

Diversity Combining via Universal Orthogonal Space-Time Transformations

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Abstract

Receiver diversity methods play a key role in combating the detrimental effects of fading in wireless communication and other applications. Commonly used linear diversity methods include maximal ratio combining, equal gain combining and antenna selection combining. A novel linear combining method is proposed where a universal orthogonal dimension-reducing space-time transformation is applied prior to quantization of the signals. The scheme may be considered as the counterpart of Alamouti modulation, and more generally of orthogonal space-time block codes. The scheme is well-suited to reduced-complexity multiple receive-antenna analog-to-digital conversion of narrowband signals. It also provides a method to achieve diversity-enhanced relaying of communication signals, for multi-user detection at a remote terminal, minimizing the required bandwidth used in the links between relays and terminal.

I. INTRODUCTION

In wireless communication, diversity methods play a central role in combating the detrimental effects of severe channel variation (fading). Of the many techniques that have been developed over the years with this goal, an important class involves the use of multiple receive antennas. With sufficient separation between the antennas, each antenna may be viewed as a branch receiving the transmitted signal multiplied by an approximately independent fading coefficient. Diversity is achieved as the probability that the signal is severely affected by fading on all branches simultaneously is greatly reduced. The number of such (roughly) independent branches is commonly referred to as the diversity order. A classical survey of receive diversity techniques is [1]. More recent accounts that also consider multiple-input multiple-output channels are [2], [3].

We introduce a new diversity-combining scheme utilizing orthogonal space-time block codes. The key difference between the proposed scheme and traditional linear combining schemes is that it is *universal*. That is, the combining weights (in the proposed scheme, the space-time transformation) do not depend on the the channel realization.

To understand the potential benefits as well as the general nature of the contribution, consider a receiver as depicted in Figure 1. A key feature of modern device architectures is the decomposition of the unit into separate functional blocks (that can be located at different physical locations, i.e., distributed processing). These blocks are connected by interfaces and a major design goal is to reduce the bandwidth between different blocks.

The proposed scheme can assist in the interface from the analog domain to the digital one, simplifying analog-to-digital conversion (ADC) and thus also reducing power consumption. The scheme can equally assist in reducing the bandwidth of the digital interface between different digital blocks. For example, in a centralized (cloud) radio access network (C-RAN) setting, this bandwidth reduction will be in the fronthaul links between the relays (remote radio head units) and the central (cloud) processing unit.

The rest of the paper is organized as follows. In Section II we describe the scheme in the most basic setup of a single-input multiple-output (SIMO) system with only two receive antennas. In Section III we describe the application of the scheme as an ADC architecture for narrowband signal acquisition. Section IV describes an application of the scheme to relaying for multi-user detection at a remote destination (cloud). The paper concludes with Section V where a discussion of possible extensions of the method to more than two antennas is presented.

II. DESCRIPTION OF THE SCHEME FOR TWO RECEIVE ANTENNAS

Consider a 2×1 SIMO channel as depicted in Figure 1. The signal received at antenna $i = 1, 2$, at discrete time t is given by

$$s_i(t) = h_i x(t) + n_i(t). \quad (1)$$

We assume that the noise $n_i(t)$ is i.i.d. over space and time with samples that are circularly-symmetric complex Gaussian with unit variance.

The scheme works on batches of two time instances and for our purposes, it will suffice to describe it for time instances $t = 1, 2$. Let us stack these four complex samples received over $T = 2$ time instances, two over each antenna, into an 8×1 real vector:

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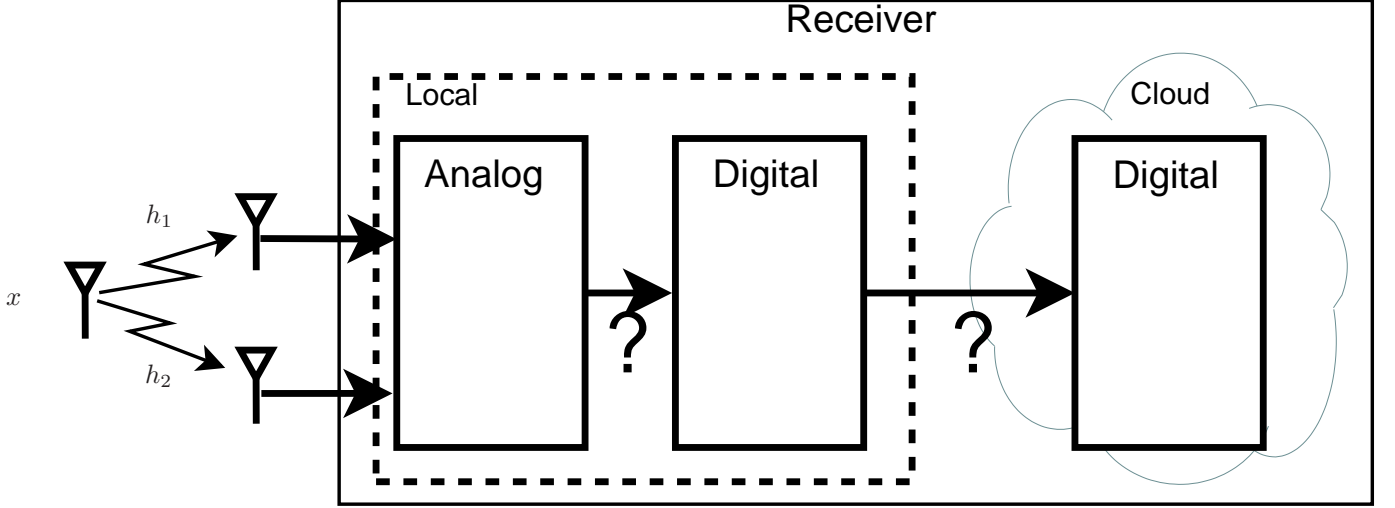


Fig. 1. Basic scenario: receiver architecture for a 2×1 SIMO channel.

$$\mathbf{s} = [s_{1R}(1)s_{1I}(1)s_{2R}(1)s_{2I}(1)s_{1R}(2)s_{1I}(2)s_{2R}(2)s_{2I}(2)]^T, \quad (2)$$

where x_R and x_I denote the real and imaginary parts of a complex number x . We similarly define the stacked noise vector \mathbf{n} . Likewise, we define

$$\mathbf{x} = [x_R(1)x_I(1)x_R(2)x_I(2)]^T. \quad (3)$$

Next, we form a 4×1 real vector \mathbf{y} by applying to the vector \mathbf{s} the transformation

$$\mathbf{y} = \mathbf{P}\mathbf{s}$$

where

$$\mathbf{P} = \frac{1}{\sqrt{2}} \begin{bmatrix} 1 & 0 & 0 & 0 & 0 & 0 & 1 & 0 \\ 0 & 1 & 0 & 0 & 0 & 0 & 0 & -1 \\ 0 & 0 & 1 & 0 & -1 & 0 & 0 & 0 \\ 0 & 0 & 0 & 1 & 0 & 1 & 0 & 0 \end{bmatrix}. \quad (4)$$

Note that unlike conventional linear diversity-combining schemes, here the combining matrix \mathbf{P} is *universal*, i.e., it does not depend on the channel coefficients.

Remark 1: We note that the transpose of \mathbf{P} is precisely the description of the linear operation performed by Alamouti modulation when expressed over the reals [4].

It is not hard to show that the following holds

$$\begin{aligned} \mathbf{y} &= \frac{\|\mathbf{h}\|}{\sqrt{2}} \mathbf{U}(h_1, h_2) \mathbf{x} + \mathbf{P}\mathbf{n} \\ &= \frac{\|\mathbf{h}\|}{\sqrt{2}} \mathbf{U}(h_1, h_2) \mathbf{x} + \mathbf{n}', \end{aligned} \quad (5)$$

where

$$\mathbf{U}(h_1, h_2) = \frac{1}{\|\mathbf{h}\|} \begin{bmatrix} h_{1R} & -h_{1I} & h_{2R} & -h_{2I} \\ h_{1I} & h_{1R} & -h_{2I} & -h_{2R} \\ h_{2R} & -h_{2I} & -h_{1R} & h_{1I} \\ h_{2I} & h_{2R} & h_{1I} & h_{1R} \end{bmatrix}. \quad (6)$$

A key observation is that $\mathbf{U}(h_1, h_2)$ is an orthonormal matrix for any h_1, h_2 :

$$\mathbf{U}^H(h_1, h_2) \mathbf{U}(h_1, h_2) = \mathbf{I}, \quad (7)$$

where \mathbf{I} is the identity matrix. Further, since the rows of \mathbf{P} are orthonormal, it follows that \mathbf{n}' is white and Gaussian with unit variance.

