PERFORMANCE OF COOPERATIVE TRANSMISSIONS IN FLAT FADING ENVIRONMENT WITH ASYNCHRONOUS TRANSMITTERS

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ABSTRACT

Channel dispersion is inevitable in asynchronous cooperative transmissions even in flat fading environment, which may greatly degrade the gain of cooperative diversity and cast doubts on their application in wireless sensor networks for energy efficiency or reliability. This paper studies the interference caused by such channel dispersion, and shows that the average SIR can be decreased to 10 *dB even in noiseless transmissions. We propose two methods to mitigate the dispersion: a simple sampling time optimization method, and an over-sampling combining method. Simulations show that they can achieve average SIR near* 30 *dB and near* 60 *dB, respectively, with two cooperative transmitters.*

1. INTRODUCTION

Recently there have been great research interests in cooperative transmissions in sensor networks and ad hoc networks, where the increased density of network communication devices can be exploited to enhance performance. For wireless sensor networks, cooperative transmissions using multiple sensors have the gain of cooperative diversity, and thus can be exploited to enhance transmission energy efficiency as well as link reliability. This is extremely important for wireless sensor networks considering the limited reliability and communication capability of each single sensor.

Since cooperative transmissions use an array of transmitters similarly as physical antenna array transmissions, the rich research results of the latter can naturally be extended for the former. Especially, space-time block codes (STBC) [1], [2] have been widely investigated for cooperative transmissions because of their efficient computation.

Nevertheless, one of the major differences between cooperative transmissions and conventional array transmissions is that cooperative transmitters need to have the same data before transmission. This raises the overhead of cooperation, and causes performance loss. As a matter of fact, this problem has been a focused research area for cooperative communications [3], [4]. Only very recently, there have been work toward another major difference, i.e., the synchronization of cooperative transmitters [5]-[11]. It is difficult, and in most cases impossible, to achieve perfect synchronization among distributed transmitters. This is even more a reality when low-cost, small-sized transmitters are used, such as tiny sensors [6]. Note that the synchronization among the transmitters is different from the conventional synchronization between the transmitter and the receiver. For the former, usually hand-shaking has to be conducted among the transmitters, which induces more synchronization overhead and eats away the benefits of cooperative diversity.

When the cooperative transmitters are not synchronized, their signals arrive at the receiver with arbitrary delays. This has two effects. First, any intended signal structure, such as the orthogonal structure of STBC encoding, is destroyed, so that the conventional receivers may not work anymore. Next, there is channel dispersion because of the pulse shaping and because universally ideal sampling time instants for all transmitter's signals do not exist. If the environment is already

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frequency selective fading, then such extra dispersion may not be significant since it only makes channels longer. However, if the environment is flat fading, as expected in wireless sensor networks, then such dispersion makes the channels fundamentally different. In contrast to flat fading, dealing with dispersive channels not only makes the receiver excessively complex and energy consuming, but also greatly reduces performance, especially if the optimal maximum-likelihood sequence estimation is not affordable [6].

It has been shown that via proper coding, the diversity order of conventional array transmissions can be preserved in asynchronous cooperative transmissions if the timing asynchronism is in integer (not fractional) number of symbol intervals [10], [11]. The underlying assumption, however, is that the channels are still flat fading. Note that the diversity order is defined as the power of the received signal-to-noise ratio (SNR) used in evaluating receiving error-rate. The problem is that, when channels become dispersive, there is intersymbol interference (ISI), and the ISI drastically reduces SNR if not mitigated. Therefore, the ISI has a detrimental effect even in case the diversity order can be preserved.

We have shown that the orthogonal structure of the STBC can be preserved by special encoding schemes conducted at the transmitters [5], or by channel equalization conducted at the receiver [6]. Nevertheless, in both cases, we still suffer from heavy performance loss compared with the ideal flat fading case. On the other hand, one may suggest to use OFDM transmission for easy equalization and timing asynchronism mitigation [7]. But OFDM has its own problems, such as residue carrier frequency offset [8], lower power amplifier efficiency, as well as more complex implementations.

Considering that it is still not very clear as to whether cooperative transmissions really provide positive overall gain in asynchronous flat fading environment, in this paper, we investigate the performance of asynchronous cooperative transmissions by studying the ISI due to channel dispersion, and by comparing its signal-to-interference ratio (SIR) against the original flat fading cases. Note that we consider only the asynchronism caused by different transmission times and propagation delays among the transmitters. Carrier frequency mismatch only makes channels time-varying, and can be addressed by adaptive techniques if the variation rate is slow enough.

