\text{s}}(t)$ have different parameters,

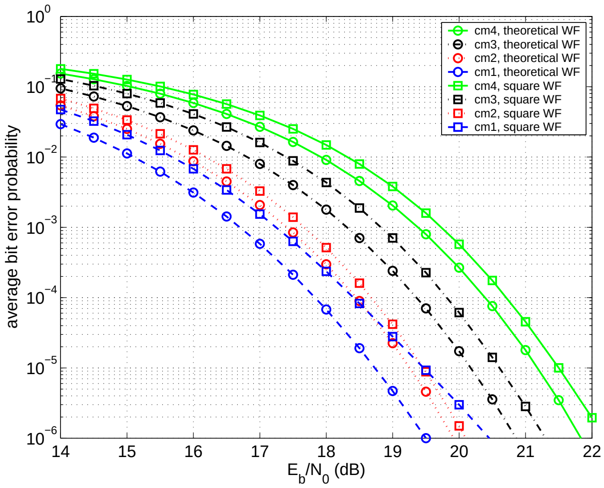


Fig. 3. BEP performance using theoretical and square weighting functions for cm1, cm2, cm3, and cm4.

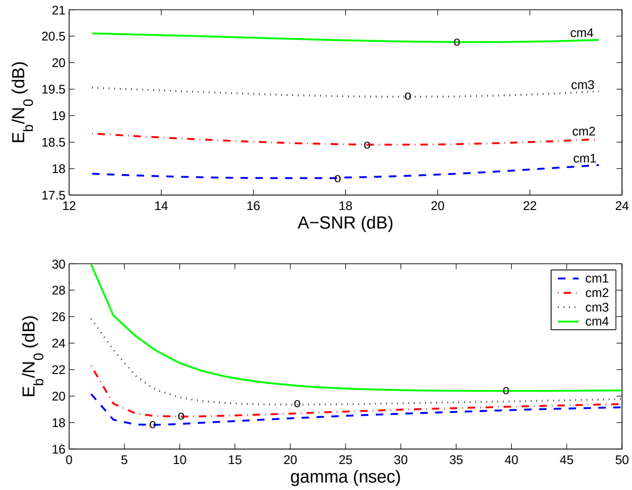


Fig. 4. $E_b/N_0$ required to achieve BEP=1e-4 for different values of A-SNR and Γ.

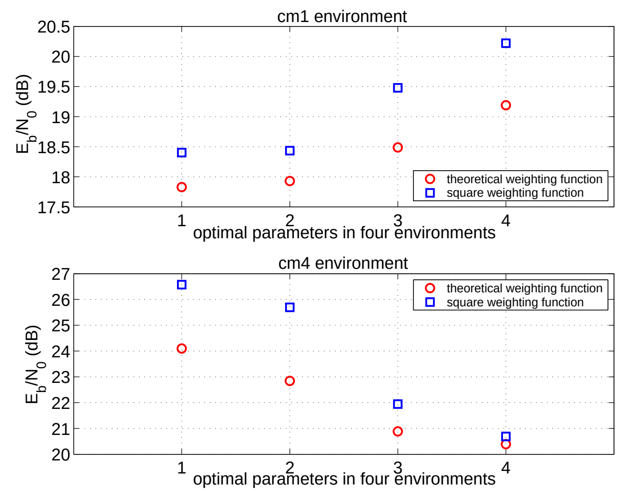


Fig. 5. Performance degradation of theoretical and square weighting functions by using non-optimal parameter values.

we have to find a fair way to compare the degradation. Table I shows the optimal values of A-SNR and $\Gamma$ for $w(t)$, and the optimal values of $T_{\text{corr}}$ for $w_s(t)$ in cm1, cm2, cm3, and cm4 environments. For the theoretical weighting function, 1, 2, 3, and 4 at the x-axis in Figure 5 represent that (A-SNR, $\Gamma$)=(17.80, 7.5), (18.45, 10.0), (19.35, 20.5), and (20.40, 39.5) which are the optimal values of A-SNR and $\Gamma$ in environments cm1, cm2, cm3, and cm4. For the square weighting function, 1, 2, 3, and 4 at the x-axis in Figure 5 represent that $T_{\text{corr}}$=20, 25, 50, and 75 which are the optimal values of $T_{\text{corr}}$ in environments cm1, cm2, cm3, and cm4. The y-axis in Figure 5 is the required $E_b/N_0$ to achieve average BEP=1e-4. Under this definition of the x-axis, the comparison is fair. In the upper figure, we can see the best performance happens while x=1 because parameters in both weighting functions are optimized for cm1 environment. For the same reason, the best performance happens while x=4 in the lower figure. Both the upper and lower figures show the performance degradation by using a theoretical weighting function is smaller than using a square weighting function when the values of parameters are not optimized. Using theoretical weighting functions makes the receiver more flexible in different environments.

## V. Conclusion

This paper derives ALRT and suboptimal receivers for single user UWB TR system in Rayleigh environments. A theoretical weighting function is obtained from the suboptimal receiver, and applied to the simple correlation receiver. Besides this theoretical weighting function, applying a square weighting function to a simple correlation receiver is also examined. Although the theoretical weighting function is derived with a Rayleigh path strength assumption, results show it can improve the receiver performance in lognormal environments as well. The most important thing is receivers using a theoretical weighting function is more flexible in different environments compared to using a square weighting function.

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