Diversity Gains Across Line of Sight and Rich Scattering Environments from Space-Polarization-Time Codes

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***Abstract*— Space-time codes built out of Alamouti components have been adopted in wireless standards such as UMTS, IEEE 802.11n and IEEE 802.16 where they facilitate higher data rates through multiplexing of parallel data streams and the addition of two or more antennas at the receiver that perform interference cancellation. This paper provides new theoretical insight into an algorithm for interference cancellation through a Bayesian analysis that expresses performance as a function of SNR in terms of the "angles" between different space-time coded data streams. Our approach provides insights into the coupling of channel coding to spatial and polarization degrees of freedom.**

I. Introduction

Information theoretic analysis by Foschini [1] and by Telatar [2] shows that multiple antennas at the transmitter and receiver enable high rate wireless communication. Space-time codes, introduced by Tarokh et al. [3], improve the reliability of communication over fading channels by correlating signals across different transmit antennas. The most famous space-time block code was discovered by Alamouti [4] and is described by a 2×2 matrix where the columns represent different time slots, the rows represent different antennas, and the entries are the symbols to be transmitted. The reason for broad commercial interest in this code is that both coherent and non coherent detection are remarkably simple. It is possible to separate the data streams transmitted from the two antennas using only linear processing at the receiver. This means that the end to end complexity of signal processing is essentially the same as single antenna systems.

The Alamouti code also facilitates higher data rates through multiplexing of parallel data streams and the addition of a second antenna at the receiver that performs interference cancellation. Data rates of 4 bits/sec/Hz have been demonstrated for several wireless channels including UMTS, GSM EDGE, IEEE 802.11n and IEEE 802.16 (see [5]). In each case the algebraic structure of the space-time block code makes it possible to implement end to end receiver functionality without going beyond the capabilities of DSPs used in second generation cellular technology. Our Bayesian analysis of interference cancellation provides new theoretical insight and is able to predict performance of different detection algorithms as a function of SNR. For transmission schemes involving multiplexing a number of independent Alamouti coded transmissions we find the performance is dominated by the "angles" between the component Alamouti channels.(see [6] for more details)

Our approach also provides insight into the value of polarization diversity at the physical layer. Standard wireless communication systems use a single transmit antenna and a single receive antenna with linear vertical polarization. Many WiFi access points exploit polarization implicitly in a way that is analogous to selection diversity and linear combining techniques applied in traditional receiver (space) diversity [7]. The signal is launched with a single polarization, the propagation medium then couples some energy into the cross-polarization plane, and the signal is received with a cross-polarized antenna [8]. Exploitation of implicit polarization diversity is suited to rich scattering (e.g. densely urban and indoor) environments [9]. By contrast, explicit polarization diversity is particularly suited to rural and remote environments where wide open spaces and flat or smooth undulating terrain give rise to line of sight (LOS) conditions between the transmitter and the receiver. The polarization diversities of transmitted signals are mostly preserved by the LOS environment. Design challenges at the transmitter include how to modulate the electromagnetic wave (carrier signal) to excite the polarization and spatial modes available. Bayesian analysis provides insight into how to multiplex using polarization and spatial degrees of freedom, specifically when to code across antennas for a given polarization, and when to code across polarizations at a given antenna.

We consider propagation models that combine direct (LOS) and scattered components. The different ways of coupling channel coding to spatial and polarization degrees of freedom lead to different distributions of the angle between channels. We are able to provide fundamental insight into the characteristics of coding schemes that have inherent stability with respect to the relative phase between a pair of spatially separated dual polarized transmitters.