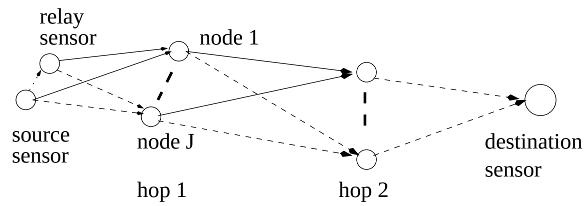

Fig. 1. Multi-hop wireless sensor network model with cooperative transmissions.

This paper is organized as follows. In Section 2, we give the asynchronous transmission model and discuss the difficulty of synchronization. Then in Section 3, we propose two methods to mitigate the channel dispersion. Their performance is investigated by extensive simulations in Section 4. Finally, conclusions are presented in Section 5.

2. SYSTEM MODEL

Consider a wireless sensor network where a source sensor needs to transmit a data packet to a destination sensor through multi-hop relaying as shown in Fig. 1. In each hop, multiple sensors may cooperatively forward the data packet to the next hop.

We consider the cooperative transmission among nodes 1 to J in hop 1. Note that all these nodes can have the knowledge of the data packet from the transmission in the previous hop. The passband signal to be transmitted by each transmitter i has a general form $\mathrm{Re}[\sum_{m=-\infty}^{\infty} s_i(m)p_b(t-mT)e^{j2\pi f_c t}]$, where $s_i(m)$ is the complex symbol transmitted within symbol interval $[mT, (m+1)T)$, $p_b(t)$ is the baseband pulse shaping filter, and f_c is the carrier frequency. After delaying with δ_i, the passband signal transmitted from the transmitter i, $1 \leq i \leq J$, and received by the receiver is $\mathrm{Re}[\sum_{m=-\infty}^{\infty} s_i(m)p_b(t-mT-\delta_i)e^{j2\pi f_c(t-\delta_i)}]$.

We assume flat-fading propagation in this paper. The received passband signal at the receiver is

$$r_p(t) = \mathrm{Re}[\sum_{i=1}^{J} \alpha_i \sum_{m=-\infty}^{\infty} s_i(m) \\ p_b(t-mT-\delta_i)e^{j2\pi f_c(t-\delta_i)} + v_p(t)], \qquad (1)$$

where α_i are (complex) fading of the propagation.

We assume asynchronism in transmission time and propagation delays, which means that δ_i are different for different transmitters.

Without loss of generality, we can demodulate (1) by $e^{-j2\pi f_c t}$ to obtain the continuous-time complex baseband signal

$$r_b(t) = \sum_{i=1}^{J} \alpha_i e^{j\theta_i} \sum_{m=-\infty}^{\infty} s_i(m) p_b(t - mT - \delta_i) + v_b(t), \tag{2}$$

where $v_b(t)$ is the equivalent baseband noise, and the phase $\theta_i = -2\pi f_c \delta_i$.

Since δ_i may be different for different transmitters, it is impossible to achieve timing synchronization or to find a universally ideal sampling time instant. Then, without loss of generality, we perform baseband sampling at time instant $nT + \tau$, which gives the sample $x(n) \triangleq r_b(nT + \tau)$, where $0 \leq \tau < T$. The sample $x(n)$ can be written as

$$x^{(\tau)}(n) = \sum_{i=1}^{J} \alpha_i e^{j\theta_i} \sum_{m=-\infty}^{\infty} s_i(m) p_b[(n-m)T + \tau - \delta_i] + v^{(\tau)}(n), \tag{3}$$

where $v^{(\tau)}(n) \triangleq v_b(nT + \tau)$.

We assume that the pulse shaping filter have support $[-KT, KT]$. Obviously, each sample $x^{(\tau)}(n)$ depends on multiple transmitted symbols, so the channel should be dispersive. On the other hand, the dispersive channel model in this case is different from that due to multipath propagation. One of the major differences is that once we know the delay δ_i and sampling timing τ, we know all dispersive channel coefficients, which are determined completely by the pulse-shaping filter $p_b(t)$.

