

RESEARCH ARTICLE

Analogue joint source-channel coding for multiple input multiple output–orthogonal frequency division multiplexing systems

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ABSTRACT

Analogue joint source-channel coding (JSCC) has been shown to approach the optimal distortion-cost trade-off when transmitting over additive white Gaussian noise channels. In this work, we consider analogue JSCC over frequency-selective channels using orthogonal frequency division multiplexing (OFDM) modulation and multiple antennas at transmission and/or reception, that is, using a multiple input multiple output (MIMO)-OFDM system. Because of its high complexity, optimal Minimum Mean Square Error (MMSE) analogue JSCC decoding is infeasible in MIMO-OFDM systems and a practical two-stage decoding approach made up of an MMSE estimator followed by a maximum likelihood decoder is proposed instead. Three different alternatives for system optimisation are considered: non-adaptive coding, adaptive coding and adaptive coding with precoding. We show that the three considered strategies approach the optimal distortion-cost trade-off, but the best performance is obtained with the adaptive coding scheme when precoding is utilised. Copyright © 2014 John Wiley & Sons, Ltd.

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1. INTRODUCTION

Source compression and channel coding are typically performed separately in most digital communication systems. This communication strategy, known as the ‘separation principle’, has been shown to be optimum for both lossless [1] and lossy compressions [2] of analogue sources. However, when digital communication systems are designed to perform close to the optimal distortion-cost trade-off, sources have to be compressed using powerful vector quantization and entropy coding methods and data have to be transmitted utilising capacity approaching digital codes that use long block lengths and introduce significant delay and high computational complexity. Moreover, full redesign of the digital system is required whenever a change is required in either the data rate or the distortion target.

Recently, discrete-time analogue communication systems based on the transmission of continuous amplitude channel symbols have been proposed as an alternative to digital communication systems. As shown in [3–5], using appropriate analogue joint source-channel coding

(JSCC) techniques, it is possible to approach the optimal distortion-cost trade-off at high data rates with very low complexity and an almost negligible delay. In addition, analogue JSCC schemes are more robust than standard digital systems to changes in the channel conditions, and they can be continuously adapted to the channel fluctuations on time-varying environments by updating the encoder parameters at the transmitter. These appealing properties make analogue JSCC strategies specially suitable for high-speed transmissions with severe constraints on delay and/or power consumption over fading channels such as real-time communications over wireless channels or sensor networks [6].

Most previous work in analogue JSCC focuses on additive white Gaussian noise (AWGN) channels while analogue JSCC over wireless channels has deserved less attention in the literature. As an example, references [7, 8] consider the use of spatial diversity to improve the performance of analogue JSCC in wireless fading channels. Another example is [9], where the transmission of analogue samples over single-carrier multiple input multiple output (MIMO) fading channels is considered. It is also

worth mentioning reference [10] where the implementation on a software-defined radio testbed of a wireless system based on analogue JSCC is presented. Excellent performance over wireless channels is attained when the encoder parameters are continuously adapted to the time-varying channel signal to noise ratio (CSNR).

In this work, we consider analogue JSCC over frequency-selective channels using orthogonal frequency division multiplexing (OFDM) modulation. When combined with MIMO transmission over multiple transmit and receive antennas, the resulting signalling method is referred to as MIMO-OFDM. This transmission method has been adopted by the last generation of broadband wireless communication systems because of its ability to achieve large spectral efficiencies while enabling low-complexity equalisation of frequency-selective channels.

This work goes several steps further than that presented in [9]. On the one hand, it considers the more realistic case of broadband frequency-selective MIMO channels and extends the design of analogue JSCC narrowband frequency-flat MIMO systems proposed in [9] to also include the transmission of OFDM symbols. Notice that complexity of broadband MIMO-OFDM systems is significantly larger than that of narrowband single-carrier MIMO ones, and this largely impacts on system optimisation. On the other hand, it considers transmission strategies not addressed in [9] and discusses their optimisation. More specifically, three different strategies are considered depending on the level of channel knowledge at the transmitter as follows: non-adaptive coding, adaptive coding and adaptive coding with linear precoding. We show that significant performance gains are obtained when linear precoding is designed according to the MMSE criterion [11–13] and utilised to exploit the channel state information at transmission. Related to system optimisation, special attention is paid to the normalisation operation required at the encoder output. In particular, the utilisation of OFDM symbols allows for using either the time or the frequency dimension to normalise the analogue JSCC symbols.

This paper is organised as follows. Section 2 reviews the basics of analogue JSCC systems. Section 3 focuses on the specifics of analogue JSCC in MIMO-OFDM systems. Section 4 explores the adaptation of the analogue encoder parameters to the channel characteristics in order to improve system performance. Section 5 presents the results of computer experiments, and Section 6 is devoted to the conclusions.

2. ANALOGUE JSCC

Figure 1 shows the block diagram of a discrete-time $N:1$ bandwidth compression analogue JSCC communication system. At the transmitter, N independent and identically distributed (i.i.d.) source symbols are grouped into the source vector $\mathbf{x} = [x_1, x_2, \dots, x_N]^T$ and compressed into one channel symbol s . The superindex T denotes

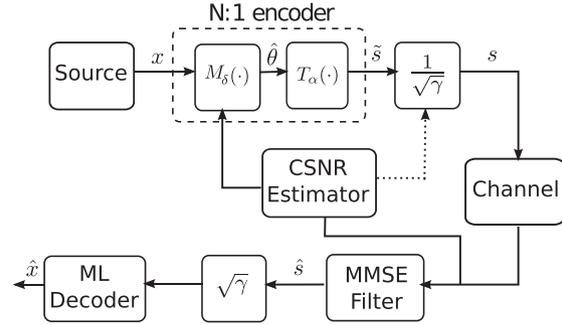


Figure 1. Block diagram of an analogue joint source-channel coding system. CSNR, channel signal to noise ratio; ML, maximum likelihood; MMSE, minimum mean square error.

transposition. The analogue encoding consists of three steps as follows: compression, $M_\delta(\cdot)$; non-linear transformation, $T_\alpha(\cdot)$; and normalisation, $1/\sqrt{\gamma}$.

Recent work on analogue JSCC [3–5, 14] proposes the use of Shannon–Kotel’nikov mappings to define the compression functions $M_\delta(\cdot)$ that map the N -dimensional source vector \mathbf{x} into a single value $\hat{\theta}$. In the particular case of $N = 2$, Akyol *et al.* [15] have shown that the optimal 2:1 mapping quite resembles a spiral-like space-filling curve. For that reason, we specifically consider the Archimedes’ spiral, whose mathematical expression is given by

$$\mathbf{z}_\delta(\theta) = \left[\text{sign}(\theta) \frac{\delta}{\pi} \theta \sin \theta, \frac{\delta}{\pi} \theta \cos \theta \right]^T \quad (1)$$

where δ is an encoder parameter that determines the distance between the two neighbouring spiral arms and θ is the angle from the origin to the point $\mathbf{z} = [z_1, z_2]^T$ on the curve. Notice that the parameter δ determines how the two-dimensional space is filled up and, therefore, how protected the source symbols are against the noise. Hence, the value of δ must be properly selected according to the level of the channel noise or, equivalently, according to the CSNR.

