Near-Far Problem in Noise-based Frequency-Offset Modulation

Bertold I. Bitachon, Ibrahim Bilal, Arjan Meijerink, and Mark J. Bentum
University of Twente, Telecommunication Engineering Group
P.O. Box 217, 7500 AE, Enschede, The Netherlands
Email: bertold.ian@gmail.com, {i.bilal, a.meijerink, m.j.bentum}@utwente.nl

***Abstract*—A multiple-access noise-based frequency offset modulation (N-FOM) system is studied in the context of near-far problem. A closed-form expression for the signal-to-noise ratio is derived and the bit error rate (BER) is evaluated. The result shows that the near-far problem compromises the performance of the multiple-access N-FOM system. Performance comparison with a standard spread-spectrum system indicates that the near-far problem has an inherently different effect on the N-FOM communication. The error rate of a node in the N-FOM system increases at a much higher rate as the number of simultaneous communication links increases. However for small number of simultaneous links, an adaptive power control scheme can help minimize the threat of the near-far problem.**

***Index Terms*—Frequency offset modulation, near-far problem, spread-spectrum, transmitted-reference, wideband communication.**

I. Introduction

Numerous studies around the world have pointed out the benefits of transmit-reference communication [1]–[6]. Transmit-reference (TR) is a spread spectrum communication technique in which the transmitter sends the spreading signal along with the spread information signal. The detection scheme is based on correlation of the received signal with itself, and does not require implementation of a complex rake receiver. This leads to a simple receiver architecture; the synchronization problem is largely eliminated, which is highly desired in low data rate systems with bursty traffic.

The concept of transmitted-reference can be traced back to the work of Basore in 1952 [7], in which a correlation-based detection schemed was employed. The template for correlation was either stored at the receiver or sent through an auxiliary channel as a *reference*. Transmitting the reference signal resulted in fast synchronization, however the noise in the auxiliary channel was seen as a limiting factor.

A couple of decades later, the technological advances and introduction of new spread-spectrum regulations helped TR scheme attract renewed interest. In 2002, Hoctor and Tomlinson proposed a TR scheme in which the transmitted signal is a combination of data-carrying pulses and reference pulses, with a small time offset between them [4]. At the receiver, the reference signal is used as a template for retrieving information from the data pulses. Since both information and reference signal are affected similarly by the channel, the communication does not require channel estimation. This results in a low-complexity receiver, making TR a viable candidate for low-powered short-ranged transmissions. Ultra-low power operation makes TR a potential spectrum-sharing communication technique; the co-existence of a TR system with other operational systems in the spectrum has been researched in [8]–[10].

In [6], we proposed a modification of the TR scheme which was inspired from our earlier work in optical communications [11]. A related study was carried out by Goeckel in [5], which strengthened the claim that TR scheme can be a promising communication scheme for wireless sensor networks (WSNs). With small synchronization times and simple receiver architecture, the TR scheme is particularly attractive to low-data-rate monitoring applications. A network of large number of sensors can be deployed with little cost and increased reliability. In [12] and [13], a variant of TR scheme was proposed. Instead of a time offset between the reference and the information signal, a frequency offset was used. This further simplifies the receiver architecture, since the wideband delay element needed to realize the time offset, is difficult to implement on a small chip.

In a TR system, the receiver only needs to know the frequency (or time offset) used at the transmitter. This means that any kind of spreading signal can be used, including pure noise. The TR system using noise-based frequency-offset modulation (N-FOM) has been investigated in [14]–[17]. The research in [15] and [16] shows robustness of the N-FOM system against narrowband interferences for low-to-moderate signal-to-interference ratios. In [17], the N-FOM system is investigated in dense frequency-selective channels, where it is shown that a communication link with acceptable error rate can be established in certain channel environments.

Multiple access (MA) is an important aspect of communication in WSNs. In N-FOM, MA can be realized by using unique values of the frequency offset for different communication links. Our previous research shows that the link performance degrades as the number of communicating users increases in the network [13]. The analytic results were derived with an unrealistic assumption that all the signals from different users are received with the same strength. In practice, the signal strength of each user degrades differently due to the relative distance of the transmitting nodes to the receiving node. This can lead to the near-far problem; a weak signal from a far-

