On Ultra-Wideband MIMO Systems with 1-bit Quantized Outputs: Performance Analysis and Input Optimization

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***Abstract*— We study the performance of multi-input multi-output (MIMO) channels with coarsely quantized outputs in the low signal-to-noise ratio (SNR) regime, where the channel is perfectly known at the receiver. This analysis is of interest in the context of Ultra-Wideband (UWB) communications from two aspects. First the available power is spread over such a large frequency band, that the power spectral density is extremely low and thus the SNR is low. Second the analog-to-digital converters (ADCs) for such high bandwidth signals should be low-resolution, in order to reduce their cost and power consumption. In this paper we consider the extreme case of only 1-bit ADC for each receive signal component. We compute the mutual information up to second order in the SNR and study the impact of quantization. We show that, up to first order in SNR, the mutual information of the 1-bit quantized system degrades only by a factor of $\frac{2}{\pi}$ compared to the system with infinite resolution independent of the actual MIMO channel realization. With Channel State Information (CSI) only at receiver, we show that QPSK is, up to the second order, the best among all distributions with independent components. We also elaborate on the ergodic capacity under this scheme in a Rayleigh flat-fading environment.**

I. Introduction

In the past few years, Ultra-Wideband communication combined with the use of multiple antennas at both sides of the transmission link has attracted much attention, as it is able to overcome the bandwidth-power tradeoff [1], [2]. The main motivation of UWB is that the capacity increases linearly with the bandwidth but only logarithmically with the power, therefore spreading it over a wide frequency band take us to the low SNR limit, where the capacity is asymptotically a linear function of the power. On the other hand, the degradation of the bandwidth-efficiency related with UWB communication can be partly compensated with the use of multiple antennas, at the cost of some additional processing complexity. However, a bigger challenge arises when we have to do with UWB systems, which concerns the analog-to-digital converters (ADCs). In fact, in order to reduce circuit complexity and save power and area, low resolution ADCs have to be employed [3], [4].

Several works studied MIMO channels in the context of UWB communication [2], [5], [6]. These studies are based on a second-order expansion of the mutual information of MIMO channels as SNR goes to zero, dealing with different classes of input and channel state information. Unfortunately, most of these contributions assume that the receiver has access to the channel data with infinite precision, and thus does not take into account the effects of quantization. In [7] and [8], the effects of quantization on the channel capacity in QPSK MIMO systems have been investigated. It turns out that the loss in channel capacity due to coarse quantization is surprisingly small.

Motivated by the second-order asymptotic approach, we study in this paper the impact of quantization on frequency flat MIMO systems at low SNR values, in terms of power penalty and optimal input distribution. We consider the extreme case of 1-bit quantized MIMO channel, with CSI at the receiver, but no CSI at the transmitter. When a single bit is used, the implementation of the all digital UWB receiver is considerably simplified [9], [10], [11]. In particular, automatic gain control (AGC) is not needed. Note that CSI can be obtained even with one-bit quantization [12]. We derive an expansion of the mutual information between the input and the quantized output up to the second order of the SNR for general input distributions. The first order term of this expansion reveals the linear growth of the capacity with the SNR, in the limit of infinite bandwidth, and is therefore a measure for the achievable performance in terms of power efficiency. On the other hand, by means of the second order term, which is negative, we can quantify how large the bandwidth should be in order to nearly reach the wideband limit. This aspect of wideband convergence is very important since the bandwidth is large but finite in practice (see [2]).

Our paper is organized as follows. Section II describes the system model and notational issues. In Section III we give the general expression of the mutual information between the inputs and the quantized outputs of the MIMO system, then we expand it into a Taylor series up to the second order of the SNR in Section IV. In Section V we derive the structure of the best independent-component input distribution at low SNR. Finally, in Section VI, we utilize these results to elaborate on the ergodic capacity in a Rayleigh flat-fading environment.

