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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]

$$s^{(k)}(t) = \sum_i \left[(1 - d_i^{(k)}) b_1^{(k)}(t - iT_s) + d_i^{(k)} b_2^{(k)}(t - iT_s) \right], \quad (1)$$

where $d_i^{(k)} \in \{0, 1\}$ is the i th data symbol and $T_s = \frac{N_s}{2} T_f$ is the symbol duration, such that T_f is the average pulse repetition period. The transmitted signal for $d_i^{(k)} = 0$ and $d_i^{(k)} = 1$ can be written, respectively, as

$$b_1^{(k)}(t) = \sum_{j=0}^{\frac{N_s}{2}-1} \sqrt{E_p} a_j^{(k)} p(t - jT_f - c_j^{(k)} T_p),$$

$$b_2^{(k)}(t) = \sum_{j=0}^{\frac{N_s}{2}-1} \sqrt{E_p} a_j^{(k)} p(t - jT_f - c_j^{(k)} T_p - \Delta), \quad (2)$$

where the parameter Δ 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 $E_p = \frac{2E_s}{N_s}$ where E_s is the symbol energy. To preclude intra-symbol interference (isi) and inter-symbol interference (ISI), we assume $\Delta \geq T_g$ and $(N_h - 1)T_p + \Delta + T_g \leq T_f$.

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

by

$$h(t) = \sum_{l=1}^L h_l \delta(t - \tau_l), \quad (3)$$

where h_l and τ_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., $|\tau_l - \tau_j| \geq T_p, \forall l \neq j$, where $\tau_l = \tau_1 + (l-1)T_p$ and $\{h_l\}_{l=1}^L$ are statistically independent r.v.'s. We can express $h_l = |h_l| \exp(j\phi_l)$, where $\phi_l = 0$ or π 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

$$\tilde{r}(t) = \sum_i \sum_{l=1}^L h_l [(1 - d_i) b_1(t - iT_s - \tau_l) + d_i b_2(t - iT_s - \tau_l)] + \tilde{n}(t), \quad (4)$$

where $\tilde{n}(t)$ represent the noise process after the BPZF.¹ 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

$$Z_{ED} = \underbrace{\sum_{j=0}^{\frac{N_s}{2}-1} \int_{jT_i+c_jT_p}^{jT_i+c_jT_p+T} (\tilde{r}(t))^2 dt}_{Z_{ED,1}} - \underbrace{\sum_{j=0}^{\frac{N_s}{2}-1} \int_{jT_i+c_jT_p+\Delta}^{jT_i+c_jT_p+T+\Delta} (\tilde{r}(t))^2 dt}_{Z_{ED,2}}, \quad (5)$$

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]

$$\mathbb{P}\{X_1 - X_2 < 0\} = \frac{1}{2} + \frac{1}{\pi} \int_0^\infty \left(\frac{1}{1+v^2} \right)^q \Re \left\{ \frac{\exp \left(\frac{-jv\mu_{X_1}}{1+jv} + \frac{jv\mu_{X_2}}{1-jv} \right)}{jv} \right\} dv. \quad (6)$$

Letting $q = q$, $X_1 = X_{ED,1}$, $X_2 = Z_{ED,2}$, $\mu_{X_1} = \mu$, and $\mu_{X_2} = 0$ in (6), and by further averaging with respect to μ , we obtain the BEP of the ED receiver for detecting binary pulse position modulation BPPM [8].

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 \mathcal{R} depends only on the area $A_{\mathcal{R}} = |\mathcal{R}|$, and is given by [11]

$$\mathbb{P}\{k \in \mathcal{R}\} = \frac{(\lambda A_{\mathcal{R}})^k}{k!} e^{-\lambda A_{\mathcal{R}}}, \quad (7)$$

where λ is the spatial density of the active devices.

¹The effect of the BPZF on $p(t)$ is considered negligible which means that no distortion is considered.

