Narrowband Interference Mitigation in Impulse Radio

Itsik Bergel, Eran Fishler, and Hagit Messer

***Abstract*—Impulse radio (IR) systems have drawn attention during the last few years. These systems are planned to coexist with narrowband systems without interfering them. Nevertheless, the narrowband systems can cause interference which may jam the IR receiver. This letter analyzes a low-complexity narrowband interference (NBI)-mitigation algorithm for IR systems, based on minimal mean-square error combining. Theoretical analysis reveals that these algorithms nearly eliminate the NBI. The concept is also extended to the case where the receiver has more correlators than channel taps.**

***Index Terms*—Impulse radio (IR), minimal mean-square error (MMSE), narrowband interference (NBI) suppression, ultra wideband (UWB).**

I. INTRODUCTION

ULTRA-WIDEBAND (UWB) systems have drawn attention, both from researchers and practitioners, in the last years. Out of the various systems suggested to span this UWB spectrum, the one that drew the most attention is the impulse radio (IR), which transmits a train of very short pulses. IR systems were suggested in the early 1990s [1], [2], and were analyzed extensively in the last few years (see, among many others, [2]–[5]).

These systems are planned to operate in coexistence with other narrowband systems over the same frequency band, making narrowband interference (NBI) cancellation extremely important. In spite of the IR system's large processing gain, in many cases, the power of an NBI will be high, and an NBI-suppression algorithm will have to be used. NBI-suppression algorithms has been investigated thoroughly for wideband systems, and mainly for code-division multiple-access (CDMA) systems. These techniques include ones that are incorporated into the despreading operation [6], [7], and others that are based on implementation of notch filters to suppress the NBI [8], [9]. However, the suggested solutions are not applicable for low-complexity IR systems because of the increase in the receiver complexity.

Low-complexity NBI suppression in IR can be implemented by a minimal mean-square error (MMSE)-RAKE receiver, as suggested in [10]–[12]. This method is based on the traditional RAKE receiver, where the correlators' outputs are linearly combined. The weights, however, are chosen to achieve NBI suppression.

So far, the performance of the MMSE-RAKE receiver was analyzed only by simulation.

In this letter, we present an analytic performance analysis of the MMSE-RAKE receiver for NBI mitigation. In addition, we extend the concept of the MMSE-RAKE receiver to the case where the number of correlators in the system is larger than the number of resolvable channel taps, and derive a lower bound on the performance in this case.

II. UWB SYSTEM MODEL

A. Received Signal Model

A common model [4], [13] for the received signal in a pulse position modulation (PPM)[1] IR system is given by

$$r(t)=\sum_{j=-\infty}^{\infty}\left\{\sum_{l=0}^{L-1}\left\{\beta_l\cdot w_{rx}\left(t-jT_f - T_j - d_{\lceil\frac{j}{N_s}\rceil}\delta-\tau_1\right)\right\}\right\}+n(t) \quad (1)$$

where T_f is the frame time, T_j is a time-hopping sequence, and N_s is the number of transmitted pulses corresponding to one information symbol. $d_k\{-1,1\}$ is the information of the kth symbol, and $w_{rx}(t)$ is the received pulse shape. The propagation channel is modeled as a tapped delay line with L taps [14], [15], where β_l and τ_l are, respectively, the amplitude and delay of the lth path arriving at the receiver. The additive noise process $n(t)$ is the superposition of two independent noise processes, both modeled as zero-mean Gaussian random processes. A white-noise term, with spectral density $N_0/2$, and an NBI with spectral density $S_{n_i}(\omega)=0\,\forall||\omega/2\pi|-f_c>(\mathrm{BW}/2)$. f_c and BW are, respectively, the interference's center frequency and bandwidth.

