On Error Analysis and Distributed Phase Steering for Wireless Network Coding over Fading Channels

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Abstract—Network coding notions promise significant gains in wireless networks’ throughput and quality of service. Future systems employing such paradigms are known to be also highly scalable and resilient to node failure and churn rates. We propose a simple framework where a single relay listens to two nodes transmitting simultaneously over the same band in the presence of Nakagami-m fading. For this multiple-access channel (MAC), we derive in closed-form the exact bit error rate of antipodal signaling with maximum-likelihood detection. As the MAC is the bottleneck in error of the overall system, this provides a good performance measure of the aggregate architecture. Using the new error expressions derived, we then propose a simple closed-loop cooperation strategy where via a ternary feedback from the relay node, significant gains in signal to noise ratio at the relay can be achieved. Our novel error analysis method is applicable to a number of other systems such as the vertical Bell labs spacetime (V-BLAST) scheme and synchronous multi-user systems.

Index Terms—Bit error rate, error analysis, limited feedback, maximum-likelihood detection, network coding, phase steering, relay networks, wireless network coding.

I. INTRODUCTION

The nodes of a network based on Network Coding (NC) protocols need to implement code-and-forward rather than simply performing buffer-and-forward. Replacing routing by coding at both source and transport nodes, simple random linear NC [1] can simultaneously achieve min-cut max-flow rates of all users. This optimality translates to best bandwidth utilization, shortest latencies, as well as decentralized operation, erasure protection, and resilience to churn rates; benefiting not only the broadcast and multicast settings but also the unicast scenarios [2], [3]. In Wireless Network Coding (WNC), the natural interference characteristics of the wireless links can be used in our favor. Recently, a number of WNC protocols need to implement code-and-forward rather than simply performing buffer-and-forward. Replacing routing by coding at both source and transport nodes, simple random linear NC [1] can simultaneously achieve min-cut max-flow rates of all users. This optimality translates to best bandwidth utilization, shortest latencies, as well as decentralized operation, erasure protection, and resilience to churn rates; benefiting not only the broadcast and multicast settings but also the unicast scenarios [2], [3]. In Wireless Network Coding (WNC), the natural interference characteristics of the wireless links can be used in our favor. Recently, a number of WNC

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This work was presented in part at the IEEE International Conference on Computer Communications and Networks (ICCN), St. Thomas, Virgin Islands, USA, Aug. 2008.

The work of A. Peng was supported by an Ontario Graduate Scholarship (OGS). The work of S. Yousefi was supported by an NSERC Discovery Grant and Queen’s University. The work of I.-M. Kim was supported in part by the Ministry of Knowledge Economy (MKE) 21st Century Frontier R&D Program in Korea and a result of subproject UCN08B3-B2-13T. Digital Object Identifier 10.1109/TWC.2009.090037

with the help of the intermediate node, called the relay, an efficient WNC was proposed and it was referred to as physical-layer NC (PNC) [4]. In PNC, the relay detects the exclusive-or of the two signals transmitted from the two end nodes [4]–[7].

Although the relay in PNC detects the exclusive-or of two signals, relays in WNC may need to separately detect the incoming signals in order to perform code-and-forward, which is a capability promised by NC and thus WNC. The separate detection has been widely considered in many WNC networks. For instance, Fig. 1 shows such a network where the two sources simultaneously transmit two signals $x_1$ and $x_2$ to the relay at the first time slot; then the relay detects the individual signals (denoted as $\hat{x}_1$ and $\hat{x}_2$) and forwards two independent linear combinations of the detected signals to multiple destination nodes at the second and third time slots. Each destination node detects $x_1$ and $x_2$ using the two independent linear combinations $\mu_1\hat{x}_1 + \mu_2\hat{x}_2$, $j = 1, 2$. Note that the network of Fig. 1 is composed of a multiple access channel (MAC) in the first phase, followed by a broadcast channel (BC) in the second phase.

In this paper, we analyze the bit error rate (BER) performance of the MAC of Fig. 1. Specifically, we assume that two source nodes simultaneously transmit two binary phase-
shift keying (BPSK)-modulated signals to the relay over fading channels and the relay detects the individual signals using an optimum maximum-likelihood (ML) detector.\textsuperscript{3} To analyze the BER, we adopt a new methodology where all the possible received signals (hypotheses) are mapped onto the so-called effective instantaneous received constellations. In the light of the resulting geometrical structure, the exact instantaneous BER with BPSK is obtained. Then, over fading channels, the average BER is discussed. In particular, the BERs of the worst and best phase-steering cases for the Rayleigh fading channel are derived in closed-form.

The derived error expressions benefit some other areas of wireless communications analysis as well. For instance, the error expression derived in this work is applicable to a two-by-one vertical Bell labs space-time (V-BLAST) system with ML detection. The structure of V-BLAST blends multiple symbols together at the receiver [8] and results in high complexity for ML detection; a situation similar to the MAC case. Initially, due to high complexity of brute-force ML, research focused only on the sub-optimal or near-optimal detectors. The optimal ML performance, however, can still be achieved with sphere decoding at a low complexity and cost [9], [10]. For simple systems such as a two-by-one V-BLAST with BPSK, the ML detection is feasible even without sphere decoding. Several researchers have analyzed the ML performance by providing Bonferroni-type (e.g., union) upper bounds on and approximations to the BER of V-BLAST systems [11], [12]. To the best of our knowledge, however, in the previous literature no exact BER expression has been reported when the ML detection is employed for even the simplest two-by-one V-BLAST configuration with BPSK. Our contribution in this work extends in that direction as well.

Furthermore, the BER analysis problem considered in this work is essentially the same as the synchronous multi-user detection (MUD) problem for two users [13]. However, the exact closed-form expression of the instantaneous BER, not to mention the average BER, was not derived in the previous literature even for the simplest two-user synchronous MUD configuration (for example, see [13] and the references therein). Instead, only some bounds were presented in the literature.\textsuperscript{4} Therefore, our work presents for the first time the closed-form exact BER solution to the two user synchronous MUD problem.