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

$$I^{(n)}(t) = \sqrt{P^I} \sum_i b_i^{(n)}(t - iN_s^I T_f^I), \quad (8)$$

where $b_i^{(n)}(t) \triangleq \sum_{j=1}^{\frac{N_s}{2}} e_i^{(n)} a_j^{(n)} p(t - jT_f - c_j^{(n)} T_p - d_i^{(n)} \Delta^I)$, $P^I \triangleq E_b^I / T_f^I N_s^I$ is the average power at the border of the near-field zone of each interfering transmitter antenna, and T_f^I 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 $T_s^I = T_f^I N_s^I$.² Note that we intentionally write (8) to account for two possible modulations, namely binary pulse amplitude modulation (BPAM) and BPPM modulation. The term $e_i^{(n)} \in \{-1, 1\}$ is the i th data symbol for BPAM modulation, $d_i^{(n)} \in \{0, 1\}$ is the i th data symbol for BPPM modulation, and Δ^I is the position modulation shift. The j th element of the random hopping and amplitude sequences are denoted by $\{c_j^{(n)}\}$ and $\{a_j^{(n)}\}$, where $0 < c_j^{(n)} < N_h^I$, N_h^I is the maximum shift associated with the hopping code, and $a_j^{(n)} \in \{-1, +1\}$ for all j and n . The average pulse repetition interval is considered long enough such that ISI and ISI can be ignored. For notational convenience, we define $\Psi^{(n)} \triangleq \{\{e_i^{(n)}\}, \{c_j^{(n)}\}, \{a_j^{(n)}\}, \{h_l^{(n)}\}\}$.

Using the spatial model in (7), the aggregate UWB interference signals received at the output of the BPZF of the desired user is given by

$$\zeta(t) = \sum_{n=1}^{\infty} \zeta^{(n)}(t), \quad (9)$$

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

$$\zeta^{(n)}(t) = \frac{e^{\sigma_I G^{(n)}}}{(R^{(n)})^\nu} \sqrt{P^I} v^{(n)}(t - D^{(n)}), \quad (10)$$

where the shadowing term $e^{\sigma_I G^{(n)}}$ follows a log-normal distribution with shadowing parameter σ_I and $G^{(n)} \sim \mathcal{N}(0, 1)$.³ According to the far-field assumption, the signal power decays as $1/(R^{(n)})^{2\nu}$, where $R^{(n)}$ is the distance between the n th UWB interferer and the desired user and ν is the amplitude 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

$$v^{(n)}(t) = \sum_i b_i^{(n)}(t - iN_s^I T_f^I) * h^{(n)}(t - iN_s^I T_f^I), \quad (11)$$

where $h^{(n)}(t) = \sum_{l=0}^{L-1} |h_l^{(n)}| e^{-j\phi_l^{(n)}} \delta(t - \tau_l^{(n)})$ is the channel impulse response of the n th UWB interferer-receiver link.⁴

²Furthermore, we assume that all UWB interferers use the same pulse waveform as the desired signal and their signals are undistorted at the output of the BPZF.

³We use $\mathcal{N}(0, \sigma^2)$ to denote a Gaussian distribution with zero-mean and variance σ^2 .

⁴For simplicity, we consider the channels from all UWB interferers have the same maximum excess delay T_g^I .

B. ED receiver

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

$$\mu_{Y_{ED,1}}^{(UWB)} \triangleq \underbrace{\frac{E_s}{N_0} \sum_{l=1}^{L_{CAP}} h_l^2}_{\triangleq \mu_{A,ED}} + \underbrace{\sum_{j=0}^{\frac{N_s}{2}-1} \sum_{m=0}^{2WT} \frac{1}{N_0} \frac{\zeta_{1,j,m}^2}{2W}}_{\triangleq \mu_{B,ED}^{(UWB)}} + \underbrace{\sum_{j=0}^{\frac{N_s}{2}-1} \sum_{m=1}^{2WT} \frac{2}{N_0} \frac{\zeta_{1,j,m} w_{j,m}}{2W}}_{\triangleq \mu_{C,ED}^{(UWB)}}, \quad (12)$$

$$\mu_{Y_{ED,2}}^{(UWB)} \triangleq \frac{1}{N_0} \sum_{j=0}^{\frac{N_s}{2}-1} \sum_{m=1}^{2WT} \frac{\zeta_{2,j,m}^2}{2W}, \quad (13)$$