II. Bayesian Approach to Interference Cancellation

Consider two co-channel users, each using the Alamouti space-time block code (STBC) [4]. If $\mathbf{c} = (c_1, c_2)^T$ and $\mathbf{s} = (s_1, s_2)^T$ are the codewords transmitted by the first and second users respectively, then the received signal vectors are $\mathbf{r}_1 =$

$(r_1, r_2)^T$ and $\mathbf{r}_2 = (r_3, r_4)^T$ where the components of $\mathbf{r}_i$ are the signals received at the antenna i over two consecutive symbols periods. We have

$$\begin{aligned} \mathbf{r}_1 &= H_1 \cdot \mathbf{c} + G_1 \cdot \mathbf{s} + \mathbf{n}_1 \\ \mathbf{r}_2 &= H_2 \cdot \mathbf{c} + G_2 \cdot \mathbf{s} + \mathbf{n}_2 \end{aligned} \tag{1}$$

where H_1 and H_2 are the channel matrices from the first user to the first and second receive antennas respectively, and the matrices G_1 and G_2 are the channel matrices from the second user to the first and second receive antennas respectively. The vectors $\mathbf{n}_1$ and $\mathbf{n}_2$ are complex Gaussian random variables with zero mean and covariance $E(\mathbf{n}_1^\dagger \mathbf{n}_1) = E(\mathbf{n}_2^\dagger \mathbf{n}_2) = \sigma^2 I_2$. Note that each of H_1, H_2, G_1 and G_2 has the structure of a 2×2 Alamouti STBC or quaternion,

$$\begin{aligned} H_1 &= \begin{pmatrix} h_{11} & h_{21} \\ -h_{21}^* & h_{11}^* \end{pmatrix}, \quad H_2 = \begin{pmatrix} h_{12} & h_{22} \\ -h_{22}^* & h_{12}^* \end{pmatrix}, \\ G_1 &= \begin{pmatrix} g_{11} & g_{21} \\ -g_{21}^* & g_{11}^* \end{pmatrix}, \quad G_2 = \begin{pmatrix} g_{12} & g_{22} \\ -g_{22}^* & g_{12}^* \end{pmatrix}. \end{aligned} \tag{2}$$

We define $\|H_1\|^2 = |h_{11}|^2 + |h_{21}|^2$.

We can rewrite (1) in matrix form as

$$\mathbf{r} = \mathbf{H}\mathbf{c} + \mathbf{G}\mathbf{s} + \mathbf{n}, \tag{3}$$

where $\mathbf{r} = (\mathbf{r}_1, \mathbf{r}_2)^T$, $\mathbf{n} = (\mathbf{n}_1, \mathbf{n}_2)^T$,

$$\mathbf{H} = \begin{pmatrix} H_1 \\ H_2 \end{pmatrix} \quad \text{and} \quad \mathbf{G} = \begin{pmatrix} G_1 \\ G_2 \end{pmatrix}. \tag{4}$$

The likelihood function of codewords $\mathbf{c}$ and $\mathbf{s}$ given the received signal $\mathbf{r}$ is given by

$$p(\mathbf{r}|\mathbf{c}, \mathbf{s}) = \frac{1}{\pi^2 \sigma^4} \exp\left(-\frac{1}{\sigma^2} \|\mathbf{r} - \mathbf{H}\mathbf{c} - \mathbf{G}\mathbf{s}\|^2 \right). \tag{5}$$

Our prior knowledge of c_1, c_2, s_1 and s_2 is that they are selected from the same constellation $\mathcal{C}$ independently, with equal probability. From Bayes' rule the posterior probability of $\mathbf{c}$ and $\mathbf{s}$ given the received data $\mathbf{r}$ is

$$p(\mathbf{c}, \mathbf{s}|\mathbf{r}) = \frac{p(\mathbf{c})p(\mathbf{s})p(\mathbf{r}|\mathbf{c}, \mathbf{s})}{\sum_{\mathbf{c}, \mathbf{s} \in \mathcal{C}^2} p(\mathbf{c})p(\mathbf{s})p(\mathbf{r}|\mathbf{c}, \mathbf{s})} \tag{6}$$

and the MAP estimate is given by

$$(\hat{\mathbf{c}}, \hat{\mathbf{s}}) = \arg\max_{\mathbf{c}, \mathbf{s} \in \mathcal{C}^2} p(\mathbf{c}, \mathbf{s}|\mathbf{r}). \tag{7}$$

