

Channel Coding for Progressive Images in a 2-D Time-Frequency OFDM Block with Channel Estimation Errors[†]

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Abstract

Coding and diversity are very effective techniques for improving transmission reliability in a mobile wireless environment. The use of diversity is particularly important for multimedia communications over fading channels. In this work, we study the transmission of progressive image bitstreams using channel coding in a 2-D time-frequency resource block in an OFDM network, employing time and frequency diversities simultaneously. In particular, in the frequency domain, based on the order of diversity and the correlation of individual subcarriers, we construct symmetric n -channel FEC-based multiple descriptions using channel erasure codes combined with embedded image coding. In the time domain, a concatenation of RCPC codes and CRC codes is employed to protect individual descriptions. We consider the physical channel conditions arising from various different coherence bandwidths and coherence times, leading to various orders of diversities available in the time and frequency domains. We investigate the effects of different error patterns on the delivered image quality due to various fade rates. We also study the

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tradeoffs and compare the relative effectiveness associated with the use of erasure codes in the frequency domain and convolutional codes in the time domain under different physical environments. Both the effects of inter-carrier interference and channel estimation errors are included in our study. Specifically, the effects of channel estimation errors, frequency selectivity and the rate of the channel variations are taken into consideration for the construction of the 2-D time-frequency block.

Index Terms

Cross-layer design, diversity, frequency diversity, time diversity, multimedia communications, multiple description coding, orthogonal frequency division multiplexing (OFDM), channel coding, progressive transmission, wireless video.

I. INTRODUCTION

In recent years, with the rapid mergers of multimedia, Internet and wireless communications, there is a growing trend of heterogeneity (in terms of channel bandwidths, mobility levels of terminals, end-user quality-of-service (QoS) requirements and so on) for the emerging integrated wired/wireless networks. Embedded/progressive source coding, allowing partial decoding at various resolution and quality levels from a single compressed bitstream, is a promising technology for multimedia communications in heterogeneous environments. However, embedded source coders are usually extremely sensitive to channel impairments which can be severe in mobile wireless links due to multipath signal propagation, delay and Doppler spreads, and other effects. Sometimes a single error can cause an unrecoverable loss in synchronization between encoder and decoder, and produce substantial quality degradation.

Early study of embedded transmission includes [1], [2]. Both papers studied the transmission of a progressively compressed bitstream employing the Set Partitioning in Hierarchical Trees (SPIHT) source coder combined with rate-compatible punctured convolutional (RCPC) codes. Coding and diversity are very effective techniques for improving the transmission reliability in a mobile wireless environment. However, time diversity achieved by channel coding plus intra-packet interleaving in a single carrier (SC) communication system becomes less effective in a slow fading environment where correlated and prolonged deep fades often result in the erasure of the whole packet or even several contiguous packets. Hence, although improvement could still be achieved due to the coding gain associated with the use of RCPC codes, the performance was not satisfactory [2].

To improve the performance against deep fades in a wireless environment, two approaches have been proposed to exploit diversity in the time domain at the physical layer for SC communication systems. One

was to add systematic Reed-Solomon (RS) codes across multiple packets [3]. Specifically, channel codes consisted of a concatenation of RCPC and CRC codes as the row codes and RS codes as the column codes. With the addition of RS codes across multiple packets, lost packets might still be recoverable due to independently faded time slots [3].

Another approach [4]–[7] uses contiguous information symbols from the progressive bitstreams, which, instead of being packed in the same packets [1], [3], are spread across multiple packets (descriptions). The information symbols are protected against channel errors using systematic RS codes with the level of protection depending on the relative importance of the information symbols. This coding scheme is sometimes referred to as symmetric n -channel FEC-based multiple description (MD) coding. Due to the individually decodable nature of the multiple packets, the source can be recoverable despite packet loss, although at a lower fidelity that depends on the number of successfully received packets. Analogous to the physical layer diversity techniques offered by channel coding, this has sometimes been referred to as *application layer diversity* [8].