II. System Model And Notation

We consider a point-to-point quantized MIMO channel where the transmitter employs M antennas and the receiver has N antennas. Fig. 1 shows the general form of a quantized MIMO system, where $\boldsymbol{H} \in \mathbb{C}^{N\times M}$ is the channel matrix. The vector $\boldsymbol{x} \in \mathbb{C}^{M}$ comprises the M transmitted symbols with zero-mean and covariance matrix $\boldsymbol{Q} = \mathrm{E}[\boldsymbol{x}\boldsymbol{x}^{\mathrm{H}}]$. The average energy of $\boldsymbol{x}$ is fixed to 1, i.e., $\mathrm{tr}(\boldsymbol{Q}) = 1$. The vector $\boldsymbol{\eta}$ refers to uncorrelated zero-mean complex circular Gaussian noise with equal one-sided power spectral density per dimension, which is assumed to be equal $N_0 = 1$ without loss of generality. $\boldsymbol{r} \in \mathbb{C}^N$ is the unquantized channel output:

$$\boldsymbol{r} = \sqrt{\mathsf{SNR}}\boldsymbol{H}\boldsymbol{x} + \boldsymbol{\eta}, \tag{1}$$

where SNR represents the average signal-to-noise ratio at each receive antenna.

In our system, the real parts $r_{i,R}$ and the imaginary parts $r_{i,I}$ of the receive signal components r_i, $1 \leq i \leq N$, are each quantized by a 1-bit resolution quantizer. Thus, the resulting quantized signals read as:

$$y_{i,c} = \text{sign}(r_{i,c}) \in \{-1, 1\}, \text{ for } c \in \{R, I\}, \ 1 \leq i \leq N. \tag{2}$$

Throughout our paper, a_i denotes the i-th element of the vector $\boldsymbol{a}$ and $[\boldsymbol{a}]_{i,c} = a_{i,c}$ with $c \in \{R, I\}$ is the real or imaginary part of a_i. The operators $(\cdot)^{\text{H}}$ and $\text{tr}(\cdot)$ stand for Hermitian transpose and trace of a matrix, respectively.

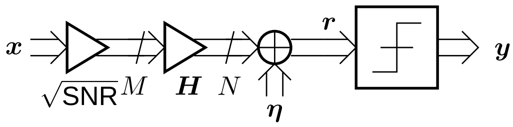

Fig. 1. One-bit Quantized MIMO System

III. Mutual Information

The mutual information (in nats/s/Hz) between the channel input and the quantized output in Fig. 1 reads as [13]

$$I(\boldsymbol{x}, \boldsymbol{y}) = \text{E}_{\boldsymbol{x}}\left[\sum_{\boldsymbol{y}} P(\boldsymbol{y}|\boldsymbol{x})\ln\frac{P(\boldsymbol{y}|\boldsymbol{x})}{P(\boldsymbol{y})}\right], \tag{3}$$

with $P(\boldsymbol{y}) = \text{E}_{\boldsymbol{x}}[P(\boldsymbol{y}|\boldsymbol{x})]$ and $\text{E}_{\boldsymbol{x}}[\cdot]$ is the expectation taken with respect to $\boldsymbol{x}$. Since all of the real and imaginary components of the receiver noise $\boldsymbol{\eta}$ are statistically independent with one-sided spectral level $\frac{1}{2}$, we can express each of the conditional probabilities as the product of the conditional probabilities on each receiver dimension

$$\begin{aligned} P(\boldsymbol{y}|\boldsymbol{x}) &= \prod_{c\in\{R,I\}}\prod_{i=1}^{N} P(y_{c,i}|\boldsymbol{x}) \\ &= \prod_{c\in\{R,I\}}\prod_{i=1}^{N} \Phi\left(\frac{y_{c,i}[\boldsymbol{H}\boldsymbol{x}]_{c,i}\sqrt{\mathsf{SNR}}}{\sqrt{1/2}}\right), \end{aligned} \tag{4}$$

with $\Phi(x) = \frac{1}{\sqrt{2\pi}}\int_{-\infty}^{x} e^{-\frac{t^2}{2}}dt$ is the cumulative normal distribution function.

IV. Second-order Expansion of the Mutual Information

In this section, we will elaborate on the second-order expansion of the input-output mutual information (3) of the considered system in Fig. 1 as the signal-to-noise ratio goes to zero.