Given a specific spiral defined by its δ value, the compression function $M_\delta(\cdot)$ calculates the value $\hat{\theta}$ corresponding to the point on the spiral that minimises the distance to \mathbf{x} , that is,

$$\hat{\theta} = M_\delta(\mathbf{x}) = \underset{\theta}{\text{argmin}} \|\mathbf{x} - \mathbf{z}_\delta(\theta)\|^2 \quad (2)$$

After the analogue mapping, a non-linear invertible function $T_\alpha(\hat{\theta}) = \text{sign}(\hat{\theta})|\hat{\theta}|^\alpha$, referred to as the stretching function in [5], is used to produce the encoded symbols \tilde{s} . Although most references in the literature [3–5] use $\alpha = 2$, analogue JSCC performance can be significantly improved if α is optimised together with δ [16].

Finally, the encoded symbol \tilde{s} is normalised by $\sqrt{\gamma}$ to ensure the average transmitted power is equal to one.

Table I. Optimal values for δ and γ for 2:1 analogue joint source-channel coding and Gaussian sources.

CSNR (dB)	0	1	2	3	4	5	6	7	8	9	10	11	12
δ	9.8	8.0	5.6	5.0	4.2	4.0	3.9	3.7	3.6	3.4	3.2	3.1	3.0
γ	0.1	0.11	0.12	0.14	0.16	0.2	0.6	1.2	1.9	2.5	3.0	3.3	3.5
CSNR (dB)	13	14	15	16	17	18	19	20	21	22	23	24	25
δ	2.9	2.7	2.5	2.3	2.2	2.1	2.0	1.8	1.7	1.5	1.4	1.3	1.2
γ	3.9	4.7	5.7	7.1	7.9	8.9	10.0	13.1	15.1	20.5	24.5	29.7	35.3
CSNR (dB)	26	27	28	29	30	31	32	33	34	35	36	37	38
δ	1.1	1.0	0.9	0.8	0.8	0.8	0.7	0.7	0.6	0.6	0.5	0.5	0.4
γ	45	56	74	99	99	99	140	140	209	209	333	333	588

CSNR, channel signal to noise ratio.

Hence, the symbol sent over the channel is given by

$$s = \frac{T_\alpha(M_\delta(\mathbf{x}))}{\sqrt{\gamma}} = \frac{\tilde{s}}{\sqrt{\gamma}} \quad (3)$$

When transmitting over an AWGN channel, the received symbols are $y = s + n$ where $n \sim \mathcal{N}(0, N_0/2)$ is a real-valued zero-mean Gaussian random variable that represents the channel noise.

At reception, the analogue source symbols are decoded from the observation y . In [17], we proposed a two-stage decoding method that first calculates an estimate of the transmitted channel symbols, \hat{s} , and then performs maximum likelihood (ML) decoding using this symbol estimate. The advantage of such approach is that performance approximates the optimal distortion-cost trade-off in the whole CSNR region while keeping complexity and delay at a minimum [17]. Recently, we have successfully applied this receiver structure to the case of MIMO transmission systems, where the use of the optimal MMSE decoding is clearly infeasible [9, 17]. In Section 3, we will also corroborate that this approach also attains excellent performance in MIMO-OFDM systems.

Analogue JSCC can also be used to satisfactorily transmit analogue symbols over fading channels. Let us assume a block of T channel symbols $s_t, t = 1, \dots, T$ is transmitted over a single input single output (SISO) flat fading channel. The block of T received symbols is given by

$$y_t = h_T s_t + n_t, \quad t = 1, \dots, T \quad (4)$$

where h_T represents the complex-valued channel response, and n_t is a block of T zero-mean complex-valued i.i.d. Gaussian random variables that represents the channel noise. We assume a block fading channel whose response remains unchanged during the block transmission. Subindex t denotes discrete-time while subindex T indicates that the channel response changes from one block to another.

In fading channels, the CSNR changes with each channel realisation and is given by $\eta_T = |h_T|^2/N_0$. For the considered analogue JSCC system to approach the optimal distortion-cost trade-off, the encoder parameters δ and

α have to be conveniently optimised. As shown in [17], using $\alpha = 1.3$ provides a good overall performance for a 2:1 compression and a wide range of CSNR and δ values. However, the encoder parameter δ has to be specifically adapted to the CSNR value the channel symbols will encounter. We have empirically determined, via computer simulations, that using $\alpha = 1.3$ and the two-stage receiver in [17] provides a good overall performance for the case of 2:1 compression in AWGN channels and a wide range of CSNR and δ values. Because choosing $\alpha \neq 2$ makes the analytical optimisation of the other encoder parameter, δ , very difficult [4], we obtained the optimum δ values for different CSNRs using off-line computer simulations.

The second row of Table I shows the optimum δ values for the range of CSNR values 0, 1, 2, \dots , 38 dB and a normalised Gaussian source. In a practical setting, the CSNR can be estimated at the receiver and sent to the transmitter via a limited feedback channel, as shown in Figure 1. We should also assume that Table I is stored at the transmitter and the receiver for both terminals to determine the correct δ to be used in each transmitted frame.

Another important issue to have in mind is that the normalisation factor, γ , corresponds to the mean square value of the symbols at the output of the analogue encoder, that is, $\gamma = \mathbb{E}[\tilde{s}^2]$. Because of the non-linear nature of the analogue JSCC mappings, the statistical description of \tilde{s} is rather difficult to characterise and strongly depends on the source distribution and the encoder parameter δ . One way to overcome this limitation is to determine off-line, via computer simulations, the mean square value of \tilde{s} for a given source distribution and a set of δ values. The third row of Table I shows the values of γ obtained for the δ values resulting from the encoding optimisation for a given CSNR. Note that in fading channels γ has to be continuously adapted, together with δ , according to the CSNR of each specific channel realisation. If we assume that the third row of Table I is also stored in the transmitter and the receiver, adaptation of γ is readily done from the CSNR information provided by the feedback channel.

An alternative normalisation approach is to determine γ from a channel encoder output block of symbols $\tilde{s}_t, t = 1, \dots, T$. The normalisation factor can thus be obtained as

$$\gamma_T = \frac{1}{T} \sum_{t=1}^T |\tilde{s}_t|^2 \quad (5)$$

We will refer to this normalisation approach as *deterministic* while the previous one will be termed *statistical*. According to ergodicity, both approaches are related by $\lim_{T \rightarrow \infty} \gamma_T = \gamma$. In practice, however, we have determined via computer simulations that both normalisations provide indistinguishable results for $T > 5$ in both AWGN and Rayleigh fading channels.

The advantage of the deterministic normalisation is that it can be calculated online independently of δ . This means that no table of γ values needs to be stored at the transmitter and the receiver and that the normalisation factor can be adapted at a rate different from δ . The drawback is that the receiver needs to know γ_T . However, γ_T is an analogue sample that could be sent together with the source symbols. Appending γ_T to the transmitted symbols will cause an overhead of $1/T$ that vanishes as the block size T increases.