Define channel coefficients (due to pulse shaping only, not the propagation flat fading) as

$$h_i(\ell) = p_b(\ell T + \tau - \delta_i), \tag{4}$$

Then the received sample can be written as

$$x^{(\tau)}(n) = \sum_{i=1}^{J} \alpha_i e^{j\theta_i} \sum_{m=-\infty}^{\infty} s_i(m) h_i(n-m) + v^{(\tau)}(n), \tag{5}$$

which can be rewritten as

$$x^{(\tau)}(n) = \sum_{i=1}^{J} \alpha_i e^{j\theta_i} \sum_{\ell=-\infty}^{\infty} h_i(\ell) s_i(n-\ell) + v^{(\tau)}(n), \tag{6}$$

Now we need to estimate the range of ℓ. Because

$$-KT \leq \ell T + \tau - \delta_i \leq KT, \tag{7}$$

we have

$$-K + \frac{\delta_i - \tau}{T} \leq \ell \leq K + \frac{\delta_i - \tau}{T}. \tag{8}$$

Define the lower range and the upper range as

$$L_{i,\tau}^{L} \triangleq \left\lceil -K + \frac{\delta_i - \tau}{T} \right\rceil, \quad L_{i,\tau}^{U} \triangleq \left\lfloor K + \frac{\delta_i - \tau}{T} \right\rfloor. \tag{9}$$

Then the channel model becomes

$$x^{(\tau)}(n) = \sum_{i=1}^{J} \alpha_i e^{j\theta_i} \left[\ h_i(L_{i,\tau}^{L}), \cdots, h_i(L_{i,\tau}^{U})\ \right] \times \begin{bmatrix} s_i(n - L_{i,\tau}^{L}) \\ \vdots \\ s_i(n - L_{i,\tau}^{U}) \end{bmatrix} + v^{(\tau)}(n). \tag{10}$$

Though there are investigations to synchronize the cooperative transmitters [9], perfect synchronization is difficult, and comes at high overhead and complexity. All the transmitters should have identical clock, carrier frequency and symbol timing in order to get perfect synchronization so as to directly use existing encoded transmission and decoding techniques. The challenge comes from the fact that the propagation delays, i.e., time required for the signal from the transmitter to reach the receiver, may be different among the transmitters. So do the carrier frequencies due to Doppler shifting when the transmitters and receivers are moving independently. As a matter of fact, synchronizing all transmitters to one receiver may increase asynchronism toward another receiver [5]. Therefore, instead of considering the more complex and resource consuming perfect synchronization, it may be more advantageous to consider directly the asynchronous transmissions instead.

3. DISPERSION INTERFERENCE AND ITS MITIGATION

Using the model (10), we need to either find ways to mitigate the channel dispersion, and hence ISI, or otherwise suffer from ISI-induced SIR reduction. In this paper, we consider only the *pre-processing* of the received samples without knowing the encoding details. We will try to restore the original flat fading channels by mitigating dispersion (though we still have delays in integer symbol intervals). Then we will evaluate the interference due to residue dispersion, and the reduction of SIR. After this, the subsequent decoding/detection of the cooperatively transmitted signals can simply use flat fading channel model based on the pre-processed samples.

As a basis for comparison, we consider first the case when the receiver just uses the flat fading channel model by assuming all other channel coefficients in (10) be interference, which we call dispersive interference. We define the SIR as

$$\mathrm{SIR} = \frac{\sum_{i=1}^{J} \max_\ell |h_i(\ell)|^2}{J \sum_{i=1}^{J} \left(\sum_\ell |h_i(\ell)|^2 - \max_\ell |h_i(\ell)|^2 \right)}. \quad (11)$$

Note that the SIR defined in (11) refers to the SIR per transmitter, averaged over the fading coefficients. We have (11) also because the signal part and the interference part have the same symbol variance.

For the ideal noisy case (without asynchronism) the signal-to-noise ratio (SNR) can be defined as

$$\mathrm{SNR} = \frac{J\sigma_s^2}{\sigma_v^2}, \quad (12)$$

which is also the average over the fading coefficients. Then, in case of asynchronism, the extra interference in (11) makes the receiver's signal-to-interference-and-noise ratio (SINR) to be

$$\mathrm{SINR} = \frac{1}{\frac{1}{J\mathrm{SIR}} + \frac{1}{\left(\sum_{i=1}^{J} \max_\ell |h_i(\ell)|^2 / J\right)\mathrm{SNR}}}. \quad (13)$$

Therefore, both SIR and the reduced value of channel coefficients $\max_\ell |h_i(\ell)|^2$ makes the SINR lower than the ideal case SNR. In particular, if SIR≪SNR, then SIR dominates and becomes the SINR floor, which means in high SNR environment, the SINR is limited to be SIR. On the other hand, if SIR≫SNR, then the interference does not degrade system performance too much.

