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Effect of Uncoordinated Network Interference on UWB Energy Detection Receiver

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Abstract—Over the last few years, there has been an emerging interest in applying ultrawideband (UWB) communications in wireless sensor networks, mainly due to the low-complexity and low-power consumption of UWB technology. In particular, low-complexity receiver like the energy detection (ED) receiver is a potential candidate for sensor network applications. However, the presence of network interference will severely degrade the performance of such receivers. In this paper, we analyze the bit error probability performance of the ED receiver in the presence of uncoordinated UWB interference. We model the network interference as an aggregate UWB interference, generated by elements of uncoordinated UWB networks scattered according to a spatial Poisson process. Our analytical framework allows a tractable performance analysis and gives sufficient insight into the effect of uncoordinated network interference on UWB systems.

I. INTRODUCTION

There has been an increasing interest in ultrawideband (UWB) technology, particularly as a strong candidate for low-power consumption sensor network applications [1], [2]. In particular, non-coherent receiver like the energy detection (ED) receiver has been considered as potential low-complexity solution in the IEEE 802.15.4a standardization process [3]. The wide spreading of sensor networks using UWB communications to ensure wireless connectivity will inevitably lead to increasing network interference (NWI), especially between uncoordinated networks. It is likely that the main NWI is contributed by a few dominant interferers at close range. As a result, the UWB NWI tends to exhibit a distribution with heavier tail than the Gaussian distribution. In addition, with the low duty-cycle of UWB transmissions, the interference presents an impulsive behavior. This complicates the modeling of UWB NWI since we can no longer use the Gaussian approximation [4], [5].

In modeling impulsive signals, the stable distribution provides a valuable mathematical tool for modeling a wide class of impulsive noise processes [6]–[8]. In the case of NWI, it is also necessary to account for the stochastic geometry of the interfering sources to obtain a more accurate statistical model of the network interference. By assuming a Poisson field of interferers, several works have analyzed the effect of narrowband interference on narrowband [6], [7] and UWB systems [8], respectively. However, to the best of our knowledge, there is hardly any results available that analyze the effect of uncoordinated UWB NWI, particularly, when non-coherent receiver structures are employed.

In this paper, we analyze the bit error probability (BEP) performance of the ED receiver in the presence of uncoordinated UWB NWI. The ED receiver is based on the binary pulse position modulation [8]. We show that multivariate stable random variables (r.v.’s) can be used to describe the statistics of the NWI. The proposed model for the aggregate interference accounts for the spatial distribution of the UWB interferers and the propagation characteristics of the interference signals.

II. SYSTEM AND CHANNEL MODELS

The transmitted signal for user $k$ can be expressed as [8]

$$s^{(k)}(t) = \sum_{i} \left[ (1 - d_{i}^{(k)}) p_{1}(t-iT_s) + d_{i}^{(k)} p_{2}(t-iT_s) \right],$$

where $d_{i}^{(k)} \in \{0, 1\}$ is the $i$th data symbol and $T_s = \frac{N}{f}$ is the symbol duration, such that $T_s$ is the average pulse repetition period. The transmitted signal for $d_{i}^{(k)} = 0$ and $d_{i}^{(k)} = 1$ can be written, respectively, as

$$b_{1}^{(k)}(t) = \sum_{j=0}^{N-1} \sqrt{E_p} a_{j}^{(k)} p(t-jT_f - c_{j}^{(k)}T_p),$$

$$b_{2}^{(k)}(t) = \sum_{j=0}^{N-1} \sqrt{E_p} a_{j}^{(k)} p(t-jT_f - c_{j}^{(k)}T_p - \Delta),$$

where the parameter $\Delta$ is the time shift between two different data symbol. In time-hopping (TH) signaling, $\{c_{j}^{(k)}\}$ is the pseudo-random sequence of the $k$th user, where $c_{j}^{(k)}$ is an integer in the range $0 \leq c_{j}^{(k)} < N_h$ and $N_h$ is the maximum allowable integer shift. The bipolar random amplitude sequence $\{a_{j}^{(k)}\}$ together with TH sequence are used to mitigate interference and to support multiple access. The energy of the transmitted pulse is $E_p = \frac{E_s}{N_p}$ where $E_s$ is the symbol energy. To preclude intra-symbol interference (isi) and inter-symbol interference (ISI), we assume $\Delta \geq T_f$ and $(N_h - 1)T_p + \Delta + T_f \leq T_s$.