The motivation behind interference cancellation is that the computation of full MAP/ML detection can be reduced if one of the signals, $\mathbf{s}$ say, can be canceled out. We would then only need to search over $\mathbf{c} \in \mathcal{C}^2$. This implies that we should attempt to marginalize the joint posterior distribution for $\mathbf{c}$ and $\mathbf{s}$ with respect to $\mathbf{s}$. This leads to

$$p(\mathbf{c}|\mathbf{r}) \propto \sum_{s_1, s_2 \in \mathcal{C}} p(\mathbf{r}|\mathbf{c}, \mathbf{s}) \tag{8}$$

and the MAP decision rule is

$$\hat{\mathbf{c}} = \arg\max_{\mathbf{c} \in \mathcal{C}^2} p(\mathbf{c}|\mathbf{r}). \tag{9}$$

It is evident that this does not help as the sum over $\mathbf{s}$ cannot be evaluated analytically, and so the evaluation of (8) requires just as many likelihood function evaluations as MAP/ML. However, if the sum in (8) were to be replaced by a Gaussian integral the marginalization could be computed analytically. In Bayesian terms, the trick is to forget something that we know in order to reduce computation. Instead of using the fact that $\mathbf{s}$ lies in some constellation, we simply assume that

$$E\{\mathbf{s}\} = 0 \quad \text{and} \quad E\{\mathbf{s}\mathbf{s}^\dagger\} = I_2. \tag{10}$$

The prior for $\mathbf{s}$ is then taken to be the maximum entropy distribution satisfying the constraints (10). This is the Gaussian distribution with zero mean and unit variance, i.e., $p(\mathbf{s}) \propto \exp\left(-\mathbf{s}^\dagger \mathbf{s}\right)$. The new prior for $\mathbf{s}$ is perfectly consistent with our partial prior knowledge of $\mathbf{s}$, it just doesn't represent all that we know. Overall this means we allow some increase in the probability of error in detecting codewords $\mathbf{c}$ (and $\mathbf{s}$), for reduced computational load. We substitute (5) and the prior $p(\mathbf{s})$ into (6) and marginalize the posterior distribution with respect to $\mathbf{s}$ to obtain

$$p(\mathbf{c}|\mathbf{r}) \propto \exp\left(-(\mathbf{c} - \tilde{\mathbf{c}})^\dagger \mathbf{H}^\dagger \mathbf{R}^{-1} \mathbf{H} (\mathbf{c} - \tilde{\mathbf{c}}) \right) \tag{11}$$

where

$$\mathbf{R}^{-1} = \frac{I_4}{\sigma^2} - \frac{\mathbf{G}\mathbf{G}^\dagger}{\sigma^2(\|\mathbf{G}\|^2 + \sigma^2)}, \tag{12}$$

and

$$\tilde{\mathbf{c}} = (\mathbf{H}^\dagger \mathbf{R}^{-1} \mathbf{H})^{-1} \mathbf{H}^\dagger \mathbf{R}^{-1} \mathbf{r}, \tag{13}$$

and we have dropped terms independent of $\mathbf{c}$. Since

$$\mathbf{H}^\dagger \mathbf{R}^{-1} \mathbf{H} = \beta_c^2 I_2, \tag{14}$$

where β_c^2 is real and positive [6], substituting (14) into (11) leads to two independent maximization problems for decoding symbols c_1 and c_2 as

$$p(\mathbf{c}|\mathbf{r}) \propto \exp\left(-\beta_c^2 |c_1 - \tilde{c}_1|^2 - \beta_c^2 |c_2 - \tilde{c}_2|^2 \right), \tag{15}$$

where

$$\tilde{\mathbf{c}} = \frac{1}{\beta_c^2} \mathbf{H}^\dagger \mathbf{R}^{-1} \mathbf{r}, \tag{16}$$

and

$$\beta_c^2 = \frac{\|\mathbf{H}\|^2}{\sigma^2} \left(1 - \frac{\|\lambda\|^2}{1 + \sigma^2 / \|\mathbf{G}\|^2} \right). \tag{17}$$

where

$$\lambda = \frac{\mathbf{H}^\dagger \mathbf{G}}{\|\mathbf{H}\| \|\mathbf{G}\|}. \tag{18}$$