Theorem 1: Consider the one-bit quantized MIMO system in Fig. 1 under an input distribution $p(\boldsymbol{x})$ with covariance matrix $\boldsymbol{Q}$, satisfying $p(\boldsymbol{x}) = p(\text{j}\boldsymbol{x}), \forall \boldsymbol{x} \in \mathbb{C}^M$ (zero-mean proper complex distribution)[1] and $\text{E}_{\boldsymbol{x}}[\|\boldsymbol{x}\|_4^{4+\alpha}] < \gamma$ for some finite constants $\alpha, \gamma > 0$. Then, to the second order, the mutual information (in nats) between the inputs and the quantized outputs is given by:

$$\begin{aligned} I(\boldsymbol{x}, \boldsymbol{y}) =& \frac{2}{\pi}\text{tr}(\boldsymbol{H}\boldsymbol{Q}\boldsymbol{H}^{\text{H}})\mathsf{SNR} - \Bigg[\frac{2}{\pi^2}\text{tr}((\text{nondiag}(\boldsymbol{H}\boldsymbol{Q}\boldsymbol{H}^{\text{H}}))^2) \\ &+ \frac{4}{3\pi}(1 - \frac{1}{\pi})\text{E}_{\boldsymbol{x}}[\|\boldsymbol{H}\boldsymbol{x}\|_4^4]\Bigg]\mathsf{SNR}^2 + \underbrace{\Delta I(\boldsymbol{x}, \boldsymbol{y})}_{o(\mathsf{SNR}^2)}, \end{aligned} \tag{5}$$

where $\text{nondiag}(\boldsymbol{A})$ is obtained from $\boldsymbol{A}$ by setting all its diagonal entries to zero, and $\|\boldsymbol{a}\|_4^4$ is the 4-norm of $\boldsymbol{a}$ taken to the power 4 defined as $\sum_{i,c} a_{i,c}^4$.

Proof: The proof is based on the Taylor series expansion of the function $\ln(t)$ and $\Phi(t)$ around $t = 0$ up to the order four. That is

$$\Phi(t) = \frac{1}{2} + \frac{1}{\sqrt{2\pi}}(t + 0 \cdot t^2 - \frac{t^3}{6} + 0 \cdot t^4 + o(t^4)) \tag{6}$$

and

$$\ln(1 + t) = t - \frac{t^2}{2} + \frac{t^3}{3} - \frac{t^4}{4} + o(t^4). \tag{7}$$

We simply apply these formulas to (4) and (3). The rest of the proof is very technical but straightforward and therefore omitted. The condition $\text{E}_{\boldsymbol{x}}[\|\boldsymbol{x}\|_4^{4+\alpha}] < \gamma$ for some finite constants $\alpha, \gamma > 0$ is necessary, so that the rest term of the expansion given by

$$\Delta I(\boldsymbol{x}, \boldsymbol{y}) = \text{E}_{\boldsymbol{x}}[o(\|\boldsymbol{x}\|_4^4\,\mathsf{SNR}^2)] \tag{8}$$

satisfies

$$\lim_{\mathsf{SNR}\to 0}\frac{\Delta I(\boldsymbol{x}, \boldsymbol{y})}{\mathsf{SNR}^2} = 0, \tag{9}$$

since

$$\begin{aligned} \Delta I(\boldsymbol{x}, \boldsymbol{y}) &= \text{E}_{\boldsymbol{x}}[o(\|\boldsymbol{x}\|_4^4\,\mathsf{SNR}^2)] \\ &\leq \text{E}_{\boldsymbol{x}}[(\|\boldsymbol{x}\|_4^4\,\mathsf{SNR}^2)^{1+\frac{\alpha'}{4}}], \text{ for some } \alpha' \in]0, \alpha] \\ &\leq \text{E}_{\boldsymbol{x}}[\|\boldsymbol{x}\|_4^{4+\alpha'}]\mathsf{SNR}^{2+\frac{\alpha'}{2}} \\ &\leq \text{E}_{\boldsymbol{x}}[\|\boldsymbol{x}\|_4^{4+\alpha}]^{\frac{4+\alpha'}{4+\alpha}}\mathsf{SNR}^{2+\frac{\alpha'}{2}} \text{(Hölder's inequality)} \\ &\leq \gamma^{\frac{4+\alpha'}{4+\alpha}}\mathsf{SNR}^{2+\frac{\alpha'}{2}} \\ &= o(\mathsf{SNR}^2). \end{aligned}$$

■

For comparison, we use the results of Prelov and Verdú [5] to express the mutual information (in nats) between the input $\boldsymbol{x}$ and the unquantized output $\boldsymbol{r}$ with the same input distribution as in *Theorem* 1:

$$\begin{aligned} I(\boldsymbol{x}, \boldsymbol{r}) = \text{tr}(\boldsymbol{H}\boldsymbol{Q}\boldsymbol{H}^{\text{H}})\mathsf{SNR} - \frac{\text{tr}((\boldsymbol{H}\boldsymbol{Q}\boldsymbol{H}^{\text{H}})^2)}{2}\mathsf{SNR}^2 + \\ + o(\mathsf{SNR}^2). \end{aligned} \tag{10}$$

Whereas the mutual information for the unquantized channel in (10), up to the second order, depends only on

[1]This restriction is simply justified by symmetry considerations.

the input covariance matrix, it depends in the quantized case (5), also on fourth order statistics of $\boldsymbol{x}$ (the fourth mixed moments of its components).