The received signal can be expressed as $r(t) = h(t) * s(t) + n(t)$, where $h(t)$ is the impulse response of the channel given...
Letting where $h_l$ and $\tau_l$ are the attenuation and the delay of the $l$th path component, respectively. The term $n(t)$ is zero-mean, white Gaussian noise with two-sided power spectral density $N_0/2$. As in [9], we consider a resolvable dense multipath channel, i.e., $|\tau_l - \tau_j| \geq T_p$ for all $j \neq l$, where $\tau_l = \tau_l + (l - 1)T_p$ and $\{h_l\}_{l=0}^L$ are statistically independent r.v.'s. We can express $h_l = |h_l| \exp(j\phi_l)$, where $\phi_l = 0$ or $\pi$ with equal probability.

The ED receiver first passes the received signal through an ideal bandpass zonal filter (BPZF) of bandwidth $W = 1/T_p$. The output of the BPZF can be written as

$$\tilde{r}(t) = \sum_{l=0}^{L} h_l [(1 - d_l)b_l(t - iT_a - \tau_l) + d_l b_2(t - iT_a - \tau_l)] + \tilde{n}(t),$$

where $\tilde{n}(t)$ represents the noise process after the BPZF. The decision variables for the ED receiver depends on the difference in energy of the received signals over the two observation variables. Mathematically it can be written as

$$Z_{ED} = \frac{1}{2} \left\{ \begin{array}{ll} 1 & \text{if } \tilde{r}(t) + \tilde{n}(t) > 0 \\ 0 & \text{otherwise} \end{array} \right.$$

and

$$Z_{ED} = \frac{1}{2} \left\{ \begin{array}{ll} 1 & \text{if } \tilde{r}(t) + \tilde{n}(t) > 0 \\ 0 & \text{otherwise} \end{array} \right.$$

where $T$ is the integration interval. For two non-central chi-squared r.v.'s $(X_1$ and $X_2$) with same degrees of freedom $q$ the probability that $X_1 - X_2 < 0$ can be expressed as [10]

$$1 - \frac{1}{2} \sum_{j=0}^{q-1} \int_{jT_a}^{(j+1)T_a} \left( \frac{\tilde{r}(t)^2}{2T_c} \right) dt - \frac{1}{2} \sum_{j=0}^{q-1} \int_{jT_a}^{(j+1)T_a} \left( \frac{\tilde{n}(t)^2}{2N_0/T_a} \right) dt + \frac{1}{2} \left( \frac{\tilde{r}(t)\tilde{n}(t)}{2T_cN_0/T_a} \right) dt,$$

where $T_a = \frac{T}{2}$. Using the spatial model in (7), the aggregate UWB interference signal received at the output of the BPZF of the desired user is given by

$$\zeta(t) = \sum_{n=1}^{\infty} \zeta^{(n)}(t),$$

and $\zeta^{(n)}(t)$ denotes the signal from the $n$th UWB interferer and it can be expressed as

$$\zeta^{(n)}(t) = \frac{e^{\sigma a^{(n)}(t)}}{R^{(n)}} \sqrt{P_{T_a}}v^{(n)}(t - D^{(n)}),$$

where the shadowing term $e^{\sigma a^{(n)}(t)}$ follows a log-normal distribution with shadowing parameter $\sigma$ and $G^{(n)} = \mathcal{N}(0,1)$. According to the far-field assumption, the signal power decays as $1/(R^{(n)})^{\alpha}$, where $R^{(n)}$ is the distance between the $n$th UWB interferer and the desired user and $\nu$ is the path loss exponent. To model time-asynchronism of the UWB interfering signals, we define $D^{(n)}$ as a uniformly distributed r.v. and $v^{(n)}(t)$ in (10) can be further expressed as

$$v^{(n)}(t) = \sum_{i} b^{(n)}_i (t - iN_i^2T_i^2) + h^{(n)}_i (t - iN_i^2T_i^2),$$

where $h^{(n)}_i(t) = \sum_{l=0}^{L-1} h_l^{(n)} [e^{-j\phi_l}] \delta(t - \tau_l)^{n(n)}$ is the channel impulse response of the $n$th UWB interferer-receiver link.