While both approaches perform well in slow fading environments, the order of diversity of the physical channel is vital to the selection of system parameters (e.g., choice of channel codes and corresponding channel code rates) as shown in [9]. Despite their importance, such factors are usually overlooked in the literature. More importantly, studies of these channel coding techniques have been limited to 1-D time domain coding in a slow fading environment [3], [7]. For fast fading, rapid channel variations due to high mobility can potentially provide a high diversity gain and significantly improve the effectiveness of channel coding in the time domain. Unfortunately, rapid channel variation also poses a significant challenge for channel estimation [10]–[12]. The accuracy of this channel state information (CSI) is particularly important in optimizing channel coding. In particular, it has been shown that imperfect CSI due to estimation errors affects the performance of communications systems designed to take advantage of the diversity opportunities [13]–[15].

In recent years, orthogonal frequency division multiplexing (OFDM) has drawn intense interest. OFDM differentiates itself from an SC communications system in many ways, such as robustness against frequency-selective fading. Frequency diversity by adding redundancy in the frequency domain can combat channel errors due to multipath fading and achieve a more reliable overall system performance. In other words, OFDM offers a unique opportunity to improve system efficiency by employing both time and frequency domain channel coding depending on the propagation environment and user's mobility. A highly scattered environment may make the frequency domain coding more effective. A highly mobile user will probably make time domain coding more compelling. Although there have been some works

investigating transmission of embedded bitstreams over OFDM networks [16]–[18], none has explicitly characterized the time and frequency domains independently and jointly optimized the coding scheme based on these different physical environments.

In this work, we study the transmission of progressively coded image bitstreams using channel coding in a 2-D time-frequency resource block in an OFDM network under different physical environments. By properly decoupling the time domain and frequency domain channel variations, we propose a 2-D channel coding scheme which employs time and frequency diversities simultaneously. In particular, in the frequency domain, based on the order of diversity, we construct FEC-based multiple descriptions using channel erasure codes combined with embedded source coding. In the time domain, concatenated RCPC codes and CRC codes protect individual descriptions. Both the effects of inter-carrier interference (ICI) and channel estimation errors, which may become severe in a fast fading environment, are taken into consideration. We use pilot symbol assisted modulation (PSAM) with pilot symbol density depending on the channel selectivities in both time and frequency. As diversity is the primary factor determining the performance of a wireless system, the results presented can provide some design criteria for other progressive transmission coding schemes over mobile wireless networks.

The remainder of this paper is organized as follows: In Section II, we give a description of the OFDM system and the channel mode. We also describe the proposed transmission system and discuss some of the issues associated with the use of channel coding in a time-frequency block. In Section IV, we describe the optimization problem. In Section V, we provide simulation results and discussion. Finally, in Section VI, we provide a summary and conclusion.

II. CHANNEL MODEL AND TIME-FREQUENCY CHANNEL CODING

The basic principle of OFDM is to split a high-rate data stream into a number of lower rate streams that are transmitted over overlapped but orthogonal subcarriers. Since the symbol duration increases for the lower rate parallel subcarriers, the relative amount of dispersion in time caused by multipath delay spread is decreased. Depending on the propagation environment and the channel characteristics, the resource block in an OFDM system can be used to exploit time and/or frequency diversities through channel coding. For time diversity, channel coding plus interleaving can be used in the time domain. However, for the technique to be effective, the time frame has to be greater than the channel coherence time $(\Delta t)_c$. The maximum time-diversity gain \mathcal{D}_t is given by the ratio between the duration of a time frame and $(\Delta t)_c$.

In addition to time diversity, frequency diversity by adding redundancy across the subcarriers can be