We may reconstruct (up to additive noise) the original samples by applying

$$\begin{aligned}\hat{\mathbf{x}} &= \mathbf{U}^H(h_1, h_2) \cdot \mathbf{y} \\ &= \frac{\|\mathbf{h}\|}{\sqrt{2}} \mathbf{x} + \mathbf{n}''\end{aligned}\quad (8)$$

where \mathbf{n}'' is also white and Gaussian with unit variance

Noting that the dimension (over the reals) of \mathbf{y} is four rather than eight which is the dimension of \mathbf{x} , we obtained a dimension-reducing linear-combining scheme that may be considered as the counterpart of Alamouti transmit-diversity precoding [4].

We note that had we eliminated the universality requirement (i.e., allowing the transformation \mathbf{P} to depend on the specific channel realization), maximal-ratio combining (MRC) could be applied, yielding sufficient statistics. As further discussed in the next section, MRC results in the effective channel

$$y_{\text{MRC}} = \|\mathbf{h}\|x + n. \quad (9)$$

It follows that the proposed method has a power loss amounting to a factor of two in signal-to-noise ratio (SNR) with respect to optimal processing.

We next describe several applications in which the proposed scheme may be beneficial.

III. APPLICATION TO ANALOG-TO-DIGITAL CONVERSION

In this section we demonstrate the applicability of the scheme to analog-to-digital conversion for power-limited receivers for narrowband signals [5]–[7].

The proposed method may be used to achieve maximal diversity order with a single radio-frequency (RF) chain and ADC, and without requiring selection and switching mechanisms that comes at substantial analog hardware complexities; see discussion of hardware aspects in [3].

Consider again the scenario of a 2×1 SIMO system as depicted in Figure 1 and described in the previous section. We note that as the fading coefficients are constants (rather than impulse responses), the model assumed is that of frequency-flat fading.

The role of diversity is easiest to understand by assuming first that the receiver front end arbitrarily processes the output of one antenna only, say, only the output of antenna 1, as depicted in Figure 2. Thus, only the signal $s_1(t)$ in (1) passes

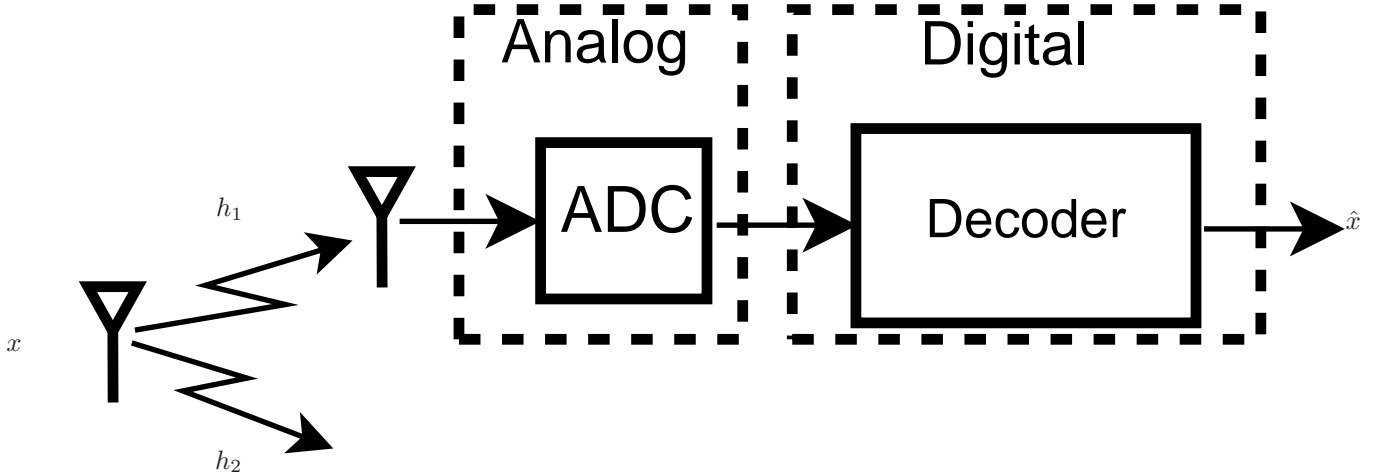


Fig. 2. Demonstration of a receiver that chooses an arbitrary branch for processing.

through the RF chain and a single ADC unit suffices. The performance however is far from robust as a fade of a single channel coefficient (h_1 with our arbitrary choice) will result in highly degraded signal-to-noise ratio (SNR). In an i.i.d. Rayleigh fading environment, the bit error rate for uncoded transmission will decay as

$$P_e \sim \text{SNR}^{-d} \quad (10)$$

where here $d = 1$, and we say that we have first order diversity. Similarly, in a system where $d = 2$ as will be described below, we have a diversity order of 2.

The best performance may be attained by quantizing (at sufficient resolution) the output of each antenna and then using MRC as depicted in Figure 3. Applying MRC amounts to forming

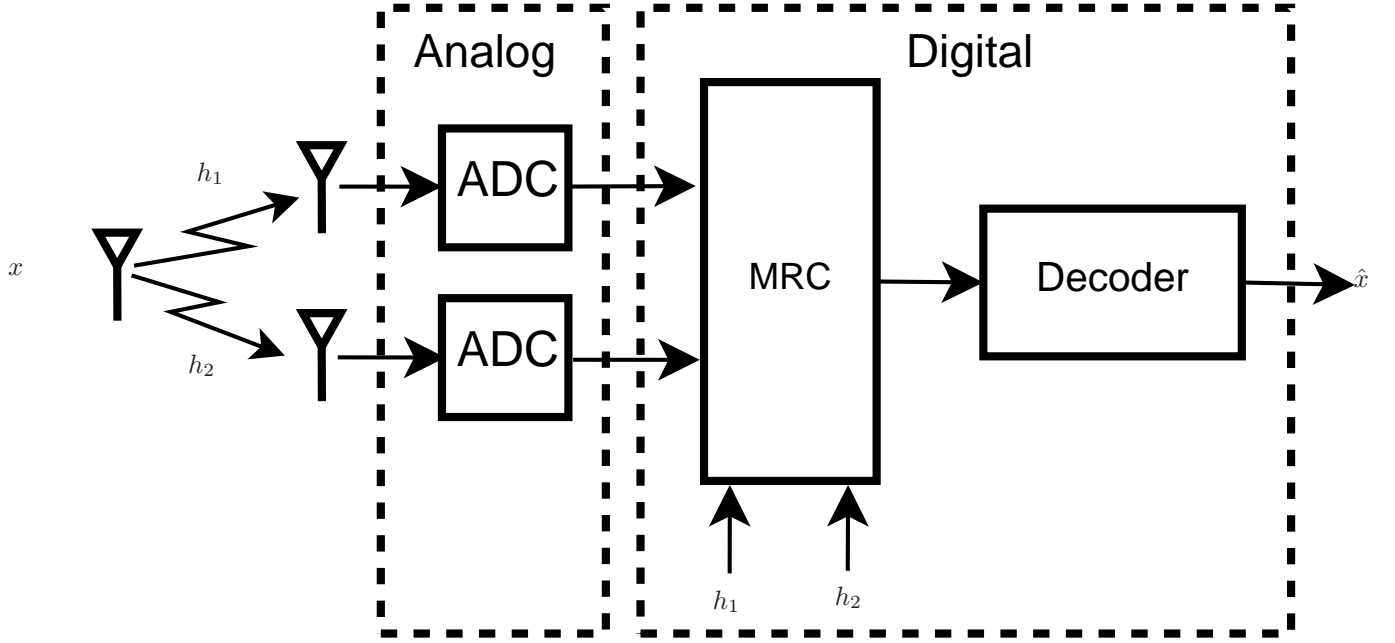


Fig. 3. A receiver employing a maximal-ratio combining front end.

$$\begin{aligned}
 y_{\text{MRC}} &= \frac{1}{\|\mathbf{h}\|} \begin{bmatrix} h_1^* & h_2^* \end{bmatrix} \begin{bmatrix} s_1 \\ s_2 \end{bmatrix} \\
 &= \frac{\|h_1\|^2 + \|h_2\|^2}{\|\mathbf{h}\|} x + \frac{h_1^* n_1 + h_2^* n_2}{\|\mathbf{h}\|} \\
 &= \|\mathbf{h}\| x + n',
 \end{aligned} \tag{11}$$

where n' is white and Gaussian with unit variance. As the variation of $\|\mathbf{h}\|$ is much smaller than that of either h_1 or h_2 , diversity is attained. This may intuitively be understood by noticing that both h_1 and h_2 have to vanish in order for h to vanish. When h_1 and h_2 are independent, we obtain a diversity order of 2. The precise performance of MRC under independent Rayleigh fading is well-known and may be found, e.g., in [1]. The major downside of such a system is that two RF chains and ADCs are needed.