Using our system model in Section II, the transmitted signal from the $n$th UWB interferer is given by

$$I^{(n)}(t) = \sqrt{P_{i}} \sum_{i} b^{(n)}_i (t - iN_i^2T_i^2),$$

where $h_l^{(n)}(t) = \sum_{j=0}^{\infty} e^{j\phi_l} a^{(n)}_i p(t - jT_i - c_j^{(n)}T_p - d_i^{(n)})],$ $P_{i} = E_{i}/T_i N_i^2$, and $\{I_t\}_{t=1}^P$ is the average power at the border of the near-field zone of each interfering transmitter antenna, and $T_i^1$ is the pulse repetition period average, such that it is assumed to be the same for all UWB interferers and all interferer signals also have the same symbol duration $T_i^1 = T_i^2 N_i^2$. Note that we intentionally write (8) to account for two possible modulations, namely binary pulse amplitude modulation (BPAM) and BPPM modulation. The term $c^{(n)}_j = \{1, -1\}$ is the $j$th data symbol for BPAM modulation, $d^{(n)}_j = \{0, 1\}$ is the $j$th data symbol for BPPM modulation, and $\Delta_i^1$ is the position modulation shift. The $j$th element of the random hopping and amplitude sequences are denoted by $\{c^{(n)}_j\}$ and $\{a^{(n)}_j\}$, where $0 < c^{(n)}_j < N_i^2$, $N_i^2$ is the maximum shift associated with the hopping code, and $a^{(n)}_j = \{1, -1\}$ for all $j$ and $n$. The average pulse repetition interval is considered long enough such that i.i.d and ISI can be ignored. For notational convenience, we define $\Psi^{(n)} = \{c^{(n)}_j\}, \{a^{(n)}_j\}, \{b^{(n)}_i\}, \{h^{(n)}_i\}$.

III. UWB INTERFERENCE

A. Multiple UWB interferers

We model the spatial distribution of the multiple UWB interferers according to a homogeneous Poisson point process in a two-dimensional plane. The probability that $k$ nodes lie inside region $R$ depends only on the area $A_R = |R|$, and is given by [11]

$$P\{k \in R\} = \left( \frac{\lambda A_R}{k!} \right)^k e^{-\lambda A_R},$$

where $\lambda$ is the spatial density of the active devices.
B. ED receiver

Conditioning on \( \{\Psi^{(n)}\}, \{c_j\}, \{a_j\}, \{h_l\} \), it can be shown that the non-centrality parameters of \( Y^{(1)} \) and \( Y^{(2)} \) for \( d_0 = 0 \) are, respectively, given by

\[
\rho^{\text{(UWB),(1)}} = \sum_{l=1}^{L} \sum_{j=0}^{2\pi W} \sum_{m=0}^{1} \frac{1}{N_0} 2 \cos(\phi_m) \sin(\phi_m) \text{Re}(1 + jT_l + c_jT_p) \text{Re}(1 + jT_l + c_jT_p + \Delta),
\]

\[
\rho^{\text{(UWB),(2)}} = \sum_{l=1}^{L} \sum_{j=0}^{2\pi W} \sum_{m=0}^{2} \frac{1}{N_0} 2 \cos(\phi_m) \sin(\phi_m) \text{Re}(1 + jT_l + c_jT_p + \Delta),
\]

where \( \zeta_{1,j,m} \) and \( \zeta_{2,j,m} \), for odd \( m \) (even \( m \)), are the real (imaginary) parts of the samples of the equivalent low-pass version of \( \zeta_j(t) \), uniformly distributed over \([0, T]\), and \( \zeta_j(t) \) and \( \zeta_j(t) \) respectively, sampled at Nyquist rate \( W \) over the interval \([0, T]\).