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 $\zeta_{1,j}(t) \triangleq \zeta(t + jT_f + c_j T_p)$ and $\zeta_{2,j}(t) \triangleq \zeta(t + jT_f + c_j T_p + \Delta)$ 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 $\bar{\zeta}_{1,j}$ which composed of WT 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 0 position. Specifically, the vector $\bar{\zeta}_{1,j}$ can be written as

$$\bar{\zeta}_{1,j} = \sum_{n=1}^{\infty} \frac{e^{\sigma_1 G^{(n)}}}{(R^{(n)})^p} \bar{\mathbf{v}}_{1,j}^{(n)}, \quad (14)$$

where $\bar{\mathbf{v}}_{1,j}^{(n)}$ is the vector of complex samples of the equivalent low-pass version of $v^{(n)}(t + c_j T_p + jT_f - D^{(n)})$, such that $\mathbf{v}_{1,j,m}^{(n)}$ 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 present in the sampling instant m , each sample can be written as

$$\mathbf{v}_{1,j,m}^{(n)} = p \left(\tau^{(n)} \right) h_m^{(n)} \sqrt{\exp(-\epsilon^I(m - T^{(n)}))} \Theta_m^{(n)}, \quad (15)$$

where $\tau^{(n)} \triangleq (D^{(n)} \bmod T_p)$ is a r.v. uniformly distributed over $[0, T_p]$, $T^{(n)}$ is a discrete r.v. uniformly distributed over $[0, 1, \dots, L-1]$, $h_m^{(n)}$ is a r.v. with variance $1/\sum_{l=1}^L \exp(-\epsilon^I(l))$ and distributed according to the small-scale fading, and $\Theta_m^{(n)} = \cos(\phi_m^{(n)}) - j \sin(\phi_m^{(n)})$ with $\phi_m^{(n)}$ uniformly distributed over $[0, 2\pi]$.⁵ Considering that the complex r.v. $\Theta_m^{(n)}$ is circularly symmetric (CS), as for the case in the presence of narrowband interference [8], $\zeta_{1,j,m}$ can be described by a stable complex distribution as follows

$$\zeta_{1,j,m} \sim S_c \left(\frac{2}{\nu}, 0, \gamma_{UWB} \right), \quad (16)$$

where $\zeta_{1,j,m}$ is the m th complex sample of $\bar{\zeta}_{1,j}$ in (14) and $\gamma_{UWB} \triangleq \lambda \pi C_{2/\nu}^{-1} e^{2\sigma_1^2/\nu^2} \mathbb{E} \{ |\Re\{\mathbf{v}_{1,j,m}\}|^{2/\nu} \}$, such that

⁵Since the low-pass equivalent version of a signal is complex, we considered the phase of each multipath component uniformly distributed over $[0, 2\pi)$.

$\mathbb{E} \{ |\Re\{\mathbf{v}_{1,j,m}\}|^{2/\nu} \} = \frac{T_f^1}{T_f^I} MFP$ and the associated parameters M, F, P are given by

$$M = \mathbb{E} \{ |h_m^{(n)}|^{2/\nu} \} \mathbb{E} \left\{ \left(\sqrt{\exp(-\epsilon T^{(n)})} \right)^{2/\nu} \right\}$$

$$F = \mathbb{E}_n \{ |e_i^{(n)} a_j^{(n)}|^{2/\nu} \} \mathbb{E} \{ |\cos(\phi_m)|^{2/\nu} \},$$

$$P = \mathbb{E} \{ |p(\tau^{(n)})|^{2/\nu} \}.$$

Note that the components of the aggregate interference vector $\bar{\zeta}_{1,j}$ in (14) are identically distributed but mutually dependent [12].⁶ To make our analysis tractable, we assume that the $S\alpha S$ vector $\bar{\zeta}_{1,j}$ is spherically symmetric since spherically symmetric vectors have the characteristic of being sub-Gaussian, which implies that they can be decomposed as

$$\bar{\zeta}_{1,j} = \sqrt{V} \bar{\mathbf{G}}_{1,j}, \quad (17)$$

where $V \sim \mathcal{S}(\alpha/2, 1, \cos(\frac{\pi\alpha}{4}))$ and $\bar{\mathbf{G}}_{1,j}$ is a multivariate Gaussian random vector with covariance matrix $\bar{\Sigma}$. Unfortunately, $\bar{\zeta}_{1,j}$ is spherically symmetric only for some scenario. For example, it can be shown that the $\bar{\zeta}_{1,j}$ is sub-Gaussian if we consider Rayleigh fading with flat power delay profile, uniform phase, and $T_g^1 = T_f^1$. To ensure the spherical symmetry of the resulting aggregate interference vector for more general scenario, we modify each received interference signal as