A classic alternative to MRC that requires only one RF chain is the method of antenna selection or “selection combining” as depicted in Figure 4.

Here, rather than choosing the antenna arbitrarily, we choose the one with the higher SNR. Thus the effective channel becomes

$$y_{\text{SC}} = \max(|h_1|, |h_2|)x + n, \tag{12}$$

where again n is Gaussian noise of unit variance. While the performance does not reach that of MRC, it does attain a diversity order of 2. The precise performance under independent Rayleigh fading of selection combining is well-known and may be found, e.g., in [1]. One downside of the selection combining method is that it requires analog detection and switching mechanisms.

The novel space-time diversity combining method described in the previous section is applied as follows. Since the processing matrix \mathbf{P} is fixed for all channels (i.e., is universal), it can be applied in the analog domain (i.e., prior to quantization). Therefore, as depicted in Figure 5, the received signals are first passed through the dimension-reducing transformation \mathbf{P} to obtain the vector

$$\mathbf{y} = [y^1, y^2, y^3, y^4]^T \tag{13}$$

as defined in (5) and (6). Then, a (component-wise) scalar uniform quantizer $Q(\cdot)$ is applied to \mathbf{y} to obtain

$$\mathbf{y}_q = Q(\mathbf{y}).$$

We denote the quantization error vector by

$$\mathbf{e} = \mathbf{y} - \mathbf{y}_q \tag{14}$$

$$= \mathbf{y} - Q(\mathbf{y}). \tag{15}$$

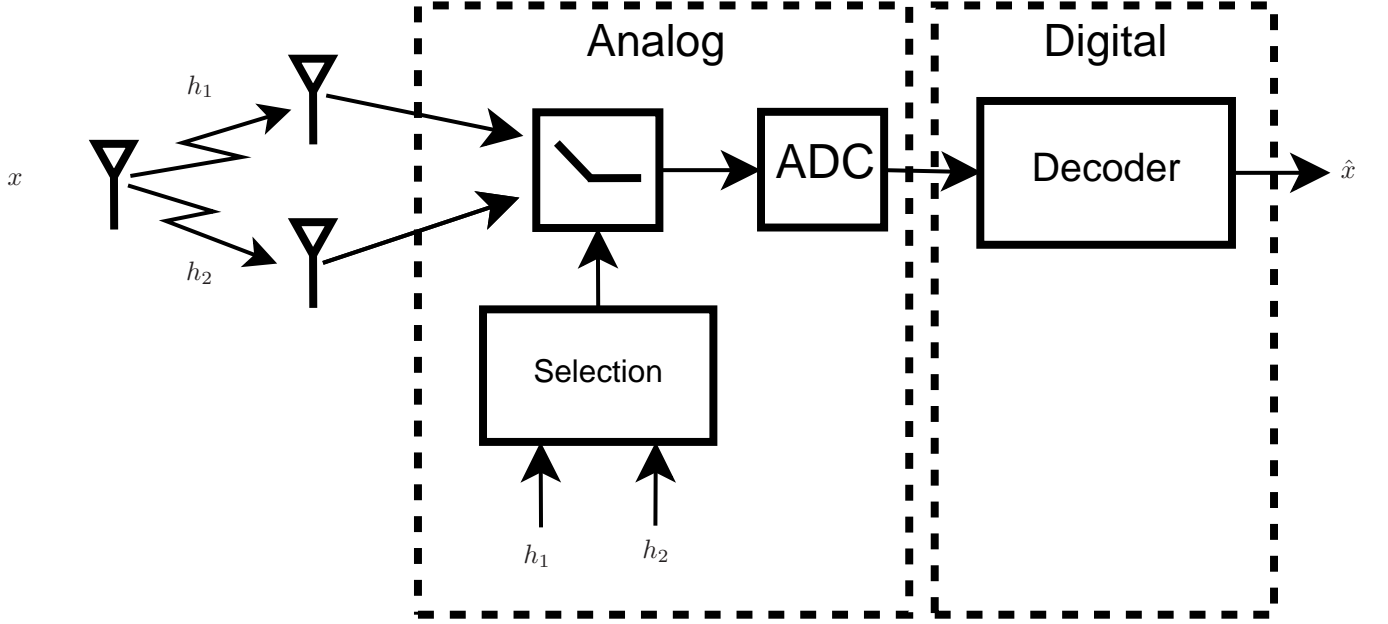


Fig. 4. A receiver employing selection-diversity combining as a front end.

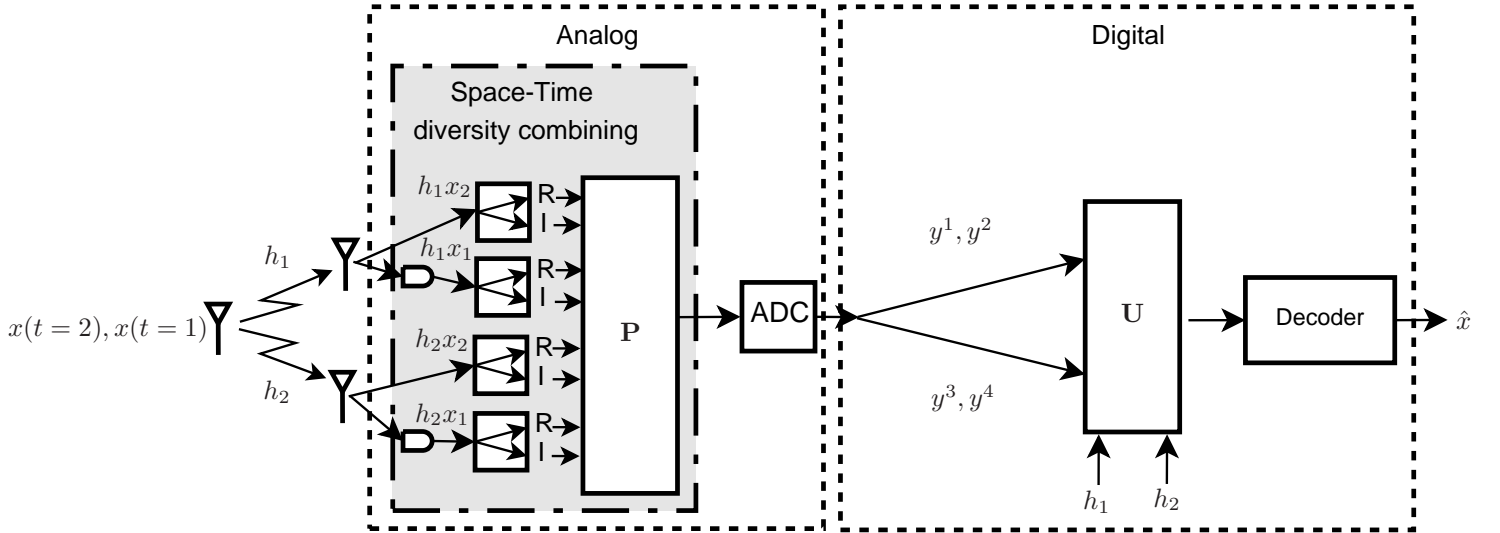


Fig. 5. Novel receiver front end employing an orthogonal space-time diversity transformation.

The sequence of quantized samples is used to reconstruct an estimation of the source vector

$$\hat{\mathbf{x}} = \begin{bmatrix} \widehat{x_R(t=1)} \\ \widehat{x_I(t=1)} \\ \widehat{x_R(t=2)} \\ \widehat{x_I(t=2)} \end{bmatrix} \quad (16)$$

by applying the transformation:

$$\hat{\mathbf{x}} = \mathbf{U}(h_1, h_2)^H \mathbf{y}_q. \quad (17)$$

Using (5) and (15), we have

$$\hat{\mathbf{x}} = \mathbf{U}(h_1, h_2)^H (\mathbf{y} - \mathbf{e}) \quad (18)$$

$$= \mathbf{U}(h_1, h_2)^H \left(\frac{\|\mathbf{h}\|}{\sqrt{2}} \mathbf{U}(h_1, h_2) \mathbf{x} + \mathbf{n}' - \mathbf{e} \right) \quad (19)$$

$$= \frac{\|\mathbf{h}\|}{\sqrt{2}}\mathbf{x} + \mathbf{n}'' - \mathbf{e}', \quad (20)$$

where \mathbf{n}'' has the same distribution as \mathbf{n} .