The receiver uses the common RAKE receiver structure, where linear combining is applied on the outputs of a set of N correlators (termed fingers) [16]. The nth correlator is delayed by $\tilde{\tau}_n$, producing for the jth pulse the output

$$y_{nj}=\int r(t)v(t-jT_f-T_j-\tilde{\tau}_n)\mathrm{d}t=d_{\lfloor\frac{j}{N_s}\rfloor}\tilde{\beta}_n+n_{nj} \quad (2)$$

and the optimal correlating function is $v(t)=w_{rx}(t-\delta)-w_{rx}(t+\delta)$. We will assume that $\tilde{\tau}_n=\tau_n$ for $n=0,\ldots,L-1$, and if $N>L$, then $\tilde{\tau}\sim U[0,T_R)$. Later, when we discuss several NBI-suppression algorithms, these choices of random delay will become clear.

The second term on the right-hand side of (2), $n_{nj}=\int_{-\infty}^{\infty}n(t+jT_f+T_j+\tilde{\tau}_n)v(t)\mathrm{d}t$, is the sampled noise, and $\tilde{\beta}_n$ is the sampled channel gain. Assuming that the correlators'

[1]For simplicity, this letter considers only binary PPM, however, all results are applicable for binary pulse amplitude modulation (PAM), as well. See [10].

Paper approved by M. Z. Win, the Editor for Equalization and Diversity of the IEEE Communications Society. Manuscript received February 3, 2003; revised June 30, 2004 and September 15, 2004. This work was supported in part by the Israeli Software Radio Consortium. This paper was presented in part at the IEEE Conference on Ultra-Wideband Systems and Technologies, Baltimore, MD, May 2002.

I. Bergel and H. Messer are with the Department of Electrical Engineering-Systems, Tel Aviv University, Tel Aviv 69978, Israel (e-mail: ibergel@eng.tau.ac.il; messer@eng.tau.ac.il).

E. Fishler was with the Department of Electrical Engineering, Princeton University, Princeton, NJ 08544 USA. He is now with The Stern School of Business, New York University, New York, NY 10003 USA (e-mail: ef485@stern.nyu.edu).

Digital Object Identifier 10.1109/TCOMM.2005.852819

relative delays are set sufficiently apart, the sampled channel gain is $\tilde{\beta}_n = \beta_n \int_{-\infty}^{\infty} w_{rx}(t-\delta)v(t)\mathrm{d}t$ for $0 \le n \le L-1$, and $\tilde{\beta}_n = 0$ for $n \ge L$.

Because of the NBI, the RAKE receiver structure is not optimal. Nevertheless, for complexity reasons, this letter focuses on this receiver with a finite number of correlators, where the only degree of freedom is the choice of the weights $\mathbf{c} \triangleq [c_0 c_1 \cdots c_{N-1}]^T$. The RAKE output for the jth pulse is, therefore, $y_j = \mathbf{c}^T(d_{\lfloor j/N_s \rfloor}\tilde{\boldsymbol{\beta}} + \mathbf{n}_j)$, where $\tilde{\boldsymbol{\beta}} = [\tilde{\beta}_0 \tilde{\beta}_1 \cdots \tilde{\beta}_{N-1}]^T$ is the channel gain vector, and $\mathbf{n}_j = [n_{0j} n_{2j} \cdots n_{N-1j}]^T$ is the sampled noise vector. The resulting signal-to-noise ratio (SNR), given the channel amplitude, is

$$\mathrm{SNR} = \frac{|\mathbf{c}^T\tilde{\boldsymbol{\beta}}|^2}{E\{|\mathbf{c}^T\mathbf{n}|^2\}} = \frac{|\mathbf{c}^T\tilde{\boldsymbol{\beta}}|^2}{\mathbf{c}^T\mathbf{R}_\mathbf{n}\mathbf{c}^2}. \tag{3}$$

B. Noise Correlation Matrix

Denote by $\mathbf{R}_n \triangleq E\{\mathbf{n}_j\mathbf{n}_j^T\}$ the correlation matrix of the sampled noise vector. We suggest an approximate statistical model for the correlation matrix that is valid for a wide range of NBI and channel models, based only on the following two assumptions. One is that the channel delay spread is small, compared with the interference coherence time 1/BW; and the second is that the channel delay spread is large, compared with $1/f_c$. The delay spread of a typical indoor channel is less than 100 ns [17], therefore, these assumptions hold for NBI with bandwidth of up to a few megahertz, and center frequency above 10 MHz.