From the BER analysis of this paper, it is observed that even with the optimal ML detection, the dominant error rate at the relay node may be still unsatisfactory. Therefore, improving the error performance with a small cost is another critical issue to be considered and this is our second objective. Assuming the presence of a very low-bandwidth feedback channel from the relay to the source nodes, we propose a near-optimum distributed beamforming or distributed phase-steering scheme. Specifically, our new scheme requires only a ternary feedback symbol and the performance is near-optimum in the sense of minimum BER. Numerical results will demonstrate that our proposed scheme provides significant performance improvement. Although many feedback techniques have been proposed to convey channel information from the receiver to the transmitter, most of the schemes in the literature have required a moderate amount of feedback bits to significantly improve the performance, whereas our scheme requires a mere ternary feedback signal. Most of all, all the previous schemes have considered only collocated classical MIMO systems, whereas our proposed scheme is designed for a WNC system with distributed terminals.

To summarize, the contributions of this paper are as follows:

1) we propose a new analysis method for the error rate of communication systems using the geometry of the so-called effective instantaneous received constellations,

2) we analyze the bit error rate of two antipodal signals received in the presence of AWGN and small-scale fading with applications to multi-user detection, V-BLAST error analysis, and wireless network coding analysis, and,

3) we propose a simple low-cost closed-loop system protocol for distributed phase steering of antipodal wireless network coding. Our scheme assumes that any user is simply cooperating with only one other user to improve fidelity to keep the complexity and overhead at low levels.

The rest of this paper is organized as follows: Section II introduces a system model. In Section III, an exact BER analysis method is proposed for BPSK transmissions from two nodes and the exact closed-form instantaneous BER is derived, followed by some simulation results. For Nakagami-\(m\) fading channels, the exact average BER is given in Section IV. Also, the average BER of the best and worst distributed phase-steering cases over Rayleigh fading channels are found in closed-form and presented in Section IV. The new distributed phase precoding and feedback design are proposed in Section V followed by an evaluation of the performance improvements for fading channels. In Section VI, the challenges of deriving the BER for multi-dimensional constellations are discussed and the paper is concluded.

II. System Model

We start by considering the system of Fig. 1, where two source nodes transmit over the same band simultaneously to a single-antenna receiver, the relay. This relay intends to decode the data for re-encoding in the subsequent communication phases. The received signal \(r\) over a Nakagami-\(m\) fading MAC is given as

\[
r = \sum_{j=1}^{2} h_j x_j + n = \sum_{j=1}^{2} |h_j| \exp(j \theta_j) x_j + n,
\]

where \(x_j\) is the transmitted symbol and \(x_j \in S = \{1, -1\}\). The noise \(n\) has a circularly symmetric complex Gaussian distribution with zero mean and variance \(\sigma_n^2\) per real dimension. The complex channel fading coefficient from the \(j\)-th source node to the relay is denoted by \(h_j = |h_j| \exp(j \theta_j)\) and \(j = \sqrt{-1}\). The phase \(\theta_j\) is uniformly distributed over \([0, 2\pi)\)

\textsuperscript{3}Alternatively, the relay may adopt a sub-optimum detector; but this may substantially degrade the performance of the system because the end-to-end error performance heavily depends on the MAC. In this paper, therefore, we consider only the optimum ML detector.

\textsuperscript{4}The author of [13] acknowledged “Even though there is no hope of evaluating (4.50) in closed form, the four quantities therein can be upper bounded quite easily.” [Section 4.3.1].
and the magnitude $|h_j|$ has a Nakagami-$m$ distribution with the probability density function (PDF) defined as (for $j = 1, 2$)

$$f_{h_j}(|h_j|) = \frac{2m^m|h_j|^{2m-1}}{\Omega^m \Gamma(m)} \exp\left(-\frac{m|h_j|^2}{\Omega}\right) U(|h_j|),$$

where $m$ is the Nakagami-$m$ fading parameter, which ranges from $\frac{1}{2}$ to $\infty$, $\Omega = E[|h_j|^2]$ is the mean-square value of $|h_j|$, $U(\cdot)$ is the unit step function, and $\Gamma(\cdot)$ is the gamma function: $\Gamma(x) = \int_0^\infty t^{x-1} \exp(-t) dt$. The Nakagami-$m$ distribution spans (via the $m$ parameter) a wide range of fading conditions ranging from severe to moderate, to light fading or no fading. For example, it includes the one-sided Gaussian distribution ($m = \frac{1}{2}$) and the Rayleigh distribution ($m = 1$) as special cases. When $m \rightarrow \infty$, the Nakagami-$m$ fading channel converges to a non-fading AWGN channel [14].

The ML estimates $\{\hat{x}_j : j = 1, 2\}$ given perfect channel state information (CSI) at the relay are determined by

$$\hat{x}_1, \hat{x}_2 = \arg \min_{x_1, x_2 \in \mathcal{S}} \left\{ r - \sum_{j=1}^2 h_j x_j \right\}^2.$$

The analysis performed in this paper is based on the above ML detection rule.

III. EXACT INSTANTANEOUS BER ANALYSIS

In this section, we obtain an exact instantaneous BER expression, which is defined as the BER given $\{h_j : j = 1, 2\}$. The proposed BER analysis method involves the derivation of the probability of a Gaussian noise vector residing in an arbitrary wedge region in a two-dimensional Euclidean space. A detailed derivation is shown for the BER expression, with an explanation on how the wedge probability functions are obtained. The section is concluded with numerical results.

A. Wedge Error Probabilities

Let us consider a BPSK constellation, i.e., $x_j \in \{\pm 1\}$. Since the transmitted symbols are points from a BPSK constellation, the signal component of the received waveform is

$$\sum_{j=1}^2 \pm h_j$$

which gives 4 possibilities for any fixed set of fading coefficients. Clearly, with one receive antenna, the received signal is always a complex number. Thus, it can be realized as a point in a two-dimensional constellation. The BER analysis method proposed in this paper is based on mapping all the possible received signals onto such two-dimensional constellations. The newly constructed constellations are referred to as the effective instantaneous received constellations (EIRC) each consisting of 4 signal points. Clearly, the positioning of the signal points depends on $\{h_j : j = 1, 2\}$. Thus, the resulting two-dimensional constellations are symmetrical yet irregular in geometrical shapes. In particular, the analysis heavily depends on the relations between $|h_j|$’s and $\theta_j$’s.