This captures the solution obtained in [10]. Similarly, symbols $\mathbf{s}$ could be decoded in a similar fashion to the decoding of symbols $\mathbf{c}$. In this case we obtain

$$\tilde{\mathbf{s}} = \frac{1}{\beta_s^2} \mathbf{G}^\dagger \mathbf{R}^{-1} \mathbf{r}, \tag{19}$$

$$\beta_s^2 = \frac{\|\mathbf{G}\|^2}{\sigma^2} \left(1 - \frac{\|\lambda\|^2}{1 + \sigma^2 / \|\mathbf{H}\|^2} \right). \tag{20}$$

The performance is then given by

$$\beta^2 = \frac{\|\mathbf{H}\|^4\|\mathbf{G}\|^4}{\sigma^8}\left(1-\frac{\|\lambda\|^2}{1+\frac{\sigma^2}{\|\mathbf{G}\|^2}}\right)^2\left(1-\frac{\|\lambda\|^2}{1+\frac{\sigma^2}{\|\mathbf{H}\|^2}}\right)^2. \tag{21}$$

The parameter λ is an inner product of two unit quaternion vectors, and measures the angle between the desired signal channel vector $\mathbf{H}$ and the interference signal channel vector $\mathbf{G}$. This parameter is fundamental in the analysis of detection performance for multiple Alamouti schemes [6] and in the next section we use it to analyse the polarization diversity.

III. Application to Polarization Diversity

In this section we apply the results obtained in Section II to analyze the performance of systems employing dual-polarization (co-located $vertical/horizontal$) antennas. We will demonstrate that our results can be used to predict, analyze and understand the performance of interference cancellation algorithms, once the channel matrices are known, without resorting to elaborate computer simulation.

Nabar et al. [11] have studied the performance of a system with one dual-polarized transmit and one dual-polarized receive antenna and Deng et al. [12] have extended to the case of two dual-polarized transmit and one dual-polarized receive antenna and proposed a particular hybrid transmission scheme which we will analyze further.

A. Dual-polarized MIMO channel model

We consider a system with one dual-polarized transmit and one dual-polarized receive antenna. During one time slot a symbol c_1 is transmitted from the one orthogonal polarization and c_2 is transmitted from the other. The signal at the receiver is

$$\mathbf{r} = H\mathbf{c} + \mathbf{n}, \tag{22}$$

where $\mathbf{c} = (c_1, c_2)^T$ and $\mathbf{n}$ are independent noise samples on the two receiver polarizations. The channel matrix

$$H = \begin{pmatrix} h_{vv} & h_{hv} \\ h_{vh} & h_{hh} \end{pmatrix}. \tag{23}$$

is a 2×2 polarization scattering matrix, where h_{vv} and h_{hh} represent path gain from vertical transmit to vertical receive antenna and from the horizontal transmit to the horizontal receive antenna, respectively. Similarly, h_{hv} and h_{vh} represent path gain from the horizontal transmit to the vertical receive antenna and vice versa.

Based on the model introduced in [11], the channel matrix H can be decomposed into the sum of an average or fixed component $\bar{H}$ which represents LOS and a scattered component $\tilde{H}$. Thus

$$H = \sqrt{\frac{K}{1+K}}\bar{H} + \sqrt{\frac{1}{1+K}}\tilde{H}, \tag{24}$$

where the quantity K will be referred to as the K-factor of the channel and the factors $\sqrt{K/(1+K)}$ and $\sqrt{1/(1+K)}$ are related to the Ricean K-factor. Note that $K = 0$ corresponds to the a pure Rayleigh fading channel and $K = \infty$ corresponds to LOS only. The elements of the matrix $\bar{H}$ are modeled as fixed complex numbers satisfying

$$|\bar{h}_{vv}|^2 = |\bar{h}_{hh}|^2 = 1, \quad |\bar{h}_{vh}|^2 = |\bar{h}_{hv}|^2 = \gamma_f, \tag{25}$$

