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Ultra-Wideband Communication using Hybrid Matched Filter Correlation Receivers

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***Abstract*— For ultra-wideband (UWB) communication, Rake receivers using matched filter detection show the best performance but they are complex to realize due to the inherent channel estimation problem and combining problem. Transmitted reference schemes, on the other hand, allow for a less complex receiver structure that is able to combine the energy from different multipath components without channel estimation. These schemes, however, show a performance loss due to non-linear operations on noise terms (generation of noise-noise crossterms) when forming the decision variable. This paper describes a receiver structure for UWB communication capable of capturing the energy from different paths without the need for channel estimation. The proposed hybrid matched filter correlation receivers reduce the performance loss, due to the noise-times-noise term.**

***Index Terms*—UWB, Impulse Radio, Transmitted Reference**

I. Introduction

Ultra-wideband (UWB) communications systems are defined as systems that have either a relative bandwidth of more than 20%, or an absolute bandwidth of more than 500MHz. The report and order [1] of the Federal Communications Commission (FCC) in the US, allowing unlicensed operation of UWB communications, has greatly increased the interest in these systems, especially in the design of low-cost transceivers for UWB communications. UWB communications have traditionally been associated with Impulse Radio (IR) [2], [3], which is well suited for low-data-rate communications, and most of the academic work in UWB of the last 10 years has concentrated on IR. The recent formation of the IEEE 802.15.4a group, which will establish a standard for such low-data-rate communications, has also created considerable commercial interest for such systems.

Impulse radio transmits information by modulating the amplitude or position of very short (on the order of 100 ps to 2 ns) pulses. In order to permit several users to communicate simultaneously, a bit is not represented by a single pulse, but rather a (pseudorandom) *sequence* of pulses, where each user is assigned a different sequence. The symbol duration is divided into N_f intervals called "frames", each of which contains one pulse. The location of the pulse within the frame is determined by a pseudorandom sequence [4]. This makes sure that there can be no "catastrophic collisions" between two simultaneously transmitting users, where the signals from those two users would completely overlap at a receiver.

Due to their wide bandwidth, impulse radio receivers can resolve many multipath components in the received signal, and have to add them up in order to "collect" all the received energy [5]. The optimum scheme to achieve this is a Rake receiver, combined with a receive filter matched to the transmit pulse [3], [6]. However, the Rake requires a fairly complicated receiver structure, with one despreader (correlator, Rake finger) for each multipath component to be received. Therefore often only the strongest, or a few of the strongest paths are used to form the decision variable. This, of course, means that the receiver does not collect all multipath components and therefore there is a performance loss compared to the ideal case. Furthermore, a Rake receiver needs to estimate the channel impulse response in order to obtain the correct Rake weights.

All these issues have led to an increased interest in so-called "transmitted reference" (TR) schemes [7], [8], [9], [10]. In a TR scheme, channel estimation and despreading is done in one simple step. Two transmitted pulses are used in each frame. The first pulse is not modulated and is called the carrier reference. The second pulse, which is modulated, is separated by a time delay T_d from the first pulse. The receiver uses pulse-pair correlators to recover the data. Each multipath components results in a peak at the output of the multiplier with the same phase (which is determined by the value of the data symbol), and therefore they can be summed by an integrator during a time, T_g. The integrator output is detected in a conventional way to make a decision on the transmitted data symbols. The scheme thus shows a fundamental similarity to differential detection. The main drawback of the TR scheme stems from a reduced signal-to-noise ratio. This reduction is partly due to "wasting" energy on the reference pulses that are non-information-bearing. More importantly, the differential detection gives rise to excess noise related to the multiplication of noise contributions in the received reference pulses with the noise contributions in the received signal pulses. Those crossterms are especially troublesome in a UWB receiver: the input to the multiplicator has a very low signal-to-noise ratio (SNR), because those are signals $before$ the despreading operation. This results in large noise-times-noise terms that are integrated over a time T_g.

Due to their simple structure, transmitted reference schemes have received considerable attention in the past two years. [7] first suggested the scheme for UWB, and also presented some experimental results in a companion paper [11]. References [8] and [9] suggested to average the reference pulses over several pulse durations, in order to reduce the noise in the reference pulses, and thus also the crossterms. Reference [8] also gives a detailed derivation of the error probability both for a conventional differential receiver, as well as one that uses averaged reference pulses.