As for the quantization error \mathbf{e} and its transformed variant \mathbf{e}' , we may invoke the standard assumption, that may be justified using subtractive dithered quantization, that it is independent of the signal (and hence of \mathbf{x}) and is white (i.e., its covariance matrix is the scaled identity).

We conclude that the input/output relationship of the proposed diversity combiner is identical to that of MRC, except for a power loss of a factor of two. In other words, we attain full diversity but no array gain, precisely as in the case of Alamouti space-time diversity transmission.

It follows that the scheme also attains a diversity order of 2 but loses precisely a factor of two in terms of SNR with respect to MRC. In comparison with selection combining (without taking into account implementation losses), there is a loss in the achieved SNR whereas an advantage is that no estimation of channel quality in the analog front end nor switching is required.

A comparison of the performance of the proposed method is shown in Figure 6 that plots the bit error rate of all three methods for uncoded QPSK transmission.

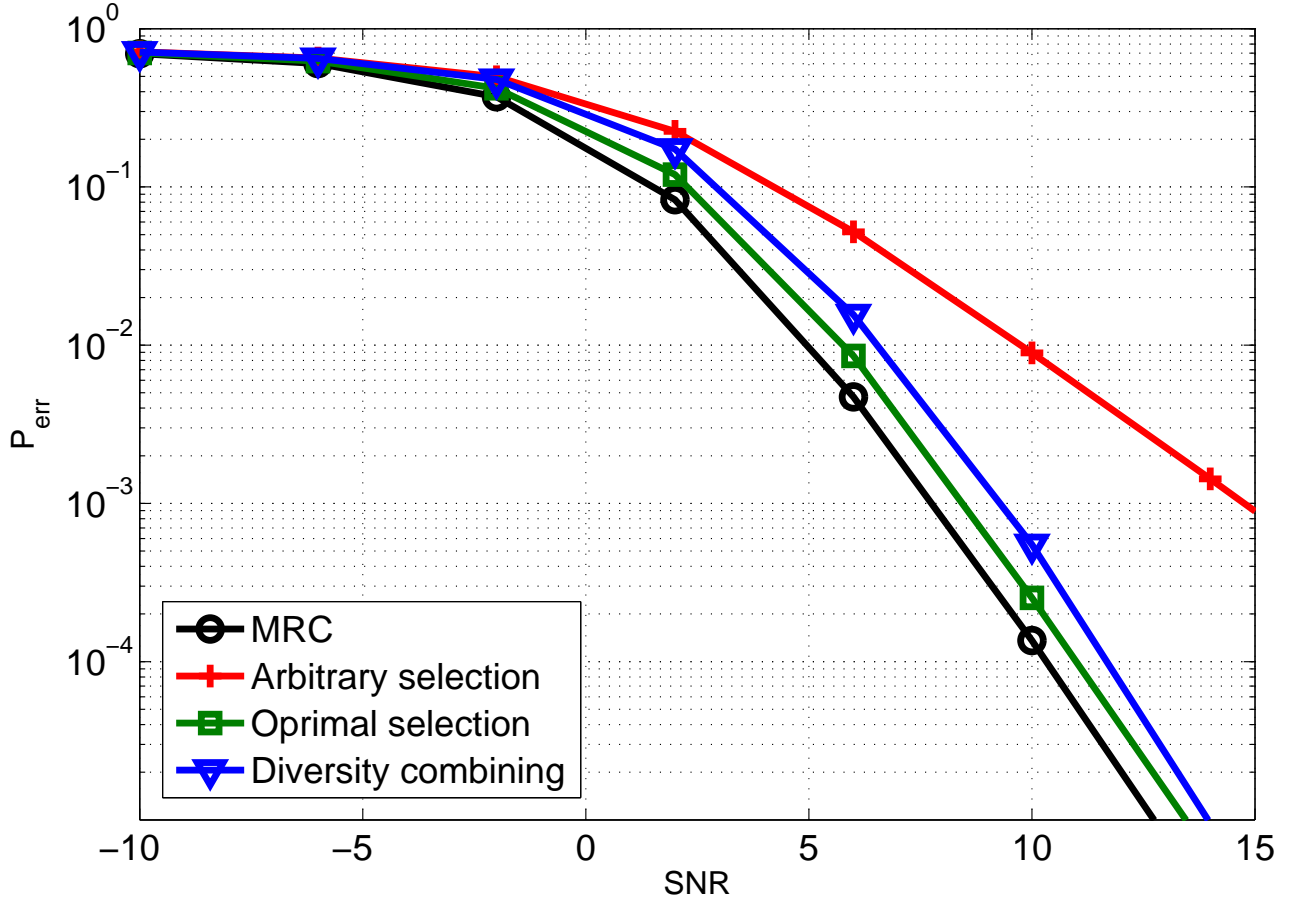


Fig. 6. Performance of bit error rate for uncoded QPSK of the proposed space-time diversity scheme and comparison with alternatives.

IV. APPLICATION TO RELAYING FOR MULTI-USER DETECTION AT A REMOTE DESTINATION

The real strength of the proposed method appears in distributed scenarios as we now exemplify in the context of 5G multi-user detection in the cloud. [8]–[11].

Unlike in the previous section, the scheme we present now operates purely in the digital domain. We assume that each antenna is sampled separately and the benefit is not in the realm of hardware simplification but rather in making efficient use of the limited bit rate available in the fronthaul link in a C-RAN setting.

A further difference is that we no longer assume frequency-flat fading. Rather, we will assume that after analog-to-digital conversion, a DFT operation is applied, thus working in the frequency domain. In other words, the static channel we will

consider is to be understood to apply to a single tone. The “time” index t will correspondingly refer to subsequent uses of the same tone, or in a practical setting could apply to adjacent tones as these typically have very similar channel coefficients.

As a baseline system, consider the system depicted in Figure 7 in which two users wish to communicate with a base station via two relays, where both users as well as the relays are equipped with a single antenna. The relays are connected to the base station via rate-constrained bit pipes.

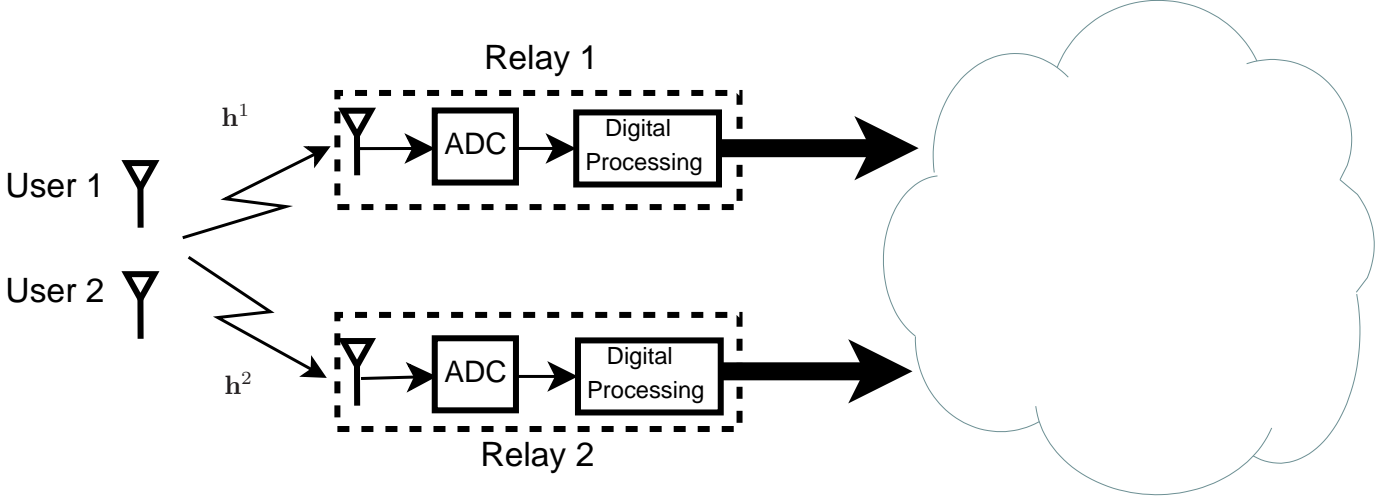


Fig. 7. Baseline uplink scenario in C-RAN: a two-user virtual MIMO system formed by two single-antenna relays connected to the cloud via rate-constrained fronthaul links.

The latter scenario has been extensively studied and we refer the reader to [11] and references therein for background. In this section we show how the space-time diversity method developed in the present work can serve as a simple and practical means to reap a substantial part of the possible gains.