By carefully examining the interrelationships of the fading coefficients, we obtain a complete list of all different types of constellations that can be possibly formed.

It is noted that for the irregular two-dimensional constellations encountered, each decision region is a combination of stripes and wedges. For the stripes, the error probability can be easily obtained using the complementary error function: $\text{erfc}(x) = 2Q(\sqrt{2}x) = \frac{1}{\sqrt{\pi}} \int_0^\infty \exp(-t^2) dt$. The expression of the probability of a point mapped into a wedge region can be found using the methods in [15], [16]. The derivation of the wedge error probability is expanded in Appendix A via Figs. 8 and 9. Specifically, for the wedge, the error probability is given by quation (3) at the top of the page where $Q_{2D}(x, y; \varrho)$ is the two-dimensional Q-function, which is defined by

$$Q_{2D}(x, y; \varrho) = \frac{1}{2\pi \sqrt{1-\varrho^2}} \int_0^\infty \int_y^\infty e^{-\frac{x^2+y^2-2\varrho xy}{2(1-\varrho^2)}} du dv.$$

In particular, when $x = y$ as in (3), $Q_{2D}(x, x; \varrho)$ can be simply expressed as [17]

$$Q_{2D}(x, x; \varrho) = \frac{1}{\pi} \int_0^{\arctan(\sqrt{\frac{\varrho}{1-\varrho}})} e^{-\frac{x^2}{2\sin^2\Phi}} d\Phi.$$

The two-dimensional Q-function has been well studied in the previous literature. As shown in [14], [17], the two-dimensional Q-function can be simplified into several simple desired forms which involve only finite single integrals. Therefore, it can be easily evaluated and tabulated and thus is often considered closed-form.

B. Instantaneous BER

For this system, in order to make the analysis easier, the number of variables can be reduced by assuming a phase compensation of $\theta_1$ at the receiver side. Therefore, the system equation can be rewritten as

$$\tilde{r} = r \exp(-j\theta_1) = |h_1|x_1 + |h_2| \cos \theta x_2 + j|h_2| \sin \theta x_2 + \tilde{n},$$

where $\theta = \theta_1 - \theta_2$ and $\tilde{n} = n \exp(-j\theta_1)$. Since $\tilde{n}$ has the same distribution as $n$, the ML metric is given by (2) with $r$ replaced by $\tilde{r}$.

Depending on $|h_1|$, $|h_2|$, and $\theta$, the possible received signals of the system can be represented by one of six different constellation types. For $\theta \in (0, \frac{\pi}{2})$, received signals can be represented by three different types of irregular two-dimensional constellations which are labeled by $k = 1, 2, 3$:

$$k = \begin{cases} 1, & \text{for } |h_1| \geq |h_2| \cos \theta, \text{ and } |h_2| \leq 2|h_1| \cos \theta; \\ 2, & \text{for } |h_1| \geq |h_2| \cos \theta, \text{ and } |h_2| > 2|h_1| \cos \theta; \\ 3, & \text{for } |h_1| < |h_2| \cos \theta. \\ \end{cases}$$

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For the special case $\theta = 0$, the received signals map onto two different quaternary Pulse Amplitude Modulation (4-PAM) constellations labeled by $k = 4$ and $5$:

$$k = \begin{cases} 4, & \text{for } |h_1| \geq |h_2|; \\ 5, & \text{for } |h_1| < |h_2|. \end{cases}$$

For the special case $\theta = \frac{\pi}{2}$, the constellation becomes a 4-QAM constellation and is labeled as $k = 6$. By studying the locus of the received points, it can be seen that the BER is symmetric about $\pi/2$. This fact is further verified via Monte-Carlo simulations presented in Figs. 2 and 3. Hence, using both the periodicity and symmetry properties of the BER functions, the consideration of $\theta \in [0, \frac{\pi}{2}]$ is sufficient and is used for the following analysis.

Considering $k = 1$, the newly constructed two-dimensional constellation is shown in Fig. 4. Let $S_i = \{x_1, x_2\} \subset \{(-1, -1), (-1, 1), (1, -1), (1, 1)\}$. The BER for this constellation can be expressed as

$$P^b_1 = \frac{1}{4} \sum_{i=1}^{4} P^b_1(E|S_i) = \frac{1}{2} P^b_1(E|S_4) + \frac{1}{2} P^b_1(E|S_3)$$

where $P^b_1(E|S_i)$ is the BER probability given $S_i$ is transmitted and $P(S_i)$ is the probability that $S_i$ is sent. Assuming equiprobable transmit symbols, (6) becomes

$$P^b_1 = \frac{1}{4} \sum_{i=1}^{4} P^b_1(E|S_i) = \frac{1}{2} P^b_1(E|S_4) + \frac{1}{2} P^b_1(E|S_3) = \frac{1}{4} \sum_{i=1}^{2} \frac{1}{2} P(\tilde{r} \in D_j|S_i) + \frac{1}{4} P(\tilde{r} \in D_3|S_4)$$

where $D_j$ is the decision region for $S_j$, $j \in \mathcal{I}_4$ and $P(D_j|S_i)$ is the probability of a received point mapped onto $D_j$ given $S_i$ is transmitted. In particular, $P(\tilde{r} \in D_j|S_i)$ is the probability of the received signal $\tilde{r}$ falling onto decision regions $D_2$ or $D_3$ given $S_4$ and it is found as

$$P(\tilde{r} \in D_2|S_4) = \frac{1}{2} \text{erfc} \left( \frac{\alpha_{1,1}}{\sqrt{2\sigma_n}} \right) - \frac{1}{2} \text{erfc} \left( \frac{\alpha_{1,2}}{\sqrt{2\sigma_n}} \right) + P(\tilde{r} \in W_1|S_4) + P(\tilde{r} \in W_2|S_4)$$

where $P(\tilde{r} \in W_j|S_i)$ is again the probability of received signal $\tilde{r}$ mapping onto the wedge region $W_j$ given $S_i$ is sent. The wedge probability derivations are rather complicated. Using a method similar to that of [15], the probability $P(\tilde{r} \in W_1|S_4)$ for $W_1$ shown in Fig. 4 can be obtained given $\beta_{1,1}, \varphi_{1,1},$ and $\psi_{1,1}$ as given at the top of next page.