3. ANALOGUE JSCC IN MIMO-OFDM

Figure 2 shows the block diagram of a MIMO-OFDM system that employs analogue JSCC. Discrete-time continuous-amplitude symbols are transmitted over a frequency-selective MIMO channel with n_T transmit antennas and n_R receive antennas using an OFDM modulation with K subcarriers.

We extend the analogue JSCC design proposed for single-carrier MIMO systems in [9] to the case of multicarrier transmissions over MIMO-OFDM. Thus, source symbols are first spatially multiplexed over the n_T transmit antennas. At each transmit antenna, a set of KN analogue source symbols is encoded into K channel symbols using the $N:1$ analogue encoding method explained in Section 2. The real-valued symbols at the encoder output are then transformed into complex-valued channel symbols with a complex interleaver. Let $\tilde{S}_{i,k}$, $i = 1, \dots, n_T$, $k = 1, \dots, K$,

denote the unnormalised encoded symbols transmitted over antenna i and subcarrier k . The corresponding normalised symbols will be represented by $S_{i,k} = \tilde{S}_{i,k}/\sqrt{\gamma_{i,k}}$. In a general setting, both the encoder parameter $\delta_{i,k}$ and the normalisation factor $\gamma_{i,k}$ may be different at each transmit antenna i and/or subcarrier k . Blocks of K normalised channel symbols are put together to be transmitted as OFDM symbols. Let $\mathcal{S}_i = [S_{i,1}, \dots, S_{i,K}]^T$, $i = 1, \dots, n_T$ be the channel symbols that constitute the OFDM symbol transmitted over antenna i . We also define the vector $\mathbf{S}_k = [S_{1,k}, \dots, S_{n_T,k}]^T$, $k = 1, \dots, K$ to represent the MIMO symbols transmitted over subcarrier k .

We assume a block-fading channel that remains unchanged during the transmission of one OFDM symbol. In the time-domain, the block-fading MIMO channel is represented by the sequence of $n_R \times n_T$ matrices $\mathbf{H}[l]$ for $l = 0, \dots, L - 1$, where L is the length of the channel impulse response. In a Rayleigh fading MIMO channel, the entries to $\mathbf{H}[l]$ are complex-valued zero-mean circularly-symmetric Gaussian random variables. In the frequency-domain, the MIMO channel response matrices can be expressed as [18]

$$\mathbf{H}_k = \sum_{l=0}^{L-1} \mathbf{R}[l]^{1/2} \mathbf{H}[l] \mathbf{T}[l]^{1/2} \exp\left(\frac{-j2\pi lk}{K}\right) \quad (6)$$

where \mathbf{H}_k is the frequency-domain $n_R \times n_T$ MIMO channel matrix response corresponding to the k -th subcarrier, $k = 1, \dots, K$. Notice that $\mathbf{H}[l]$ entries are i.i.d. random variables, whereas $\mathbf{R}[l]$ and $\mathbf{T}[l]$ represent the receive and transmit spatial-correlation matrices, respectively.

In order to eliminate the channel intersymbol interference, an inverse fast Fourier transform (IFFT) transformation is applied to the vector of channel symbols \mathcal{S}_i , $i = 1, \dots, n_T$ and a cyclic prefix larger than the channel impulse response is appended at the beginning. These two stages produce the discrete-time representation of the OFDM symbols to be transmitted over the MIMO channel. At reception, the inverse operations of

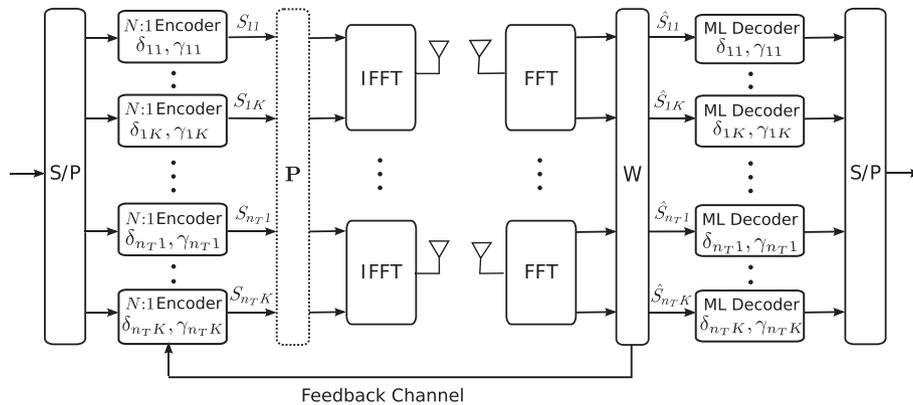


Figure 2. Block diagram of an analogue joint source-channel coding multiple input multiple output-orthogonal frequency division multiplexing system. IFFT, inverse fast Fourier transform; FFT, fast Fourier transform.

fast Fourier transform (FFT) transformation and cyclic prefix removal are applied. Elaborating the signal model, the received observations \mathbf{Y}_k at subcarrier k are given by

$$\mathbf{Y}_k = \mathbf{H}_k \mathbf{S}_k + \mathbf{N}_k, \quad k = 1, \dots, K \quad (7)$$

where \mathbf{N}_k is an i.i.d. circularly symmetric complex Gaussian random vector that represents the additive spatially and temporally white channel noise.

In analogue JSCC, MMSE estimation of the source symbols is the optimal decoding strategy. When considering a MIMO-OFDM system, optimal decoding consists in the calculation, at each subcarrier k , of the MMSE estimate of the Nn_T transmitted source symbols $\mathbf{x}_k = [x_{1,1}, \dots, x_{N,1}, \dots, x_{1,n_T}, \dots, x_{N,n_T}]^T$ from the received symbol vector \mathbf{Y}_k , that is,

$$\begin{aligned} \hat{\mathbf{x}}_{k,\text{MMSE}} &= \mathbb{E}[\mathbf{x}_k | \mathbf{Y}_k] = \int \mathbf{x}_k p(\mathbf{x}_k | \mathbf{Y}_k) d\mathbf{x}_k \\ &= \frac{1}{p(\mathbf{Y}_k)} \int \mathbf{x}_k p(\mathbf{Y}_k | \mathbf{x}_k) p(\mathbf{x}_k) d\mathbf{x}_k \end{aligned} \quad (8)$$

where $\mathbb{E}[\cdot]$ denotes the expectation operator. Because of the conditional probability, $p(\mathbf{Y}_k | \mathbf{x}_k)$, involves the mapping function $M_\delta(\cdot)$, which is discontinuous and highly non-linear, the integral in (8) can only be calculated numerically. This implies that the discretization of the set of all possible source values, \mathbf{x}_k , is needed. If Q discrete-points are selected per source dimension, we would have to calculate Q^{Nn_T} values for $p(\mathbf{Y}_k | \mathbf{x}_k)$ and $p(\mathbf{x}_k)$ and then compute the integral in (8). This is infeasible in MIMO-OFDM even for a small number of transmit antennas and subcarriers.