The received signal at relay $i = 1, 2$ is given by

$$s^i(t) = h_{11}^i \cdot x_1(t) + h_{12}^i \cdot x_2(t) + n_1^i(t), \quad (21)$$

where h_{jk}^i denotes the channel gain between user k and antenna j in relay i . Since we are considering at this point single-antenna relays, $j = 1$.

We employ the most basic variant of the quantize-and-forward protocol where correlation between the received signals is ignored and each relay simply quantizes the received signal, possibly after applying linear processing, and then sends the quantized output via its bit pipe. We further assume that only linear equalization is performed at the cloud prior to decoding.

To simplify the exposition and since the quantization noise will again behave in much the same manner in all schemes to be considered, we do not account for it in the system description in the sense that we assume that it is small with respect to the Gaussian noise (although it plays an important role in determining the rate needed to be supported by the bit pipes).

The base station thus receives at each time instance (the time index plays no role at this point) the vector

$$\mathbf{y} = [s_1(t), s_2(t)]^T.$$

We assume that it also knows (is able to estimate) the channel coefficients and thus can form the combined effective matrix

$$\mathbf{G} = \begin{bmatrix} h_{11}^1 & h_{12}^1 \\ h_{11}^2 & h_{12}^2 \end{bmatrix}, \quad (22)$$

where the upper part of the matrix corresponds to the signal originating from relay 1 and the bottom part corresponds to relay 2.

It next applies a linear equalizer to obtain a soft estimation of the source vector. For instance, it may use a zero forcing equalizer,

$$\begin{bmatrix} \widehat{x_1(t)} \\ \widehat{x_2(t)} \end{bmatrix} = \mathbf{G}^{-1} \mathbf{y}, \quad (23)$$

or a linear minimum mean square error (MMSE) equalizer. Assuming uncoded transmission for simplicity, the latter vector is then fed into a slicer.

Now consider the same scenario except that each relay is now equipped with two antennas, rather than one, as depicted in Figure 8. Thus, received signal at relay $i = 1, 2$ and antenna $j = 1, 2$ is given by

$$s_j^i(t) = h_{j1}^i \cdot x_1(t) + h_{j2}^i \cdot x_2(t) + n_j^i(t), \quad (24)$$

Thus, the channel matrix of relay i is

$$\mathbf{H}^i = \begin{bmatrix} h_{11}^i & h_{12}^i \\ h_{21}^i & h_{22}^i \end{bmatrix}. \quad (25)$$

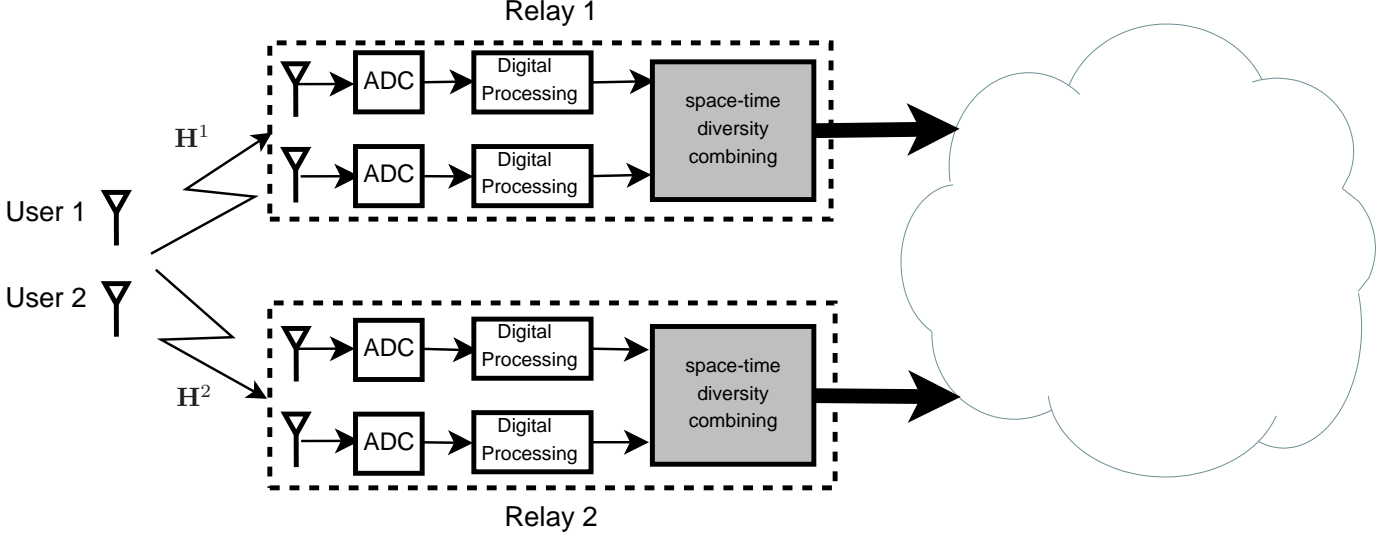


Fig. 8. Multi-antenna uplink scenario in C-RAN: two-user virtual MIMO system formed by two two-antenna relays connected to the cloud via rate-constrained fronthaul links.

The question now arises as to how best to utilize the finite number of bits available per sample in quantizing the output of the two antennas. Due to the distributed nature of the problem, both MRC and selection combining are inapplicable as the base station is interested in recovering both signals. A natural question is therefore: Can linear processing prior to quantization be useful at all?

We now demonstrate that while keeping the bit rate fixed (with respect to the case of a single-antenna relay), each relay can nonetheless provide diversity gains to *both users* using the novel combining method, precisely since it makes no use of channel state information at the linear combining stage, rather only in the reconstruction stage.

Assuming both relays use the proposed space-time diversity combining scheme, the signal passed to the cloud from relay i is given by

$$\mathbf{y}^i = \mathbf{U}(h_{11}^i, h_{21}^i) \mathbf{x}_1 + \mathbf{U}(h_{12}^i, h_{22}^i) \mathbf{x}_2 + \mathbf{n}'^i, \quad (26)$$

where \mathbf{x}_j represents the real representation of the signal transmitted by user j over the two time instances according to the notation in (3). Thus, at the cloud we obtain the effective channel

$$\begin{bmatrix} \mathbf{y}^1 \\ \mathbf{y}^2 \end{bmatrix} = \underbrace{\begin{bmatrix} \mathbf{U}(h_{11}^1, h_{21}^1) & \mathbf{U}(h_{12}^1, h_{22}^1) \\ \mathbf{U}(h_{11}^2, h_{21}^2) & \mathbf{U}(h_{12}^2, h_{22}^2) \end{bmatrix}}_{\mathcal{G}} \begin{bmatrix} \mathbf{x}_1 \\ \mathbf{x}_2 \end{bmatrix} + \begin{bmatrix} \mathbf{n}'^1 \\ \mathbf{n}'^2 \end{bmatrix}. \quad (27)$$

Note that the effective matrix \mathcal{G} has the desirable property that each of the four submatrices is orthogonal. Thus, it is expected that applying linear equalization to the effective channel followed by a slicer (or in general, a decoder) will perform better than the baseline scheme.

The performance of the proposed scheme is demonstrated in Figures 9 and 10 where substantial improvement may be seen with respect to the baseline (arbitrary selection) scheme. Figure 9 shows the bit error rate achieved by the scheme for uncoded QPSK transmission as well as the baseline scheme, which may be thought of as choosing one antenna at arbitrary. As another benchmark, we also plot the performance obtained for the same receiver structure (linear MMSE equalization followed by a slicer) where the central receiver obtains the output of all four antennas (two from each relay).

Figure 10 compares the three schemes with the same (linear MMSE) equalization but coupled with ideal coding over many i.i.d. realization of the channel. Specifically, it depicts the ergodic capacity associated with each combining method.