Now, using (5) and (10), we deduce the mutual information penalty in the wideband regime incurred by quantization

$$\lim_{\mathsf{SNR}\to 0} \frac{I(\boldsymbol{x},\boldsymbol{y})}{I(\boldsymbol{x},\boldsymbol{r})} = \frac{2}{\pi}, \qquad (11)$$

which is independent of the channel and the chosen distribution. Since the capacity is the supremum of the mutual information over the set of all input distributions, the same ratio holds for the capacity

$$\lim_{\mathsf{SNR}\to 0} \frac{C_{\text{1-bit}}}{C_{\infty\text{-bit}}} = \frac{2}{\pi}. \qquad (12)$$

This generalizes the result known for the AWGN channel [14].

Fig. 2 illustrates the mutual information for a randomly generated 4×4 channel[2] with QPSK signalling, computed exactly using (3), and its first and second-order approximations from (5). For comparison, the mutual information without quantization (using i.i.d. Gaussian input) is also plotted. Fig. 2 shows that the ratio $\frac{2}{\pi}$ holds for low to moderate SNR.

For a high number of antennas, the inner summation in (3) may be intractable. In this case the second-order approximation in (5) is advantageous to overcome the high complexity of the exact formula.

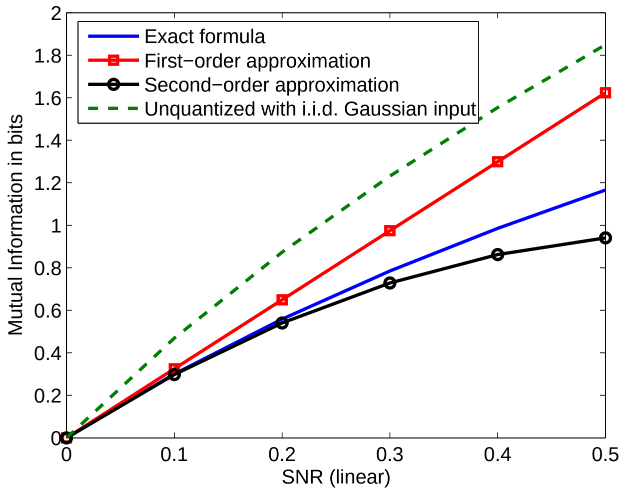

Fig. 2. Mutual information of a 1-bit quantized 4×4 QPSK MIMO system and its first and second-order approximations. For comparison the mutual information without quantization is also plotted.

V. Capacity with Independent-Component Input

Lacking knowledge of the channel (or its statistics), the transmitter assigns the power evenly over the components $x_{i,c}$ of the input vector $\boldsymbol{x}$, i.e. $\mathrm{E}[x_{i,c}^2] = \frac{1}{2M}$, in order to achieve good performance in average. Furthermore these components are assumed to be independent from each other.[3] Thus, the probability density function of the input vector $\boldsymbol{x}$ is $p(\boldsymbol{x}) = \prod_{i,c} p_{i,c}(x_{i,c})$.[4] Now, with

$$[\boldsymbol{H}\boldsymbol{x}]_{i,R} = \left(\sum_j [h_{i,j,R}x_{j,R} - h_{i,j,I}x_{j,I}]\right), \qquad (13)$$

$$\mu_{j,c} = \frac{\mathrm{E}[x_{j,c}^4]}{\mathrm{E}[x_{j,c}^2]^2} = 4M^2\mathrm{E}[x_{j,c}^4], \qquad (14)$$