Since $V(\omega) \triangleq \mathcal{F}(v(t))$ (where $\mathcal{F}$ denotes the Fourier transform) is very wide and well behaved, the (m,n) element of $\mathbf{R}_\mathbf{n}$ can be approximated as

$$\begin{aligned}[\mathbf{R}_\mathbf{n}]_{mn} &= \int_{-\infty}^{\infty}\int_{-\infty}^{\infty} R_n(t_1-\tau_m-t_2+\tau_n)v(t_1)v(t_2)\mathrm{d}t_1\mathrm{d}t_2 \\ &\simeq |V(\omega_c)|^2 R_{n_i}(\tau_n-\tau_m) + \frac{N_0}{2}R_v(\tau_n-\tau_m) \end{aligned} \tag{4}$$

where $R_{n_i}(\tau) = \mathcal{F}^{-1}(S_{n_i}(\omega))$ is the interference autocorrelation function, and $R_v(\tau) \triangleq \int_{-\infty}^{\infty} v(t)v(t-\tau)\mathrm{d}t$ is the correlation function of $v(t)$. Assuming that the correlators are separated enough, $R_v(\tau_n-\tau_m) = 0$ for $m \ne n$, and using assumption (1), for $\tau \ll (1/\mathrm{BW})$, the autocorrelation function can be approximated as $R_{n_i}(\tau) \approx P_I\cos(\omega_c\tau)$, where $P_I = \int_{-\infty}^{\infty} S_{n_i}(\omega)\mathrm{d}\omega$ is the interference power.

Define the vector $\mathbf{z} = [e^{j\phi 0} \cdots e^{j\phi L-1}]^T$, where $\phi_j = (2\pi f_c\tilde{\tau}_j)_{2\pi}$, and $(\cdot)_x$ denotes the modulo x operation. Then, $\mathbf{R}_\mathbf{n}$ can be approximated by

$$\mathbf{R}_\mathbf{n} \approx \frac{N_0}{2}R_v(0)\left(\mathbf{I} + \eta\Re(\mathbf{z}\mathbf{z}^H)\right) \tag{5}$$

where η is the interference-to-noise power ratio (INR) given by

$$\eta = \frac{P_I}{\frac{N_0}{2}R_v(0)}|V(f_c)|^2. \tag{6}$$

Adding to assumption (2) the assumption that $T_R \gg 1/f_c$ allows approximating the distribution of $\{\phi_j\}$ as independent identically distributed (i.i.d.) random variables, uniformly distributed in $[0, 2\pi)$, which complete the statistical model of the sampled noise vector.

III. MMSE-RAKE Receiver

The MMSE-RAKE receiver uses weights that are optimized for achieving maximum instantaneous SNR per pulse. In this case, the weights are given by $\mathbf{c} = \alpha\mathbf{R}_\mathbf{n}^{-1}\tilde{\boldsymbol{\beta}}$, where α is any positive constant [18, p. 298]. The resulting instantaneous SNR, given the channel state, is $\mathrm{SNR} = \tilde{\boldsymbol{\beta}}\mathbf{R}_\mathbf{n}^{-1}\tilde{\boldsymbol{\beta}}$.

This weight vector is optimal when the transmitter uses one pulse per symbol ($N_s = 1$). When $N_s > 1$, the RAKE receiver output for the ith bit is $\tilde{y}_i = \sum_{j=iN_s}^{(i+1)N_s-1} y_j$. In this case, achieving the optimal SNR per bit requires consideration of the correlation among the various y_j. Since the correlation changes from bit to bit due to the random time-hopping sequence, the combiner weights have to be recalculated at the bit rate. Such receiver is impractical, and therefore, we still consider the receiver that maximizes the instantaneous SNR for each pulse separately. It can also be shown [19] that ignoring the interpulse correlation results in a lower bound on the system performance.