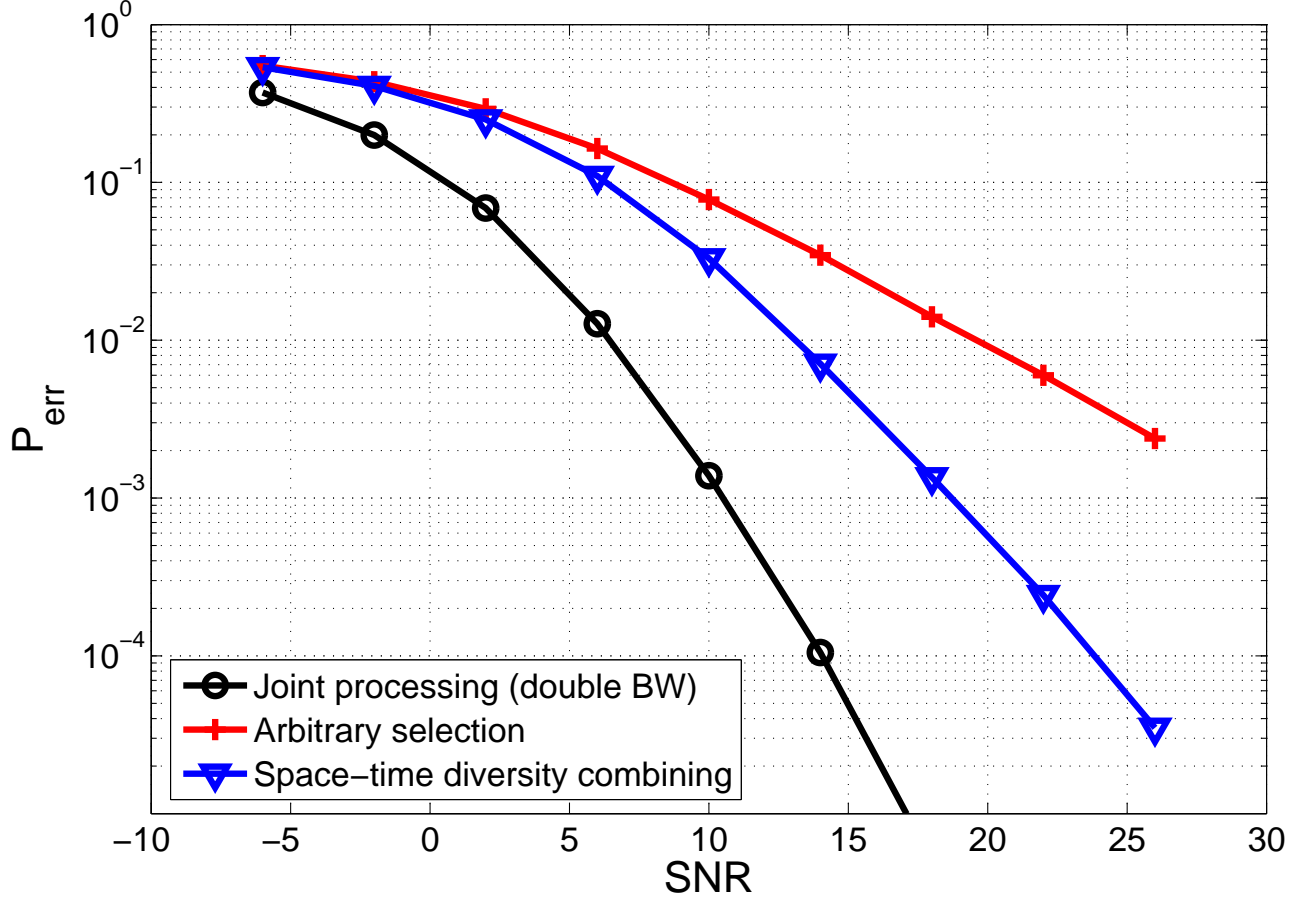


Fig. 9. Bit error rate for uncoded QPSK with quantize-and-forward relaying using the proposed linear combining method and comparison with alternatives.

V. EXTENSIONS TO MORE THAN TWO ANTENNAS

As in the case of space-time modulation for channel coding, extension to more receive antennas is possible, albeit with some loss.

A natural approach is to try utilizing the theory of orthogonal designs. It should be noted however that it is well known that the decoding delays (number of time instances stacked together) roughly grows exponentially with the number of antennas. Another possible avenue is to try to follow the approach of quasi-orthogonal space-time codes as developed in [12]–[14].

Attempting to apply orthogonal designs, one immediately confronts a basic obstacle due to the fact that rate-1 complex orthogonal designs do not exist beyond the case of two antennas. We next demonstrate the problem that arises and also show how it may be resolved by judiciously combining balanced rate-1/2 orthogonal designs [15] (which includes the four basic OSTBCs described in [16] for 2-8 antennas) with repeated quantization used in conjunction with multiplicative dithering. For the sake of concreteness and ease of exposition, we demonstrate the extension of the method, for the case of four receive antennas, in the context of analog-to-digital conversion (as discussed in Section III). This extension translates in a straightforward manner also to the relaying scenario described in Section IV.

The received signals are given by (1) where now $i = 1, \dots, M$ (with $M = 4$). We proceed by stacking $T = 8$ time instances of the received signal from the 4 antennas and build an effective real-valued vector by decomposing each entry into its real and imaginary components, just as is done in (2). This yields for $M = 4$, a vector \mathbf{s} of dimension $2 \times 4 \times 8 = 64$.

By reinterpreting the rate-1/2 orthogonal design of 4 transmit antennas (see [16]), we arrive at a 8×64 transformation matrix \mathbf{U} as displayed in (36).

Next, we form a 8×1 real vector \mathbf{y} by applying to the effective received vector \mathbf{s} , formed in the manner described in (2), the transformation

$$\mathbf{y} = \mathbf{U}\mathbf{s}$$

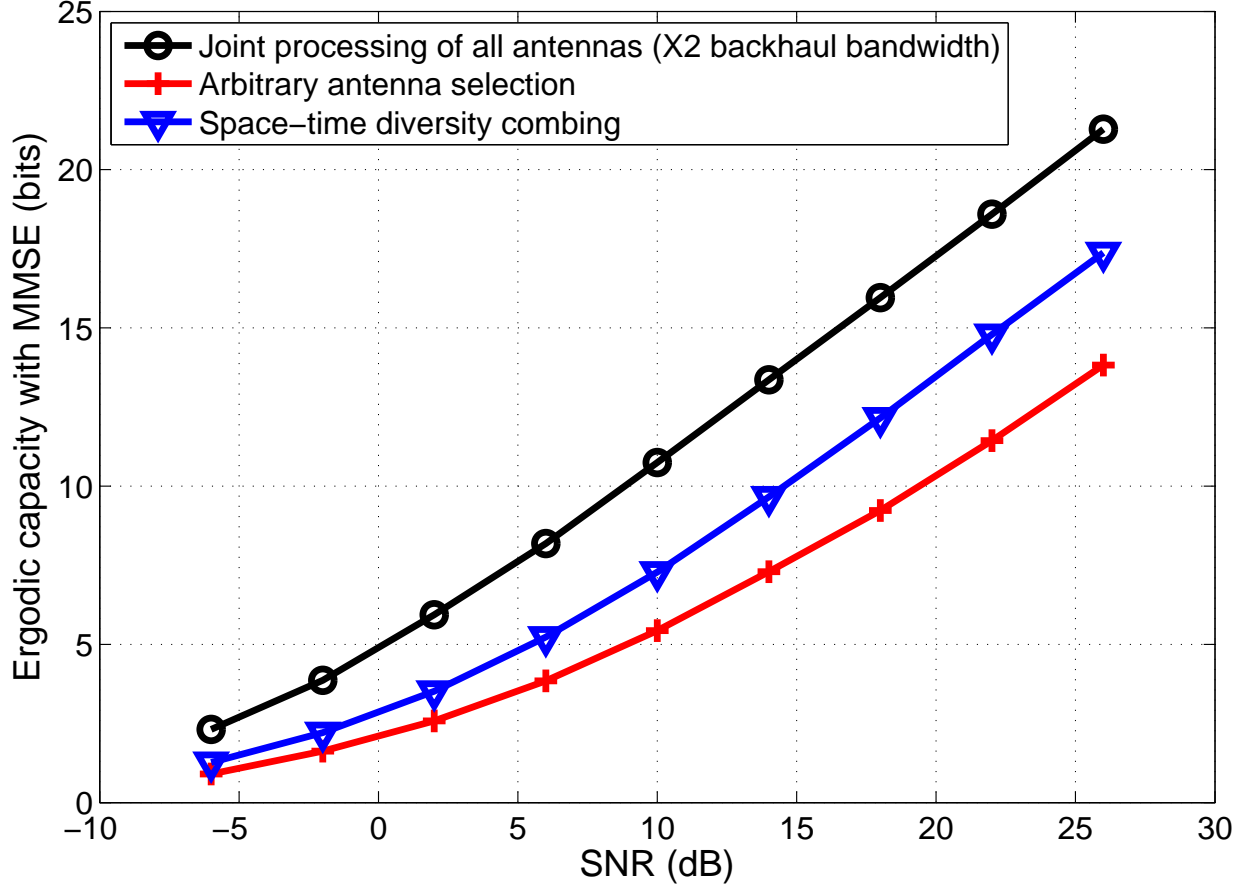


Fig. 10. Comparison of the ergodic capacity of quantize-and-forward relaying coupled with linear MMSE equalization and decoding in the cloud, for the proposed combining method and comparison with alternatives.