In this paper, we introduce new differential receiver structure that shows performance similar to that of [8], while being easier to implement. We also present a performance analysis; specifically our contributions include:

- we present a receiver structure that requires only $symbol$-rate sampling, not $frame$-rate sampling
- we analyze the effect of the non-Gaussian nature of the noise-noise- crossterms
- we analyze a system that uses BPSK as modulation format, and polarity randomization of the pulse sequence. BPSK gives a better receiver SNR [12], and the polarity randomization results in a transmit spectrum that better exploits the FCC mask [13]
- we simulate the performance of a TR scheme in the standardized IEEE 802.15.3a channel models

The paper is organized the following way: Section II describes our new hybrid receiver structure, and discusses the implementability of the scheme. Next, we set up the mathematical model of the transceiver structure, and derive the signal, noise, and noise-noise crossterms. Section IV presents simulation results for both the AWGN (additive White Gaussian Noise) channel and for delay-dispersive channels. A summary and conclusions wrap up this paper.

II. Theory

A. System model

The transmit signal uses time-hopping impulse radio (TH-IR) [3] as multiple access format, and transmitted-reference BPSK as modulation format. The transmit signal can thus be written as

$$s_{TX}(t) = \sqrt{\frac{E_s}{2N_f}} \left[\sum_{j=-\infty}^{\infty} d_j w_{tx}\left(t - jT_f - c_jT_c\right) + \sum_{j=-\infty}^{\infty} b_{\lfloor j/N_f \rfloor} d_j w_{tx}\left(t - jT_f - c_jT_c - T_d\right) \right] \quad (1)$$

where T_c denotes the chip duration, T_f the frame duration, and T_d the delay between the reference pulse and the data-carrying pulse. The c_j denote a (pseudo-)random integer sequence with values between 0 and $N_c - 1$, which determines the time-hopping sequence. The d_j denote a pseudorandom sequence of $+1$ and -1 that ensures a zero-mean output and is also helpful in the shaping of the transmit spectrum [13] according to the FCC rules [1]. The function $w_{tx}(t)$ denotes the transmit waveform; in the following, we assume that its support extends only over one chip duration. E_s is the energy per transmitted symbol. Note that $N_fT_f = T_s$, the symbol duration.

When transmitted over an AWGN channel, the received signal can then be written as

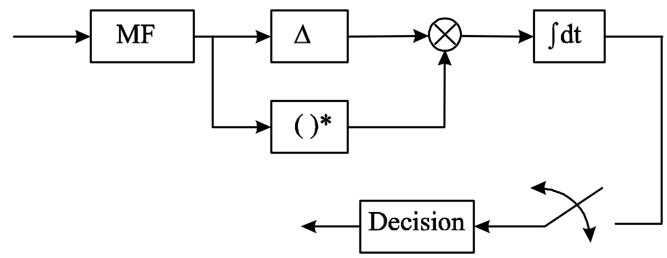

Fig. 1. Building blocks for the basic hybrid receiver. Note that the sampling circuit performs symbol rate sampling.

$$r_{RX}(t) = \sqrt{\frac{E_b\alpha}{2N_f}} \left[\sum_{j=-\infty}^{\infty} d_j w_{RX}\left(t - jT_f - c_jT_c\right) + \sum_{j=-\infty}^{\infty} b_{\lfloor j/N_f \rfloor} d_j w_{RX}\left(t - jT_f - c_jT_c - T_d\right) \right] + \sigma_n n(t) \quad (2)$$

where α is the channel attenuation, σ_n^2 is the noise variance, and $n(t)$ is a unit-variance Gaussian process. Depending on whether we consider a baseband or bandpass filter, $n(t)$ is a real or complex Gaussian process, respectively.

B. Hybrid receiver principle

As mentioned above, a main performance degradation of the hybrid receiver is due to the noise-noise crossterms created by the multiplication process in a conventional TR receiver. As the input SNR of a spread spectrum UWB receiver is low, the noise-noise crossterms are non-negligible - in contrast to conventional differential detection [12]. The key idea of our new receiver is to perform a despreading $before$ the multiplication operation. Thus, the SNR of the inputs to the multiplier is higher, and the relative impact of the noise-noise crossterms is lower.