Alternatively, the two-stage receiver proposed in [17] for analogue JSCC decoding in SISO channels can also be applied to MIMO-OFDM channels. This receiving strategy consists of a first stage where the received symbols are filtered with the aim of minimising the MSE between the encoder output and the decoder input, and a second step where ML decoding is applied to the filtered symbols to obtain an estimate of the transmitted source symbols.

Assuming \mathbf{H}_k is perfectly known at the receiver, the linear filter \mathbf{W}_k that minimises the MSE between the channel symbol vector \mathbf{S}_k and the estimated symbol vector $\hat{\mathbf{S}}_k = \mathbf{W}_k \mathbf{Y}_k$ is given by

$$\mathbf{W}_k = (\mathbf{H}_k^H \mathbf{H}_k + n_T N_0 \mathbf{I}_{n_T})^{-1} \mathbf{H}_k^H \quad (9)$$

where the superindex H represents conjugate transposition. Then, the set of estimated symbols $\hat{\mathbf{S}}_k = [\hat{S}_{1,k}, \dots, \hat{S}_{n_T,k}]^T$ can be denormalised, transformed into the corresponding real-valued symbols and finally input to a bank of ML decoders to calculate an estimate $\hat{\mathbf{x}}_{i,k}, i = 1, \dots, n_T$, of the source symbols transmitted over antenna i and subcarrier k , that is,

$$\hat{\mathbf{x}}_{i,k} = \mathbf{z}_\delta \left(\tilde{\theta}_{i,k} \right) \quad (10)$$

where

$$\tilde{\theta}_{i,k} = T_\alpha^{-1} \left(\sqrt{\gamma} \hat{S}_{i,k} \right) = \text{sign} \left(\hat{S}_{i,k} \right) \left| \sqrt{\gamma} \hat{S}_{i,k} \right|^{-\alpha} \quad (11)$$

4. ADAPTIVE ANALOGUE JSCC

As explained in Section 2, optimal encoders have to be used for the analogue JSCC system to approximate the optimal distortion-cost trade-off. In the case of MIMO-OFDM systems, the optimisation procedure is specially important, because symbols transmitted over different antenna i and subcarrier k experience in general a different CSNR, $\eta_{i,k}$. This means that different encoder parameters $\delta_{i,k}$ (and correspondingly, different normalisation factors, $\gamma_{i,k}$) should be used when encoding the symbols transmitted over antenna i and subcarrier k .

If no information about the channel is available at the transmitter, the same δ value should be used to encode all the analogue source symbols. In this case, it is sensible to use the δ value that corresponds to the average expected CSNR. This fixed approach will perform adequately in frequency-flat and quasi-static channels where the CSNR remains approximately the same at all antennas and all subcarriers during the transmission of several OFDM symbols. However, it can lead to serious performance degradation in practical MIMO-OFDM channels where each subcarrier is expected to have a different time-varying CSNR.

Better performance is obtained when following an adaptive coding strategy where the optimal $\delta_{i,k}$ values are used according to the instantaneous CSNR at each subcarrier and transmit antenna, $\eta_{i,k}$. In a practical setting, this implies that the system be equipped with a feedback channel that regularly sends the $\eta_{i,k}$ values to the transmitter.

In order to appropriately calculate $\eta_{i,k}$, the detector has to be taken into account. When MMSE detection is considered, the filter \mathbf{W}_k does not completely cancel the spatial interference of the MIMO channel. If we consider the residual spatial interference as noise that adds to the thermal noise, it can be shown that the CSNR at the detector output corresponding to the symbols transmitted over the i -th antenna and the k -th subcarrier can be expressed as [19]

$$\eta_{i,k} = \frac{\mu_{i,k}^2}{\mu_{i,k} - \mu_{i,k}^2} = \frac{\mu_{i,k}}{1 - \mu_{i,k}} \quad (12)$$

where $\mu_{i,k} = (\mathbf{W}_k \mathbf{H}_k)_{ii}$ is the i -th diagonal entry of the equivalent MIMO channel $\mathbf{W}_k \mathbf{H}_k$.

In summary, the MMSE detector transforms a MIMO-OFDM channel into a set of n_T SISO-OFDM parallel channels, each one with an equivalent CSNR per subcarrier given by (12). At the transmitter, we encode the symbols to be transmitted through each antenna and subcarrier using the appropriate $\delta_{i,k}$ parameter. This parameter is selected from the fed-back $\eta_{i,k}$ values and using Table I, which is assumed to be stored at the transmitter and the receiver.

4.1. Analogue JSSC MIMO-OFDM with linear precoding.

If the feedback channel is able to provide the transmitter with the MIMO channel matrices, \mathbf{H}_k , $k = 1, \dots, K$ (and not just the $\eta_{i,k}$ values) further performance improvements can be obtained if channel symbols are precoded prior to their transmission. In digital systems, the water-filling algorithm is the optimal strategy to distribute the transmit power among the streams corresponding to different antennas because it maximises the channel capacity. However, the performance of analogue JSSC systems is commonly measured in terms of the signal distortion. For that reason, following [11], we propose to jointly design a linear precoder and a linear detector suitable for analogue JSSC in MIMO-OFDM according to the MMSE criterion.

Recall the MIMO-OFDM signal model defined by (7). Let us assume the output encoder symbols are linearly precoded with a rectangular $n_T \times n_T$ matrix \mathbf{P}_k per subcarrier. Hence the transmitted symbols are $\mathbf{P}_k \mathbf{S}_k$. As in Section 3, \mathbf{W}_k represents the MIMO linear detector per subcarrier. Hence, the channel symbol estimates obtained at the detector output are given by

$$\hat{\mathbf{s}}_k = \mathbf{W}_k (\mathbf{H}_k \mathbf{P}_k \mathbf{S}_k + \mathbf{n}_k) \quad (13)$$

and the error between the estimated and transmitted symbols per subcarrier is

$$\mathbf{e}_k = \mathbf{S}_k - \hat{\mathbf{s}}_k = \mathbf{S}_k - \mathbf{W}_k (\mathbf{H}_k \mathbf{P}_k \mathbf{S}_k + \mathbf{n}_k) \quad (14)$$

The MMSE linear precoder and detector are obtained after solving the following constrained optimisation problem

$$\arg \min_{\mathbf{P}_k, \mathbf{W}_k} \sum_{k=1}^K \mathbb{E} [\text{tr} (\mathbf{e}_k \mathbf{e}_k^H)] \quad (15)$$

$$\text{subject to } \sum_{k=1}^K \text{tr} (\mathbf{P}_k \mathbf{P}_k^H) \leq P_{\text{tx}} \quad (16)$$

where $\text{tr}(\cdot)$ denotes the trace operator, and P_{tx} is the total power available at the transmitter.