978-1-4673-9907-4/15/$31.00

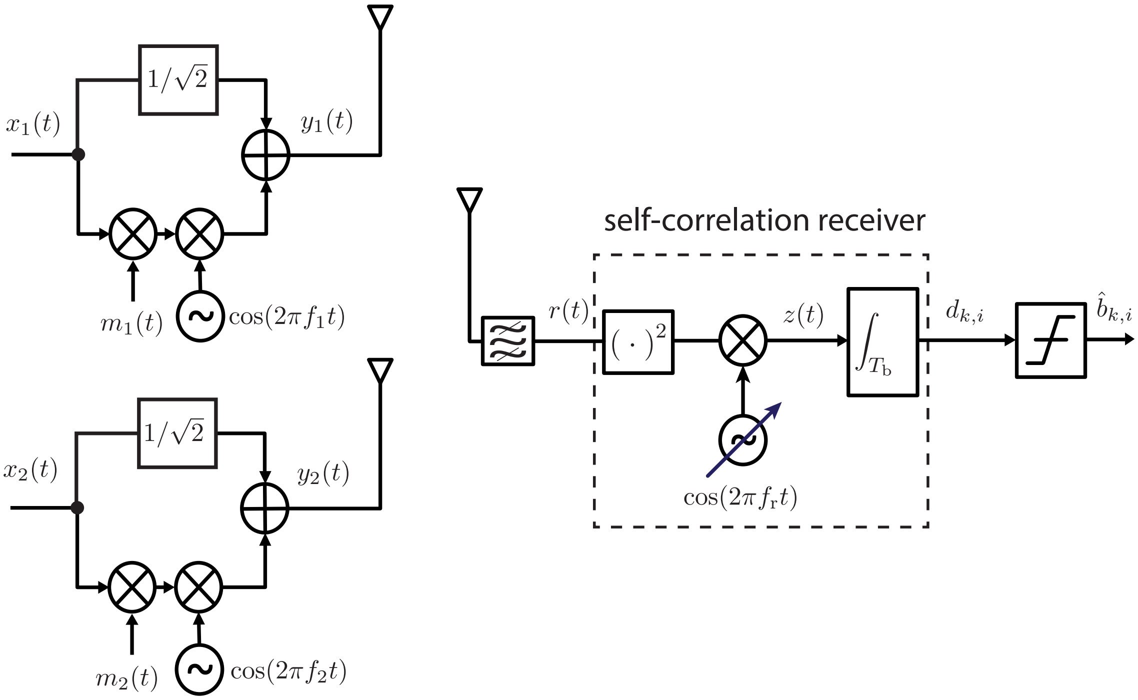

Fig. 1: MA N-FOM system: Two transmitting users are simultaneously communicating with a single receiver.

away user will be masked by a strong signal from a user which is much closer to the receiver. Potentially, the strong user can completely inhibit the weak signal. The goal of this paper is to study the MA N-FOM system in a scenario where the near-far problem arises. A closed-form expression of the signal-to-noise ratio for each link in the MA network will be derived. This will help analyze the system performance and suggest potential problem-solving strategies.

The paper begins with a description of the MA N-FOM system in Section II. The link performance is derived in Section III, followed by a numerical example and discussion in Section IV. Finally, Section V ends with concluding remarks.

II. System Model

Consider a multi-user TR communication system as shown in Fig. 1 where two transmitting nodes are simultaneously communicating with a single receiver node. Without loss of information, we express the signals in their complex envelopes. Each transmitting user locally generates a bandpass spreading signal with a complex envelope $x_i(t)$ where $i \in \{1, 2\}$ in this example. The spreading signals $\{x_1(t), x_2(t)\}$ are modeled as independent wide-sense stationary (WSS) Gaussian processes. Using properties of the WSS bandpass processes [18], the power spectral density (PSD) of the complex envelope is

$$S_{x_i x_i}(f) = \begin{cases} \dfrac{2P_{x_i}}{B_x}, & |f| \leq B_x/2, \\ 0, & \text{otherwise}, \end{cases} \tag{1}$$

where P_{x_i} is the mean power in the bandpass signal of user i and B_x is the bandwidth. The signals $x_i(t)$ are mutually independent since they are generated independently by the transmitters. Each user has a narrowband message signal $m_i(t)$, modeled as a non-return-to-zero (NRZ) signal with a rectangular pulse shape; i.e.,

$$m_i(t) = \sum_k b_{k,i} p(t - kT_s), \tag{2}$$

where $p(t)$ is the unit-amplitude rectangular pulse of duration T_s and $\{b_{k,i}\}$ is a binary bit sequence of user i, taking values $\{+1, -1\}$ with equal probability. At each transmitter, the message signal is spread by the spreading signal $x_i(t)$. The transmitter then shifts the spectrum of the spread message by a small frequency offset $f_i \ll B_x$. The signal transmitted over the wireless channel is a linear combination of the spreading signal and the *information* signal, and has a complex envelope

$$y_i(t) = x_i(t)\left(m_i(t)\cos(2\pi f_i t) + \frac{1}{\sqrt{2}}\right), \text{ for } i = 1, 2. \tag{3}$$

where the scaling factor of $1/\sqrt{2}$ is chosen so that both the information and spreading signal in $y_i(t)$ have the same mean power. The frequency offset f_i is different for each transmitting node and is already known to the receiver. For correct detection, f_i must be an integer multiple of the symbol rate, i.e., $f_i = m/T_s$, where $m \in \mathbb{Z}_+$. This specific requirement on the frequency-offset will be discussed in Sec. III.