From (12) and (13), it can be observed that we still need to derive some statistical model for the aggregate UWB interference. In the following, we define the complex vector \( \zeta_{1,j} \), which composed of \( W \) samples of \( \zeta(t) \), defined in (9).

The samples of the signal are taken at the Nyquist rate \( W \) in the interval \([0, T] \) within the \( j \)th signal frame of the bit position. Specifically, the vector \( \zeta_{1,j} \), can be written as

\[
\zeta_{1,j} = \sum_{n=1}^{\infty} e^{j\pi G(n)} (R(n)) \bar{v}_{1,j}^{(n)},
\]

where \( \bar{v}_{1,j}^{(n)} \) is the vector of complex samples of the equivalent low-pass version of \( v^{(n)}(t + c_jT_p + jT_l - D(n)) \), such that \( \bar{v}_{1,j}^{(n)} \) at the sampling instant \( m \) are a sequence of i.i.d. r.v.'s in \( n \). If the signal of the \( n \)th UWB interferer is in the sampling instant \( m \), each sample can be written as

\[
v_{1,j}^{(n)} = \sum_{m=1}^{T^2} G(n) \mu P_m(t - mT_p),
\]

where \( G(n) \), uniformly distributed over \([0, T^2] \), is a r.v. uniformly distributed over \([0, 1, \ldots, L - 1] \), \( \mu P_m(t - mT_p) \) is a r.v. with variance \( 1/\sum_{d=1}^{L} \exp(-e^d(t)) \) and \( \mu P_m(t - mT_p) \) is a r.v. uniformly distributed over \([0, 2\pi) \).

Considering that the complex r.v. \( \Theta_m \) is circularly symmetric (CS), for the case in the presence of narrowband interference [8], \( \zeta_{1,j,m} \) can be described by a stable complex distribution as follows

\[
\zeta_{1,j,m} \sim \Sigma \left( \frac{2}{\nu}, \gamma^{\text{UWB}} \right),
\]

where \( \gamma^{\text{UWB}} \) is the \( n \)th complex sample of \( \zeta_{1,j} \) in (14) and \( \gamma^{\text{UWB}} = \lambda \pi C_{1/\nu}^{2} e^{2aT^2}/\nu \nu^2 \text{Re}(\{\bar{v}_{1,j,m}\}) \), such that

\[
E\{\text{Re}(\{\bar{v}_{1,j,m}\})^{2/\nu}\} = \frac{T^2}{\nu} MFP \text{ and the associated parameters } M, F, P \text{ are given by}
\]

\[
M = \mathbb{E}\{|h_{m}^{(n)}|^{2/\nu}\} \mathbb{E}\left\{\left(\exp(-e^{T^{(n)}})\right)^{2/\nu}\right\},
\]

\[
F = \mathbb{E}\{|e^{(n)}|^{2/\nu}\} \mathbb{E}\left\{\left|\cos(\phi_m)\right|^{2/\nu}\right\},
\]

\[
P = \mathbb{E}\{|p|^{2/\nu}\}.
\]

Note that the components of the aggregate interference vector \( \zeta_{1,j} \) in (14) are identically distributed but mutually dependent [12]. To make our analysis tractable, we assume that the \( S\alpha S \) vector \( \zeta_{1,j} \) is spherically symmetric since spherically symmetric vectors have the characteristic of being sub-Gaussian, which implies that they can be decomposed as

\[
\zeta_{1,j} = \sqrt{V} \bar{G}_{1,j},
\]

where \( V \sim S(\alpha/2, 1, \cos(\alpha)) \) and \( \bar{G}_{1,j} \) is a multivariate Gaussian random vector with covariance matrix \( \Sigma \). Unfortunately, \( \zeta_{1,j} \) is spherically symmetric only for some scenario. For example, it can be shown that the \( \zeta_{1,j} \) is sub-Gaussian if we consider Rayleigh fading with flat power delay profile, uniform phase, and \( T_{s} = T_{f} \). To ensure the spherical symmetry of the resulting aggregate interference vector for more general scenario, we modify each received interference signal as