$$\mathbf{v}_{1,j}'^{(n)}(t) = z^{(n)} d_\alpha^{-\alpha} \sum_{m=1}^{WT_f^1} \mathbf{G}_{1,j,m} p(t - mT_p), \quad (18)$$

where $d_\alpha^\alpha = 2^{\alpha/2} \pi^{-1/2} (\Gamma(\frac{\alpha+1}{2}))^{-1}$ corresponds to $\mathbb{E} \{ |\mathbf{G}_{1,j,m}|^\alpha \}$, $\{\mathbf{G}_{1,j,m}\}_{m=1}^{WT_f^1}$ is a sequence of i.i.d complex Gaussian r.v.'s with zero mean and unit variance, and

$$\mathbb{E} \{ |z^{(n)}|^\alpha \} = \frac{T_f^1}{T_f^I} M \times F. \quad (19)$$

Note that each interfering UWB signal now covers the entire frame interval T_f^1 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 $1 - \frac{T_f^1}{T_f^I}$. 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 $n_1 T_f^1 = \Delta$ and $n_2 T_f^1 = T_f$ such that n_1 and n_2 ($n_2 > n_1$) are integers. For simplicity, we consider no modulation is used and no random amplitude sequences and hopping code sequences are used.⁷ If the vector representing

⁶In fact, the aggregate interference vector in (14) is symmetric alpha stable ($S\alpha S$)

⁷With 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

$$\mu_{Y_{ED,1}}^{(UWB)} \quad (20)$$

$$= \underbrace{\frac{E_s}{N_0} \sum_{l=1}^{L_{CAP}} h_l^2}_{\triangleq \mu_{A,ED}} + \underbrace{\frac{P_1 N_s}{4WN_0} 2\gamma_{UWB}^\nu V C_1^{(1)}}_{\triangleq \mu_{B,ED}^{(UWB)}} + \underbrace{\sum_{j=0}^{\frac{N_s}{2}-1} \sum_{m=1}^{2WT} \frac{\zeta_{1,j,m} w_{j,m}}{2WN_0}}_{\triangleq \mu_{C,ED}^{(UWB)}}, \quad (21)$$

$$\mu_{Y_{ED,2}}^{(UWB)} = \mu_{B,ED}^{(UWB)},$$

where $C_1^{(1)} = \sum_{m=1}^{2WT} G_{1,j,m}^2$ is a chi-square r.v. with $2WT$ 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 $\mu_{C,ED}^{(UWB)}$ negligible compared to the other first two terms in (20) to obtain the conditional BEP as [8]

$$P_{e,ED|C_1^{(1)}}^{(UWB)} \simeq \frac{1}{2} + \frac{1}{\pi} \int_0^\infty \left(\frac{1}{1+v^2} \right)^{q_{TR}} \times \Re \left\{ \frac{\psi_{\mu_{ED}^{(1)}} \left(\frac{-jv}{1+jv} \right) \psi_{A^2|C_1^{(1)}} \left(g_{ED}^{(1)}(jv) \right)}{jv} \right\} dv, \quad (22)$$

where $\psi_{A^2|C_1^{(1)}}(jv)$ is given by

$$\psi_{A^2|C_1^{(1)}}(jv) = \exp \left[-(2C_1^{(1)})^{1/\nu} \gamma_{UWB} \cos \left(\frac{\pi}{2\nu} \right) |jv|^{1/\nu} \times \left(1 - \frac{jv}{|jv|} \tan \left(\frac{\pi}{2\nu} \right) \right) \right], \quad (23)$$

and

$$g_{ED}^{(1)}(jv) = \frac{P_1 N_s}{4WN_0} \left(\frac{-jv}{1+jv} + \frac{jv}{1-jv} \right). \quad (24)$$

In addition, using Gamma distribution as an approximation of the process $(C_1^{(1)})^{1/\nu}$ in (23) [8], the approximate BEP of the ED receiver for detecting BPPM in the presence of UWB Type 1 interference is given by

$$P_{e,ED}^{(UWB)} \simeq \frac{1}{2} + \frac{1}{\pi} \int_0^\infty \left(\frac{1}{1+v^2} \right)^{q_{ED}} \times \Re \left\{ \frac{\psi_{\mu_{ED}^{(1)}} \left(\frac{-jv}{1+jv} \right) \psi_{A^2} \left(g_{ED}^{(1)}(jv) \right)}{jv} \right\} dv. \quad (25)$$