The wireless link between the transmitter nodes and the receiver node is modeled as an additive white Gaussian noise (AWGN) channel. At front-end of the receiver, an ideal bandpass filter $H(f)$ is assumed which perfectly suppresses all out-of-band signals without any distortion to the desired signal. For investigation of the near-far problem, transmitter 1 is considered to be located in the near vicinity of the receiver whereas transmitter 2 is a faraway user. The TR signal from user 2 faces severe attenuation compared to that of user 1. A *scaling factor* $0<\alpha_i\leq 1$ characterizes this attenuation. At the output of the filter, the bandpass-received-signal $\tilde{r}(t)$ can be written as:

$$\tilde{r}(t)=\sqrt{\alpha_1}\tilde{y}_1(t)+\sqrt{\alpha_2}\tilde{y}_2(t)+\tilde{n}(t), \tag{4}$$

where $\tilde{y}_i(t)$ is the bandpass signal of user i and $\tilde{n}(t)$ is bandpass filtered noise, independent from $x_i(t)$. The noise is modeled as a zero-mean WSS Gaussian bandpass process with PSD $N_0/2$ centered around a carrier frequency $f_\mathrm{c}\gg B_x$. The complex envelope of the noise has a PSD,

$$S_{nn}(f)=\begin{cases} 2N_0, & |f|\leq B_x/2\\ 0\ , & \text{else.}\end{cases} \tag{5}$$

The receiver is simultaneously listening to both transmitting users. The message bit of a user is retrieved by correlating the received signal $\tilde{r}(t)$ with a frequency-shifted version of itself. This is achieved by implementation of a squaring block, a tunable local oscillator (LO), followed by an integrate-and-dump filter (IDF). In order to detect the message bit of user i, the frequency f_r of the LO has to be same as the frequency offset of that user, i.e, $f_\mathrm{r}=f_i$. This is essential for successful detection and maximizing SNR. At the output of the IDF, the decision samples are passed through a comparator which gives a bit estimate, $\hat{b}_{k,i}$. In practice, the receiver stores the received signal $\tilde{r}(t)$; it first detects the message bit of user 1 by self-correlation operation with $f_\mathrm{r}=f_1$. It then sets $f_\mathrm{r}=f_2$ and decodes $b_{k,2}$ using similar operation.

In traditional multiple-access (MA) spread-spectrum communication, signals from the communicating nodes raise the noise level. This is inherently different than N-FOM. The squaring operation at the receiver produces multiple cross-terms; the noise level is raised by the self and cross mixing products of the signals from the communicating nodes. The strength of these mixing products varies differently with system parameters. In this sense, the near-far problem has a different effect on the N-FOM system compared to the standard SS system. This will be explained with numerical example in Sec. IV.

III. Performance Analysis

In this section, we derive a closed-form expression for the SNR of user 1 in the MA N-FOM as shown in Fig. 1. We further calculate the bit error rate (BER). Without loss of generality, the derivation is carried out for $b_{0,1}$, i.e., $k=0$. At any given time t, the receiver obtains a bandpass signal $\tilde{r}(t)$. After filtering, it is squared and shifted by the frequency offset f_r. At the mixer output we have:

$$\begin{aligned} z(t)&=\tilde{r}^2(t)\cos(2\pi f_\mathrm{r}t)\\ &=\tfrac{1}{2}|r(t)|^2\cos(2\pi f_\mathrm{r}t)+\tfrac{1}{2}\mathrm{Re}\Big[r^2(t)e^{j4\pi f_\mathrm{c}t}\Big]\cos(2\pi f_\mathrm{r}t)\end{aligned}$$

where the second term in the last equation is a high frequency term which will be suppressed by the integrate-and-dump filter (IDF). We can therefore ignore this term. Substituting the baseband equivalent of $\tilde{r}(t)$ from (4), we can write

$$z(t)=z_{y_1y_1}(t)+z_{y_2y_2}(t)+z_{y_1y_2}(t)+z_{y_1n}(t)+z_{y_2n}(t)+z_{nn}(t),$$