All other $P(\tilde{r} \in W_j|S_i)$’s can be found similarly. Applying the same method to all other $P(\tilde{r} \in D_j|S_i)$’s to obtain $P^b_k(E|S_i)$, we finally derive the expression for $P^b_1$. Similar
\[ P(\hat{r} \in W_1|S_d) = P_w \left( \frac{\beta_{1,1}}{2\sigma_n} + \phi_{1,1} \right) + \frac{1}{2\sqrt{\pi}} \int_{-\infty}^{\infty} \frac{1}{\sqrt{1 + \tan^2 \theta}} \left( \phi_{1,1} \right) \, d\theta, \quad \phi_{1,1} = \arctan \left( \frac{|h_1| \tan \theta + 2|h_2| \sin \theta}{|h_1| - |h_2| \cos \theta + |h_2| \sin \theta \tan \theta} \right), \]

where \( \phi_1 = \arcsin \left( \frac{\tan \theta \sqrt{(|h_1| + 2|h_2| \cos \theta)^2 + (|h_2| \sin \theta - |h_1| \tan \theta)^2}}{|h_1| - |h_2| \cos \theta - |h_2| \sin \theta \tan \theta + |h_1| \tan^2 \theta} \right) \). \tag{10} \]

derivations can be performed on \( P^b_k \) for \( k = 2, 3, \ldots, 6 \). In summary, the BER in the \( k \)-th case can be written as

\[ P^b_k = \frac{1}{2} \sum_{i=1}^{6} b_{k,i} P_w \left( \frac{\beta_{k,i}}{2\sigma_n} \right) \tag{11} \]

\[ \sum_{i=1}^{6} b_{k,i} P_w \left( \frac{\beta_{k,i}}{2\sigma_n}, \phi_{k,i}, \psi_{k,i}, \right), \quad k \in T \]

where \( a_{1,1} = a_{1,4} = a_{2,1} = a_{2,4} = a_{3,1} = a_{3,4} = a_{4,2} = a_{5,1} = a_{5,2} = \frac{1}{2}, a_{1,2} = a_{2,2} = a_{3,3} = \frac{2}{3}, a_{1,3} = a_{3,3} = a_{4,4} = a_{5,4} = -\frac{2}{5}, \) and \( a_{6,3} = a_{6,4} = 0; b_{1,1} = b_{1,3} = b_{2,2} = b_{2,4} = b_{3,1} = b_{3,2} = 0; b_{1,2} = b_{2,1} = b_{2,4} = b_{3,4} = 0, b_{1,5} = b_{3,5} = 0, b_{1,6} = b_{2,5} = b_{2,6} = b_{3,6} = -0.5, \) and \( b_{4,i} = b_{5,i} = 0; a_{1,1} = a_{4,1} = 0.6, b_{1,2} = |h_2|, a_{1,3} = |h_2| - \frac{2|h_1|}{\sqrt{1 + \tan^2 \theta}}, a_{1,4} = a_{2,4} = a_{3,4} = a_{4,3} = \frac{1}{2}|h_2| \cos \theta, a_{2,1} = a_{2,2} = a_{3,3} = \frac{1}{2}|h_2| |h_2| \cos \theta, a_{4,2} = a_{5,1} = a_{5,2} = a_{5,4} = 2|h_1| + 2|h_2| \cos \theta, a_{5,2} = a_{5,4} = 2|h_1| |h_2| \cos \theta, a_{6,3} = a_{6,4} = 0. \) The other terms, \( \beta_{k,i}, \phi_{k,i}, \) and \( \psi_{k,i} \) are more complicated which are included in Appendix B.

\[ \sum_{k=1}^{6} \int_{\theta=0}^{\pi/2} \int_{\phi=0}^{\pi} \int_{d=0}^{\infty} P^b_{\theta} f_{\theta}(\theta)|h_1|d|h_2|d\theta |d|h_2|, \tag{12} \]

where \( \theta = \theta_2 - \theta_1, D_{\theta}^k \) is the integration interval for \( |h_1| \) in the characterization of the \( k \)-th constellation, and \( D^b = [-2\pi, 2\pi] \). The PDFs, \( f_{\theta}(\theta) \), of the phase difference is also uniformly distributed over \([0, 2\pi])\). In the following subsections, two special cases are considered for \( m = 1 \) (i.e., Rayleigh fading) with \( \theta \) fixed to \( \frac{\pi}{2} \) and 0.

A. Special Case I: \( \theta = \frac{\pi}{2} \) (lower bound on error)

With \( \theta = \frac{\pi}{2} \) (or its equivalent values, i.e., odd multiples of \( \pi \)), the EIRC is reduced to a simple rectangular 4-QAM. From the Monte-Carlo simulation results in Fig. 3, it is noticed that BER is minimum at \( \theta = \frac{\pi}{2} \). This is the best BER case and the BER, \( P^b_{\theta=\frac{\pi}{2}} \), can be analytically obtained by averaging \( P^b_6 \) in (11) over \(|h_1| \) and \(|h_2| \) as follows:

\[ \sum_{h_2=0}^{\infty} \int_{|h_1|=0}^{\infty} P^b_{\theta=\frac{\pi}{2}} f_{h_1}(|h_1|) f_{h_2}(|h_2|) |d|h_1|d|h_2|, \tag{13} \]

Solving the double integration of (13), we have

\[ P^b_{\theta=\frac{\pi}{2}} = \frac{1}{2} \left( 1 - \frac{\sqrt{\gamma}}{2 + \gamma} \right), \tag{14} \]

where \( \gamma = \frac{\pi^2}{\sigma^2} \) is the average received SNR. As expected, the average BER in (14) is consistent with the average BER for rectangular 4-QAM in previous literature [19]. Note that the best case BER in (14) is exact. In Fig. 5, the analytical and simulation results are compared, which are exactly the same.

B. Special Case II: \( \theta = 0 \) (upper bound on error)

From Monte-Carlo simulation results in Fig. 3 it is observed that the worst BER is achieved at \( \theta = 0 \). The worst case BER at \( \theta = 0, P^b_{\theta=0} \), can be analytically obtained by averaging \( P^b_4 \) in (11) over \(|h_1| \) and \(|h_2| \) as given in (15).

