Performance analysis of three kinds of non-coherent detectors for ultra-wideband Communications

Jian Pan*, Sheng Zhang, Benzhou Jin, Xiaokang Lin
Department of Electronic Engineering, Tsinghua University
Beijing, China
*Email: jianlonger@foxmail.com

Abstract—This paper discusses the basic signal detection scheme and error performances of three kind receiver architecture, i.e., the peak detector, the square-law detector and the direct-integral receiver. The theoretical bit error probabilities are derived through Gaussian approximation that is well satisfied in UWB systems. Simulations are performed in the additive white Gaussian (AWGN) channel model and results shows that the theoretical curve derived from the analytical model is in accordance with the experimental curve. This paper indicates that the bit error rate (BER) performance of direct-integral detectors is slightly better than energy detectors (EDs) when E_b/N_0 is small (less than 6.5db). Besides, this paper indicates that, compare with coherent detectors, Direct-integral detector has a potential to achieve an appropriate tradeoff between error rate performance and system complexity.

Keywords- Direct-integral receiver; quare-law detector; Trigger; ultra-wideband.

I. INTRODUCTION

Former UWB system designers mostly focused on using coherent architectures, typically Rake schemes, at the receiver end in order to achieve high data rate and good error rate performance. However, it has been claimed that the optimum receivers face many challenges in practical implementation due to UWB channel characteristics and signal spectral densities constrains[1][2].

The energy-detection-based UWB receiver architecture known as EDs has been attracting much attention recently because of its good performance on low-complexity, low-power and low cost solutions for practical applications such as UWB wireless sensor networks. The ED receiver is consisted by an square-law detector which measures the energy in the received waveform over an observation time window, though this kind receiver has much merit ,when we focus on low SNR(signal noise rate) condition, its performance is no better than an simple receiver we proposed which can be defined as direct-integral receiver. A direct-integral receiver is formed by a linear detector which measures the integral of received waveform over an observation time window. In this paper, we only discuss the performance of these receivers over no-fading, no-multipath channel, because in short distance (0.1-1m) communicating, this assume is reasonable.

Our contribution in this paper is twofold. First, we offer the detailed deduce from the theoretical analysis and obtain

expressions for the error probability of detection over AWGN channel. Second, and more importantly, the influence of decreasing observation time duration is discussed. The simulations show that, the ED receiver is inferior to direct-integral receiver in low SNR and superior to direct-integral receiver in high SNR.

The paper is organized as follows: The System models for the two receiver are described in Section II; Section III is devoted to deduce the expression of error performance for Direct-integral receiver; In Section IV, the BER performance for ED Receiver is discussed; In Section V, simulation results under different conditions are presented; Finally, comparisons are demonstrated and analyzed in Section VI, and Conclusions are offered in Section VII.

II. SYSTEM MODELS

While our analysis applies to either low-pass (LP) or band pass (BP) systems, we here focus on the LP representation. The received signal at the receiver end can be represented as

$$r_0(t) = a_i \bullet s_L(t) + n(t) \quad a_i \in \{0,1\} \quad (1)$$

Here, a_i is decided by the sending symbol '1' and '0'. $s_L(t)$ is the baseband signal, $n(t)$ is an AWGN process with a zero mean and a known flat power spectral density (PSD). Let N_0 denote the one-sided noise PSD and B the signal bandwidth (i.e., positive spectrum support).

Fig.1. shows the proposed direct-integral receiver architecture. The received signal is first prefiltered by an ideal LP filter. Then, the output of this filter is integrated over a time interval T_c to finally produce a measure of the electric quantity of the received waveform. Fig.2. shows the peak detector, which employ the peak of received waveform to trigger the sampler and utilize this value for symbol decision. Fig.3. refers to the ED receiver.