where $z_{y_iy_i}$, z_{y_in}, $z_{y_1y_2}$ and z_{nn} are the cross products produced by the mixing operation; and are defined as

$$z_{y_iy_i}(t):=\tfrac{1}{2}\alpha_i\,|y_i(t)|^2\cos(2\pi f_\mathrm{r}t), \tag{6a}$$

$$z_{y_1y_2}(t):=\tfrac{1}{2}\sqrt{\alpha_1\alpha_2}\Big[y_1(t)y_2^*(t)+y_2^*(t)y_1(t)\Big]\cos(2\pi f_\mathrm{r}t), \tag{6b}$$

$$z_{y_in}(t):=\tfrac{1}{2}\sqrt{\alpha_i}\big(y_i^*(t)n(t)+y_i(t)n^*(t)\big)\cos(2\pi f_\mathrm{r}t), \tag{6c}$$

$$z_{nn}(t):=\tfrac{1}{2}\,|n(t)|^2\cos(2\pi f_\mathrm{r}t).\quad \text{for } i=1,2. \tag{6d}$$

The desired information is contained in the signal-signal mixing product $z_{y_1y_1}(t)$; the rest of the terms add noise to the decision variable. At the output of the integrate-and-dump filter, the decision sample for the user 1 is:

$$d_{0,1}=\int_{-T_\mathrm{s}/2}^{T_\mathrm{s}/2}z(t)\mathrm{d}t. \tag{7}$$

The bandwidth of both $y_i(t)$ and $n(t)$ is much larger than the inverse of the integration period (T_s), therefore the decision filter has a very small bandwidth compared to the input signal $z(t)$. Using Central Limit Theorem approximation, the decision variable can be considered a Gaussian distributed variable, so that it is fully described by its expected value and variance. The mean of the decision variable is

$$\mathbb{E}\left[d_{0,1}|b_{0,1}\right]=\int_{-T_\mathrm{s}/2}^{T_\mathrm{s}/2}E\left[z(t)|b_{0,1}\right]\mathrm{d}t. \tag{8}$$

We can solve the integration by separately calculating the contribution of each term in (6). The signal-signal mixing product $z_{y_1y_1}(t)$ contains the desired despread information-bearing term and an undesired wideband noise term. The information-bearing term can be retrieved only if the frequency of the receiver LO is same as that used by the transmitter, i.e., $f_\mathrm{r}=f_1$.

$$\begin{aligned}&\int_{-T_\mathrm{s}/2}^{T_\mathrm{s}/2}\mathbb{E}\left[z_{y_1y_1}\,|\,b_{0,1}\right]dt\\ &=\int_{-T_\mathrm{s}/2}^{T_\mathrm{s}/2}\frac{\alpha_1}{2}\mathbb{E}\left[|x_1(t)|^2\right]\begin{pmatrix} m_1(t)\cos(2\pi f_1t)\\ +1/\sqrt{2}\end{pmatrix}^2\cos(2\pi f_\mathrm{r}t)\mathrm{d}t\\ &=\alpha_1P_{x_1}\int_{-T_\mathrm{s}/2}^{T_\mathrm{s}/2}\Big[\tfrac{1}{4}m_1^2(t)\cos\big(2\pi(f_\mathrm{r}\pm 2f_1)t\big)\\ &\quad+\tfrac{1}{\sqrt{2}}m_1(t)\cos\big(2\pi(f_\mathrm{r}\pm f_1)t\big)+\tfrac{1}{2}\big(1+m_1^2(t)\big)\cos(2\pi f_\mathrm{r}t)\Big]\mathrm{d}t\\ &=\tfrac{1}{\sqrt{2}}\,\alpha_1b_{0,1}P_{x_1}T_\mathrm{s},\end{aligned}$$

where the second equality follows from (1) and the last equality follows only if $f_\mathrm{r}=f_1$. It can be seen that in order to maximize the SNR, the frequency offset of one user should never be equal to twice the frequency offset of another user,

$$\mathrm{SNR}_l = \frac{8\gamma_l^2}{\left(25\gamma_l^2 + 17\sum_{\substack{i=1\\i\neq l}}^{N}\gamma_i^2 + 20\gamma_l\sum_{\substack{i=1\\i\neq l}}^{N}\gamma_i + 16\sum_{\substack{i=1\\i\neq l}}^{N-1}\sum_{\substack{j=i+1\\j\neq l}}^{N}\gamma_i\gamma_j\right)\frac{1}{S} + \left(20\gamma_l + 16\sum_{\substack{i=1\\i\neq l}}^{N}\gamma_i\right) + 8S}, \text{ where } \gamma_l = \frac{E_{\mathrm{b},l}}{N_0}. \quad (16)$$