The BER expression in (15) is exact, but not in closed-form. A closed-form BER expression, which is very accurate, is obtained using the exponential approximation in [20]:

\[ \text{erfc}(x) \approx \frac{1}{6} \exp(-x^2) + \frac{1}{2} \exp\left(-\frac{4x^2}{3}\right). \tag{16} \]
\[ P_{\theta=0}^h = \int_{|h_2|=0}^{\infty} \int_{|h_1|=|h_2|} P_{f|h_1|}^h(|h_1|) f_{|h_2|}(|h_2|) |d|h_1| |d|h_2| + \int_{|h_2|=0}^{\infty} \int_{|h_1|=0}^{\infty} P_{f|h_1|}^h(|h_1|) f_{|h_2|}(|h_2|) |d|h_1| |d|h_2|. \]  

\[ \tilde{P}_{\theta=0}^h \approx \frac{1}{4} - \frac{\sqrt{\gamma}}{4\sqrt{4+\gamma}} + \frac{1}{24+12\gamma} + \frac{\gamma \arctan \frac{1+\gamma}{24+1+\gamma}}{24+1+\gamma} + \frac{1}{24+1+\gamma} + \frac{3}{24+16\gamma} \]

\[ - \frac{\sqrt{3\gamma} \arctan \sqrt{1+\frac{4}{\gamma}}}{4(3+4\gamma)(3+2\gamma)} - \frac{3\gamma}{12(1+2\gamma)(4+9\gamma)} + \frac{\arctan \frac{2+5\gamma}{2\gamma}}{12(1+3\gamma)} + \frac{\gamma(2+5\gamma)(4+17\gamma+18\gamma^2)}{6(2+5\gamma)(4+17\gamma+18\gamma^2)} \]

\[ + \frac{3\gamma}{12(2+5\gamma)^1.5} - \frac{\sqrt{3\gamma} \arctan \frac{3+10\gamma}{\sqrt{3(1+4\gamma)}}}{2(3+10\gamma)(6+34\gamma+48\gamma^2)} - \frac{\arctan \frac{3+10\gamma}{\sqrt{3(1+4\gamma)}}}{2(3+10\gamma)^1.5} \]

Using the approximation in (16) the BER can be derived in closed-form as in (17)

The analytical values are plotted along with the actual simulation results to confirm the accuracy in Fig. 5. It can be seen that the closed-form worst case BER approximation in (17) is very accurate.

V. DISTRIBUTED PHASE PRECODING AND FEEDBACK DESIGN

Examining the BER expressions in (14) and (17), also the numerical and simulation results in Figs. 2 and 3, it is observed that for different \( \theta \) values, the performance may change significantly. Motivated by this observation, a new distributed phase precoding scheme is proposed for two users with BPSK. The ideal feedback renders conveying the exact values of \( \theta_1 \) and \( \theta_2 \) to both \( S_1 \) and \( S_2 \), such that the source terminals are able to enforce \( \theta = \frac{\pi}{2} \) and achieve the best performance given in (14). However, sending exact values of \( \theta_1 \) and \( \theta_2 \) requires infinite number of feedback bits, which is unrealistic and infeasible. Hence, instead of controlling the \( \theta \) exactly at \( \frac{\pi}{2} \), the system performance can be improved by forcing the phase to particular ranges where the resulting overall BER value is small.

A. Distributed Phase Precoding

The proposed precoding scheme design is a quantization problem. From the periodicity, the phase difference \( \theta \) is considered only for \( [0, \pi] \) which can be quantized into \( I + 1 \) regions. At node \( S_1 \) no precoding is performed\(^5\). Depending on which region \( \theta \) resides in, different amount of phase shifting in node \( S_2 \) is needed. We define \( \theta_i^h \) as the \( i \)-th threshold value such that for \( \theta \in [\theta_{i-1}^h, \theta_i^h) \), \( \theta \) requires a shift by adding \( \Delta \theta_i \) and for \( \theta \in [\pi - \theta_{i-1}^h, \pi - \theta_i^h) \) the shifting is a subtraction of \( \Delta \theta_i \). Fig. 6 is an example of the proposed precoding scheme with five different regions. If \( \theta \in [\theta_{i-1}^h, \theta_i^h) \) such as for points \( P_1 \) and \( P_2 \) shown in Fig. 6, nodes \( S_1 \) and \( S_2 \) cooperate and shift both \( P_1 \) and \( P_2 \) by adding \( \Delta \theta_2 \). For \( \theta \in [0, \theta_1^h) \) as the example \( P_3 \) in Fig. 6, the transmitter of \( S_2 \) performs a shift by adding \( \Delta \theta_1 \), whereas for \( \theta \in [\pi - \theta_1^h, \pi) \) such as \( P_4 \), the shift is performed by a subtraction of \( \Delta \theta_1 \). In this example, the precoding scheme can be implemented as a quinary feedback scheme.

The cost function of the quantization problem is the exact instantaneous BER obtained in Section III. The instantaneous BER is rewritten as \( P_{k}^h = P_{k}^h(\theta) \) to emphasize the fact that

\(^5\) The distinction of the nodes can be established outside of the physical (PHY) layer; e.g., in medium access control layer. If we use, say, user one as the reference, then the system distinguishes between the two users and applies the correct phase shift with that same reference.
and the instantaneous BER values depend on \( \theta \). Generally when \( \log_2 l \) feedback bits are available, the system can be optimized through (18) where \( \theta_0 = 0, f_\theta(\hat{\theta}) = \frac{1}{\pi}, \hat{\theta} \in [0, \pi] \) is the PDF for \( \hat{\theta} = \theta \mod \pi \), and \( P_k^h(\theta) \) is given by (11).