Fig. 1 shows a block diagram of the receiver. The receiver consists of filters matched to all the base signals, including the reference signal. It is critical that this filter is matched to the whole *pulse sequence* representing one symbol, and not just the basis pulse within a frame. The output of the filter matched to the reference signal is then delayed by a delay T_d and multiplied with the outputs of the filters matched to the modulated signals. This product is then integrated over an finite interval T_{int}, determined by the excess delay and signal duration to achieve maximum SNR. After the integration and at the correct decision instant, the outputs of the integrators are compared and a decision is made in favor of the largest one. The hybrid detector scheme is similar to the synchronization scheme proposed in [14] for preamble-based synchronization in OFDM systems and shares many of its advantages.

The receiver of the hybrid scheme differs from the conventional transmitted reference scheme [8] in that correlation in the matched filters is performed prior to multiplication of the reference signal and the modulated signal. In this way, the terms in the multiplication have a much higher SNR, and the strong influence from the noise-times-noise terms can be decreased or almost eliminated. The SNR of the terms is increased by a factor N_f, where N_f is the number of monocycles per symbol, compared to the conventional transmitted reference scheme. Note that the *effect* of that averaging is similar to averaging over several reference pulses as suggested in [8], i.e., reducing the noise-noise crossterms. However, when comparing the receiver $structure$ to that of [8], we find important differences. The Choi/Stark receiver samples the signal at least once in each frame to allow the averaging of the reference pulses. The sampling, as well as the subsequent A/D conversion and digital processing of the sample values, thus occurs at the *frame* rate. Our scheme performs a filtering matched to the *transmit sequence*, followed by the multiplication, and sampling at the *data rate*. As the data rate is typically orders of magnitude lower than the frame rate, operating at such low rates allows large savings in the complexity and especially the power consumption of the transceiver.

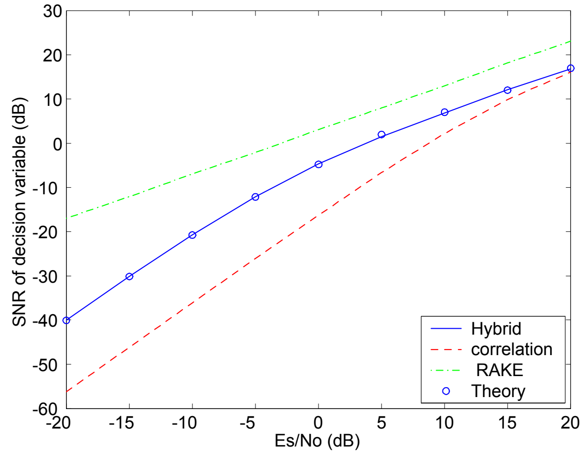

Fig. 2. SNR of the decision variable in AWGN, $N_f = 40$.

C. Received signal statistics

In our new hybrid receiver, the incoming signal is first filtered with a filter that is matched to the transmit waveform $(1/\sqrt{N_f})\Sigma d_j w_{TX}\left(t - jT_f - c_jT_c\right)$. Note that the use of $w_{RX}(t) = w_{TX}(t)$ is not necessarily optimum in this context. However, it is the best practical implementation, as a generator for this waveform is already available in a transceiver. In the following, we will neglect a possible interchip interference, and make the simplifying assumption that the support of the autocorrelation function $R_w(t)$ of w_{TX} is limited to one chip duration. We also assume that each frame contains a guard interval of at least length T_d so that the data pulse cannot "spill over" into the adjacent frame. The output from this matched filter is then

$$
\begin{aligned}
p(t) = {} & \sqrt{\frac{E_b\alpha}{2N_f^2}}\left[\sum_{j=-\infty}^{\infty}\sum_{j'=0}^{N_f-1} d_j d_{j'}\right. \\
& R_w\left(t - [(j'-j)T_f + (c_{j'} - c_j)T_c]\right) + \\
& \sum_{j=-\infty}^{\infty}\sum_{j'=0}^{N_f-1} b_{\lfloor j/N_f \rfloor} d_j d_{j'} \\
& \left. R_w\left(t - T_d - [(j'-j)T_f + (c_{j'} - c_j)T_c]\right)\right] + \\
& \sqrt{\sigma_n^2 \frac{1}{N_f}} \sum_{k'=0}^{N_f-1} w_{RX}(t - kT_f - c_kT_c) \otimes n(t)
\end{aligned}
\tag{3}
$$