It can be shown that the following holds

$$\begin{aligned}
 \mathbf{y} &= \frac{\sqrt{2}\|\mathbf{h}\|}{\sqrt{8}} \mathbf{U}(h_1, h_2, h_3, h_4) \mathbf{x} + \mathbf{U} \mathbf{n} \\
 &= \frac{\|\mathbf{h}\|}{2} \mathbf{U}(h_1, h_2, h_3, h_4) \mathbf{x} + \mathbf{n}',
 \end{aligned} \tag{28}$$

where $\mathbf{U}(h_1, h_2, h_3, h_4)$ is defined in (29). Here, the vector \mathbf{x} is the 16-dimensional real representation of the transmitted signal over $T = 8$ time instances, formed analogously to (3).

$$\sqrt{2}\|\mathbf{h}\| \cdot \mathbf{U}(h_1, h_2, h_3, h_4) =$$

$$\begin{bmatrix}
h_{1R} & h_{1I} & h_{2R} & h_{2I} & h_{3R} & h_{3I} & h_{4R} & h_{4I} \\
-h_{1I} & h_{1R} & -h_{2I} & h_{2R} & -h_{3I} & h_{3R} & -h_{4I} & h_{4R} \\
h_{2R} & h_{2I} & -h_{1R} & -h_{1I} & h_{4R} & h_{4I} & -h_{3R} & -h_{3I} \\
-h_{2I} & h_{2R} & h_{1I} & -h_{1R} & -h_{4I} & h_{4R} & h_{3I} & -h_{3R} \\
h_{3R} & h_{3I} & -h_{4R} & -h_{4I} & -h_{1R} & -h_{1I} & h_{2R} & h_{2I} \\
-h_{3I} & h_{3R} & h_{4I} & -h_{4R} & h_{1I} & -h_{1R} & -h_{2I} & h_{2R} \\
h_{4R} & h_{4I} & h_{3R} & h_{3I} & -h_{2R} & -h_{2I} & -h_{1R} & -h_{1I} \\
-h_{4I} & h_{4R} & -h_{3I} & h_{3R} & h_{2I} & -h_{2R} & h_{1I} & -h_{1R} \\
h_{1R} & -h_{1I} & h_{2R} & -h_{2I} & h_{3R} & -h_{3I} & h_{4R} & -h_{4I} \\
-h_{1I} & -h_{1R} & -h_{2I} & -h_{2R} & -h_{3I} & -h_{3R} & -h_{4I} & -h_{4R} \\
h_{2R} & -h_{2I} & -h_{1R} & h_{1I} & h_{4R} & -h_{4I} & -h_{3R} & h_{3I} \\
-h_{2I} & -h_{2R} & h_{1I} & h_{1R} & -h_{4I} & -h_{4R} & h_{3I} & h_{3R} \\
h_{3R} & -h_{3I} & -h_{4R} & h_{4I} & -h_{1R} & h_{1I} & h_{2R} & -h_{2I} \\
-h_{3I} & -h_{3R} & h_{4I} & h_{4R} & h_{1I} & h_{1R} & -h_{2I} & -h_{2R} \\
h_{4R} & -h_{4I} & h_{3R} & -h_{3I} & -h_{2R} & h_{2I} & -h_{1R} & h_{1I} \\
-h_{4I} & -h_{4R} & -h_{3I} & -h_{3R} & h_{2I} & h_{2R} & h_{1I} & h_{1R}
\end{bmatrix}^T \quad (29)$$

We note that rows of $\mathbf{U}(h_1, h_2, h_3, h_4)$ are orthogonal for any values of h_1, \dots, h_4 . From this, it also follows that \mathbf{n}' is white (and Gaussian with unit variance).

The problem with using a non-rate 1 orthogonal design now becomes clear. Unlike $\mathbf{U}(h_1, h_2)$ (see (6)) which is square, $\mathbf{U}(h_1, h_2, h_3, h_4)$ on the other hand is non-square and hence is non-invertible.

We overcome this obstacle by passing the same observation vector \mathbf{s} via a “dithered” version of \mathbf{U} , such that another set of 8 mutually orthogonal measurement rows is attained. We will refer to this approach as “dithered OSTBC”.

Specifically, let us define a 4 dimensional vector $\mathbf{d} = (d_1, d_2, d_3, d_4)$ where d_i are complex numbers of unit magnitude (pure phases). We form a dithered version of the antenna outputs as

$$\tilde{s}_i(t) = d_i \cdot s_i(t), \quad (30)$$

where d_i does not depend on t . We assume that the d_i are drawn at random as i.i.d. uniform phases.

We may associate with $\tilde{s}_i(t)$, $t = 1, \dots, T = 8$, the effective 64-dimensional real vector $\tilde{\mathbf{s}}$. Next, we obtain another 8-dimensional real vector $\tilde{\mathbf{u}}$ by applying to the vector $\tilde{\mathbf{s}}$ the transformation

$$\tilde{\mathbf{y}} = \mathbf{U}\tilde{\mathbf{s}} \quad (31)$$

We therefore obtain

$$\tilde{\mathbf{y}} = \frac{\|\mathbf{h}\|}{2} \mathbf{U}(d_1 h_1, d_2 h_2, d_3 h_3, d_4 h_4) \mathbf{x} + \mathbf{n}'' \quad (32)$$

where \mathbf{n}'' is distributed as \mathbf{n}' (and of course is a function of it).

Note that the dithers drawn implicitly (via (30) and (31)) define a “dithered” combining matrix \mathbf{P}_{Dit} . Combining (28) and (32), we have

$$\underbrace{\begin{bmatrix} \mathbf{y} \\ \tilde{\mathbf{y}} \end{bmatrix}}_{\mathbf{y}_{\text{eff}}} = \underbrace{\frac{\|\mathbf{h}\|}{2} \begin{bmatrix} \mathbf{U}(h_1, h_2, h_3, h_4) \\ \mathbf{U}(d_1 h_1, d_2 h_2, d_3 h_3, d_4 h_4) \end{bmatrix}}_{\mathcal{F}} \mathbf{x} + \begin{bmatrix} \mathbf{n}' \\ \mathbf{n}'' \end{bmatrix} \quad (33)$$

Finally, we apply component-wise quantization to obtain

$$\mathbf{y}_q = Q(\mathbf{y}_{\text{eff}}). \quad (34)$$

We may then recover an estimate of \mathbf{x} by applying the inverse of \mathcal{F} to \mathbf{y} or a linear MMSE estimator.

As mentioned above, another approach is to borrow ideas from quasi-orthogonal space-time codes. As an example for a quasi-orthogonal space-time linear combining matrix, we construct a matrix \mathbf{P}_{Quas} by taking half of the columns of \mathbf{P} (as given in (36)), specifically columns 1 – 16 and 49 – 64, scaling by $\sqrt{2}$ to maintain orthonormality. Using \mathbf{P}_{Quas} induces the following square quasi-orthogonal effective channel which provides similar performance to the “dithered OSTBC” construction described above.

$$\|\mathbf{h}\| \cdot \mathbf{U}_{\text{Quasi}}(h_1, h_2, h_3, h_4) = \begin{bmatrix}
h_{1R} & h_{1I} & h_{2R} & h_{2I} & h_{3R} & h_{3I} & h_{4R} & h_{4I} \\
-h_{1I} & h_{1R} & -h_{2I} & h_{2R} & -h_{3I} & h_{3R} & -h_{4I} & h_{4R} \\
h_{2R} & h_{2I} & -h_{1R} & -h_{1I} & h_{4R} & h_{4I} & -h_{3R} & -h_{3I} \\
-h_{2I} & h_{2R} & h_{1I} & -h_{1R} & -h_{4I} & h_{4R} & h_{3I} & -h_{3R} \\
h_{3R} & -h_{3I} & -h_{4R} & h_{4I} & -h_{1R} & h_{1I} & h_{2R} & -h_{2I} \\
-h_{3I} & -h_{3R} & h_{4I} & h_{4R} & h_{1I} & h_{1R} & -h_{2I} & -h_{2R} \\
h_{4R} & -h_{4I} & h_{3R} & -h_{3I} & -h_{2R} & h_{2I} & -h_{1R} & h_{1I} \\
-h_{4I} & -h_{4R} & -h_{3I} & -h_{3R} & h_{2I} & h_{2R} & h_{1I} & h_{1R}
\end{bmatrix}^T \quad (35)$$

We tested the performance attained with both combining matrices in the scenario considered in Section III. Specifically, Figure 11 depicts the performance achieved for uncoded QPSK transmission when using different linear-combining schemes, for the case of four receive antennas. As both variants of space-time diversity combining do not achieve orthogonality, the gap from optimal combining (MRC) is larger. Optimal antenna selection has a gap of 3.5 dB from MRC. Yet as the number of antennas increases, the complexity of performing optimal selection increases as well.