B. A Ternary Feedback Scheme

The system with a ternary feedback is considered as a special case. The values of \( \theta_{1_{th, opt}} \) and \( \Delta \theta_{1_{opt}} \) vary with SNR and \( \rho \); and they are prohibitively complex to obtain analytically. Hence, \( \theta_{1_{th, opt}} \) and \( \Delta \theta_{1_{opt}} \) are precalculated “offline” by an exhaustive search and tabulated for immediate usage in a real system. For different SNRs and \( \rho \)'s, \( \theta_{1_{th, opt}} \) and \( \Delta \theta_{1_{opt}} \) are found and listed in Table I. It is observed that \( \theta_{1_{th, opt}} \) and \( \Delta \theta_{1_{opt}} \) are very closely fixed at 60°. The slight differences in the results are due to the accumulated numerical errors associated with evaluating the integrals numerically. Therefore, the conjecture made earlier is further verified. Comparing the BERs with the ternary feedback, the performance is very close to the best case.

Specifically, the protocol suggests the use of a feedback symbol \( B \) as follows:

- \( B = -1 \): the receiver sends feedback \( B = -1 \), if \( 0 < \theta < \theta_{1_{th, opt}} \). When \( B = 0 \) is received, the transmitter at \( S_2 \) shifts \( x_2 \) by adding \( \Delta \theta_{1_{opt}} \).
- \( B = 1 \): the receiver sends feedback \( B = 1 \), if \( \pi - \theta_{1_{th, opt}} < \theta < \pi \). When \( B = 1 \) is received, the transmitter at \( S_2 \) shifts \( x_2 \) by subtracting \( \Delta \theta_{1_{opt}} \).
- No feedback (\( B = 0 \)): nothing is fed back from the receiver to the transmitter if \( \theta_{1_{th, opt}} < \theta < \pi - \theta_{1_{th, opt}} \).

Note that the signaling symbol, \( B \), should be fed back from the relay for every coherence time interval to both \( S_1 \) and \( S_2 \). For the wireless systems with high transmission rates and very low mobilities, which are very typical in commercial Wireless Local Area Network (WLAN) systems, the coherence time can be much larger than the symbol duration. Therefore the overall feedback overhead is very low because in the ternary feedback scheme the \( B = 0 \) signal is mostly required. This in essence means that a differential limited-feedback distributed phase steering scheme is proposed. In such a scheme, when no feedback bit is sent (\( B = 0 \)), the transmitters continue with the same phase correction dictated by the very last feedback bit received. When the phase difference in the received signals is outside the range \( [\theta_{1_{th, opt}}, \pi - \theta_{1_{th, opt}}] \) due to the channel variation, a new feedback bit \( B = 1 \) or \( B = -1 \) is sent to the transmitters. Then the transmitter at \( S_2 \) adds the new phase shifting value \( (\pm \Delta \theta_{1_{opt}}) \) to the previous one; hence the existence of memory and the term differential phase steering. This feedback scheme requires a very small power and bandwidth overhead.

C. Numerical Results

For fading channels with \( \theta_{1_{th, opt}} = \Delta \theta_{1_{opt}} = \frac{\pi}{3} \), Monte-Carlo simulations are performed to compare the average BER performance of systems with and without feedback. Also, simulations are done for \( \theta \) fixed to \( \frac{\pi}{3} \) which is the optimal case. The results are shown in Figs. 5 and 7. It is observed that when the ternary feedback the performance is very close to the optimal. Thus, the new distributed precoding scheme is extremely efficient. It is noticed that the improvements are substantial and they increase with SNR or \( \rho \). Compared to the limited feedback designs proposed in the previous literature, our scheme is able to achieve a BER performance that is very close to the best case BER in (14), at a small cost.

VI. CONCLUSIONS, DISCUSSIONS, AND FURTHER WORKS

In this paper, a new error analysis methodology is proposed. By mapping the possible received signals onto the effective received constellations, the exact BER at the relay node is derived for a two-user NC system (MAC) with BPSK. Using the BER expressions, a new distributed phase steering strategy is presented. Even with a simple ternary feedback, the system performance significantly improves in the presence of fading. Our work is applicable to and enhances similar efforts such as COPE architecture proposed recently in [5] for wireless mesh networks. Such protocols are also applicable to many other cases such as sensor as well as cellular networks. Our protocol respects the reality that in sensor settings, at any moment of time only a fraction of sensor nodes are awake and capable of opportunistic listening. In cellular cases, an
Fig. 6. A demonstration of a distributed phase precoding scheme based on quinary feedback.

Fig. 7. The performance comparison with BPSK system in Nakagami-m fading channel with \( m = 2, 3 \) and 4.

An efficient way to increase coverage is to deploy relay nodes that intervene between base stations and mobile users creating a multi-hop cellular back-bone. Our simple protocol allows for all of the above increasing system throughput. Ericsson has recently proposed a similar framework limited to full duplex flows [24]. Due to the complexity of the resulting geometrical formulations, for more general systems with more than two users and/or higher-order constellations, bounding techniques will be advantageous to exact analysis [21]–[23].

APPENDIX A
DERIVATION OF THE WEDGE PROBABILITY

Given the fading coefficients \( \{ h_j : j = 1, 2 \} \) and AWGN with zero mean and variance \( \sigma_n^2 \) per real dimension, the probability of a point \( S_o \) mapping into a wedge region with vertex \( V_o \) as shown in Figs. 8 and 9, is derived in this section. The normalized distance between \( S_o \) and \( V_o \) is given as

\[
A = \frac{|S_o - V_o|^2}{2\sigma_n^2}.
\]

There are two types of wedge error probabilities to be considered. For \( \Phi_1 \Phi_2 \geq 0 \) as presented in Fig. 8, using a similar method to that in [15], the wedge probability is derived as in (A.1) simplifying further to (A.2). Similarly, for \( \Phi_1 \Phi_2 < 0 \) as shown in Fig. 9, the error probability for the wedge is as in (A.4).