i.e., we should have $f_i > 2f_j$, for $i \neq j$. These are the key requirements for efficient detection in a MA N-FOM communication network. In this example, these conditions ensure that the IDF filters out the despread information of user 2. In other words, it can be shown that the contribution from $z_{y_2y_2}(t)$ to $\mathbb{E}[d_{0,1}]$ is zero as long as $f_2 = m/T_\mathrm{s} > 2f_1$. Furthermore, following similar analysis as that in [16] and [17], we can show that the terms $\{z_{y_1y_2}(t), z_{y_1n}(t), z_{y_2n}(t), z_{nn}(t)\}$ have no contribution to the mean of the decision variable. This results in

$$\mathbb{E}\left[d_{0,1}|b_{0,1}\right] = \int_{T_\mathrm{s}} \mathbb{E}\left[z_{y_1y_1}(t)|b_{0,1}\right]\mathrm{d}t = \tfrac{1}{\sqrt{2}}\,\alpha_1 b_{0,1}P_{x_1}T_\mathrm{s}. \quad (9)$$

We see that the expected value of the decision variable contains the despread information bit. The noise in the MA N-FOM communication is calculated as:

$$\begin{aligned}\sigma^2_{d_{0,1}|b_{0,1}} &= \mathbb{E}\left[\left(d_{0,1} - E\left[d_{0,1}|b_{0,1}\right]\right)^2\right]\\ &= \int_{-T_\mathrm{s}/2}^{T_\mathrm{s}/2}\int_{-T_\mathrm{s}/2}^{T_\mathrm{s}/2} C_{zz}(t_1,t_2)\mathrm{d}t_1\mathrm{d}t_2,\end{aligned}$$

where $C_{zz}(t_1,t_2)$ is the covariance function of $z(t)$. The covariance function contains multiple cross-covariance and auto-covariance terms. However, due to the mutual independence of $\{x_1(t), x_2(t), n(t)\}$, one can prove that $C_{zz}(t_1,t_2)$ is the sum of the auto-covariance functions of each term in (6), i.e.,

$$\begin{aligned}C_{zz}(t_1,t_2) &= C_{z_{y_1y_1}}(t_1,t_2) + C_{z_{y_2y_2}}(t_1,t_2) + C_{z_{y_1y_2}}(t_1,t_2)\\ &\quad + C_{z_{nn}}(t_1,t_2) + C_{z_{y_1n}}(t_1,t_2) + C_{z_{y_2n}}(t_1,t_2). \quad (10)\end{aligned}$$

An extensive calculation of the covariance function for single-user transmission was carried out in [16]. Following similar calculation steps, the variance can be shown to be:

$$\begin{aligned}\sigma^2_{d_{0,1}|b_{0,1}} \approx &\left(\frac{25}{16}\alpha_1^2P_{x_1}^2 + \frac{17}{16}\alpha_2^2P_{x_2}^2 + \frac{5}{4}\alpha_1\alpha_2P_{x_1}P_{x_2}\right)\frac{T_\mathrm{s}}{B_x}\\ &+ \left(\frac{5}{4}\alpha_1P_{x_1} + \alpha_2P_{x_2}\right)N_0T_\mathrm{s} + \frac{1}{2}N_0^2B_xT_\mathrm{s}. \quad (11)\end{aligned}$$

This is the total noise power in MA N-FOM communication network where two users are simultaneously communicating with a receiver node. The first term in (11) represents the mean noise power of cross products $z_{y_iy_j}(t)$, for $i,j = 1,2$. The second term represents the mean noise power of signal-noise cross products $z_{y_in}(t)$; and the last terms is from noise-noise mixing product $z_{nn}(t)$. The SNR of the decision variable of user 1 is calculated by the following relationship,

$$\mathrm{SNR}_1 = \frac{\mathbb{E}^2\left[d_{0,1}|b_{0,1}\right]}{\sigma^2_{d_{0,1}|b_{0,1}}}. \quad (12)$$

The average received bit energy of user i is expressed as:

$$\begin{aligned}E_{\mathrm{b},i} &= \mathbb{E}\left[\int_{-T_\mathrm{s}/2}^{T_\mathrm{s}/2}\alpha_i\tilde{y}_i^2(t)dt\right]\\ &= \tfrac{1}{2}\,\alpha_i\mathbb{E}\left[|x_i(t)|^2\right]\int_{T_\mathrm{s}}\left[1/\sqrt{2} + m(t)\cos(2\pi f_\mathrm{r}t)\right]^2\mathrm{d}t\\ &= \alpha_iP_{x_i}T_\mathrm{s}. \quad (13)\end{aligned}$$