where $0 \le \gamma_f \le 1$ describes the cross polarization discrimination (XPD) or separation of orthogonal polarization for the LOS component. For LOS the XPD is a property of the polarimetric transmit and receive antennas and represents the degree to which the two orthogonally polarized components of the antennas mutually couple. A small (large) value of γ_f corresponds to good (poor) separation of orthogonal polarizations. The elements of the matrix $\tilde{H}$ are modeled as complex Gaussian random variables with zero-mean and

$$E\{|\tilde{h}_{vv}|^2\} = E\{|\tilde{h}_{hh}|^2\} = 1, \tag{26}$$

$$E\{|\tilde{h}_{vh}|^2\} = E\{|\tilde{h}_{hv}|^2\} = \gamma_s, \tag{27}$$

where $0 < \gamma_s \le 1$, describes the XPD for the variable component of the channel. γ_s is a composite of the properties of the antennas and the scattering environment. The transmit and receive correlation coefficients are defined as in [11]

$$\begin{aligned} t &= \frac{E\{\tilde{h}_{vv}\tilde{h}^*_{hv}\}}{\sqrt{\gamma_s}} = \frac{E\{\tilde{h}_{vh}\tilde{h}^*_{hh}\}}{\sqrt{\gamma_s}}, \\ r &= \frac{E\{\tilde{h}_{vv}\tilde{h}^*_{vh}\}}{\sqrt{\gamma_s}} = \frac{E\{\tilde{h}_{hv}\tilde{h}^*_{hh}\}}{\sqrt{\gamma_s}}, \end{aligned} \tag{28}$$

and

$$E\{\tilde{h}_{vv}\tilde{h}^*_{hh}\} = E\{\tilde{h}_{vh}\tilde{h}^*_{hv}\} = 0.$$

Consider a two user system with two dual-polarized transmit and one dual-polarized receive antenna, where each user employs the Alamouti STBC over a dual-polarized transmit antenna. The polarization matrices from the first user to the first and second receivers, can be represented as

$$H_1 = \begin{pmatrix} h_{vv} & h_{hv} \\ -h^*_{hv} & h^*_{vv} \end{pmatrix}, \quad \text{and} \quad H_2 = \begin{pmatrix} h_{vh} & h_{hh} \\ -h^*_{hh} & h^*_{vh} \end{pmatrix}. \tag{29}$$

We consider three different transmission schemes. The first is a uni-polarized system as described in the previous sections, i.e., four vertical transmit and two vertical receive antennas, all spatially separated. The channel matrices H_1, H_2, G_1 G_2 are defined in (2) and all the elements in the channel matrices are co-polarized (i.e.,$\gamma_s = 1$). The LOS components of channel matrices are modeled as

$$\bar{H}_1 = \begin{pmatrix} 1 & \mu_1 \\ -\mu_1^* & 1 \end{pmatrix}, \quad \bar{G}_1 = \begin{pmatrix} \mu_4 & \mu_5 \\ -\mu_5^* & \mu_4^* \end{pmatrix}, \tag{30}$$

$$\bar{H}_2 = \begin{pmatrix} \mu_2 & \mu_3 \\ -\mu_3^* & \mu_2^* \end{pmatrix}, \quad \bar{G}_2 = \begin{pmatrix} \mu_6 & \mu_7 \\ -\mu_7^* & \mu_6^* \end{pmatrix}, \tag{31}$$

where $\mu_1, \mu_2, \cdots, \mu_7$ are arbitrary chosen unimodular complex numbers that represent the different LOS paths.