\[
\beta_{1,c} = \left\{ \begin{array}{c}
|h_1|^2 + 2|h_2| \cos \theta - |h_1||h_2| \cos \frac{\theta}{2} + |h_2| \sin \theta)^2,
|h_1|^2 + 2|\frac{h_2}{\tan \frac{\theta}{2}}| \cos \theta - \frac{|h_1||h_2|}{\tan \frac{\theta}{2}} \sin \theta)^2, \\
|h_1|^2 + 2\frac{h_2}{\tan \frac{\theta}{2}} \cos \theta - |h_2| \sin \theta)^2,
|\frac{h_1}{\tan \frac{\theta}{2}}| \cos \theta - |\frac{h_1}{\tan \frac{\theta}{2}}| \sin \theta)^2.
\end{array} \right\
\]

\[
\beta_{2,c} = \left\{ \begin{array}{c}
|h_1|^2 + 2|h_2| \cos \theta - |h_1||h_2| \cos \frac{\theta}{2} + |h_2| \sin \theta)^2,
|h_1|^2 + 2|\frac{h_2}{\tan \frac{\theta}{2}}| \cos \theta - \frac{|h_1||h_2|}{\tan \frac{\theta}{2}} \sin \theta)^2, \\
|h_1|^2 + 2\frac{h_2}{\tan \frac{\theta}{2}} \cos \theta - |h_2| \sin \theta)^2,
|\frac{h_1}{\tan \frac{\theta}{2}}| \cos \theta - |\frac{h_1}{\tan \frac{\theta}{2}}| \sin \theta)^2.
\end{array} \right\}
\]
\[
P_w(A, \Phi_1, \Phi_2) = \frac{1}{2\pi} \int_0^{\Phi_1} \exp \left( -A \sin^2 \Phi_2 \right) d\Phi_2 + \frac{1}{2\pi} \int_0^{\Phi_2} \exp \left( -A \sin^2 \Phi_1 \right) d\Phi_1 + \frac{1}{2\pi} \int_0^{\Phi_2} \exp \left( -A \sin^2 \Phi_2 \right) d\Phi_2 - \frac{1}{2\pi} \int_0^{\Phi_1} \exp \left( -A \sin^2 \Phi_1 \right) d\Phi_1 - \frac{1}{2\pi} \int_0^{\Phi_2} \exp \left( -A \sin^2 \Phi_2 \right) d\Phi_2 + \frac{1}{2\pi} \int_0^{\Phi_1} \exp \left( -A \sin^2 \Phi_1 \right) d\Phi_1 .
\]

\[
P_w(A, \Phi_1, \Phi_2) = \frac{1}{2} Q_{2D} \left( \sqrt{2A \sin \Phi_1}, \sqrt{2A \sin \Phi_2}; \frac{\tan^2 \Phi_2 - 1}{\tan^2 \Phi_2 + 1} \right) - \frac{1}{2} Q_{2D} \left( \sqrt{2A \sin \Phi_1}, \sqrt{2A \sin \Phi_2}; \frac{\tan^2 \Phi_1 - 1}{\tan^2 \Phi_1 + 1} \right) .
\]

**REFERENCES**


\[ \varphi_{1,i} = \begin{cases} 
\arctan\left(\frac{|h_1| \tan \theta + 2|h_2| \sin \theta}{|h_1| - |h_2| \cos \theta + |h_2| \sin \theta \tan \theta}\right), & i = 1; \\
\arctan\left(\frac{|h_1| - |h_2| \cos \theta + |h_2| \sin \theta \tan \theta}{2|h_2| \sin \theta - |h_1| \tan \theta}\right), & i = 2; \\
\arctan\left(\frac{|h_1| - |h_2| \cos \theta + |h_2| \sin \theta \tan \theta}{|h_1| - |h_2| \cos \theta - |h_1| \tan \theta}\right), & i = 3, 2|h_2| \cos \theta \leq |h_1|; \\
\arctan\left(\frac{|h_2| \sin \theta \tan \theta + |h_1| - |h_2| \cos \theta}{|h_1| \tan \theta}\right), & i = 4, |h_1| \cos \theta \geq |h_2|; \\
\arctan\left(\frac{|h_2| \sin \theta \tan \theta - |h_1| - |h_2| \cos \theta}{2|h_2| \sin \theta - |h_1| \tan \theta}\right), & i = 5; \\
\arctan\left(\frac{|h_2| \sin \theta \tan \theta + |h_1| - |h_2| \cos \theta}{2|h_2| \sin \theta - |h_1| \tan \theta}\right), & i = 6, 2|h_2| \cos \theta \leq |h_1|; \\
\arctan\left(\frac{|h_2| \sin \theta \tan \theta - |h_1| - |h_2| \cos \theta}{2|h_2| \sin \theta - |h_1| \tan \theta}\right), & i = 6, 2|h_2| \cos \theta \geq |h_1|. 
\end{cases} \]

\[ \varphi_{2,i} = \begin{cases} 
\arctan\left(\frac{|h_1| \tan \theta}{|h_1| - |h_2| \cos \theta + |h_2| \sin \theta \tan \theta}\right), & i = 1; \\
\pi - \arctan\left(\frac{|h_1| - |h_2| \cos \theta + |h_2| \sin \theta \tan \theta}{|h_1| \tan \theta - 2|h_2| \sin \theta}\right) - \phi_2, & i = 2; \\
\arctan\left(\frac{|h_1| - |h_2| \cos \theta + |h_2| \sin \theta \tan \theta}{2|h_2| \sin \theta - |h_1| \tan \theta}\right) + \phi_2, & i = 3; \\
\pi - \arctan\left(\frac{|h_2| \cos \theta - |h_1| + |h_2| \sin \theta \tan \theta}{|h_1| \tan \theta}\right), & i = 4, 2|h_2| \cos \theta \geq |h_1|; \\
\pi - \arctan\left(\frac{|h_2| \cos \theta - |h_1| + |h_2| \sin \theta \tan \theta}{2|h_2| \sin \theta - |h_1| \tan \theta}\right) - \phi_2, & i = 5; \\
\phi_2 - \arctan\left(\frac{|h_2| \cos \theta - |h_1| + |h_2| \sin \theta \tan \theta}{|h_1| \tan \theta}\right), & i = 6, 2|h_2| \cos \theta \leq |h_1|; \\
\arctan\left(\frac{|h_2| \sin \theta \tan \theta - |h_1| - |h_2| \cos \theta}{2|h_2| \sin \theta - |h_1| \tan \theta}\right), & i = 6, 2|h_2| \cos \theta \geq |h_1|. 
\end{cases} \]