Substituting (9) and (11) in (12), the SNR of the user 1 in MA N-FOM communication network shown in Fig. 1 can be written as

$$\mathrm{SNR}_1 = \frac{8\gamma_1^2}{(25\gamma_1^2 + 17\gamma_2^2 + 20\gamma_1\gamma_2)/S + (20\gamma_1 + 16\gamma_2) + 8S}, \quad (14)$$

where we have defined $\gamma_i = E_{\mathrm{b},i}/N_0$ as the received bit SNR of user i and $S = B_xT_\mathrm{s}$ as the spreading factor. The BER can be shown to follow the relation

$$\mathrm{BER}_1 = \mathrm{Q}\left(\sqrt{\mathrm{SNR}_1}\right), \quad (15)$$

where $\mathrm{Q}(x)$ is the Gaussian tail probability,

$$\mathrm{Q}(x) = \frac{1}{\sqrt{2\pi}}\int_x^\infty \exp\left(-\frac{u^2}{2}\right)\mathrm{d}u.$$

Note that the SNR of user 2 can be found from (14) by replacing γ_1 by γ_2 and vice versa. For more than two transmitting nodes, the number of cross-terms increase quadratically. This results in elevated noise levels. The SNR of a desired user l in N number of communicating nodes can be derived using similar steps and is given in (16) at the top of this page.

One can observe from (16) that the SNR is not a monotonic function of S. Thus for a fixed value of γ_i, it is possible to find the optimum value of S that results in the highest SNR. The optimum value of the spreading factor for user l is calculated by finding the zero of the derivative of (16) with respect to S, resulting in:

$$S_{\mathrm{opt},l} = \sqrt{\frac{25}{8}\gamma_l^2 + \frac{17}{8}\sum_{\substack{i=1\\i\neq k}}^{N}\gamma_i^2 + 2\sum_{\substack{i=1\\i\neq k}}^{N-1}\sum_{\substack{j=i+1\\j\neq k}}^{N}\gamma_i\gamma_j + \frac{5}{2}\gamma_l\sum_{\substack{i=1\\i\neq l}}^{N}\gamma_i}.$$

The maximum SNR can be found by directly substituting S in (16) with S_opt. It must be noted that in a practical scenario, it is difficult to choose optimum value of the spreading factor since the parameters that make up S_opt are usually unknown.

IV. Results and Discussion

In the last section, we carried out the performance analysis of a MA N-FOM communication system. As indicated by the denominator of (14), the SNR of a particular user degrades as the received signal energy of other users increases. In this section, we study numerical examples to understand the effect of the near-far problem in an N-FOM communication network.

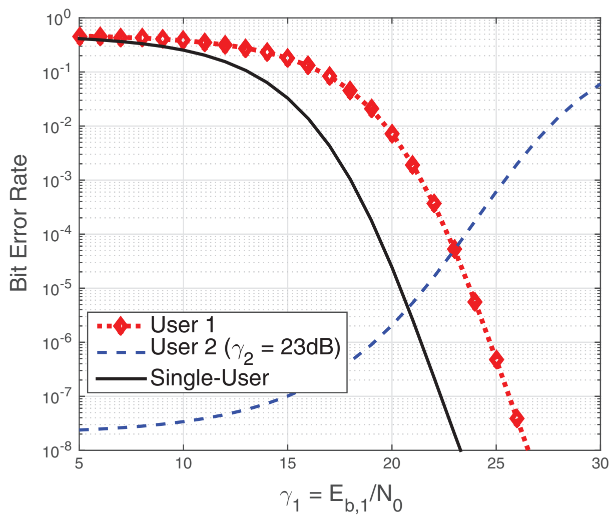

Fig. 2: BER of two users as a function of $\gamma_1 = E_{\mathrm{b},1}/N_0$, with $S = 200$.

We consider the scenario in which the receiver is simultaneously listening to two transmitting users, as shown in Fig. 1. The receiver detects the message bit of user 1 and user 2, one after another using the self-correlation operation with tunable LO. In Fig. 2, we fix the received SNR per bit of user 2 (γ_2) and plot the BER of both users as a function of γ_1. As a reference we also plot the BER of the N-FOM system in single user point-to-point communication which can be obtained by substituting $\gamma_2 = 0$ in (14). The graph clearly shows the effect of the near-far problem. The error rate of user 2 deteriorates as that of user 1 improves. The strong signal from user 1 shadows the weak signal from user 2, making it difficult for the receiver to detect. Although user 2 is operating at decent SNR levels ($\gamma_2 = 23\,\mathrm{dB}$), the performance degradation is very sharp as signal level of user 1 increases. This is because unlike in traditional spread-spectrum, the noise power in N-FOM depends on multiple cross terms produced by the mixing operation which raise the noise level. Finally, we observe that user 1 and user 2 have the same error rates when they have the same signal levels ($\gamma_1 = \gamma_2$). This crossing point characterizes the compromise between the performance of the two users which can be achieved by using adaptive power control (APC).