Substituting the error expression (14) in (15), differentiating with respect to \mathbf{P}_k and \mathbf{W}_k , and using the Karush–Kuhn–Tucker conditions, we arrive at the following equations to obtain the optimal \mathbf{P}_k and \mathbf{W}_k matrices:

$$\mathbf{P}_k = (\lambda \mathbf{I}_{n_T} + \mathbf{H}_k^H \mathbf{W}_k^H \mathbf{W}_k \mathbf{H}_k)^{-1} (\mathbf{H}_k^H \mathbf{W}_k^H) \quad (17)$$

$$\mathbf{W}_k = (\mathbf{P}_k^H \mathbf{H}_k^H) (n_T N_0 \mathbf{I}_{n_R} + \mathbf{H}_k \mathbf{P}_k \mathbf{P}_k^H \mathbf{H}_k^H)^{-1} \quad (18)$$

where N_0 is the AWGN noise variance and $\lambda \geq 0$ is the Lagrange multiplier that ensures the total transmit power is equal to P_{tx} . Unfortunately, both equations depend on each other, so we need to iterate between them to calculate the precoder and the detector [11]. The iterative

algorithm starts assuming an initial precoder equal to the identity matrix and, at each iteration, both the precoder and the detector are sequentially updated using Equations (17) and (18). Notice that the Lagrange multiplier λ is a scalar value, that is, recalculated at each iteration using Newton's method to ensure that the transmit power constraint is still satisfied. Although the convergence of this algorithm has not been mathematically analysed, it has been shown to converge after a few iterations in practice for the whole range of CSNRs.

It is important to note that linear precoders change the signal to noise ratio (SNR) of the equivalent channel that corresponds to the symbols transmitted over antenna i and subcarrier k . As a consequence, the encoder parameter $\delta_{i,k}$ should be adapted accordingly. Notice that the CSNR values $\eta_{i,k}$ when linear precoding is used are easy to obtain if we take into account that linear precoders simply transform the MIMO channel \mathbf{H}_k into another one given by $\mathbf{H}_k \mathbf{P}_k$. Hence, the transmitter can calculate the corresponding $\eta_{i,k}$ values using equation (12) and then choosing $\delta_{i,k}$ according to Table I.

4.2. Symbol normalisation in MIMO-OFDM.

An important issue regarding the optimisation of an adaptive analogue JSSC MIMO-OFDM system is the normalisation of the transmitted symbols $S_{i,k}$. In a non-adaptive system, all subcarriers in the OFDM symbol transmitted over the i -th antenna can be normalised using the same factor γ_i , $i = 1, \dots, n_T$ per antenna. However, when considering adaptive coding, source symbols are encoded with different $\delta_{i,k}$ values that are selected according to the CSNR $\eta_{i,k}$ of the equivalent channel corresponding to antenna i and subcarrier k .

As in Section 2, either a statistical or a deterministic approach can be followed to determine the normalisation factors, $\gamma_{i,k}$. In the statistical approach, $\gamma_{i,k}$ should be the mean square value of the unnormalised encoded symbols $\tilde{S}_{i,k}$, that is, $\gamma_{i,k} = \mathbb{E} [|\tilde{S}_{i,k}|^2]$. For a given source statistics and encoder parameter $\delta_{i,k}$, these mean square values can be estimated off-line via computer simulations and stored in a table such as Table I.

Alternatively, the normalisation factors can be obtained following a deterministic approach. Let us assume a block of T OFDM symbols $S_{i,k,t}$, $t = 1, \dots, T$, are transmitted over a block fading channel, that is,

$$\mathbf{Y}_{k,t} = \mathbf{H}_{k,T} \mathbf{S}_{k,t} + \mathbf{N}_{k,t}, \quad k = 1, \dots, K; \quad t = 1, \dots, T \quad (19)$$

where the MIMO channel response per subcarrier $\mathbf{H}_{k,T}$ (and hence the equivalent CSNR $\eta_{i,k}$) remains unchanged. The normalisation factors can thus be obtained as

$$\gamma_{i,k,T} = \frac{1}{T} \sum_{t=1}^T |\tilde{S}_{i,k,t}|^2 \quad (20)$$

These normalisation factors must be known at reception for correct analogue decoding. In a practical setup, they can be sent using specific OFDM symbols. This will imply an overhead of $1/T$ that vanishes as the block size T increases.

Contrarily to the single carrier case, the requirement that the channel remains unchanged during the transmission of a block of T OFDM symbols is more difficult to satisfy in practical situations, specially if the FFT size is large. Hence, although the previous deterministic normalisation performs well for block sizes as small as $T = 5$, it may be unfeasible in many practical scenarios. This limitation can be overcome by formulating a different deterministic normalisation approach where symbols transmitted along subcarriers with similar SNR values are used to estimate their mean square value.

Let us define the set $\mathcal{K}_i = \{1, \dots, K\}$ that contains all subcarriers when transmitting over antenna i . Let us divide the set \mathcal{K}_i into a number of B nonoverlapping subsets $\mathcal{K}_{i,1}, \mathcal{K}_{i,2}, \dots, \mathcal{K}_{i,B}$. The number of elements in these subsets will be represented by $K_{i,1}, K_{i,2}, \dots, K_{i,B}$. Hence, $\mathcal{K}_i = \mathcal{K}_{i,1} \cup \mathcal{K}_{i,2} \cup \dots \cup \mathcal{K}_{i,B}$ and $K_i = K_{i,1} + K_{i,2} + \dots + K_{i,B}$. Recall that in adaptive coding, channel symbols $S_{i,k}$ are obtained with an analogue encoder with a parameter $\delta_{i,k}$, that is, selected according to the CSNR, $\eta_{i,k}$. Hence, symbols transmitted over subcarriers with similar SNR values will have similar statistics and can be used to estimate their corresponding normalisation factor.

Indeed, let us assume that for a given channel realiation and a transmit antenna i , the SNRs across the different subcarriers belong to the interval $[\hat{\eta}_{\min}, \hat{\eta}_{\max}]$, that is, $\hat{\eta}_{\min} \leq \eta_{i,k} \leq \hat{\eta}_{\max}$. Next, we define the subsets $\mathcal{K}_{i,b}, b = 1, \dots, B$ as follows:

$$\begin{aligned} \mathcal{K}_{i,1} &= \{k : \hat{\eta}_{\min} \leq \eta_{i,k} < \hat{\eta}_{i,1}\} \\ \mathcal{K}_{i,2} &= \{k : \hat{\eta}_{i,1} \leq \eta_{i,k} < \hat{\eta}_{i,2}\} \\ &\vdots \\ \mathcal{K}_{i,B} &= \{k : \hat{\eta}_{i,B-1} \leq \eta_{i,k} \leq \hat{\eta}_{\max}\} \end{aligned}$$

that is, each subset $\mathcal{K}_{i,b}, b = 1, \dots, B$ contains those subcarriers with similar SNR for a given channel realisation.