and the *kurtosis* of the random component $x_{j,c}$ defined as

$$\kappa_{j,c} = \mu_{j,c} - 3, \qquad (15)$$

we get

$$\mathrm{E}_{\boldsymbol{x}}[([\boldsymbol{H}\boldsymbol{x}]_{i,R})^4] = \frac{1}{4M^2}\left(3\sum_{\substack{j,c,j',c'\\(j,c)\neq(j',c')}} h_{i,j,c}^2 h_{i,j',c'}^2 + \sum_{j,c}\mu_{j,c}h_{i,j,c}^4\right)$$
$$= \frac{1}{4M^2}\left(3\left(\left[\boldsymbol{H}\boldsymbol{H}^{\mathrm{H}}\right]_{i,i}\right)^2 + \sum_{j,c}\kappa_{j,c}h_{i,j,c}^4\right). \qquad (16)$$

Similar results hold for the other components of the vector $\boldsymbol{H}\boldsymbol{x}$. Plugging this result and $\boldsymbol{Q} = \frac{\mathbf{I}}{M}$ into (5), we obtain the expression of the mutual information with independent-component input and $\boldsymbol{Q} = \frac{\mathbf{I}}{M}$ up to the second order

$$I^{\text{ind}}(\boldsymbol{x},\boldsymbol{y}) \approx \frac{2}{\pi}\mathrm{tr}(\boldsymbol{H}\boldsymbol{H}^{\mathrm{H}})\frac{\mathsf{SNR}}{M} - \left[\frac{2}{\pi^2}\mathrm{tr}((\mathrm{nondiag}(\boldsymbol{H}\boldsymbol{H}^{\mathrm{H}}))^2)\right.$$
$$\left. + \frac{2}{3\pi}(1-\frac{1}{\pi})\left(3\ \mathrm{tr}((\mathrm{diag}(\boldsymbol{H}\boldsymbol{H}^{\mathrm{H}}))^2) + \sum_{i,j,c}\kappa_{j,c}h_{i,j,c}^4\right)\right]\frac{\mathsf{SNR}^2}{M^2}, \qquad (17)$$

where $\mathrm{diag}(\boldsymbol{A})$ is a diagonal matrix containing the diagonal elements of the matix $\boldsymbol{A}$.
Now, we state a theorem on the structure of the near-optimal input distribution under these assumptions.

Theorem 2: To the second order, QPSK is uniquely best among all distributions with independent components. The achieved capacity up to the second order is then

$$C_{\mathrm{Q}} \approx \frac{2}{\pi}\mathrm{tr}(\boldsymbol{H}\boldsymbol{H}^{\mathrm{H}})\frac{\mathsf{SNR}}{M} - \left[\frac{2}{\pi^2}\mathrm{tr}((\mathrm{nondiag}(\boldsymbol{H}\boldsymbol{H}^{\mathrm{H}}))^2)\right.$$
$$\left. + \frac{2}{3\pi}(1-\frac{1}{\pi})\left(3\ \mathrm{tr}((\mathrm{diag}(\boldsymbol{H}\boldsymbol{H}^{\mathrm{H}}))^2) - 2\sum_{i,j,c}h_{i,j,c}^4\right)\right]\frac{\mathsf{SNR}^2}{M^2}. \qquad (18)$$

Proof: Since $\mathrm{E}[x_{i,c}^4] \geq \mathrm{E}[x_{i,c}^2]^2$, we have $\kappa_{i,c} = \frac{\mathrm{E}[x_{i,c}^4]}{\mathrm{E}[x_{i,c}^2]^2} - 3 \geq -2, \forall i,c$. Obviously, the QPSK distribution is the unique distribution with independent-component input that can achieve all these lower bounds simultaneously, i.e., $\kappa_{i,c} = \kappa_{\mathrm{QPSK}} = -2\ \forall i,c$, and thus maximize $I^{\text{ind}}(\boldsymbol{x},\boldsymbol{y})$ in (17) up to the second order. ∎

[2] The generated entries $h_{i,j}$ of $\boldsymbol{H}$ are uncorrelated and $h_{i,j} \sim \mathcal{N}(0,1)$.

[3] Whether this is the optimal strategy for quantized MIMO systems, is an open problem.

[4] Note that $p_{i,c}(x_{i,c})$ have to be even functions and $p_{i,R}(x_{i,R}) = p_{i,I}(x_{i,I})\ \forall i$, due to the symmetry (see *Theorem* 1).