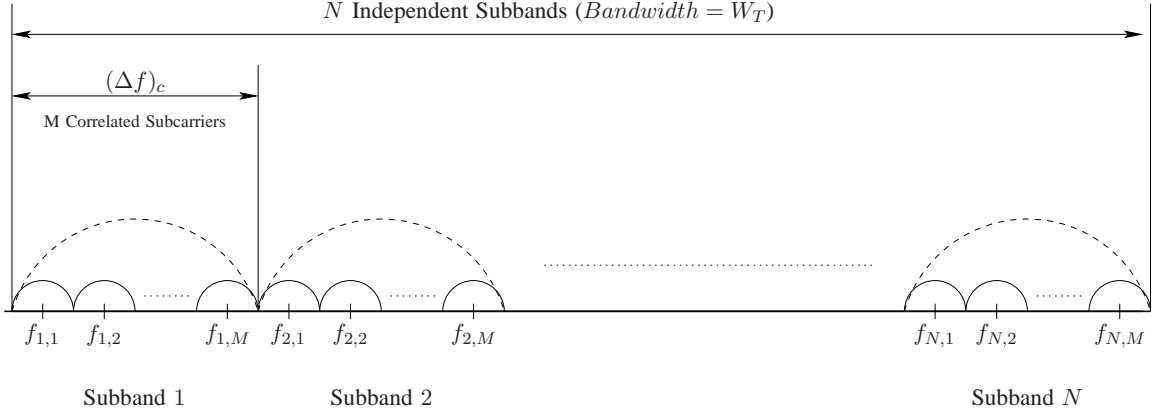


Fig. 1. Subcarrier spectrum assignment.

applied to combat channel errors. Generally, the maximum achievable frequency diversity \mathcal{D}_f is given by the ratio between the overall system bandwidth W_T and the coherence bandwidth $(\Delta f)_c$.

In this work, we consider a frequency-selective environment and use a block fading channel model to simulate the frequency selectivity [19]. In this model, the spectrum is divided into blocks of size $(\Delta f)_c$. Subcarriers in different blocks are considered to fade independently; subcarriers in the same block experience identical fades. As illustrated in Fig. 1, we assume an OFDM system with an overall system bandwidth W_T , such that we can define N independent subbands. Each subband consists of M correlated subcarriers spanning a total bandwidth of $(\Delta f)_c$. The total number of subcarriers in the OFDM system is NM . In the time domain, we assume the channel experiences Rayleigh fading. We use the modified Jakes' model [20] to simulate different fading rates, resulting in different time diversity orders.

Fig. 2 illustrates the proposed scheme for transmission of an embedded bitstream over a mobile channel characterized by a doubly selective environment. In the frequency domain, $S_{tot} = NM$ symmetric descriptions of approximately equal importance are constructed in which contiguous information from the embedded bitstream is spread across the multiple descriptions/packets [4], [5]. The information symbols are protected by systematic (n, k) RS codes, with the level of protection depending on the relative importance of the information symbols, as well as on the order of diversity available in the frequency domain. Generally, an (n, k) MDS erasure code can correct up to $n - k$ erasures. Hence, if *any* g out of n descriptions are received, those codewords with minimum distance $d_{min} \geq n - g + 1$ can be decoded. As a result, decoding is guaranteed at least up to distortion $D(R_g)$, where $D(R_g)$ refers to the distortion achieved with R_g information symbols.