Having in mind the previous subcarrier clustering, we can modify our previous adaptive analogue coding approach so that all symbols to be transmitted over the subcarriers in subset $\mathcal{K}_{i,b}$ are encoded with the same representative parameter $\delta_{i,b}$. This is reasonable because all subcarriers in $\mathcal{K}_{i,b}$ have similar CSNR. Hence, all channel symbols $S_{i,k}$ with $k \in \mathcal{K}_{i,b}$ have the same statistical properties and their mean square value can be estimated as

$$\gamma_{i,b} = \frac{1}{K_{i,b}} \sum_{k \in \mathcal{K}_{i,b}} |S_{i,k}|^2, \quad i = 1, \dots, n_T, \quad b = 1, \dots, B$$

The choice of the optimal partitioning of the CSNR range into the corresponding B blocks is not trivial, because it is necessary to find the set of block limits and representative $\delta_{i,b}$ values that minimise the signal distortion at the receiver. It is not feasible to solve this problem mathematically because the impact of the δ parameter on the overall distortion can be only determined for the case of ML decoding and $\alpha = 2$, as shown [5]. Otherwise, it is rather difficult to characterise the distribution of the coded symbols at the output of the analogue encoder and, hence, it is not possible to establish a closed-form expression that relates the value of δ and the observed distortion. For that reason, an exhaustive search by computer simulations has been carried out to evaluate the impact of using non-optimal values for δ on the overall distortion and to determine the subcarrier subsets and the representative $\delta_{i,b}$ values of these subsets for different number of blocks. Table II shows the subcarrier subsets and their corresponding representative values $\delta_{i,b}$ obtained for $B = 2, 4$ and 8 assuming that the CSNR values lie within an interval between $\hat{\eta}_{\min} = 0$ dB and $\hat{\eta}_{\max} = 40$ dB.

It is important to note that, similar to other deterministic normalisations, the normalisation factors have to be known at reception for the decoders to perform adequately. In a practical setting, the normalisation factors can be sent

Table II. Subcarrier partitioning for deterministic normalisation.

Number of intervals	channel signal to noise ratio range (dB)	$\delta_{i,b}$ values
$B = 2$	[0–20]	2.3
	[21–40]	0.8
$B = 4$	[0–8]	8.0
	[9–19]	2.2
	[20–29]	0.9
	[30–40]	0.5
$B = 8$	[0–7]	8.0
	[8–14]	3.0
	[15–20]	2.0
	[21–25]	1.3
	[26–30]	0.8
	[31–34]	0.7
	[25–37]	0.5
	[38–40]	0.4

over certain subcarriers within the same OFDM symbols that are specifically reserved to this aim. This will be the approach followed in the sequel. The maximum number of transmitted normalisation factors in a limiting situation, where the channel changes at each OFDM symbol is B , and the maximum overhead that normalisation factor transmission implies is B/K . This overhead tends to be negligible as the number of intervals B decreases and/or the FFT size K increases.

In practice, however, the specific number of normalisation factors to be transmitted depends on the channel realisation and is often less than B . Indeed, in standard MIMO-OFDM channels it is often the case in which sub-carrier subsets of a given channel realisation are empty and the normalisation factor corresponding to this subset need not to be transmitted. This issue is further analysed in the ensuing section.

4.3. Complexity analysis

The overall complexity of the analogue JSCC MIMO-OFDM system is determined by the filtering operations, because the complexity and delay introduced by the analogue encoding and ML decoding are practically negligible. On one hand, the analogue encoding consists in mapping the bidimensional point given by two source symbols into the closest point on the spiral and applying the stretching function to the corresponding angle from the origin to that point. Efficient search algorithms can be employed to determine the closest point on the spiral with low complexity. On the other hand, the ML decoding basically inverts the encoding operations and the complexity is hence identical.

The overall complexity depends on the transmission strategy considered in the analogue JSCC MIMO-OFDM system. In the case of non-adaptive and adaptive coding, a linear MMSE detector is used for filtering the received symbols. The complexity of such filtering operation is $\mathcal{O}(n_T^3)$ because it involves several products of matrices and especially the inverse of an $n_T \times n_T$ -sized matrix. Notice that the complexity of both transmission strategies is the same because the main difference of the adaptive coding with respect to the non-adaptive method is the use of information about the estimated SNR to select the optimal encoder parameters at the transmitter.

In the case of the adaptive coding with linear precoding, the optimal linear MMSE filters for each subcarrier are individually calculated by an iterative algorithm that sequentially updates the precoder and the corresponding detector. On one hand, the algorithm complexity at each iteration is $\mathcal{O}(M^3)$, where $M = \max\{n_T, n_R\}$, because the expressions for the transmit and receive filters also involve the inverse of a matrix of size $n_R \times n_R$ and $n_T \times n_T$, respectively. On the other hand, the number of iterations required for the iterative algorithm to converge basically depends on the CSNR. In the high CSNR region, it can be shown that the MSE cost function is almost flat and, thus, a large number of iterations is required to converge whereas for

low CSNRs the algorithm converges after very few iterations. In practice, the computation of the optimal MMSE filters never exceeds 50 iterations. Notice that these filters are obtained once for each channel realisation and must only be recalculated when the channel changes.

5. SIMULATION RESULTS

Computer simulations were carried out to assess the performance of the analogue JSCC MIMO-OFDM systems considered in previous sections. Three different configurations were evaluated: non-adaptive coding, adaptive coding and adaptive coding with linear precoding.

System performance is measured in terms of the signal to distortion ratio (SDR) with respect to the average CSNR. If the MSE between decoded and source analogue symbols, that is,

$$\text{MSE} = \frac{1}{N} \mathbb{E} \left\{ \left\| \mathbf{x} - \hat{\mathbf{x}}^2 \right\|^2 \right\}$$

is selected as the distortion metric, the SDR in dB can be calculated as

$$\text{SDR}_{(dB)} = 10 \log_{10} \left(\frac{\sigma_x^2}{\text{MSE}} \right) \quad (21)$$

where σ_x^2 is the source variance. The optimal distortion-cost trade-off is the maximum attainable SDR for a given CSNR. In the literature, this theoretical limit is known as the optimum performance theoretically attainable (OPTA) and is calculated by equating the rate distortion function to the channel capacity [20]. For $N:1$ compression of Gaussian sources over a generic stochastic $n_R \times n_T$ channel matrix \mathbf{H}_k and assuming that the channel is not known at the transmitter, the OPTA is given by

$$n n_T \log(\text{SDR}) = \mathbb{E}_{\mathbf{H}_k} \left[\log \det \left(\mathbf{I}_{n_R} + \frac{\eta}{n_T} \mathbf{H}_k \mathbf{H}_k^H \right) \right] \quad (22)$$

where $\mathbb{E}_{\mathbf{H}_k}[\cdot]$ represents expectation with respect to \mathbf{H}_k and η is the average CSNR. In a system where the channel is known at the transmitter, capacity is maximised by the water-filling solution at each channel realisation. In our case, however, channel knowledge at transmission is not exploited to maximise capacity but to precode transmitted symbols with the linear MMSE precoding matrices \mathbf{P}_k , $k = 1, \dots, K$, given by (17). Therefore, OPTA is calculated in this case by replacing \mathbf{H}_k with the equivalent channel $\mathbf{H}_k \mathbf{P}_k$ in (22).

Let us start by considering the case $n_T = n_R = 1$, that is, SISO-OFDM. We considered a source of i.i.d. normalised Gaussian random variables and $K = 64$ subcarriers. In a first computer experiment we considered real wireless channels measured in an indoor scenario (an office) by using a hardware testbed which was jointly designed and implemented by two research groups from the University of Cantabria and University of A Coruña in Spain for

the practical evaluation of multiuser multiantenna transmission techniques. This testbed consists of three transmit nodes and three receive nodes each equipped with MIMO capabilities. For a detailed description of the Grupo de Tecnología Electrónica y Comunicaciones (GTEC) MIMO testbed, see the URL of the COMONSENS project [21].