Fig.1 Architecture of Direct-integral receiver

Fig. 2 Architecture of peak detector

Fig. 3 Architecture of Direct-integral receiver

III. DETECTION AND ERROR RATE PROBABILITIES OVER AWGN CHANNELS FOR THREE KIND RECEIVER

A. BER for Direct-integral receiver

As shown in Fig.1. $r_1(t)$ denotes the demodulated signal of $r_0(t)$, and as the figure shows, the bandwidth of LP is identical to the efficient bandwidth of signal $s_L(t)$, hence the filter outputs the whole signal while strains the noise beyond its passband. Thus the noise output is $n_L(t)$, which is a gauss process with Zero mean its mean 0 and variance of BN_0 .

The autocorrelation of $n_L(t)$ can be expressed as

$$R_n(\tau) = \frac{N_0}{2} \bullet \frac{\sin(\pi B\tau)}{\pi} = N_0 B \bullet \text{sinc}(\pi B\tau) \quad (2)$$

The output of integrator denoted by y acts as a test statistic to test the two hypotheses H_1 (when sending symbol '1') and H_0 (when sending symbol '0'). Thus y can be expressed as

$$y = \int_0^{T_c} a_i \bullet s_{L(t)} + n_{L(t)} dt \quad (3)$$

$$m_0 = E(y/H_0) = 0 \quad (4)$$

$$m_1 = E(y/H_1) = \int_0^{T_c} s_{L(t)} dt \approx \Delta \sum_{i=1}^M s_L(i\Delta) \quad (5)$$

$$\begin{aligned} \sigma^2 &= \text{var}(y/H_1) = \int_0^{T_c} \int_0^{T_c} E[n_{L(t_1)} n_{L(t_2)}] dt_1 dt_2 \\ &= \int_0^{T_c} \int_0^{T_c} R_n(t_1 - t_2) dt_1 dt_2 = \frac{N_0}{2} \int_0^{T_c} \int_0^{T_c} \frac{\sin \pi B(t_1 - t_2)}{\pi(t_1 - t_2)} dt_1 dt_2 \\ &= N_0 B \int_0^{T_c} \int_0^{T_c} \text{sinc}[\pi B(t_1 - t_2)] dt_1 dt_2 \end{aligned} \quad (6)$$

Where M is the number of samples in per code pulse (we finally have terms to sum over), $T_s = T_c/M$ is the sample interval. Then under H_1 , y follows a non-central gauss distribution with variance σ^2 , and m_1 as its mean. Now, the probability density function (PDF) of y can be expressed as

$$f_Y(y) = \begin{cases} \frac{1}{2\pi\sigma} \exp(-\frac{y^2}{2\sigma^2}) & H_0 \\ \frac{1}{2\pi\sigma} \exp[-\frac{(y-m_1)^2}{2\sigma^2}] & H_1 \end{cases} \quad (7)$$

Under the Gaussian approximation, error performances for OOK and 2-PPM systems can be derived since the mean and variance of Y is known.

In OOK, there are two frame errors:

- i) Miss alarm error: It means no pulse is detected when a pulse frame is transmitted.
- ii) False alarm error: This happens when an empty frame is transmitted but a pulse is detected.

Using the statistical characteristics of Y , we have

$$P_{1|0} = \frac{1}{\sqrt{2\pi}\sigma} \int_V^\infty e^{-x^2/2\sigma^2} dx \quad (8)$$

$$P_{0|1} = \frac{1}{\sqrt{2\pi}\sigma} \int_{-\infty}^V e^{-(x-m_1)^2/2\sigma^2} dx \quad (9)$$

Where, $P_{1|0}$ and $P_{0|1}$ represent the false alarm error probability and the Leak-detecting error probability respectively. m_1 , see expression (5), is the mean of Y when the binary symbol a_i , and σ^2 is the variance of Y . In order to obtain the best error performance, the optimum threshold V can be solved as $V = m_1/2$.