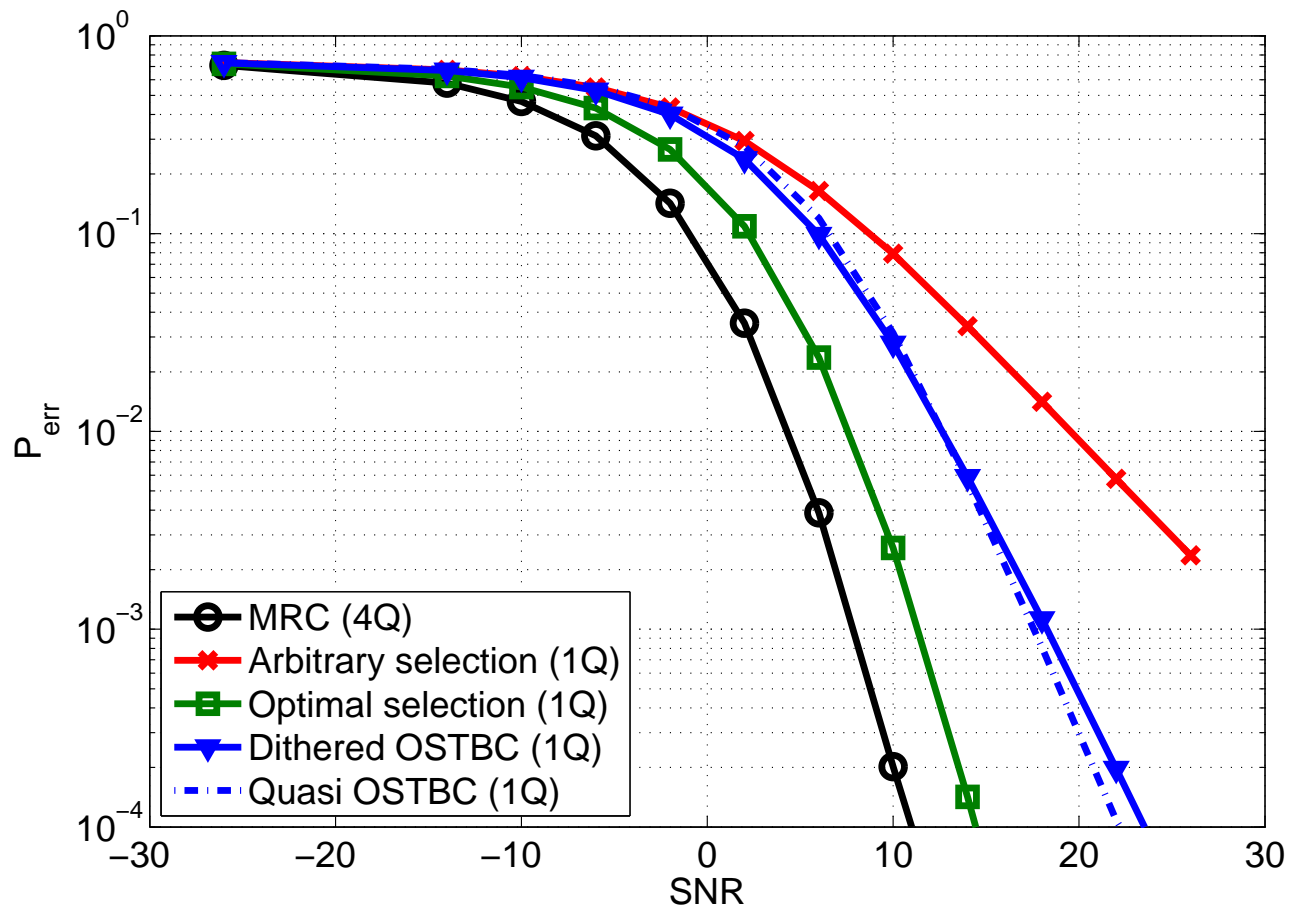


Fig. 11. Bit error rate for uncoded QPSK transmission with reception employing the proposed space-time diversity scheme, using P_{Dit} and P_{Quas} , and comparison with alternatives for the case of four receive antennas.

REFERENCES

- [1] D. G. Brennan, "Linear diversity combining techniques," *Proceedings of the IRE*, vol. 47, no. 6, pp. 1075–1102, 1959.
- [2] S. Sanayei and A. Nosratinia, "Antenna selection in MIMO systems," *IEEE Communications Magazine*, vol. 42, no. 10, pp. 68–73, 2004.
- [3] A. F. Molisch and M. Z. Win, "MIMO systems with antenna selection," *IEEE microwave magazine*, vol. 5, no. 1, pp. 46–56, 2004.
- [4] S. M. Alamouti, "A simple transmit diversity technique for wireless communications," *IEEE Journal on selected areas in communications*, vol. 16, no. 8, pp. 1451–1458, 1998.
- [5] N. Mangalvedhe, R. Ratasuk, and A. Ghosh, "NB-IoT deployment study for low power wide area cellular IoT," in *Personal, Indoor, and Mobile Radio Communications (PIMRC), 2016 IEEE 27th Annual International Symposium on*. IEEE, 2016, pp. 1–6.
- [6] Y.-P. E. Wang, X. Lin, A. Adhikary, A. Grovlen, Y. Sui, Y. Blankenship, J. Bergman, and H. S. Razaghi, "A primer on 3GPP narrowband internet of things," *IEEE Communications Magazine*, vol. 55, no. 3, pp. 117–123, 2017.
- [7] M. R. Palattella, M. Dohler, A. Grieco, G. Rizzo, J. Torsner, T. Engel, and L. Ladid, "Internet of things in the 5G era: Enablers, architecture, and business models," *IEEE Journal on Selected Areas in Communications*, vol. 34, no. 3, pp. 510–527, 2016.
- [8] P. Rost, C. J. Bernardos, A. De Domenico, M. Di Girolamo, M. Lalam, A. Maeder, D. Sabella, and D. Wübben, "Cloud technologies for flexible 5G radio access networks," *IEEE Communications Magazine*, vol. 52, no. 5, pp. 68–76, 2014.
- [9] D. Wübben, P. Rost, J. S. Bartelt, M. Lalam, V. Savin, M. Gorgoglione, A. Dekorsy, and G. Fettweis, "Benefits and impact of cloud computing on 5G signal processing: Flexible centralization through cloud-ran," *IEEE signal processing magazine*, vol. 31, no. 6, pp. 35–44, 2014.
- [10] F. Boccardi, R. W. Heath, A. Lozano, T. L. Marzetta, and P. Popovski, "Five disruptive technology directions for 5G," *IEEE Communications Magazine*, vol. 52, no. 2, pp. 74–80, 2014.
- [11] S.-H. Park, O. Simeone, O. Sahin, and S. S. Shitz, "Fronthaul compression for cloud radio access networks: Signal processing advances inspired by network information theory," *IEEE Signal Processing Magazine*, vol. 31, no. 6, pp. 69–79, 2014.

- [12] O. Tirkkonen, A. Boariu, and A. Hottinen, "Minimal non-orthogonality rate 1 space-time block code for 3+ tx antennas," in *Spread Spectrum Techniques and Applications, 2000 IEEE Sixth International Symposium on*, vol. 2. IEEE, 2000, pp. 429–432.
- [13] H. Jafarkhani, "A quasi-orthogonal space-time block code," *IEEE Transactions on Communications*, vol. 49, no. 1, pp. 1–4, 2001.
- [14] N. Sharma and C. B. Papadias, "Improved quasi-orthogonal codes through constellation rotation," *IEEE Transactions on Communications*, vol. 51, no. 3, pp. 332–335, 2003.
- [15] S. S. Adams, J. Davis, N. Karst, M. K. Murugan, B. Lee, M. Crawford, and C. Greeley, "Novel classes of minimal delay and low PAPR rate 1/2 complex orthogonal designs," *IEEE Transactions on Information Theory*, vol. 57, no. 4, pp. 2254–2262, 2011.
- [16] V. Tarokh, H. Jafarkhani, and A. R. Calderbank, "Space-time block codes from orthogonal designs," *IEEE Transactions on Information theory*, vol. 45, no. 5, pp. 1456–1467, 1999.