In this first experiment, the real channel measurements were specifically obtained for 64 subcarriers using one single transmit antenna and one single receive antenna, which are at a distance of approximately 9 m with direct line-of-sight. Figure 3 plots the power delay profile of such measured indoor channels. The corresponding RMS delay spread is 6 ns. In addition, it has been observed that the channel frequency response values at each subcarrier approximately follow a Rayleigh distribution although with a different variance at each subcarrier.

Figure 4 plots the obtained results. It can be seen that the three proposed analogue JSCC techniques approach

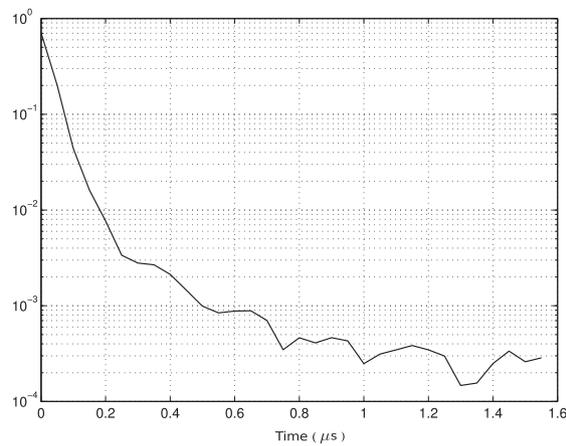


Figure 3. Power delay profile of the measured real indoor channel.

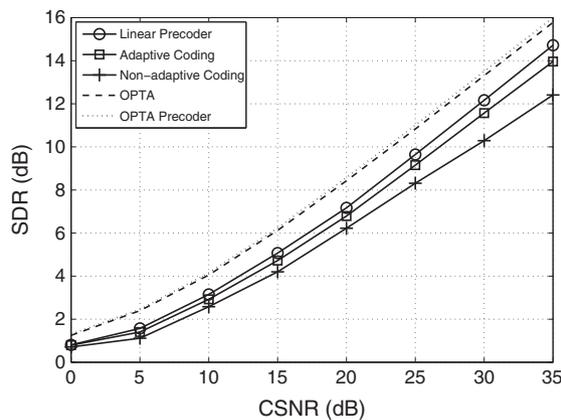


Figure 4. Performance of 2:1 analogue joint source-channel coding single-input single-output orthogonal frequency division multiplexing systems over measured real channels. CSNR, channel signal to noise ratio; SDR, signal to distortion ratio; OPTA, optimum performance theoretically attainable.

the OPTA in the whole SNR region. As expected, the worst performance is obtained when no channel state information is available at the transmitter (i.e. non-adaptive coding and no precoding), specially when the CSNR is high in which case performance is almost 4.0 dB below the OPTA. Performance is significantly improved (about 1.8 dB at high SNR) when adaptive coding is considered and further improvement can be obtained if MMSE linear precoding is utilised. This latter improvement, however, is relatively small because we are in a SISO scenario and not much performance gain should be expected from linear precoding. Stochastic normalisation was used in the adaptive coding approaches, but the same results were obtained with per-subcarrier deterministic normalisation and $B = 8$ blocks.

In a second computer experiment for SISO-OFDM, we considered the International Telecommunication Union (ITU)-Pedestrian B model. For a complete description of the delay and Doppler power profiles of such model see reference [22]. Doppler shift is assumed low enough so that the channel remains unchanged during the transmission of an OFDM symbol and no intercarrier interference arises. In the ITU-Pedestrian B channel model, this condition is met for a reasonable large number of practical situations. Nevertheless, the channel changes from one OFDM symbol to another according to the Doppler power profile of the model. In the adaptive coding approaches, the encoder parameters are adjusted at each OFDM symbol and stochastic normalisation was used. Figure 5 shows the obtained results, which are similar to those presented in Figure 4, although the performance gains when considering adaptive coding and linear precoding are slightly higher.

We now shift our focus to MIMO-OFDM channels. Let us start by considering real 4×4 MIMO-OFDM fading channels measured using the GTEC MIMO testbed

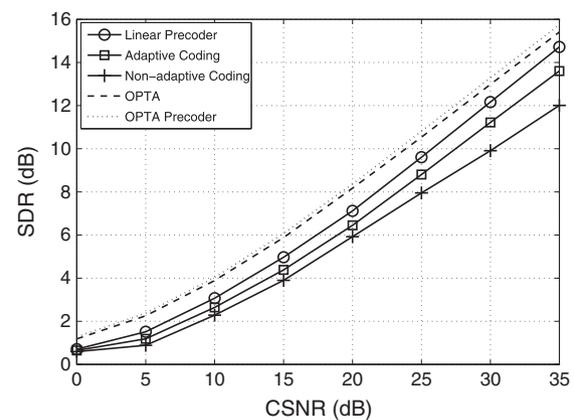


Figure 5. Performance of 2:1 analogue joint source-channel coding single-input single-output orthogonal frequency division multiplexing systems over a Pedestrian B model. CSNR, channel signal to noise ratio; SDR, signal to distortion ratio; OPTA, optimum performance theoretically attainable.

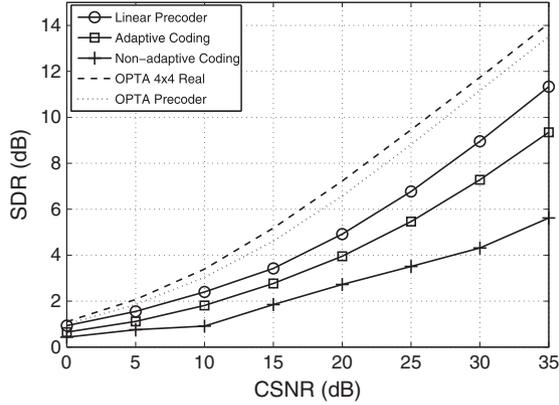


Figure 6. Performance of analogue joint source-channel coding 4×4 multiple-input multiple-output orthogonal frequency division multiplexing systems over measured real channels. CSNR, channel signal to noise ratio; SDR, signal to distortion ratio; OPTA, optimum performance theoretically attainable.

described above in the same indoor scenario and for $K = 64$. Stochastic normalisation was used in the adaptive coding approaches, but the same results were again obtained with per-subcarrier deterministic normalisation and $B = 8$ blocks. Figure 6 plots the obtained results. The non-adaptive coding strategy exhibits the worst performance. It is 8.0 dB below the OPTA at high channel SNR values. A significant improvement in performance is obtained (3.5 dB at high SNR) when considering adaptive coding. Different to the SISO case, performance also improves significantly (2 dB at high SNR) when incorporating linear MMSE precoding. It is interesting to note from Figure 6 that the OPTA with precoding is actually lower than that without precoding. This is because the precoder has been designed to minimise the MSE and not to maximise capacity. Nevertheless, notice that the actual performance of the precoded system is better than that of the system without precoding.