\[
\bar{v}_{1,j}^{(n)}(t) = \sum_{m=1}^{T^2} G_{1,j,m} p(t - mT_p),
\]

where \( d_{s}^{\alpha} = 2^{-2/\nu} \left(1 + \frac{1}{\nu}\right)^{-1} \) corresponds to \( E\{|\bar{z}_{(n)}^{(n)}|\}^{2/\nu} \), \( \{G_{1,j,m}\}_{m=1}^{W^2} \) is a sequence of i.i.d complex Gaussian r.v.'s with zero mean and unit variance, and

\[
E\{|\bar{z}_{(n)}^{(n)}|\} = \frac{T_{s}^{2}}{T_{f}^{2}} M \times F.
\]

Note that each interfering UWB signal now covers the entire frame interval \( T_{f}^{2} \) and the effect of the duty cycle, channel fading, and channel power delay profile (PDP) are captured in the statistics of \( z(n) \), where \( z(n) = 0 \) with probability \( 1 - T_{f}^{2} / T_{s}^{2} \). It can be shown that the statistics of the aggregate interference obtained by using the interference model in (18) is in good agreement with the empirical statistics generated via simulation when realistic conditions are considered.

IV. BEP of the ED Receiver in the Presence of UWB Interference

A. Type 1 interference

We consider \( n_{1} T_{f} = \Delta \) and \( n_{2} T_{f} = T_{f} \) such that \( n_{1} \) and \( n_{2} \) (\( n_{1} > n_{2} \)) are integers. For simplicity, we consider no modulation is used and no random amplitude sequences and hopping code sequences are used. If the vector representing

Footnotes:

4In fact, the aggregate interference vector in (14) is symmetric alpha stable (\( S\alpha S \)).

5With this assumption, the Type 1 interference is periodic, which allows us to obtain a BEP expression that does not require numerical averaging.
the aggregate interference can be expressed as in (17), the non-centrality terms of $Y_{ED,1}$ and $Y_{ED,2}$ for $d_0 = 0$ are, respectively, given by
\begin{equation}
\mu^{(UWB)}_{Y_{Ed,1}} = \frac{1}{2} \sum_{l=0}^{L_{CAP}} \sum_{j=1}^{N_0} h_l^2 + \frac{P_1 N_s}{4 W N_0} 2 \gamma_{UWB} V C_1^{(1)} + \sum_{j=0}^{\infty} C_{1,j,m} w_{j,m} \sum_{j=0}^{\infty} \sum_{m=1}^{N_0} \frac{\gamma_{UWB} V C_1^{(1)}}{2 W N_0},
\end{equation}
where $C_1^{(1)} = \sum_{m=1}^{2 WT} C_{1,j,m}$ is a chi-square r.v. with $2 WT$ degrees of freedom. To evaluate the BEP of the ED receiver for BPPM detection, we can numerically average (6) with respect to the r.v’s that appear in (20) and (21). Alternatively, we can use the approximate analytical approach, which assumes $\mu^{(UWB)}_{C,ED}$ negligible compared to the other first two terms in (20) to obtain the conditional BEP as [8]
\begin{equation}
P_{e,ED}^{(UWB)}(1) \approx \frac{1}{2} \frac{1}{\pi} \int_0^{\infty} \left( \frac{1}{1 + v^2} \right) \psi_{AP}(1) \psi_{ED}(jv) dv \times \text{Re} \left\{ \psi_{ED}(jv) \psi_{AP}(1) \psi_{ED}(jv) \right\},
\end{equation}
where $\psi_{AP}(1) = \exp \left\{ -\left( 2 C_1^{(1)} \right) \frac{1}{2} \right\}$ and $\psi_{ED}(jv)$ is given by
\begin{equation}
\psi_{ED}(jv) = \frac{P_1 N_s}{4 W N_0} \left( \frac{-jv}{1 + jv} \right) \left( \frac{1}{1 - jv} \right).
\end{equation}
In this section, we evaluate the performance of the ED receiver in the presence of uncoordinated UWB NWI. For the desired signal, we consider a bandpass UWB system with pulse duration $T_p = 0.5$ ns, symbol interval $T_s = 3200$ ns, and $N_a = 32$. For simplicity, $\Delta$ is set such that there is no ISI or isi in the system, i.e., $T_1 = 2 \Delta$. We consider a TH sequence of all ones ($c_j = 1$ for all $j$) and $N_0 = 2$. The desired signal is affected by a dense 10 channel, where each multipath amplitude is Nakagami distributed with fading severity index $m$ and average power $E_0 \{ h_t^2 \}$.