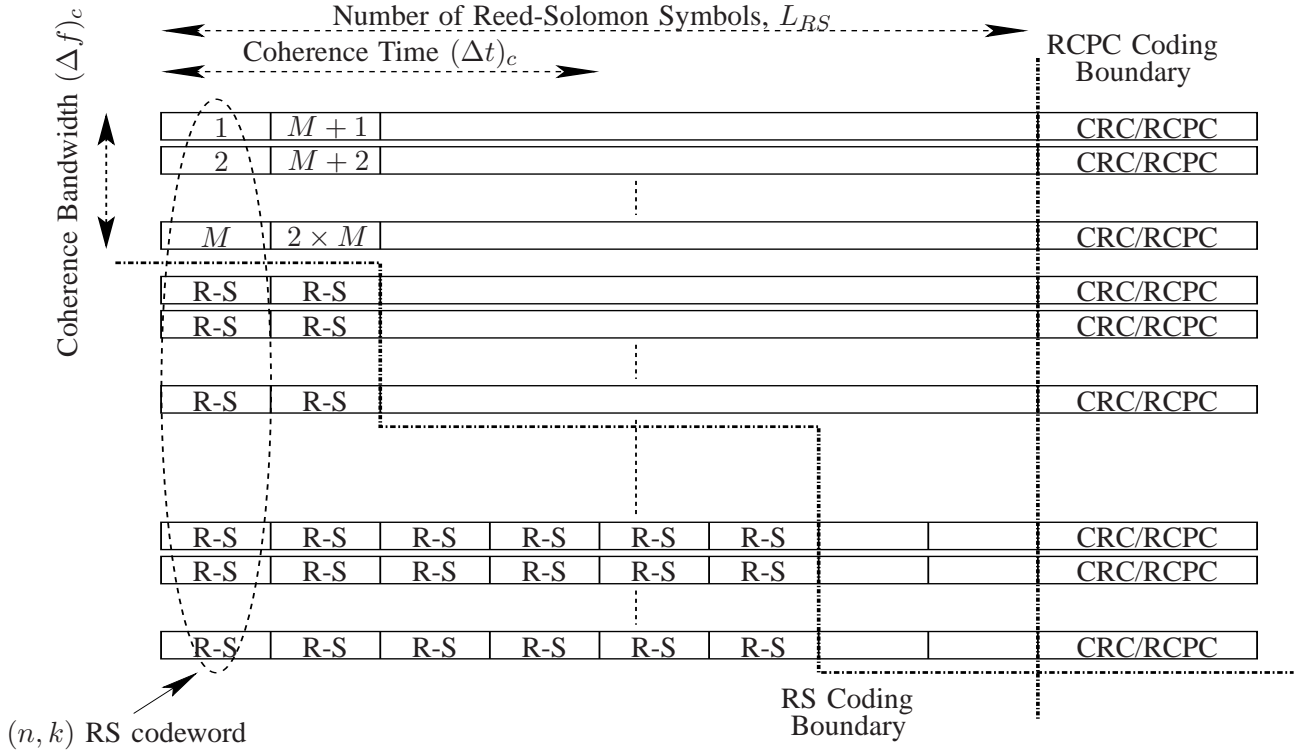


Fig. 2. The transmission of embedded bitstreams over OFDM mobile wireless networks; note that the CRC/RCPC parity symbols are interleaved with the RS symbols in the actual system.

The individual descriptions are then mapped to the $S_{tot} = NM$ subcarriers. A concatenation of CRC codes and RCPC codes, for possible diversity and coding gains in the time domain, are applied to each description. Since the descriptions are approximately equally important, RCPC codes with the same channel code rate can be applied to protect each individual description. This results in a vertical boundary (RCPC coding line), as illustrated in Fig. 2. The symbols on the left of the boundary are the RS symbols, while those on the right are CRC/RCPC parity symbols. It should be noted that the multiple description RS symbols and RCPC parity symbols would be interleaved in an actual system. However, for illustration, we show the de-interleaved version throughout the paper so that the relative amounts of RCPC parity symbols and RS symbols can be clearly indicated.

Since both forms of diversity are not necessarily simultaneously available at any given instant of time, the channel coding scheme should be designed to synergistically exploit the available diversity. For example, in a slow fading environment, channel coding plus interleaving is usually ineffective, especially for delay-sensitive applications such as real-time multimedia services. Hence, in this case, frequency

Pros	Cons
Higher orders of time domain diversity	Larger channel estimation errors
⇒ higher coding gain	⇒ Lower channel decoding efficiency
⇒ higher diversity gain	Higher level crossing rates
	⇒ Errors scattered across multiple packets
	⇒ Lower application layer throughput

TABLE I

FACTORS AFFECTING THE APPLICATION LAYER QUALITY-OF-SERVICE (QoS) IN A FAST FADING ENVIRONMENT.

diversity techniques may be more effective than time diversity techniques.

As stated previously, traditional studies of progressive transmission have concentrated on slow fading channels. In fact, in addition to the performance differences in channel coding efficiencies and channel estimation accuracies, the error patterns for different fade rates also affects the application layer throughputs and hence the end-user delivered quality. In particular, in a fast fading environment, the errors are more scattered among multiple packets due to the higher level crossing rate which measures how often the fading crosses some threshold [21]. However, for a slow fading environment, the errors appear more bursty. Consequently, the application layer throughput, measured by the number of successively transmitted packets, of a fast fading environment can be dramatically lower than that of a slow fading system. In Table I, we summarize the factors affecting the selection of an optimal channel coding scheme and end-user performance due to different fading rates.

On the other hand, information on frequency diversity can assist a source-channel codec in selecting a more robust source-channel coding scheme [9]. For example, while unequal error protection (UEP) is considered as primarily important for robustness for some of the progressive transmission schemes proposed in the literature (e.g., [7]), it was shown that in a highly frequency selective environment, UEP only provides marginal improvement over equal error protection (EEP), while in a frequency diversity deficient system, UEP can greatly improve the performance of progressive transmission over an OFDM system.