In this letter, we use the average SNR as the figure of merit for the system performance. The average SNR reflects very well the performance differences between a system that does use an NBI-suppression algorithm and those that do not. In addition, the average SNR can produce a lower bound on the uncoded bit-error rate (BER), given by

$$\begin{aligned}\mathrm{BER} &= E\left\{Q\left(\sqrt{\frac{1-\rho}{2}\mathrm{SNR}}\right)\right\} \\ &\ge Q\left(\sqrt{\frac{1-\rho}{2}E\{\mathrm{SNR}\}}\right)\end{aligned} \tag{7}$$

where ρ is the correlation between the modulated signals, and the inequality results from the easily verified convexity of $Q(\sqrt{x})$ for $x \ge 0$.

The following proposition evaluates the average SNR of the MMSE-RAKE receiver.

Proposition 1: The average SNR of the MMSE-RAKE receiver is given by

$$\overline{\mathrm{SNR}}_{\mathrm{MMSE}} = \frac{2E_s}{N_0}\epsilon^2(N,\eta) \tag{8}$$

where $E_s = E\{\tilde{\boldsymbol{\beta}}^T\tilde{\boldsymbol{\beta}}/R_v(0)\}$ is the effective symbol energy, and $0 \le \epsilon^2(N,\eta) \le 1$ is a function of the number of correlators N and the INR η.

Proof of Proposition 1: The average SNR is

$$\begin{aligned}\overline{\mathrm{SNR}}_{\mathrm{MMSE}} &= \mathbf{E}_{\tilde{\boldsymbol{\beta}}}\left\{E\{\mathrm{SNR}|\tilde{\boldsymbol{\beta}}\}\right\} \\ &= E\left\{\tilde{\boldsymbol{\beta}}^T\mathbf{R}_\mathbf{n}^{-1}\tilde{\boldsymbol{\beta}}\right\} \\ &= \mathbf{E}_{\tilde{\boldsymbol{\beta}}}\left\{\tilde{\boldsymbol{\beta}}^T E\left\{\mathbf{R}_\mathbf{n}^{-1}\right\}\tilde{\boldsymbol{\beta}}\right\}.\end{aligned} \tag{9}$$

The proof is based on showing that $E\{\mathbf{R}_\mathbf{n}^{-1}\}$ is a scaled identity matrix, which, combined with the definition of E_s, proves the proposition. It can be verified by direct computation that $\mathbf{R}_{\mathbf{n}}^{-1}$ is given by

$$\mathbf{R}_{\mathbf{n}}^{-1} = \frac{2}{N_0 R_v(0)} \times \left(\mathbf{I} - \frac{\left(\frac{1}{\eta} + 0.5\mathbf{z}^H\mathbf{z}\right)\Re(\mathbf{z}\mathbf{z}^H) - 0.5\Re(\mathbf{z}\mathbf{z}^H\mathbf{z}^*\mathbf{z}^T)}{\left(\frac{1}{\eta} + 0.5\mathbf{z}^H\mathbf{z}\right)^2 - 0.25|\mathbf{z}^H\mathbf{z}^*|^2} \right). \quad (10)$$

Inspecting one of the off-diagonal terms in the matrix, we can write

$$[\mathbf{R}_{\mathbf{n}}^{-1}]_{mn} = -\frac{2}{N_0 R_v(0)} \left(\frac{\left(\frac{1}{\eta} + 0.5N\right)\Re\left(z_m z_n^*\right)}{\left(\frac{1}{\eta} + 0.5N\right)^2 - 0.25\left|\sum_j \left(z_j^*\right)^2\right|^2} - \frac{0.5\Re\left(z_m z_n \sum_j \left(z_j^*\right)^2\right)}{\left(\frac{1}{\eta} + 0.5N\right)^2 - 0.25\left|\sum_j \left(z_j^*\right)^2\right|^2} \right) \quad \forall m \neq n. \quad (11)$$