\begin{equation}
\int_0^{2 \pi} \int_0^{\infty} \left( \frac{1}{1 + v^2} \right) \psi_{ED}(jv) \psi_{AP}(1) \psi_{ED}(jv) dv \times \text{Re} \left\{ \psi_{ED}(jv) \psi_{AP}(1) \psi_{ED}(jv) \right\},
\end{equation}

\begin{equation}
B. Type 2 interference

We assume that the vector $\xi$ is sub-Gaussian, representing the aggregate interference signal along the entire symbol duration $T_s$. The covariance matrix of the Gaussian vector $\xi$ is still diagonal. The non-centrality terms of $Y_{ED,1}$ and $Y_{ED,2}$ for $d_0 = 0$ are, respectively, given by
\begin{equation}
\mu^{(UWB)}_{Y_{Ed,2}} = \frac{1}{2} \sum_{l=0}^{L_{CAP}} \sum_{j=1}^{N_0} h_l^2 + \frac{P_1 N_s}{2 W N_0} 2 \gamma_{UWB} V C_1^{(1)} + \sum_{j=0}^{\infty} C_{1,j,m} w_{j,m} \sum_{j=0}^{\infty} \sum_{m=1}^{N_0} \frac{\gamma_{UWB} V C_1^{(1)}}{2 W N_0},
\end{equation}
where $C_1^{(1)} = \sum_{m=1}^{2 WT} C_{1,j,m}$ are chi-square distributed r.v’s with $2 WT$ degrees of freedom, respectively. Differently from Type 1 interference, the interference observed in the integration intervals associated with bit 0 and bit 1 are no longer the same. Using the approximate analytical approach, the approximate BEP conditioned on $C_1^{(1)}$ and $C_1^{(2)}$ can be expressed as
\begin{equation}
P_{e,ED}^{(UWB)}(2) \approx \frac{1}{2} \frac{1}{\pi} \int_0^{\infty} \left( \frac{1}{1 + v^2} \right) \psi_{ED}(jv) \psi_{AP}(2) \psi_{ED}(jv) dv.
\end{equation}

V. Numerical results

In Fig. 1, the BEP performance of the ED receiver is plotted as a function of $WT$ for $E_b/N_0 = 20$ dB, $\text{SNR}_T = -20$ dB, and $\lambda = 0.01$. It can be noticed that the interference channel PDP with a higher $\epsilon$ results in lower performance degradation. This can be explained by the fact that with a steeper PDP, the interference signal energy is effectively concentrated in fewer multipath components and, thus, leads to a lower probability of collision. In Fig. 2, the effect of pulse repetition $T_p$ on the BEP performance of ED receiver. From these figures, we can clearly observe that better BEP performance is obtained for lower repetition rate due to lower probability of collision, given by $\frac{T_p}{T_s}$.
of the interference signals, and the signaling parameters of the interference systems. Using our statistical UWB NWI model, we evaluated the BEP performance of the ED receiver in different types of UWB NWI interference. Our proposed analytical framework allows a tractable BEP performance analysis and still provides valuable insight when planning the coexistence of UWB systems in wireless networks.

VI. CONCLUSIONS

In this paper, we investigated the effect of uncoordinated UWB NWI on the ED receiver. We first derived a statistical model of the aggregate NWI based on multivariate stable distribution, which takes into consideration the spatial distribution of the interference nodes, the propagation characteristics

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