As a result, by examining SIR instead of SINR, we can reduce complexity while are still able to show the impact of the asynchronism-induced interference. Therefore, to simplify the problem, we consider noiseless transmission in this paper so as to focus on the interference.

3.1. Use fixed sampling timing

We refer (10) as the "fixed" sampling scheme, and use it as a basis for comparison. With the fixed scheme, without loss of generality, we can use $\tau = 0$. From

$$h_i(\ell) = p_b(\ell T + \tau - \delta_i) = p_b(\ell T - \delta_i), \quad (14)$$

we can decompose the delay δ_i into an integer value $\delta_i^{(1)}$ and a fractional value $\delta_i^{(2)}$ so that

$$\delta_i = \delta_i^{(1)} T + \delta_i^{(2)} T, \quad (15)$$

and

$$-\frac{1}{2} < \delta_i^{(2)} \le \frac{1}{2}. \quad (16)$$

In addition, we assume that the delays of the transmitters are independent and uniformly distributed, which means $\delta_i^{(2)}$ is a uniform random variable with probability density function (pdf)

$$f(\delta_i^{(2)}) = 1, \quad \text{for} \quad -\frac{1}{2} < \delta_i^{(2)} \le \frac{1}{2}. \quad (17)$$

Based on the model (14)-(15), we can rewrite the SIR (11) as

$$\mathrm{SIR} = \frac{\sum_{i=1}^{J} |p_b(\delta_i^{(2)} T)|^2}{J \sum_{i=1}^{J} \sum_k |p_b(kT + \delta_i^{(2)} T)|^2}, \quad (18)$$

where the range of the integer k in the denominator of (18) is

$$\lceil -K - \delta_i^{(2)} \rceil \le k \le \lfloor K - \delta_i^{(2)} \rfloor, \quad k \ne 0. \quad (19)$$

3.2. Mitigate interference by optimizing sampling timing

Since the delays are random, the SIR of (10), or the "fixed" sampling scheme, may not be good enough. A simple enhancement scheme is to look for the optimal time instant τ_0, where $0 \leq \tau_0 < T$, such that

$$\tau_0 = \arg \max_{\tau \in [0,T)} \text{SIR}. \tag{20}$$

With the optimized τ_0, we have channel coefficients

$$h_i(\ell) = p_b(\ell T + \tau_0 - \delta_i). \tag{21}$$

Similarly, we can define $\tau_0 - \delta_i = d_i T + z_i T$, where d_i is an integer, and $z_i \in (-1/2,\ 1/2]$. Then the SIR expression (11) becomes

$$\text{SIR} = \frac{\sum_{i=1}^{J} |p_b(z_i T)|^2}{J \sum_{i=1}^{J} \sum_k |p_b(kT + z_i T)|^2}, \tag{22}$$

where the range of the integer k in the denominator of (22) is

$$\lceil -K - z_i \rceil \leq k \leq \lfloor K - z_i \rfloor, \quad k \neq 0. \tag{23}$$

The major difference between (18) and (22) is that $\delta_i^{(2)}$ has uniform distribution, but z_i has a distribution more concentrated around 0, so that (22) is more likely larger than (18).

To see this point clearly, we assume the following simple (maybe suboptimal) choice of τ_0, i.e.,

$$\tau_0 = \frac{1}{J} \sum_{i=1}^{J} \delta_i^{(2)}, \tag{24}$$

which means τ_0 is the average of the fractional portion of the delays.

For $J = 2$, it is easy to show that z_i has a marginal pdf

$$f(z_i) = \begin{cases} 2(1 + 2z_i), & \text{for } -1/2 < z_i \leq 0 \\ 2(1 - 2z_i), & \text{for } 0 \leq z_i < 1/2 \\ 0, & \text{else.} \end{cases} \tag{25}$$

For larger J, $f(z_i)$ can be approximated as normal distribution $\mathcal{N}(0, \frac{J-1}{12J})$, with the value of z_i limited within range $-(J-1)/J < z_i < (J-1)/J$.

Obviously, z_i has higher probability to be near 0 (or to be small), which means the SIR in (22) has higher probability to be larger than that in (18). In Section 4, we will use numerical methods to evaluate the distribution of SIR over all possible delays. For example, the SIR averaged over all delays can be evaluated from

$$\int \text{SIR} f(x) dx, \tag{26}$$

where x denotes J random variables z_i, $i = 1, \cdots, J$.