The second scheme consists of two dual-polarized transmit and one dual-polarized receive antenna and employs Alamouti STBC on each pair of dual-polarized transmit antenna.This scheme will be referred to as the dual-polarized system and the LOS components are modeled as

$$\bar{H}_1 = \begin{pmatrix} 1 & \sqrt{\gamma_f} \\ -\sqrt{\gamma_f} & 1 \end{pmatrix}, \quad \bar{G}_1 = \begin{pmatrix} \mu & \sqrt{\gamma_f}\mu \\ -\sqrt{\gamma_f}\mu^* & \mu^* \end{pmatrix},$$
$$\bar{H}_2 = \begin{pmatrix} \sqrt{\gamma_f} & 1 \\ -1 & \sqrt{\gamma_f} \end{pmatrix}, \quad \bar{G}_2 = \begin{pmatrix} \sqrt{\gamma_f}\mu & \mu \\ -\mu^* & \sqrt{\gamma_f}\mu^* \end{pmatrix}, \quad (32)$$

where μ is a unimodular complex number. Under the assumption that the receiver is in the far field of the transmit antenna, then $\mu = \exp\left(-\frac{2\pi i}{\lambda}\mathbf{n}\cdot\mathbf{\Delta}\right)$, where λ is the carrier wavelength, $\mathbf{n}$ is the unit vector in the direction of receiver and Δ is the position of antenna 2 relative to antenna 1. Furthermore, since the two orthogonal components of each of the dual polarized antennas are co-located their relative path difference is zero.

The third scheme is that proposed by Deng et al. [12]. It also consists of two dual-polarized transmit antennas and one dual-polarized receive antenna, but employs Alamouti STBC on the same polarization across the two transmit antennas. It will be referred to as the dual-polarized hybrid scheme. In this case the LOS components of the channel matrices are modeled as follows

$$\bar{H}_1 = \begin{pmatrix} 1 & \mu \\ -\mu^* & 1 \end{pmatrix}, \quad \bar{G}_1 = \begin{pmatrix} \sqrt{\gamma_f} & \sqrt{\gamma_f}\mu \\ -\sqrt{\gamma_f}\mu^* & \sqrt{\gamma_f} \end{pmatrix},$$
$$\bar{H}_2 = \begin{pmatrix} \sqrt{\gamma_f} & \sqrt{\gamma_f}\mu \\ -\sqrt{\gamma_f}\mu^* & \sqrt{\gamma_f} \end{pmatrix}, \quad \bar{G}_2 = \begin{pmatrix} 1 & \mu \\ -\mu^* & 1 \end{pmatrix}, \quad (33)$$

where μ is as in (32) above.

B. Analysis and Simulation

We now analyze the three coding schemes described above in tems of the results derived in Section II. We consider both Rayleigh and Ricean channels. For the Ricean channel $K = 10$ was chosen to represent a typical suburban environment [11]. In the simulation results reported here the data symbols were taken from an 8-PSK constellation and decoded with the algorithm (16) and (19). For the polarization schemes we set the XPD parameters to $\gamma_s = 0.4$ and $\gamma_f = 0.3$ and, for correlated channels, the transmit and receive correlation coefficients, as defined in (28), were chosen to be $t = 0.5$ and $r = 0.3$.

Figure 1 illustrates how the angles between the channel matrices $\|\lambda\|$ are distributed for an uncorrelated Rayleigh fading channel, for the three different transmission schemes. Figure 2 shows the symbol error rate obtained under the same channel conditions. Figure 3 and Figure 4 show the corresponding results for the case of an uncorrelated Ricean channel. It is clear that the presence of the LOS component causes the means of the $\|\lambda\|$ distributions to shift and the distributions to become more concentrated around the mean. The effect is quite pronounced even for relatively small values of K. This behavior holds generally and leads to the conclusion that, when present, the angle between LOS paths is the dominant factor in the performance of various schemes.

Figure 5 shows the $\|\lambda\|$ distributions for different LOS paths for correlated channels. It is evident that the $\|\lambda\|$ distributions

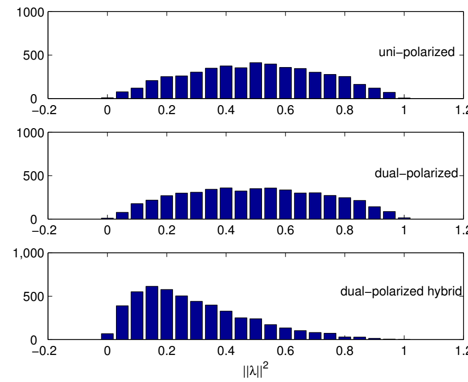

Fig. 1. Distribution of the angle between channels for different transmission schemes over an uncorrelated Rayleigh channel.