\[ \varphi_{3,i} = \begin{cases} 
\frac{\pi}{2} - \arctan\left(\frac{|h_2| \cos \theta - |h_1| + |h_2| \sin \theta \tan \theta}{|h_1| \tan \theta + 2|h_2| \sin \theta}\right) - \phi_3, & i = 1; \\
\arctan\left(\frac{|h_2| \cos \theta - |h_1| + |h_2| \sin \theta \tan \theta}{|h_1| \tan \theta}\right), & i = 2; \\
\arctan\left(\frac{|h_2| \cos \theta - |h_1| + |h_2| \sin \theta \tan \theta}{2|h_2| \sin \theta - |h_1| \tan \theta}\right), & i = 3; \\
\pi/2 - \arctan\left(\frac{|h_2| \cos \theta - |h_1| + |h_2| \sin \theta \tan \theta}{|h_1| \tan \theta}\right), & i = 4; \\
\arctan\left(\frac{|h_2| \cos \theta - |h_1| + |h_2| \sin \theta \tan \theta}{|h_1| \tan \theta}\right), & i = 5; \\
\arctan\left(\frac{|h_1| \tan \theta}{|h_2| \cos \theta - |h_1| + |h_2| \sin \theta \tan \theta}\right) - \theta, & i = 6. 
\end{cases} \]

\[ \psi_{1,i} = \begin{cases} 
\arctan\left(\frac{|h_1| \tan \theta + 2|h_2| \sin \theta}{|h_1| - |h_2| \cos \theta + |h_2| \sin \theta \tan \theta}\right) + \phi_1, & i = 1; \\
\arctan\left(\frac{|h_1| - |h_2| \cos \theta + |h_2| \sin \theta \tan \theta}{|h_2| \tan \theta}\right) + \phi_1, & i = 2; \\
- \arctan\left(\frac{|h_2| \sin \theta \tan \theta + |h_1| - |h_2| \cos \theta}{2|h_2| \sin \theta - |h_1| \tan \theta}\right), & i = 3, 2|h_2| \cos \theta \leq |h_1|; \\
\arctan\left(\frac{|h_2| \sin \theta \tan \theta + |h_1| - |h_2| \cos \theta}{|h_1| \tan \theta}\right) + \phi_1, & i = 4, |h_1| \cos \theta \geq |h_2|; \\
\arctan\left(\frac{|h_1| - |h_2| \cos \theta - |h_2| \sin \theta \tan \theta}{2|h_2| \sin \theta + |h_1| \tan \theta}\right) + \phi_1, & i = 4, |h_1| \cos \theta \leq |h_2|; \\
\arctan\left(\frac{|h_2| \sin \theta \tan \theta - 2|h_2| \sin \theta}{|h_1| \tan \theta + |h_2| \cos \theta}\right) + \arctan\left(\frac{|h_2| \sin \theta}{|h_1| - |h_2| \cos \theta}\right), & i = 5; \\
\arctan\left(\frac{|h_1| \sin \theta \tan \theta - |h_1| - |h_2| \cos \theta}{|h_2| \sin \theta + |h_1| \cos \theta}\right) + \arctan\left(\frac{|h_2| \sin \theta}{|h_1| - |h_2| \cos \theta}\right), & i = 6, 2|h_2| \cos \theta \leq |h_1|; \\
- \arctan\left(\frac{|h_2| \sin \theta \tan \theta - |h_1| - |h_2| \cos \theta}{|h_1| \tan \theta + |h_2| \cos \theta}\right) + \arctan\left(\frac{|h_2| \sin \theta}{|h_1| - |h_2| \cos \theta}\right), & i = 6, 2|h_2| \cos \theta \geq |h_1|. 
\end{cases} \]


ψ_{2,i} = \begin{align*}
\arctan\left(\frac{|h_2| \cos \theta - |h_1|}{|h_1| \tan \theta + |h_2| \sin \theta} \right) + \phi_1, & \quad i = 1; \\
\arctan\left(\frac{|h_2| \cos \theta - |h_1|}{|h_1| \tan \theta + |h_2| \sin \theta} \right), & \quad i = 2; \\
\arctan\left(\frac{|h_2| \cos \theta - |h_1|}{|h_1| \tan \theta + |h_2| \sin \theta} \right), & \quad i = 3; \\
\arctan\left(\frac{|h_2| \cos \theta - |h_1|}{|h_1| \tan \theta + |h_2| \sin \theta} \right) - \arctan\left(\frac{|h_1|}{|h_2| \sin \theta} \right), & \quad i = 4; \\
\arctan\left(\frac{|h_2| \cos \theta - |h_1|}{|h_1| \tan \theta + |h_2| \sin \theta} \right) - \arctan\left(\frac{|h_1|}{|h_2| \sin \theta} \right), & \quad i = 5; \\
\arctan\left(\frac{|h_2| \cos \theta - |h_1|}{|h_1| \tan \theta + |h_2| \sin \theta} \right) - \arctan\left(\frac{|h_1|}{|h_2| \sin \theta} \right), & \quad i = 6.
\end{align*}

ψ_{3,i} = \begin{align*}
\arctan\left(\frac{|h_2| \cos \theta - |h_1|}{|h_1| \tan \theta + |h_2| \sin \theta} \right) + \phi_1, & \quad i = 1; \\
\arctan\left(\frac{|h_2| \cos \theta - |h_1|}{|h_1| \tan \theta + |h_2| \sin \theta} \right) + \phi_1, & \quad i = 2; \\
\arctan\left(\frac{|h_2| \cos \theta - |h_1|}{|h_1| \tan \theta + |h_2| \sin \theta} \right) + \phi_1, & \quad i = 3; \\
\sqrt{|(h_2^2 - |h_1|^2) \cos^2 \theta + (h_1^2 - |h_2|^2) \sin^2 \theta|}, & \quad i = 4; \\
\sqrt{|(h_2^2 - |h_1|^2) \cos^2 \theta + (h_1^2 - |h_2|^2) \sin^2 \theta|}, & \quad i = 5; \\
\sqrt{|(h_2^2 - |h_1|^2) \cos^2 \theta + (h_1^2 - |h_2|^2) \sin^2 \theta|}, & \quad i = 6.
\end{align*}

1049-1056, June 2006.


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