Then, the bit error probability for OOK is

$$\begin{aligned} P_{e1_OOK} &= \frac{1}{2} P_{1|0} + \frac{1}{2} P_{0|1} = Q\left(\frac{m_1 - V}{\sigma}\right) \\ &= Q\left(\frac{m_1}{2\sigma}\right) \end{aligned} \quad (10)$$

Since a maximum likelihood decoder can be adopted in 2-PPM systems, it requires no threshold.

$$P_{e1_PPM} = p\{m_1 + n_1 < n_2\} \quad (11)$$

Where, m_1 is the same as in equation (5). n_1, n_2 is given as follow

$$n_1 = \int_0^{T_c} n_{L(t)} dt \quad n_2 = \int_{T_c}^{2T_c} n_{L(t)} dt \quad (12)$$

$$\begin{aligned} \text{cov}(n_1, n_2) &= E[n_1 n_2] = E\left[\int_{T_c}^{2T_c} \int_0^{T_c} n_{L(t_1)} n_{L(t_2)} dt_1 dt_2\right] \\ &= \int_{T_c}^{2T_c} \int_0^{T_c} R_n(t_1 - t_2) dt_1 dt_2 = N_0 B \int_{T_c}^{2T_c} \int_0^{T_c} \text{sinc}[\pi B(t_1 - t_2)] dt_1 dt_2 \end{aligned} \quad (13)$$

The bit error probability can be easily derived as

$$\begin{aligned}
P_{e1_PPM} &= p\{m_1 + n_1 < n_2\} \\
&= Q\left(\frac{m_1}{\sqrt{2\sigma^2 - 2\text{cov}(n_1, n_2)}}\right)
\end{aligned} \quad (14)$$

B. BER for peak detector

According to the architecture of peak detector, its bit error probability can be easily derived as

$$P_{e2_PPM} = Q\left(\frac{A}{\sqrt{2N_0B \bullet [1 - \text{sinc}(\pi BT_c)]}}\right) \quad (15)$$

Where, A is the amplitude value of the peak point of the received waveform. That is

$$A = \max_{t \in T} \{r_1(t)\} = \max_{t \in T} \{s(t) + n_L(t)\} \quad (16)$$

C. BER Approximation for ED receiver

The architecture of square-law receiver (the ED receiver) is different from the Direct-integral-receiver and peak detector, for its nonlinear character, it is difficult to acquire analytic solution.

However, by using Shannon's sampling theorem, one can show that the energy in a finite time interval can be described as a sum of the square of a number of statistically independent Gaussian variables if the noise input is Gaussian and has a flat spectral density over a limited bandwidth [3].

Under this architecture, our analysis is obtained from [4], according to the Fig.3. y can be expressed as

$$y = \int_0^{T_s} (a_i s_i(t) + n_{L_i}(t))^2 dt \approx \frac{1}{T_s} \sum_{k=1}^M (a_i s_k + n_k)^2 \quad (17)$$

Where, M is the number of samples per component (we finally have M terms to sum over), and generally s_k and n_k respectively, denote the k -th samples of $s(t)$ and $n_L(t)$.

It then follows that y satisfies chi-square distribution, with its variance $\sigma_k^2 = 1$, non-centrality parameter $\mu = 2\alpha$, $\alpha = E_b/N_0$ and M degrees of freedom (DOF), E_b is the energy of signal $s(t)$. Likewise, for a given $s(t)$, y will be central chi-square distributed. Now, the probability density function (PDF) of y can be expressed as

$$f_Y(y) = \begin{cases} \frac{y^{M/2-1}}{\sigma_k^M 2^{M/2} \Gamma(M/2)} e^{-\frac{y}{2\sigma_k^2}} & a_i = 0 \\ \frac{1}{2\sigma_k^2} \left(\frac{y}{2\alpha}\right)^{\frac{M-2}{4}} e^{-\frac{2y+y}{2\sigma_k^2}} I_{\frac{M}{2}-1}\left(\frac{\sqrt{2\alpha y}}{\sigma_k^2}\right) & a_i = 1 \end{cases} \quad (18)$$

Where $\Gamma(\bullet)$ is the gamma function and $I_\nu(\bullet)$ is the ν -th order modified Bessel function of the first kind.