The expectation of $[\mathbf{R}_{\mathbf{n}}^{-1}]_{mn}$ for $m \neq n$ can be written as

$$E\left\{[\mathbf{R}_{\mathbf{n}}^{-1}]_{mn}\right\} = \mathrm{E}_{z_1,\ldots,z_{n-1},z_{n+1},\ldots,z_N} \left\{\mathrm{E}_{z_n}\left\{[\mathbf{R}_{\mathbf{n}}^{-1}]_{mn} | z_1,\ldots,z_{n-1},z_{n+1},\ldots,z_N\right\}\right\}. \quad (12)$$

It is easily seen from (11) that

$$[\mathbf{R}_{\mathbf{n}}^{-1}]_{mn} | z_1,\ldots,z_{n-1},z_n,z_{n+1},\ldots,z_N = -[\mathbf{R}_{\mathbf{n}}^{-1}]_{mn} | z_1,\ldots,z_{n-1},-z_n,z_{n+1},\ldots,z_N \quad (13)$$

and since z_n is uniformly distributed on the unit circle, $E_{z_n}\{[\mathbf{R}_{\mathbf{n}}^{-1}]_{mn} | z_1,\ldots,z_{n-1},z_{n+1},\ldots,z_N\} = 0$, and hence, $E\{[\mathbf{R}_{\mathbf{n}}^{-1}]_{mn} = 0$ for $m \neq n$. This completes the first part of the proof, that is all the off-diagonal elements of $E\{\mathbf{R}_{\mathbf{n}}^{-1}\}$ are zero. Next, we check the mth diagonal element

$$E\left\{[\mathbf{R}_{\mathbf{n}}^{-1}]_{mn}\right\} = \frac{2}{N_0 R_v(0)}\epsilon^2(N,\eta) = \frac{2}{N_0 R_v(0)} \times \left(1 - E\left\{\frac{\frac{1}{\eta} + 0.5N - 0.5\Re\left(z_m^2 \sum_j \left(z_j^*\right)^2\right)}{\left(\frac{1}{\eta} + 0.5N\right)^2 - 0.25\left|\sum_j \left(z_j^*\right)^2\right|^2}\right\}\right) \quad (14)$$

and since z_m are identically distributed for all m, we conclude that all the diagonal elements of $E\{\mathbf{R}_{\mathbf{n}}^{-1}\}$ are equal, which completes the proof of the proposition.

Proposition 1 characterizes the degradation in the average SNR due to the existence of an NBI. When only white noise exists, the average SNR is equal to $2E_s/N_0$. As such, $\epsilon^2(N,\eta)$ represents the portion of the SNR lost due to the existence of the additional NBI.

Although there is no analytic solution for (14), it is important to note that $\epsilon^2(N,\eta))$ is a function of only two variables, and as such, it is easy to evaluate it numerically. Fig. 1 depicts $\epsilon^2(N,\eta)$ versus the number of correlators for several values of η. Additionally, we can analyze it in the extreme cases, as follows.

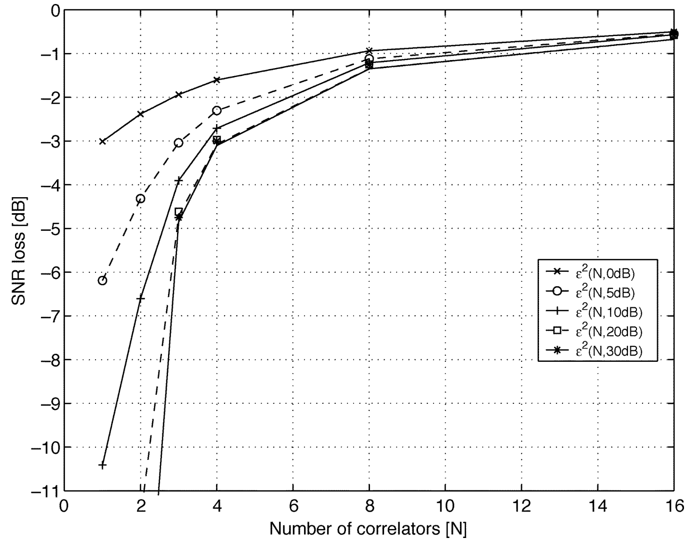

Fig. 1. Average SNR degradation due to the existence of an NBI ($\epsilon^2(N,\eta)$) versus the number of correlators used (N), for different values of the INR (η) [calculated from (14)].