B. Type 2 interference

We assume that the vector $\bar{\zeta}$ is sub-Gaussian, representing the aggregate interference signal along the entire symbol duration T_s . The covariance matrix of the Gaussian vector $\bar{\zeta}$ is still diagonal. The non-centrality terms of $Y_{ED,1}$ and $Y_{ED,2}$ for $d_0 = 0$ are, respectively, given by

$$\mu_{Y_{ED,1}}^{(UWB)} = \underbrace{\frac{E_s}{N_0} \sum_{l=1}^{L_{CAP}} h_l^2}_{\triangleq \mu_{A,ED}} + \underbrace{\frac{P_1}{2WN_0} 2\gamma_{UWB}^\nu V C_1^{(2)}}_{\triangleq \mu_{B,ED}^{(UWB)}} + \underbrace{\sum_{j=0}^{\frac{N_s}{2}-1} \sum_{m=1}^{2WT} \frac{\zeta_{1,j,m} w_{j,m}}{2WN_0}}_{\triangleq \mu_{C,ED}^{(UWB)}}, \quad (26)$$

$$\mu_{Y_{ED,2}}^{(UWB)} = \frac{P_1}{2WN_0} 2\gamma_{UWB}^\nu V C_2^{(2)}, \quad (27)$$

where $C_1^{(2)} = \sum_{j=0}^{N_s/2-1} \sum_{m=1}^{2WT} G_{1,j,m}^2$ and $C_2^{(2)} = \sum_{j=0}^{N_s/2-1} \sum_{m=1}^{2WT} G_{2,j,m}^2$ are chi-square distributed r.v.'s with $\frac{N_s}{2} 2WT$ 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 $C_1^{(2)}$ and $C_2^{(2)}$ can be expressed as

$$P_{e,ED|C_1^{(2)}, C_2^{(2)}}^{(UWB)} \simeq \frac{1}{2} + \frac{1}{\pi} \int_0^\infty \left(\frac{1}{1+v^2} \right)^{q_{ED}} \times \Re \left\{ \frac{\psi_{\mu_{ED}^{(2)}} \left(\frac{-jv}{1+jv} \right) \psi_V \left(g_{ED|C_1^{(2)}, C_2^{(2)}}^{(2)}(jv) 2\gamma_{UWB}^\nu \right)}{jv} \right\} dv, \quad (28)$$

where $\psi_V(jv)$ is the CF of the stable r.v. V and

$$g_{ED|C_1^{(2)}, C_2^{(2)}}^{(2)}(jv) = \frac{P_1}{2WN_0} V \left[C_1^{(2)} \frac{-jv}{1+jv} + C_2^{(2)} \frac{jv}{1-jv} \right] \quad (29)$$

V. NUMERICAL RESULTS

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 $N_s = 32$. For simplicity, Δ is set such that there is no ISI or ISI in the system, i.e., $T_f = 2\Delta$ with $\Delta > T_g - N_h T_p$. We consider a TH sequence of all ones ($c_j = 1$ for all j) and $N_h = 2$. The desired signal is affected by a dense resolvable multipath channel, where each multipath amplitude is Nakagami distributed with fading severity index m and average power $\mathbb{E}\{h_l^2\}$, where $\mathbb{E}\{h_l^2\} = \mathbb{E}\{h_1^2\} \exp[-\epsilon(l-1)]$, for $l = 1, \dots, L$, are normalized such that $\sum_{l=1}^L \mathbb{E}\{h_l^2\} = 1$. For simplicity, the fading severity index m is assumed to be identical for all paths. The average power of the first arriving multipath component is given by $\mathbb{E}\{h_1^2\}$ and ϵ is the channel power decay factor. With this model, we denote the channel characteristic of the desired signal by (L, ϵ, m) for convenience. For the UWB interferers, they use the same waveform as the signal of interest with Nakagami fading channels and severity index m^I and average power $\mathbb{E}\{|h_l^I|^2\}$.

A. Type 1 interference

In Fig. 1, the BEP performance of the ED receiver is plotted as a function of WT for $E_b/N_0 = 20$ dB, $SIR_T = -20$ dB, and $\lambda = 0.01$. It can be noticed that the interference channel PDP with a higher ϵ^I results in lesser 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_f^I 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 $\frac{T_f^I}{T_f}$.