An approximate expression for the probability of bit error

over AWGN channels was presented in [3]. Based on the statistics of y , when V is the decision threshold, assume the symbol '0,'1' are sent in equal probability, then the error probability p_e can be evaluated as

$$p_e = \frac{\Gamma\left(\frac{M}{2}, \frac{V}{2\sigma_k^2}\right)}{2\Gamma\left(\frac{M}{2}\right)} + \frac{1}{2} Q_{M/2}\left(\sqrt{\frac{2\alpha}{\sigma_k^2}}, \sqrt{\frac{V}{\sigma_k^2}}\right) \quad (19)$$

where $\Gamma(\bullet, \bullet)$ is the incomplete gamma function. $Q_{M/2}(\bullet, \bullet)$ is the generalized Marcum-function [5].

IV. SIMULATIONS

Theoretical curves of direct-integral receiver, peak detector and correlation-receiver and theirs simulation results over the AWGN channel are given in Fig.4. And Fig.5 gives the BER performance comparison between direct-integral receiver and ED, and theirs performance when the time integral interval decreased. In both figures, we adopt the Gaussian pulse with a bandwidth B of 1GHz as the received signal forms in the simulation.

Fig.4. shows that the simulation results is well agree with the theoretical curves, compare with peak detector the BER performance of Direct-integral receiver is better with a power advantage which is approach to 2.45dB with respect to E_b/N_0 at BER of 10^{-2} . Compare with a correlation receiver, direct-integral receiver has approximately a power weakness of 3.6dB when BER is at 10^{-3} .

It is of interest to compare a direct-integral receiver with a square-law receiver. This comparison is shown in Fig.5 square-law receiver has approximately a power advantage of 3dB when BER is at 10^{-3} . But under small E_b/N_0 (less than 6.5db), the BER performance of direct-integral receiver is better. Besides, in this simulation for both square-law receiver and direct-integral receiver theirs BER performance deteriorate significantly when the integral time duration is halved.

Fig. 4 Bit error rates for theoretical and simulation of three kind detectors

Fig.5 Bit error rates when integral window changed

V. DISCUSSION AND ANALYSIS

On the one hand, if the form of a signal to be detected is unknown, it appears appropriate to consider an energy detector as a device for deciding whether or not the signal is present. Since an energy detector does not care about any-thing but the amount of energy in the given observation time, the form of the signal does not affect the conditional probability that a threshold will be exceeded when the signal is present. However, the square-law also increases the effect of noise accumulation, thus the SNR output of the decision variable not increases in the same pace.

The direct-integral receiver has its own merits, first it has better error rate performance than square-law receiver in small E_b/N_0 , its integral module can offset some influence brought by noise while accumulating the effect of signal we wanted, for the integral module has the effect of summarizing, as long as the noise is zero mean Gaussian process, when summed up it remains zero mean Gaussian variable, thus, the SNR output of the decision variable of receiver is little higher compare with the square-law case. Second, direct-integral receiver detectors have different features from the EDs such as the piecewise linear property and the utilization of the integration of signal amplitude for signal detection. Those characteristics deserve further investigation of new signal processing techniques which make it potential for Direct-integral receiver to achieve an appropriate tradeoff between error performance and system complexity compared with EDs.

On the other hand, the BER performance also depends on the integral time duration T_c the longer T_c is, the better the BER performance is. This is another way of stating the obvious conclusion that the observation time should be as long as possible, for beyond the observation window the SNR of received waveform can not be neglected. Meanwhile

A none-negative signal is better than a signal with both positive and negative peak.

VI. CONCLUSIONS

The theoretical error performance of Direct-integral receiver and peak detector are studied in this paper. It shows that simulation results in accordance with the theoretical curves. Compared with the ED, the Direct-integral receiver has a little power advantage when E_b/N_0 is small (less than 6.5db), whereas the square-law receiver has approximately a power advantage of 3dB when BER is at 10^{-3} . With the time duration of received signal shortened, the BER performance of the detector deteriorated greatly.

In this study, we only focus on the basic detection architecture and signal processing of direct-integral receiver and peak detector. Future works need to be done considering more factors.

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