If the number of correlators is one ($N = L = 1$), the weight vector is a scalar that multiplies the single correlator output, and as such, cannot affect the SNR. Indeed, using (14), $\epsilon^2(1,\eta) = 1/(1+\eta)$. Thus, no interference suppression is achieved.

On the other hand, if the number of correlators grows to infinity ($N \to \infty$), we get $\epsilon^2(N,\eta) \to 1$, that is, complete NBI suppression. This result is not surprising, as we note that a notch filter can achieve the same result by suppressing the interference with negligible effects on the signal.

The case where the number of correlators is equal to the number of channel taps is the common MMSE-RAKE receiver. In this case, $N = L$, and the average SNR is predicted very accurately by $2E_s\epsilon^2(L,\eta)/N_0$. Alternatively, if $N > L$, the $N - L$ additional delays are determined by the receiver. This receiver is termed generalized MMSE-RAKE (GMMSE-RAKE). In this case, the analysis result $2E_s\epsilon^2(N,\eta)/N_0$ is a lower bound on the achievable performance, since we assumed random selection of the additional delays.

IV. Simulations

Several simulations were performed in order to confirm the performance predicted by the theoretical calculations. Fig. 2 depicts the average SNR degradation due to the existence of an NBI as a function of the number of channel taps used by the MMSE-RAKE receiver. The graph depicts the theoretical, as well as the empirical, degradation in the average SNR, compared with the case where no interference exists. In the simulations, the NBI center frequency is 1.9 GHz, and the interference power spectrum is rectangular; the pulse shape was $w_{rx}(t) = \sqrt{8/3t_n}[1 - 4\pi(t/t_n)^2]e^{-2\pi(t/t_n)^2}$, with $t_n = 0.4472$ ns, as in [20]. The channel model is the one described in [17], slightly modified to allow for simulation of NBI by adding small jitter to the location of each channel tap. The SNR for each channel realization was calculated using (3), and then averaged over 10 000 channel realizations.

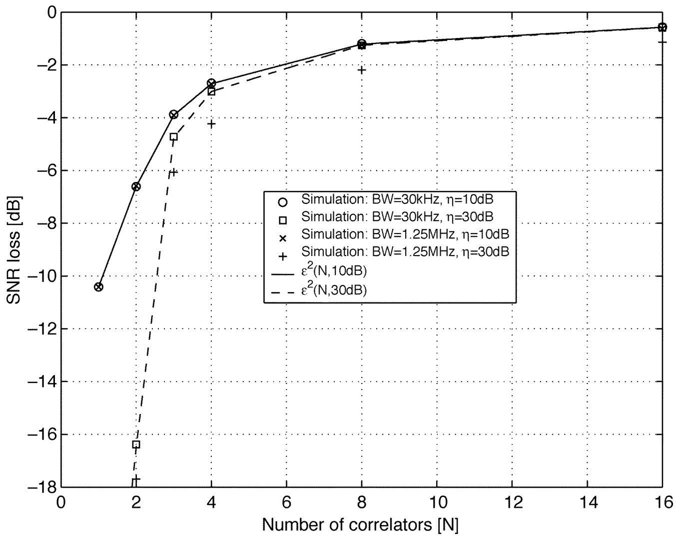

Fig. 2. Degradation in the average SNR of an MMSE-RAKE receiver due to the existence of NBI as a function of the number of correlators used.