It can be interesting to see if APC helps to achieve an acceptable BER. In APC, the receiver node negotiates transmitting power between the transmitting users. In steady state, the APC ensures that the received signals from all simultaneous links have equal strength. In Fig. 3, we plot again the BER as a function of γ for several transmitting nodes where APC has been implemented; i.e., the communicating users have same received signal strength. The figure shows that adequate performance can be achieved if there are a small number of transmitting users in the MA N-FOM network. However for large number of users, APC does not help in achieving acceptable error rates. This result was previously analyzed and discussed in [13].

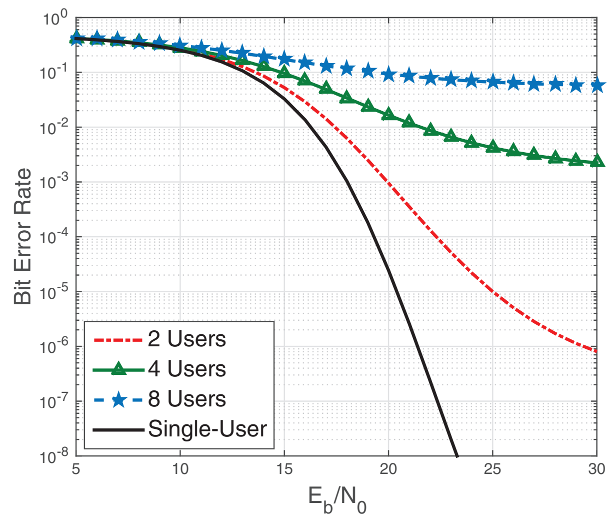

Fig. 3: BER of a MA N-FOM system with 2, 4 or 8 transmitting users, in a scenario where APC is implemented.

In Sec. II, we claimed that the near-far problem has a different effect on the N-FOM system compared to a standard SS system. We now compare the two systems against the near-far problem and quantify the differences. For a fair comparison, we assume the SS system to use pure noise instead of a pseudo-random (PN) sequence as the spreading signal. The SNR of a user, using noise-based traditional SS receiver, in a MA network is given by

$$\mathrm{SNR}_l^{\mathrm{ss}} = \frac{\gamma_l}{1 + \dfrac{1}{S}\left(\gamma_l + \displaystyle\sum_{i \neq l}^{N} \gamma_i\right)}. \tag{17}$$

The proof of this equation follows similar analysis as in Sec. III. In Fig. 4, we plot the BER of the two systems in a scenario where APC is in steady state. The figure shows a large difference between the performance of the two systems which increases with the number of users in the MA. This is because unlike the N-FOM receiver, the standard SS receiver does not produce multiple cross-terms and hence has a much lower noise level. It is clearly evident from (17), which has fewer terms in the denominator compared to that of (16), thus resulting in a much better performance. However, the improved performance of the traditional SS (rake) receiver comes at the cost of complex circuitry and longer synchronization times, which is not wanted in low-data-rate applications with bursty traffic.

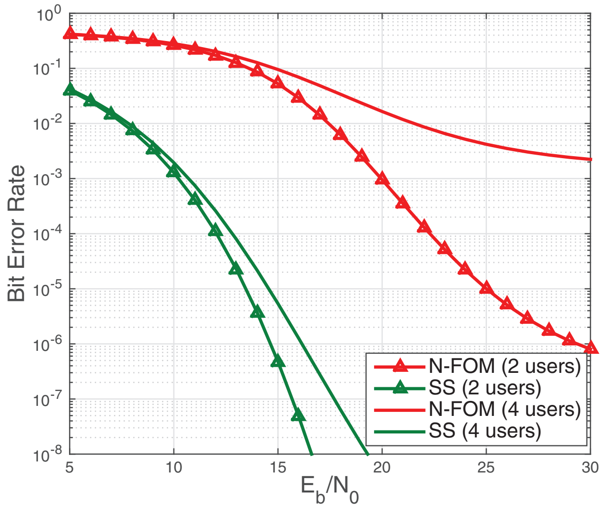

Fig. 4: BER comparison of N-FOM and traditional SS system, in a scenario where APC is implemented.