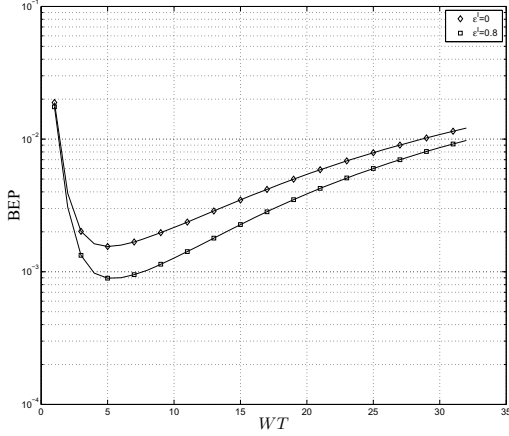


Fig. 1. BEP performance of the ED receiver in the presence of Type 1 interference as a function of WT for $E_b/N_0 = 20$ dB, $SIR_T = -20$ dB, $(L, \epsilon, m) = (32, 0.4, 3)$, $(L^I, \epsilon^I, m^I) = (32, \epsilon^I, 3)$, $T_f^I = 50$ ns, $\lambda = 0.01$, $\nu = 1.5$, and $\sigma_I = 1.6$ dB.

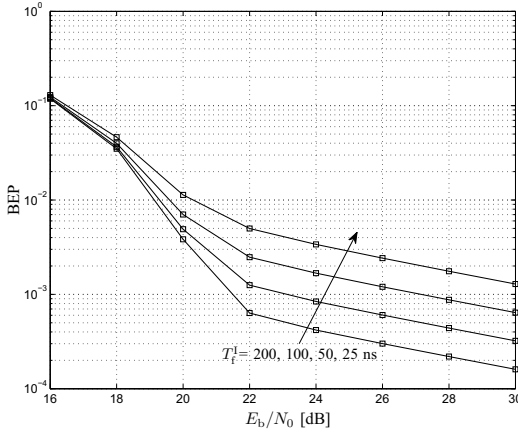


Fig. 2. Effect of pulse repetition interval T_f^I on the BEP performance of the ED receiver in the presence of Type 1 interference for $E_b/N_0 = 20$ dB, $SIR_T = -20$ dB, $(L, \epsilon, m) = (32, 0, 3)$, $(L^I, \epsilon^I, m^I) = (32, 0, 3)$, $\lambda = 0.01$, $\nu = 1.5$, and $\sigma_I = 1.6$ dB.

B. Type 2 interference

The numerical results below are obtained by averaging over many realizations of the variables $C_1^{(2)}$ and $C_2^{(2)}$. In Fig. 3, the performance of ED receiver is presented for $\lambda = 0.1$. It can be seen that when the effect of interference becomes dominant, Type 2 interference rapidly leads to the saturation of the BEP curves. This phenomenon can be explained by the fact that, in the case of Type 2 interference, the integrated interference energy in the bit positions 0 and 1 are no longer equivalent and this increases the BEP, especially when the interference effect is dominant. From the figure, we see that the BEP performance is better for Type 1 interference compared to Type 2 interference.

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

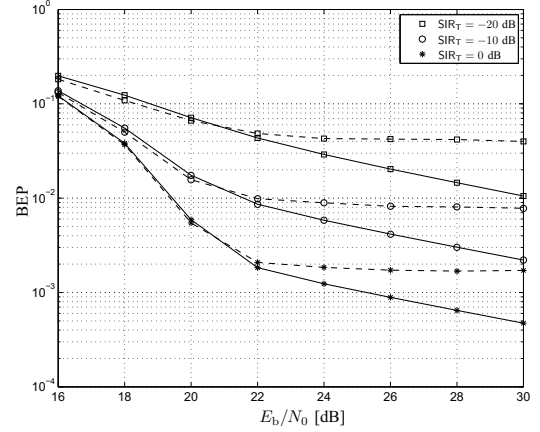


Fig. 3. Comparison of the effect of Type 1 interference (solid line) and Type 2 interference (dashed line) on the BEP performance of the ED receiver for $(L, \epsilon, m) = (32, 0, 3)$, $(L^I, \epsilon^I, m^I) = (32, 0, 3)$, $T_f^I = 50$ ns, $\nu = 1.5$, and $\sigma_I = 1.6$ dB.

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.

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