We also carried out computer experiments considering a standard model such as the Intelligent Multi-element Transmit and Receive Antennas case D (IMETRA-D) channel model described in [23]. The IMETRA-D model only specifies the spatial correlation of a MIMO channel, that is, the matrices $\mathbf{R}[l]$ and $\mathbf{T}[l]$ in (6). For the delay and Doppler power profile, we chose the same parameters as in the ITU-pedestrian B model [22]. Figure 7 plots the obtained results which highlight the importance of the adaptive coding strategies. Indeed, without adapting the analogue encoder parameters, system performance is very far away from the OPTA (9.5 dB at high SNR). With adaptive coding the gap to the OPTA reduces significantly (5.0 dB at high SNR) while the distance to the OPTA is even less (less than 3 dB at high SNR) when adaptive coding with linear precoding is utilised.

It is important to note that the deterministic normalisation based on temporal averaging cannot be used when transmitting over this type of time varying channels. In

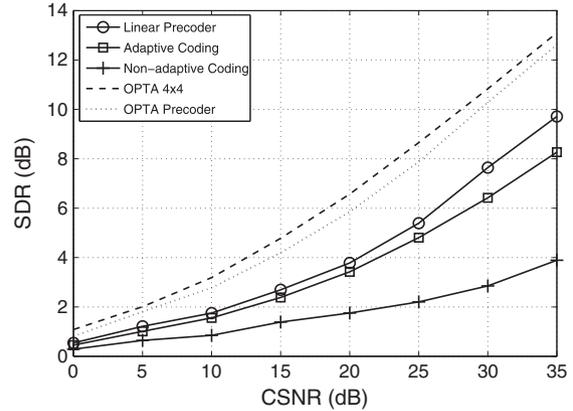


Figure 7. Performance of analogue joint source-channel coding 4×4 multiple-input multiple-output orthogonal frequency division multiplexing systems over an Intelligent Multi-element Transmit and Receive Antennas case D channel. CSNR, channel signal to noise ratio; SDR, signal to distortion ratio; OPTA, optimum performance theoretically attainable.

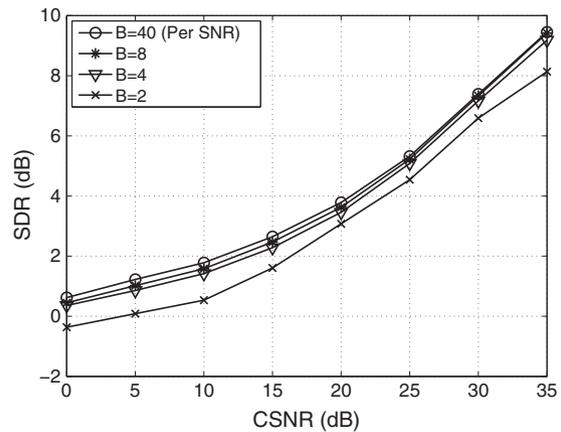


Figure 8. Performance with respect to B of an analogue joint source-channel coding 4×4 multiple-input multiple-output orthogonal frequency division multiplexing system with deterministic normalisation over an Intelligent Multi-element Transmit and Receive Antennas case D channel. CSNR, channel signal to noise ratio; SDR, signal to distortion ratio; SNR, signal to noise ratio.

such cases, the subcarrier clustering deterministic normalisation described in Section 4.2 should be used. Figure 8 plots the performance of an analogue JSCC 4×4 MIMO-OFDM system with adaptive coding and precoding when transmitting over an IMETRA-D channel for different values of B , that is, the number of subsets in which we divided the CSNR range $[\hat{\eta}_{\min}, \hat{\eta}_{\max}]$. Similar behaviour as a function of B was observed for the case of adaptive coding without precoding. The subset limits and the encoder representant values $\delta_{i,b}$ are those in TABLE II. Figure 8 also plots the limiting case $B = 40$ in which there is a subset per CSNR integer value in dB (Table I). The results shown

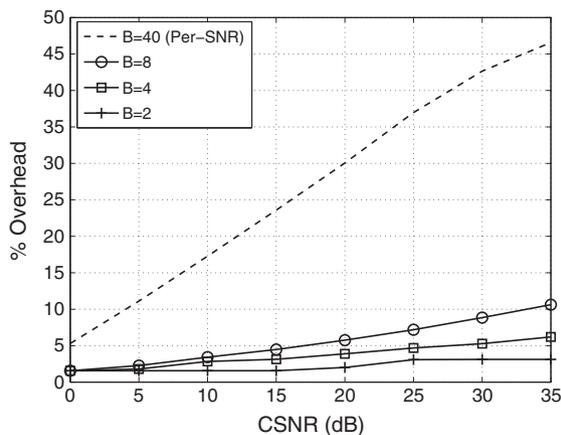


Figure 9. Overhead caused by deterministic normalisation in an analogue joint source-channel coding 4×4 multiple-input multiple-output orthogonal frequency division multiplexing system over an Intelligent Multi-element Transmit and Receive Antennas case D channel. CSNR, channel signal to noise ratio; SDR, signal to distortion ratio; SNR, signal to noise ratio.

in Figure 8 indicate that the SDR decreases very slightly when $B \geq 4$ (less than 0.5 dB) with respect to the per-SNR normalisation. Performance starts degrading significantly when B is less than 4.

As explained before, deterministic normalisation has the inconvenience that the normalisation factors have to be transmitted, hence, increasing the system overhead. Nevertheless, the fact that the number of subcarrier subsets can be significantly reduced without degrading performance enables the system to reduce the number of normalisation factors to be transmitted. Figure 9 illustrates this issue by plotting the overhead percentage for different values of CSNR and B . Notice that the per-SNR normalisation ($B = 40$) demands a significant overhead, which can be as high as 50 per cent at high CSNR values. At lower SNRs, the overhead is less, because the number of empty SNR subsets increases, and it is not necessary to send their normalisation factors. As shown in Figure 9, overhead reduces significantly for $B = 2, 4$ and 8.

In summary, according to Figures 8 and 9, the choice $B = 4$ attains a good trade-off between performance and overhead. The number of normalisation factors to be sent can be greatly reduced, with a negligible impact on performance, from almost 50 per cent to just 6 per cent in the worst case.

6. CONCLUSIONS

We have studied the analogue transmission of discrete-time coded samples using a MIMO-OFDM system. Source symbols are analogue JSCC encoded and sent as OFDM symbols over frequency-selective MIMO fading channels. As an alternative to optimal MMSE decoding, we proposed a more practical two-stage receiver made up of

a MMSE estimator followed by a ML decoder. This approach exhibits a satisfactory performance in the whole SNR region while keeping complexity and delay at a minimum. We studied three different alternatives for system optimisation as follows: non-adaptive coding, adaptive coding and adaptive coding with precoding. Simulation results show that the three analogue JSCC transmission strategies approach the optimal distortion-cost trade-off (i.e. the OPTA), and the best performance is obtained when using the adaptive coding together with precoding. We have also paid special attention to the normalisation of the analogue encoded symbols. A deterministic normalisation approach has been proposed that can be applied to each OFDM symbol individually while significantly reducing the overhead caused by the transmission of the normalisation factors.

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