In practice, we may over-sample the signal, and then select one of the sampling timing as τ_0 by optimizing (20), with the knowledge about the delays and the channel coefficients. On the other hand, because the channels and the optimization (20) are determined completely by delays δ_i and τ, the receiver may directly analyze and obtain τ_0 before sampling, such as using (24).

3.3. Mitigate interference by over-sampling and combining

The basic idea is that the received signals have a rich structure because of different delays, so the over-sampled signals contain new information.

Consider over-sampling by a factor N, which means we have $N+1$ samples in each symbol interval at time instants τ_k, $k = 0, \cdots, N$, where $0 \leq \tau_k < T$. Then similarly to (10), for each τ_k, we have the corresponding sample

$$x^{(\tau_k)}(n) = \sum_{i=1}^{J} \alpha_i e^{j\theta_i} \left[\ h_i(L_{i,\tau_k}^L), \cdots, h_i(L_{i,\tau_k}^U)\ \right] \times \begin{bmatrix} s_i(n - L_{i,\tau_k}^L) \\ \vdots \\ s_i(n - L_{i,\tau_k}^U) \end{bmatrix} + v^{(\tau_k)}(n). \tag{27}$$

Note that for different τ_k, we may have different channel length and index ranges $[L_{i,\tau_k}^L, L_{i,\tau_k}^U]$, which means that the sample may involve slightly different symbols. Define

$$L_i^L = \min_{\tau_k} L_{i,\tau_k}^L, \ \ L_i^U = \max_{\tau_k} L_{i,\tau_k}^U. \tag{28}$$

We stack $N+1$ samples of each symbol interval to obtain the following vector model

$$\mathbf{x}(n) = \sum_{i=1}^{J} \alpha_i e^{j\theta_i} \mathbf{H}_i \mathbf{s}_i(n) + \mathbf{v}(n), \tag{29}$$

where

$$\mathbf{x}(n) = \begin{bmatrix} x^{(\tau_0)}(n) \\ \vdots \\ x^{(\tau_N)}(n) \end{bmatrix}, \ \ \mathbf{s}_i(n) = \begin{bmatrix} s_i(n - L_i^L) \\ \vdots \\ s_i(n - L_i^U) \end{bmatrix},$$

$$\mathbf{v}(n) = \begin{bmatrix} v^{(\tau_0)}(n) \\ \vdots \\ v^{(\tau_N)}(n) \end{bmatrix},$$

and the $(N+1)\times(L_i^U-L_i^L+1)$ dimensional channel matrices

$$\mathbf{H}_i =$$

$$\begin{bmatrix} \mathbf{0}_{L^L_{i,\tau_0}-L^L_i} & h_i(L^L_{i,\tau_0}) & \cdots & h_i(L^U_{i,\tau_0}) & \mathbf{0}_{L^U_i-L^U_{i,\tau_0}} \\ \vdots & & & & \vdots \\ \mathbf{0}_{L^L_{i,\tau_N}-L^L_i} & h_i(L^L_{i,\tau_N}) & \cdots & h_i(L^U_{i,\tau_N}) & \mathbf{0}_{L^U_i-L^U_{i,\tau_N}} \end{bmatrix}$$

Note that $\mathbf{0}_k$ denotes a $1\times k$ dimensional zero vector.

Then we can combine the samples by an $N+1$ dimensional vector $\mathbf{f}$ to obtain

$$y(n) = \mathbf{f}^H\mathbf{x}(n). \quad (30)$$

Ideally (with zero-forcing criterion) we expect

$$y(n) \approx \sum_{i=1}^{J}\alpha_i e^{j\theta_i} g_i s_i(n-d_i) + w(n), \quad (31)$$

which becomes a flat fading channel model, although there may have integer delays d_i. The key point is that the problems with respect to dispersive channels are gone [10],[11].

Using zero-forcing criterion means that

$$\mathbf{f}^H\mathbf{H}_i \approx g_i\mathbf{e}^T_{d_i}, \quad i=1,\cdots,J, \quad (32)$$

where $\mathbf{e}_{d_i}$ is a unit vector with value 1 in the d_i^{th} entry and zeros elsewhere.

Under the assumption of extremely low noise, one way to calculate the vector $\mathbf{f}$ is

$$\mathbf{f}^H = \left[\ \mathbf{e}^T_{d_1} \ \ \cdots \ \ \mathbf{e}^T_{d_J}\ \right]\left[\ \mathbf{H}_1 \ \ \cdots \ \ \mathbf{H}_J\ \right]^+. \quad (33)$$

Note that $(\cdot)^T$, $(\cdot)^H$, and $(\cdot)^+$ denote transpose, Hermitian, and pseudo-inverse, respectively. In order to find the optimal $\mathbf{f}^H$ using (33), we still need to consider all possible integer delays $d_i \in [L_i^L, L_i^U]$, for $i=1,\cdots,J$. But we can also use the information of δ_i to reduce search complexity.