III. ICI AND CHANNEL ESTIMATION ERRORS

The assumptions on perfect channel estimation and orthogonality between subcarriers cannot be considered accurate for fast fading environments. Rapid channel variations may cause severe ICI [22]–[24]

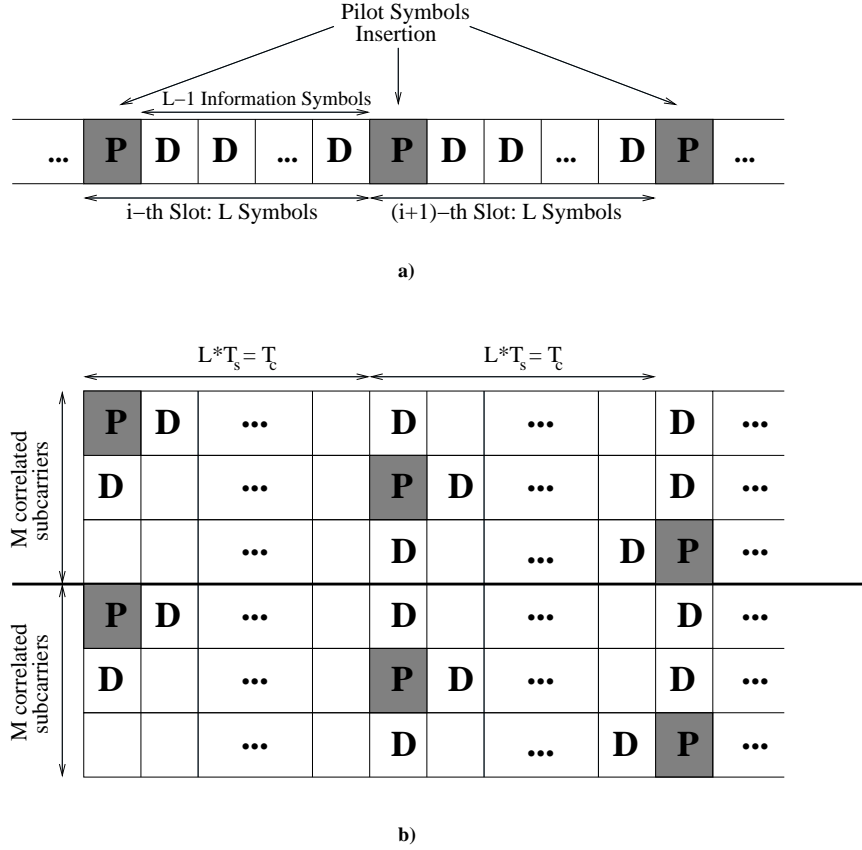


Fig. 3. Pilot insertion scheme for systems with: (a) one correlated subcarrier; (b) M correlated subcarriers.

and channel estimation errors, thereby degrading overall system performance. In this work, we model the ICI as in [22], i.e., a zero mean Gaussian random process with variance σ_{ICI}^2 expressed as

$$\sigma_{ICI}^2 = E_s - \frac{E_s}{N_t^2} \left\{ N_t + 2 \sum_{i=1}^{N_t-1} (N_t - i) J_0(2\pi f_{nd} i) \right\} \quad (1)$$

where E_s is the modulated symbol energy, N_t is the number of subcarriers in the OFDM system, f_{nd} is the normalized Doppler spread and $J_0(\cdot)$ is the zero-order Bessel function of the first kind. The ICI varies directly with the Doppler frequency.

In addition to ICI, channel variations in the time domain may also increase the difficulty in channel estimation. The accuracy of this channel state information (CSI) is particularly important for coherent demodulation and channel decoding. We adopt pilot symbol assisted modulation (PSAM), commonly used in practical OFDM networks [11], [12], [25]–[29]. We refer the reader to [25] for details of PSAM and the analysis of channel estimation errors.