Due to the small support of the autocorrelation function, only terms with $j = j'$ give nonzero contributions. [1]

D. Output SNR in AWGN channels

The signal component of $p(t)p(t - T_d)$ is given as

$$\frac{E_b\alpha}{2} R_w^2(0) \tag{4}$$

[1] We use here the approximation that the autocorrelation function $R(t)$ has a support of width T_c.

while the energy of the signal-times-noise contribution is $E_b\alpha\sigma_n^2 R_w^3(0)$, so that (without the noise-noise crossterms), the SNR is

$$SNR = \frac{E_b\alpha}{4\sigma_n^2} R_w(0) \tag{5}$$

This result can be easily explained intuitively: it is the SNR of an ideal Rake receiver, plus a 6 dB penalty. A 3 dB penalty arises from the fact that half of the signal energy is used for the reference pulse. The other half arises from the fact that the "local oscillator" is noise, and actually contributes as much noise to the receiver output as the "desired signal".

In addition, we have a noise-times-noise crossterm. The output of the matched filter when excited with Gaussian noise is again Gaussian noise, with a variance $\sigma_n^2 R_w(0)$. This term is multiplied by another (independent) Gaussian variable with the same variance. The probability density function (pdf) of the product of two Gaussian variables is given by [15]:

$$p_x(x) = \frac{1}{\pi\sigma^2} K_0\left(\frac{|x|}{\sigma^2}\right) \tag{6}$$

where $K_0(x)$ is the modified Bessel function zero-th order, second kind [16] and σ^2 is the variance of the underlying Gauss process. Note that the structure of this noise is different from the one occurring in [8]. In that reference, the noise-noise-crossterm within each frame follows a pdf of the form 6. Such contributions are summed over all frames, resulting in a total output pdf that is closer to Gaussian than the terms we observe. In Fig. 2 the variance of the decision variable in AWGN is presented for different values of E_s/N_0 for a one-tap RAKE receiver, the hybrid receiver and a correlating receiver without averaging of the reference pulses. For the hybrid receiver, theoretical values of the SNR using (6) and (5) are also presented. The effect of the noise-noise-crossterms can clearly be seen for low E_s/N_0. The breakpoint where these cross terms become dominant is determined by N_f and is shifted towards low E_s/N_0 for a large N_f .

E. Output SNR in multipath channels

In this section, we consider a multipath channel with the impulse response

$$h(t) = \sum_{l=0}^{L-1} a_l \delta(t - \tau_l) \tag{7}$$

This is a simplified model that assumes that all multipath components have delays that are integer multiples of the chip duration. While not exact, this model can reasonably approximate real UWB channels [17]. We use the normalization $\Sigma a_l^2 = 1$. We furthermore assume that the duration of each frame is much larger than the delay spread of the channel, so that inter-frame interference (IFI) is negligible.[2] The more general case of finite IFI is left for future work [18].

Under these assumptions, the output of the matched filter can be written as

$$p(t) = \sqrt{\frac{E_b \alpha}{2N_f^2}} \left[\sum_{j=-\infty}^{\infty} \sum_{l=0}^{L-1} a_l R_w(t-\tau_l) + \right. \tag{8}$$
$$\left. \sum_{j=-\infty}^{\infty} \sum_{l=0}^{L-1} b_{\lfloor j/N_f \rfloor}\, a_l R_w(t - \tau_l - T_d) \right] +$$
$$\sqrt{\sigma_n^2 \frac{1}{N_f}} \sum_{k'=0}^{N_f-1} w_{RX}(t - kT_f - c_k T_c) \otimes n(t)$$

As the signal suffers delay dispersion due to the properties of the channel, the output of the multiplier must be integrated over a longer period of time $T_{int} = QT_c$