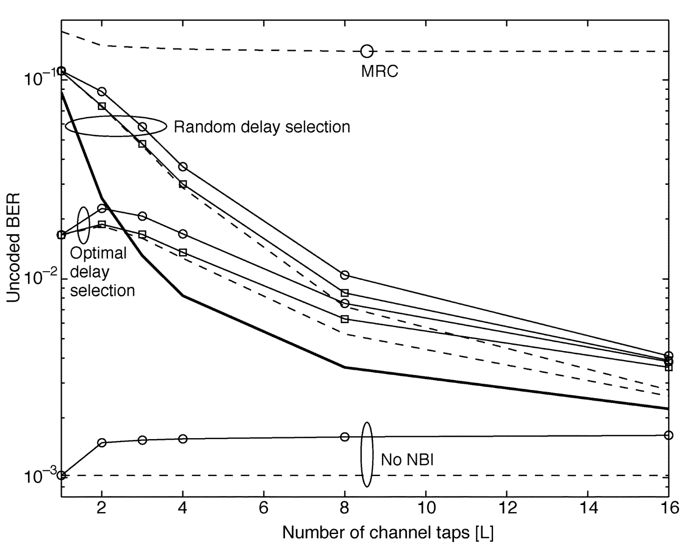

Fig. 3. BER of practical GMMSE systems with one extra correlator in the presence of NBI as a function of the number of channel taps. Figure shows performance of system with exact estimation of channel and NBI (dashed lines), wrong NBI parameters (square markers), imperfect channel estimates (circle markers), and average SNR bound (bold solid line).

It is easily seen from the figure that the theoretical results accurately predict the performance degradation due to the NBI. For very high interference levels ($\eta = 30$ dB) and large bandwidth, the difference between the theoretical results and the empirical ones is no more than 1 dB. Also, it is seen that by using four or more correlators and the MMSE weights, the NBI is almost completely suppressed.

In Fig. 3, we consider the GMMSE-RAKE receiver with one extra correlator $(N = L + 1)$ in the presence of strong NBI ($\eta = 10$ dB). As mentioned earlier, the first L correlators are assigned to the channel taps. The remaining correlators use either random delay or optimally selected delay. The figure depicts the uncoded BER as a function of the number of channel taps, in systems with perfect and imperfect channel and NBI information. For reference, the figure also shows the BER achieved with the conventional maximum ratio combiner receiver and the BER in the absence of NBI. The NBI in the simulations had a raised-cosine spectrum with rolloff factor 0.22, BW $= 1.25$ MHz, and $f_c = 1.9$ GHz. The systems with imperfect knowledge of this NBI had assumed a rectangular spectrum and $\hat{f}_c = f_c + 1$ MHz. It can be seen that both the random and the optimal delay selection had performed well, and shown low sensitivity to errors in the NBI's parameters. The imperfect channel estimates were modeled as a Gaussian error in the channel-taps amplitude estimation, with a variance of 10% of each channel tap power. Again, both receivers suffered only minor performance degradation, comparable to the performance degradation seen in the absence of NBI. We conclude that the GMMSE-RAKE receiver has better performance than MMSE-RAKE. In particular, when the number of channel taps is small, the addition of even one finger results in a significant performance improvement, which is even larger if this delay is selected in an optimal manner.

Fig. 3 also depicts the lower bound (solid line) calculated using (7) for the MMSE-RAKE receiver from the average SNR. As can be seen, this bound, although not tight, gives a good characterization of the performance of a GMMSE receiver with random delay selection.

V. Discussion and Summary

In this letter, the performance of the MMSE-RAKE receiver and the GMMSE-RAKE receiver for IR systems were analyzed in the presence of NBI. The analysis is valid for a wide range of channel models and NBI. Results show that as long as the receivers have enough correlators, the receivers achieve near-absolute NBI suppression. Both receivers require only a small modification of the traditional RAKE receiver to gain the capability to practically suppress any level of NBI.

Simulations also pointed out that optimal delay assignment in GMMSE-RAKE has significant advantage when the number of correlators is small. This receiver still requires detailed analysis.