In this work, as shown in Fig. 3a, pilot symbols are periodically inserted in the transmitted data symbols with a spacing equal to L modulated symbols. At the receiver, a linear minimum mean square error (MMSE) channel estimator [30] is adopted to estimate the fading coefficient using the following procedures: First, pilot symbols are extracted from the received sequence and the associated channel coefficients are evaluated. Then, the channel coefficient at the l -th data time ($l \neq jL$) is estimated by interpolating the K nearest pilot symbols with a Wiener filter. In particular, we consider $K = 2$, i.e., each channel parameter is estimated by interpolating the two closest pilot samples. Let us denote the received pilot symbol of the generic i -th slot¹ as $\tilde{h}[iL]$ and assume that the l -th data symbol is transmitted in this i -th slot. Hence $\tilde{\mathbf{h}} = [\tilde{h}[iL], \tilde{h}[(i+1)L]]^T$ is the set of two pilot symbols interpolated to estimate the l -th channel coefficient $h[l]$. Defining $\hat{h}[l]$ as the estimator of $h[l]$, the channel estimation errors can be expressed as

$$\varepsilon[l] = h[l] - \hat{h}[l], \quad l = iL \dots (i+1)L - 1. \quad (2)$$

The quality of the estimation is expressed in terms of the mean square error $\sigma_e^2 = E[\varepsilon^2]$, where $E[\cdot]$ denotes the expectation operator. Defining $\mathbf{w}[l] = E[\tilde{\mathbf{h}}h^*[l]]$ and $\mathbf{R} = E[\tilde{\mathbf{h}}\tilde{\mathbf{h}}^\dagger]$, the variance of the channel estimation errors can be expressed as [25] [12]²

$$\sigma_e^2[l] = 1 - \mathbf{w}^\dagger[l]\mathbf{R}^{-1}\mathbf{w}[l], \quad (3)$$

where superscripts $*$, T and \dagger stand for conjugate, transpose and transpose conjugate respectively. From (3), it can be seen that the estimation error variance depends on the channel correlation function. Recalling that we use a Jakes' model, the time correlation function is $r[n] = E[h[l]h^*[l+n]] \propto J_0(2\pi n f_{nd})$. This means that both the correlation function and the estimation error variance depend on the normalized Doppler frequency (f_{nd}). In particular, the channel estimation gets worse when the Doppler frequency increases. It is worthwhile to notice that the variance $\sigma_e^2[l]$ depends also on the received pilot samples $\tilde{\mathbf{h}}$, and thus on the signal-to-noise-ratio (SNR).

In this work, the pilot and data symbols are transmitted at the same power level. Since a frequent pilot insertion improves channel estimation, at the cost of reduced throughput, a fixed pilot scheme for

¹The transmitted bitstream is divided into slots of length equal to the pilot spacing, i.e., L symbols. The first symbol of each slot is a pilot symbol, the other $(L-1)$ symbols are data, as shown in Fig. 3.

²A multiplicative coefficient (the total average power of the channel impulse response) has been set equal to 1 and therefore ignored.

different Doppler frequency environments is not the best solution. Thus, we consider a pilot spacing equal to the coherence time. It follows that for a slow fading channel, the number of pilot symbols is negligible, leading to a high transmitted throughput. In high Doppler systems, to achieve good channel estimation, we have to reduce significantly the number of transmitted data symbols in each packet. Since under the block fading model, correlated subcarriers experience the same fading channel in the frequency domain, pilot symbols inserted once every coherence time are distributed among correlated subcarriers, as shown in Figure 3b. Thanks to this pilot scheme, the number of inserted pilot symbols decreases drastically in systems with low frequency diversity order.

IV. PROBLEM FORMULATION

In this section, we describe the optimization problem to be solved. Consider N i.i.d. subbands, each with M subcarriers and packet size equal to L_{RS} code symbols before channel coding using RCPC/CRC codes. Since each vertical column corresponds to one RS codeword, there are altogether L_{RS} RS codewords. The constraint on the bit budget/packet can then be written as

$$(L_{RS} \times B_{RS} + B_{CRC})/R_{rcpc} \leq B_{tot}, \quad (4)$$

where B_{CRC} is the bit budget allocated for the CRC codes and R_{rcpc} is the channel code rate of the RCPC codes. B_{RS} is the number of bits-per-RS symbol and B_{tot} is the total bit budget of the RB.