V. Conclusion

This paper studies a multiple-access N-FOM communication system, where multiple users are communicating in the network. Theoretical value of the BER is derived and numerical examples are discussed. Results reveal that the near-far problem critically limits the system performance of the N-FOM system, especially in comparison with a standard spread-spectrum system. The BER of a particular user faces severe degradation in presence of a strong user. This sharp degradation is mainly attributed to the operation of the self-correlation receiver which produces multiple noise terms. Implementation of adaptive power control in the system is essential to help solve the near-far problem. However, this alone is insufficient to reduce the noise floor when multiple users are present. Evidently, it demands further research in the modulation format for efficient multiple-access communication.

Acknowledgment

This work is supported by the Dutch Technology Foundation (STW) through the WALNUT project, STW project no. 11317.

References

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

[2] J. Choi and W. Stark, "Performance of ultra-wideband communications with suboptimal receivers in multipath channels," *IEEE Journal on Selected Areas in Communications*, vol. 20, no. 9, pp. 1754–1766, Dec. 2002.

[3] M. Win, "A unified spectral analysis of generalized time-hopping spread-spectrum signals in the presence of timing jitter," *IEEE Journal on Selected Areas in Communications*, vol. 20, no. 9, pp. 1664–1676, Dec. 2002.

[4] R. Hoctor and H. Tomlinson, "Delay-hopped transmitted-reference RF communications," in *IEEE Conference on Ultra Wideband Systems and Technologies*, Baltimore, MD, May 2002, pp. 265–269.

[5] H. Zhang and D. Goeckel, "Generalized transmitted-reference UWB systems," in *IEEE Conference on Ultra Wideband Systems and Technologies*, Nov 2003, pp. 147–151.

[6] J. C. Haartsen, X. Shang, J. W. Balkema, A. Meijerink, and J. L. Tauritz, "Novel wireless modulation technique based on noise," in *11th Annual Symp. of the IEEE/SCVT Benelux*, Ghent, Belgium, Nov. 2004, p. 8.

[7] B. L. Basore, "Noise-like signals and their detection by correlation," Ph.D. dissertation, Massachusetts Institute of Technology, Cambridge, MA, USA, May 1952.

[8] A. Giorgetti, M. Chiani, and M. Win, "The effect of narrowband interference on wideband wireless communication systems," *IEEE Transactions on Communications*, vol. 53, no. 12, pp. 2139–2149, Dec. 2005.

[9] T. Quek, M. Win, and D. Dardari, "Unified analysis of UWB transmitted-reference schemes in the presence of narrowband interference," *IEEE Transactions on Wireless Communications*, vol. 6, no. 6, pp. 2126–2139, June 2007.

[10] M. Chiani and A. Giorgetti, "Coexistence between UWB and narrow-band wireless communication systems," *Proceedings of the IEEE*, vol. 97, no. 2, pp. 231–254, Feb. 2009.

[11] A. Meijerink, R. Taniman, and W. van Etten, "Coherence-Multiplexed optical RF feeder networks," *Journal of Lightwave Technology*, vol. 25, no. 11, pp. 3396–3406, Nov 2007.

[12] D. Goeckel and Q. Zhang, "Slightly frequency-shifted reference ultra-wideband (UWB) radio," *IEEE Transactions on Communications*, vol. 55, no. 3, pp. 508–519, March 2007.

[13] J. C. Haartsen, X. Shang, J. W. Balkema, A. Meijerink, and J. L. Tauritz, "A new wireless modulation scheme based on frequency-offset," in *12th Annual Symp. of the IEEE/SCVT Benelux*, Enschede, the Netherlands, May 2005, pp. 1–7.

[14] A. Meijerink, S. L. Cotton, M. J. Bentum, and W. G. Scanlon, "Noise-based frequency offset modulation in wideband frequency-selective fading channels," in *16th Annual Symp. of the IEEE/SCVT Benelux*, Louvain-la-Neuve, Belgium, Nov. 2009.

[15] I. Bilal, A. Meijerink, and M. J. Bentum, "Single-tone interference in noise-based frequency-offset modulation," in *21st Annual Symp. of the IEEE/SCVT Benelux*, Delft, The Netherlands, Nov. 2014.

[16] ——, "Analysis of in-band interference in noise-based frequency-offset modulation," in *IEEE 25th Annual Symp. on PIMRC*, Washington D.C., USA, Sep. 2014.

[17] ——, "Performance analysis of noise-based frequency-offset modulation in dense frequency-selective fading channels," in *9th International Conference on Signal Processing and Communication Systems (ICSPCS'2015)*, Cairns, Australia, Dec. 2015, [Accepted for Publication].

[18] W. C. van Etten, *Introduction to Random Signals and Noise*. Sussex, UK: Wiley, 2006.