References

[1] P. Withington and L. W. Fullerton, "An impulse radio communication system," in *Proc. Int. Conf. Ultra-Wideband, Short-Pulse Electromagn.*, Oct. 1992, pp. 113–120.
[2] R. A. Scholtz, "Multiple access with time-hopping impulse modulation," in *Proc. IEEE Military Commun. Conf.*, vol. 2, Boston, MA, Oct. 1993, pp. 447–450.
[3] F. R. Mireles, "Performance of ultra-wideband SSMA using time hopping and M-ary PPM," *IEEE J. Sel. Areas Commun.*, vol. 19, no. 6, pp. 1186–1196, Jun. 2001.
[4] M. Z. Win and R. A. Scholtz, "Impulse radio: How it works," *IEEE Commun. Lett.*, vol. 2, no. 2, pp. 36–38, Feb. 1998.
[5] M. Z. Win, "A unified spectral analysis of generalized time-hopping spread-spectrum signals in the presence of timing jitter," *IEEE J. Sel. Areas Commun.*, vol. 20, no. 9, pp. 1664–1676, Dec. 2002.
[6] F. Amoroso, "Adaptive A/D converter to suppress CW interference in DSPN spread spectrum communications," *IEEE Trans. Commun.*, vol. COM-31, no. 10, pp. 1117–1123, Oct. 1983.
[7] R. Vijayan and H. V. Poor, "Nonlinear techniques for interference suppression in spread-spectrum systems," *IEEE Trans. Commun.*, vol. 38, no. 7, pp. 1060–1065, Jul. 1990.
[8] L. B. Milstein, "Interference rejection techniques in spread spectrum communications," *Proc. IEEE*, vol. 76, no. 6, pp. 657–671, Jun. 1988.
[9] L.-M. Li and L. B. Milstein, "Rejection of narrowband interference in PN spread spectrum systems using transversal filters," *IEEE Trans. Commun.*, vol. COM-30, no. 5, pp. 925–928, May 1982.

[10] I. Bergel, E. Fishler, and H. Messer, "Narrowband interference supression in time hopping impulse radio systems," in *Proc. IEEE Conf. Ultra Wideband Syst. Technol.*, Baltimore, MD, May 2002, pp. 303–308.

[11] L. Qinghua and L. A. Rusch, "Multiuser detection for DS-CDMA UWB in the home environment," *IEEE J. Sel. Areas Commun.*, vol. 20, no. 9, pp. 1701–1711, Dec. 2002.

[12] N. Boubaker and K. B. Letaief, "A low-complexity MMSE-RAKE receiver in a realistic UWB channel and in the presence of NBI," in *Proc. IEEE Wireless Commun. Netw.*, vol. 1, New Orleans, LA, Mar. 2003, pp. 233–237.

[13] C. J. Le-Martret and G. B. Giannakis, "All-digital PPM impulse radio for multiple access through frequency-selective multipath," in *Proc. IEEE Global Telecommun. Conf.*, vol. 1, San Francisco, CA, Nov. 2000, pp. 77–81.

[14] M. Z. Win and R. A. Scholtz, "Ultra-wide bandwidth time-hopping spread-spectrum impulse radio for wireless multiple-access communications," *IEEE Trans. Commun.*, vol. 48, no. 4, pp. 679–689, Apr. 2000.

[15] M. Z. Win and R. A. Scholtz, "Characterization of ultra-wide bandwidth wireless indoor channels: A communication-theoretic view," *IEEE J. Sel. Areas Commun.*, vol. 20, no. 9, pp. 1613–1627, Dec. 2002.

[16] J. G. Proakis, *Digital Communications*. New York: McGraw-Hill, 1995.

[17] M. Z. Win, D. Cassioli, and A. F. Molisch, "The ultra-wide bandwidth indoor channel: From statistical model to simulations," *IEEE J. Sel. Areas Commun.*, vol. 20, no. 6, pp. 1247–1257, Aug. 2002.

[18] S. Verdu, *Multiuser Detection*, 1st ed. Cambridge, U.K.: Cambridge Univ. Press, 1998.

[19] I. Bergel, E. Fishler, and H. Messer, "Low complexity narrow-band interference supression in impulse radio," in *Proc. Int. Workshop Ultra Wideband*, Oulu, Finland, Jun. 2003, [CD-ROM].

[20] H. Lee, B. Han, Y. Shin, and S. Im, "Multipath characteristics of impulse radio channels," in *Proc. IEEE Conf. Veh. Technol.*, vol. 3, Tokyo, Japan, May 2000, pp. 2487–2491.