In that case, the signal energy at the output of the multiplier becomes

$$r_{out} = \frac{E_b^2 \alpha^2}{4} R_w^2(0) \tag{9}$$
$$\left[b \left(\sum_{l=0}^{Q-1} a_l^2 + a_{l-\Delta} a_{l+\Delta} \right) + \right.$$
$$\left. \sum_{l=0}^{Q-1} a_{l+\Delta} a_l + a_l a_{l-\Delta} \right]$$

where Δ is T_d in units of the chip duration, and channel coefficients with negative index are understood to be zero. We note that the signal consists of four terms: a term that multiplies the desired bit polarity (first line), and an offset that depends only on the channel state. The latter term could lead to a deterioration of the bit error rate (BER); however, it is possible to compensate for it by adjusting the decision threshold of the detector - as the channel can usually be assumed to be quasistatic. We also note that the desired symbol is multiplied by a term that contains $\sum_{l=0}^{Q-1} a_l^2 + a_{l-\Delta} a_{l+\Delta}$ The the latter term occurs only if the delay between reference pulse is smaller than the maximum excess delay. This effect leads to additional fading, and thus a worse BER. Its magnitude can be controlled by appropriately choosing the value of Δ.

[2]This condition can be strictly assured by having a guard interval of length $D + T_{ds}$, where T_{ds} is the channel delay spread.

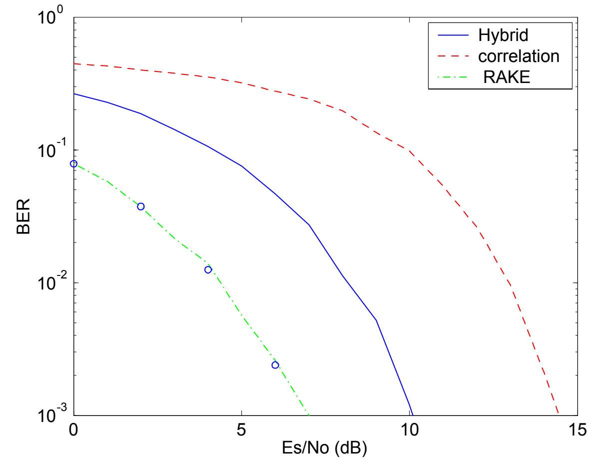

Fig. 3. Siulated BER for different E_s/N_0 in AWGN, $N_f = 40$

As we are now summing over Q samples, the noise power (both for the Gaussian and the non-Gaussian noise) increases by a factor Q. The statistics of the sum of the noise-signal crossterms are again Gaussian, with the variance being larger by a factor of Q. For the noise-noise crossterms, the addition of Q terms makes the statistics more Gaussian-like. If Q is odd, $Q = 2m + 1$, the pdf of the output can be written as

$$p_X(x) = \frac{\left[|x|/2\sigma^2 \right]^m}{\sqrt{\pi}\Gamma(m+0.5)\sigma^2} K_m \left(\frac{|x|}{\sigma^2} \right) \tag{10}$$

where $K_m(x)$ is the modified Bessel function of the $m-$th order, second kind, and Γ is Euler's Gamma function. For m even, a similar equation holds [15]. Note that the variance of this term increases linearly with Q, but the statistics change. If Q is large, the fading statistics can be approximated as Gaussian, with a variance of $Q\sigma^4$. The optimum integration time can then be easily obtained by maximizing the output SNR.

III. Performance Evaluation

To evaluate the performance we have also simulated the BER for an UWB system with antipodal signalling and a data rate of 500 kbit/s, $N_f = 40$ monocycles per symbol, $T_d = 850$ ns delay between the reference and modulated signal, a frame time of $T_f = 20$ ns, and a chipduration which is assumed to be small. Fig. 3 shows the BER for different E_s/N_0 in AWGN. For the Hybrid receiver and the correlation receiver there is, again, a 3 dB performance loss due to the energy spent on the reference signal and a loss due to the noisy reference signals. The presented correlation receiver does not use averaging and therefore this has quite a large loss. If averaging over one symbol period is performed, the performance would actually be equal to that of the Hybrid receiver in AWGN channels. To evaluate the performance in multipath channels we use the channel model specified in IEEE 802.15.3a [19]. This channel model specifies a set of 100 impulse responses for different environments. Channel CM1 is an indoor short range channel with low delay spread whereas channel CM4 has the highest delay spread. The duration of the monocycles was assumed to be small. In Fig. 4 the