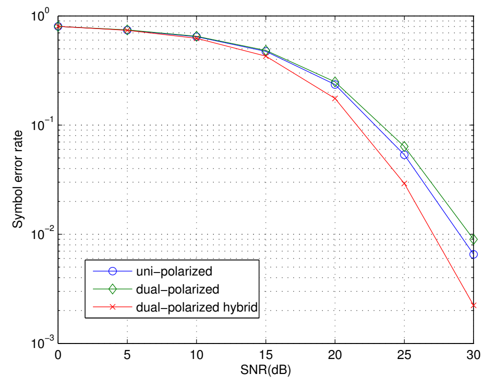

Fig. 2. Performance of three different transmission schemes over an uncorrelated Rayleigh channel.

for the uni-polarized and dual-polarized schemes have considerable variability as the relative phases in the LOS components vary. However, what stands out is the stability of the $\|\lambda\|$ distribution for the dual-polarized hybrid transmission scheme. The distribution, and hence the performance of this scheme, is insensitive to the relative phases of the LOS components of the channel.

We can understand this behavior by considering the angles between the LOS components of the channels which we denoted by λ_{LOS}. From (18) and (24), we have

$$\lambda_{\text{LOS}} = \frac{\bar{H}^\dagger \bar{G}}{\|\bar{H}\|\|\bar{G}\|}. \quad (34)$$

For the dual-polarized hybrid system

$$\|\lambda_{\text{LOS}}\| = \frac{2\sqrt{\gamma_f}}{1+\gamma_f}, \quad (35)$$

and for the dual-polarized system we have

$$\|\lambda_{\text{LOS}}\| = \sqrt{\cos^2\phi + \frac{4\gamma_f}{(1+\gamma_f)^2}\sin^2\phi}, \quad (36)$$

where $\phi = \angle\mu$. We immediately note from (35), that $\|\lambda_{\text{LOS}}\|$ for dual-polarized hybrid scheme depends only on γ_f. This explains why the dual-polarized hybrid scheme is robust to

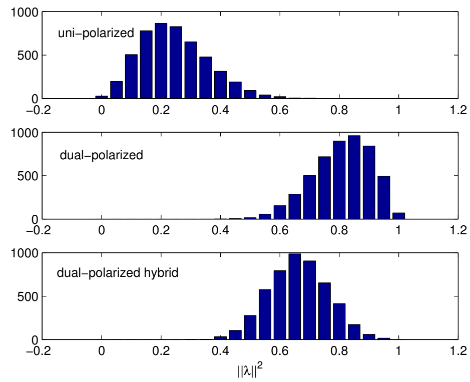

Fig. 3. Distribution of the angle between two channels for different transmission schemes over an uncorrelated Ricean channel.

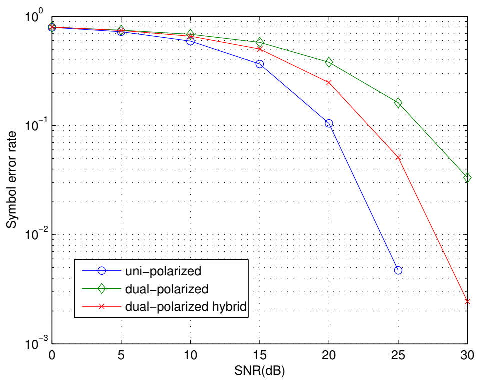

Fig. 4. Performance of three different transmission schemes over an uncorrelated Ricean channel.

changes in the relative path lengths between the two dual polarized transmitters, as observed in Figures 5. We note that for dual-polarized scheme [6]

$$\frac{2\sqrt{\gamma_f}}{1+\gamma_f} \leq \|\lambda_{\mathrm{LOS}}\| \leq 1. \tag{37}$$

IV. Conclusion

We have presented a Bayesian analysis of interference cancellation in combination with space-time block coding. We considered the problem of two co-channel users, each with a space-time block code employing two transmit antennas and a receiver with two antennas. The Bayesian approach is based on giving up the information that the symbol lies in some constellation. Instead we assume that the symbol is Gaussian distributed with zero mean and unit variance and allow some increase in the probability of error in detection. Our results are consistent with those previously obtained by [10]. However, the Bayesian analysis also leads to closed form expressions that provide new theoretical insight and are able to predict performance of different decoding algorithms as a function of signal to noise ratio.