VI. Ergodic Capacity under QPSK Scheme

Now, the channel $\boldsymbol{H}$ is assumed to be ergodic with i.i.d. Gaussian components $h_{i,j} \sim \mathcal{N}(0,1)$. The ergodic capacity achieved by QPSK reads as

$$C_{\mathrm{Q}}^{\mathrm{erg}} = \mathrm{E}_{\boldsymbol{H}}[C_{\mathrm{Q}}]. \tag{19}$$

We apply the expectation over $\boldsymbol{H}$ to the second order expansion of C_{Q} in (18). Then, it is easy to show that

$$\begin{aligned}
\mathrm{E}_{\boldsymbol{H}}\left[\mathrm{tr}(\boldsymbol{H}\boldsymbol{H}^{\mathrm{H}})\right] &= MN & (20)\\
\mathrm{E}_{\boldsymbol{H}}\left[\mathrm{tr}\left((\mathrm{nondiag}(\boldsymbol{H}\boldsymbol{H}^{\mathrm{H}}))^2\right)\right] &= MN(N-1) & (21)\\
\mathrm{E}_{\boldsymbol{H}}\left[\mathrm{tr}\left((\mathrm{diag}(\boldsymbol{H}\boldsymbol{H}^{\mathrm{H}}))^2\right)\right] &= MN(M+1) & (22)\\
\sum_{i,j,c}\mathrm{E}_{\boldsymbol{H}}\left[h_{i,j,c}^4\right] &= \frac{3}{2}MN. & (23)
\end{aligned}$$

Finally, using (18) we obtain the following second-order expression for the ergodic capacity of one-bit quantized Rayleigh fading channels under QPSK scheme

$$C_{\mathrm{Q}}^{\mathrm{erg}} \approx N\frac{2}{\pi}\mathsf{SNR} - \frac{N(N+(\pi-1)M-1)}{2M}\left(\frac{2}{\pi}\mathsf{SNR}\right)^2. \tag{24}$$

Compared to the ergodic capacity in the unquantized case achieved by i.i.d. Gaussian input (or even by QPSK up to the second order) [2]

$$C^{\mathrm{erg}} \approx N\mathsf{SNR} - \frac{N(N+M)}{2M}\mathsf{SNR}^2, \tag{25}$$

the ergodic capacity of one-bit quantized MIMO under QPSK $C_{\mathrm{Q}}^{\mathrm{erg}}$ incorporates a power penalty of almost $\frac{\pi}{2}$ (1.96 dB), when considering only the linear term that characterizes the capacity in the limit of infinite bandwidth.

On the other hand, the second order term quantifies the convergence of the capacity function to the wideband limit by increasing the bandwidth [2]. Therefore, it can be observed from

$$1 < \frac{N+(\pi-1)M-1}{N+M} < \pi - 1, \tag{26}$$

that the quantized channel converges to this limit slower than the unquantized channel. Nevertheless, for $M = 1$ or $M \ll N$, this difference in the convergence behavior vanishes almost completely, since both second-order expansions (24) and (25) become nearly the same up to the factor $2/\pi$ in SNR.

In addition, the ergodic capacity of the quantized channel $C_{\mathrm{Q}}^{\mathrm{erg}}$ in (24) increases linearly with the number of receive antennas N and only sublinearly with the number of transmit antennas M, which holds also for C^{erg}. For the special case of one receive antenna, $N = 1$, $C_{\mathrm{Q}}^{\mathrm{erg}}$ does not depend on the number of transmit antennas M up to the second order, contrary to C^{erg}.

VII. Conclusion

We derived an expression for the second-order expansion of the mutual information of MIMO channels with one-bit ADC for low SNR and general input distribution. Based on this, we showed that the power penalty due to the 1-bit quantization is approximately equal $\frac{\pi}{2}$ (1.96 dB) at low SNR. This shows that mono-bit ADCs may be used to save system power without an excessive degradation in performance, and confirms the significant potential of the coarsely quantized UWB MIMO channel. Contrary to the unquantized channel, the second order term in the Taylor expansion depends not only on the covariance matrix of the input signal but also on its fourth order statistics. In the absence of channel knowledge at the transmitter, QPSK turned out to be the best among all distributions with independent components. We studied this special case in details and obtained the ergodic capacity for the i.i.d. Rayleigh channel model under QPSK scheme. An interesting topic for the future is to elaborate on the capacity-achieving input distribution for low SNR with and without Channel State Information (CSI) at the transmitter.

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