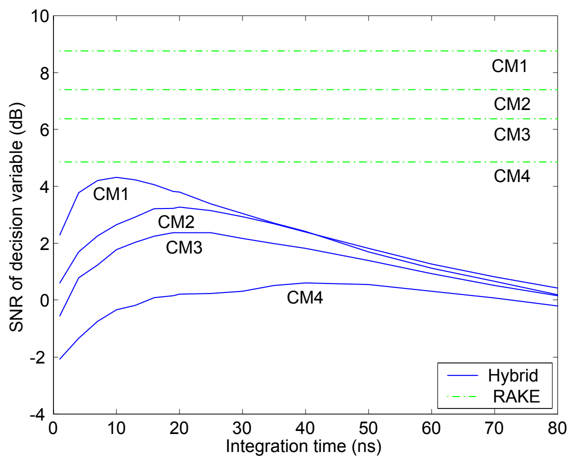

Fig. 4. SNR of the decision variable for the hybrid scheme and a one tap RAKE reciever in multipath channels, $N_f = 40$ and $E_s/N_0 = 10$dB.

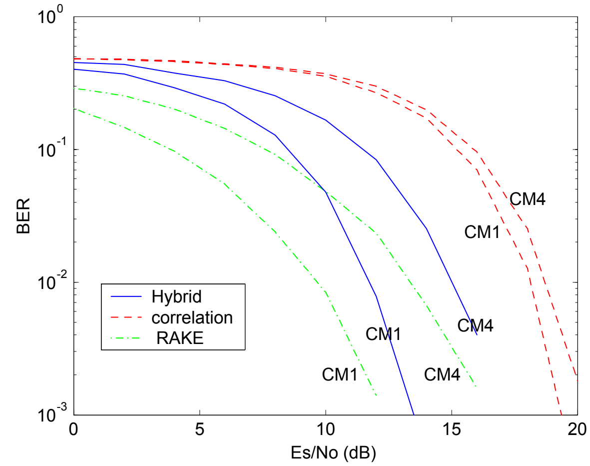

Fig. 5. Siulated BER for different E_s/N_0 in multipath channels, $N_f = 40, .T_{int} = 15$ ns.

influence of the integration time, T_{int}, on the SNR of the decision variable is shown, using the same parameters as before, for $E_s/N_0 = 10$ dB. A short integration time means that the receiver is capable to capture only a small part of the received energy. A too large T_{int}, on the other hand, means contributions from noisy terms with only weak signal components. For this particular setup an integration time of around 15 ns seems appropriate. Though the Hybrid receiver has the ability to capture energy from many multipath components the one-tap RAKE has a better performance when it uses the strongest tap only. Finally, the simulated BER is shown for CM1 and CM4 using an integration length $T_{int} = 15$ ns. For high values of E_s/N_0 the BER of the Hybrid receiver approaches that of the RAKE receiver and there is a loss of approximately 1 dB. As expected, the Hybrid receiver clearly outperforms the conventional correlation receiver, with a gain for this setup of approximately 6 dB. In the simulations we have used a delay, T_d, that is large compared to the excess delay of the channels. It should be noted, however, that the receiver is able to handle smaller delays as well.

IV. Conclusions

In the paper we have presented a novel structure for an UWB-receiver. The proposed hybrid matched filter correlation receiver performs matched filter detection before correlating (multiplying) the data symbol with the reference symbol. Therefore the SNR in this multiplication operation is much better compared to conventional correlation receiver and the proposed receiver does not suffer from large noise-times-noise products. After multiplication the signals from different multipath components have the same phase and therefore they can be integrated to capture signal energy from the whole, or a large portion of, the impulse response. We then evaluated the performance in UWB channels specified in IEEE 802.15. The simulations showed a BER performance superior to that of the conventional correlation receiver without a significant increase in the complexity.

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