The performance expressions obtained make it possible to analyze different architectures that combine polarization and spatial diversity we are able to predict performance as a function of propagation environment and to see greater robustness in the dual polarized hybrid.

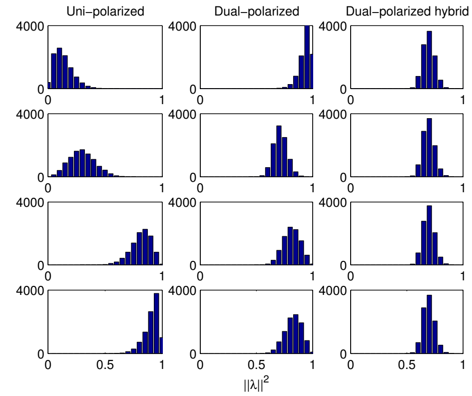

Fig. 5. Distribution of angle between channels for different LOS over correlated channels. Each row shows three different transmission schemes under the same LOS with $\gamma_f = 0.3$.

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References

[1] G. J. Foschini, "Layered space-time architecture for wireless communication in a fading environment when using multi-element antennas," *Bell Labs Technical Journal*, vol. 1, pp. 41–59, 1996.

[2] E. Telatar, "Capacity of multi-antenna gaussian channels," *European Transactions on Communications*, vol. 10, pp. 585–596, 1999.

[3] V. Tarokh, N. Seshadri, and A. R. Calderbank, "Space-time codes for high data rate wireless communications: Performance criteria and code construction," *IEEE Trans. Inform. Theory*, vol. 44, pp. 744–765, 1998.

[4] S. Alamouti, "Space-block coding: A simple transmitter diversity technique for wireless communications," *IEEE J. Select. Areas Commun.*, pp. 1451–1458, 16 (1998).

[5] S. N. Diggavi, N. Al-Dhahir, A. Stamoulis, and A. R. Calderbank, "Great expectations: The value of spacial diversity to wireless networks," *Proc. IEEE*, pp. 219–270, Feb. 2004.

[6] S. Sirianunpiboon, A. R. Calderbank, and S. D. Howard, "Space-polarization-time codes for diversity gains across line of sight and rich scattering environments," *submitted to IEEE Trans. Inform. Theory*, 2006.

[7] J. D. Gibson (Ed), *The Mobile Communications Handbook, 2nd edition*. IEEE Press, 1999.

[8] A. M. D. Turkmani, A. A. Arowojulu, P. A. Jefford, and C. J. Kellett, "An experimental evaluation of the performance of two-branch space and polarisation diversity schemes at 1800 MHz," in *Proc. IEEE Trans. on Vehicular Technology*, pp. 290–294, May 1995.

[9] R. G. Vaughan, "Polarization diversity in mobile communications," in *Proc. IEEE Trans. on Vehicular Technology*, pp. 177–186, Aug. 1990.

[10] A. R. Calderbank and A. F. Naguib., *Orthogonal designs and third generation wireless communication, Surveys in Combinatorics 2001*. pp.75-107, edited by J.W.P. Hirschfeld, London Math. Soc. Lecture Note Series 288, Cambridge Univ. Press, 2001.

[11] R. U. Nabar, H. Bölsckei, V. Erceg, D. Gesbert, and A. J. Paulraj, "Performance of multiantenna signaling techniques in the presence of polarization diversity," *IEEE Trans. Signal Processing*, vol. 50, no. 10, pp. 2553–2562, 2002.

[12] Y. Deng, A. Burr, and G. White, "Performance of MIMO systems with combined polarization multiplexing and transmit diversity," in *Proc. IEEE Trans on Vehicular Technology*, pp. 869–873, May 2005.