Obviously, in most practical situations, there is still residue dispersion in (30). The SIR of $y(n)$ can be calculated similarly as (11), but with the new composite vectors $\mathbf{f}^H\mathbf{H}_i$ in place of $h_i(\ell)$.

4. SIMULATIONS

In this section, we use simulations to compare the SIR of the three schemes: *fixed* sampling time (Section 3.1), *selective* optimized sampling time (Section 3.2), and linear *combining* (Section 3.3).

Fig. 2 shows that larger roll-off factor (when using raised-cosine pulse shaping filter) is better for interference mitigation. The fixed method has SIR between 10 dB to 21 dB, depending on roll-off factor γ. The selective method can enhance the SIR by 10 dB, while the combining method by almost 40 dB. In addition, the Monte-Carlo simulated results fit well with the analysis results.

Fig. 3 and Fig. 4 show the cumulative distribution of SIR. Specifically, we see that for fixed method, the SIR has 80% probability of being lower than 15 dB. But for the selective method, the SIR has 80% probability to be higher than 15 dB. The combining method has 99% probability higher than 30 dB.

5. CONCLUSIONS

In this paper, we investigate the performance of asynchronous cooperative transmissions in flat fading environment, where the channel becomes dispersive due to pulse shaping. We show that without any dispersion mitigation techniques, the SIR can only be very low. With our proposed dispersion mitigation techniques, we can greatly remove dispersion interference, and enhance SIR by 10 to 40 dB in two-transmitter cooperation.

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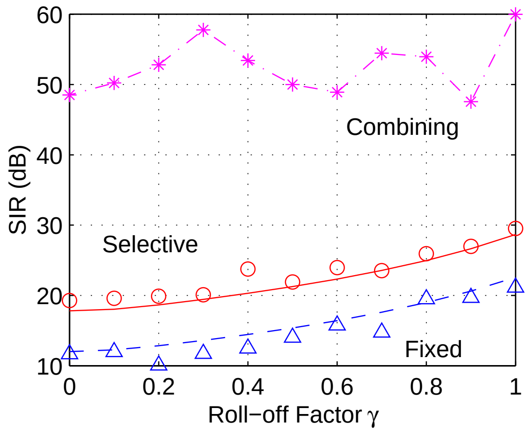

Fig. 2. SIR as function of raised-cosine roll-off factor. Solid line and dashed line are numerical evaluation of analysis results (26). The others are Monte-Carlo simulations for random δ_i. $J = 2$.

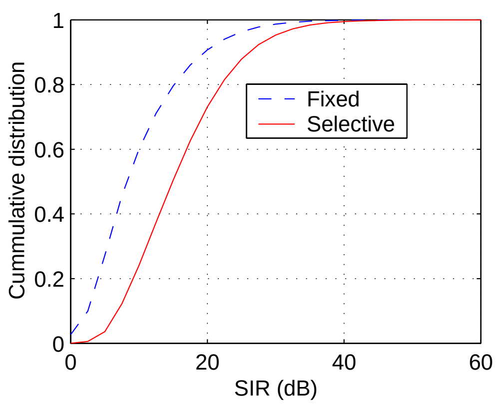

Fig. 3. Cumulative distribution of SIR, obtained by numerical evaluation of (26). $J = 2$, $\gamma = 1$.

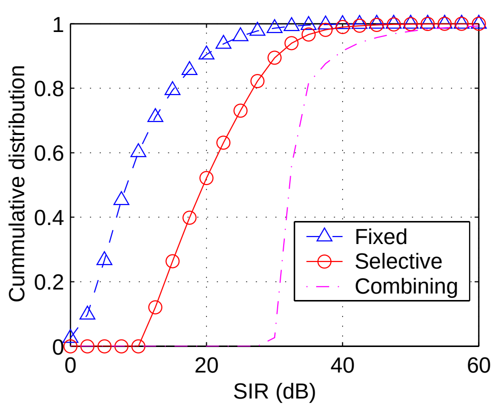

Fig. 4. Cumulative distribution of SIR, obtained as Monte-Carlo simulation. $J = 2$, $\gamma = 1$.

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