We assume that for RS codeword l , where $l \in [1, L_{RS}]$, c_l code symbols are assigned to information data symbols. Hence, the number of RS parity symbols assigned to codeword l is

$$f_l = S_{tot} - c_l \quad l \in [1, L_{RS}]. \quad (5)$$

Let ϕ_{th} be the minimum number of descriptions that a decoder needs to reconstruct the source, and g be the number of correctly received packets. The reception of any number of packets $g \geq \phi_{th}$ leads to improving image quality $D(R_g)$, where R_g is the allocated bit budget for the information symbols,

$$R_g = \sum_{\{l: c_l \leq g\}} c_l \times B_{RS}. \quad (6)$$

Hence, the overall RS channel code rate equals $R_{rs} = R_{S_{tot}}/(S_{tot} \times L_{RS} \times B_{RS})$. Given the source code rate-distortion curve $D(R_g)$ and the packet loss probability mass function $P_{\mathcal{J}}(j)$, where $j = S_{tot} - g$ is the number of lost packets, we can minimize the expected distortion as follows:

$$E^*[D] = \min_{\{c_l, R_{rcpc}\}} \left\{ \sum_{j=0}^{S_{tot}-\phi_{th}} P_{\mathcal{J}}(j) D(R_{S_{tot}-j}) + \sum_{j=S_{tot}-\phi_{th}+1}^{S_{tot}} P_{\mathcal{J}}(j) D_0 \right\}, \quad (7)$$

subject to the constraint on the overall bit budget

$$\frac{R_{S_{tot}}/R_{rs} + B_{CRC}}{R_{rcpc}} \leq B_{tot} \quad (8)$$

where D_0 corresponds to the distortion when fewer than ϕ_{th} descriptions are received and so the decoder must reconstruct the source without being able to use any of the transmitted information. For a still image, this typically means reconstructing the entire image at the mean pixel value.

The packet loss probability mass function $P_{\mathcal{J}}(j)$ depends on $(\Delta f)_c$, $(\Delta t)_c$ and R_{rcpc} . Although $P_{\mathcal{J}}(j)$ can be found analytically for uncorrelated fading channels, due to the correlated fading in both time and frequency domains of the mobile environment considered here, we use simulations to find $P_{\mathcal{J}}(j)$. We use the iterative procedure described in [5] to solve the optimization problem (7).

V. RESULTS AND DISCUSSION

We carried out simulations on the 512×512 gray-scale images Lena, Peppers and Goldhill. Similar results were obtained for all three. Hence, in this paper, we only present the results using the Lena image. The image was encoded using the SPIHT [31] algorithm to produce an embedded bitstream. The serial bitstream was converted into 128 parallel bitstreams using the FEC-based multiple description encoder. The 128 descriptions were mapped to the OFDM system with 128 subcarriers. We used RS codes in the frequency domain and there were 8 bits per RS symbol. The packet size was set equal to 512 bits. We used QPSK modulation and considered both perfect and imperfect CSI. The RCPC codes of rates $R_{rcpc} = \frac{8}{9}, \frac{8}{10}, \dots, \frac{8}{24}$, were obtained by puncturing an $R_c = 1/3$ mother code with $K = 7$, $p = 8$ and generator polynomials $(133, 165, 171)_{octal}$ with the puncturing table given in [32].

In the following figures, we illustrate the proposed channel coding scheme under different fading environments and study the effects of channel estimation on the selection by comparing performance of systems with perfect CSI to systems with imperfect CSI and ICI. From here onwards, for systems with imperfect CSI and ICI, ICI is omitted from the notation for sake of brevity, although it is considered as well. We begin by studying the optimized construction of RS information and parity symbols for

- Different values of RCPC coding rate,
- Both perfect and imperfect CSI, and
- Different frequency diversity orders and different fading rates.

Then we study how the received image PSNR varies for different Doppler spreads, for perfect and imperfect CSI, and for different frequency diversity orders.

In Fig. 4, we show the optimized construction of RS information symbols, RS parity symbols and RCPC parity symbols for different R_{rcpc} 's for $(N, M) = (4, 32)$ and normalized Doppler spread $f_{nd} = 10^{-3}$

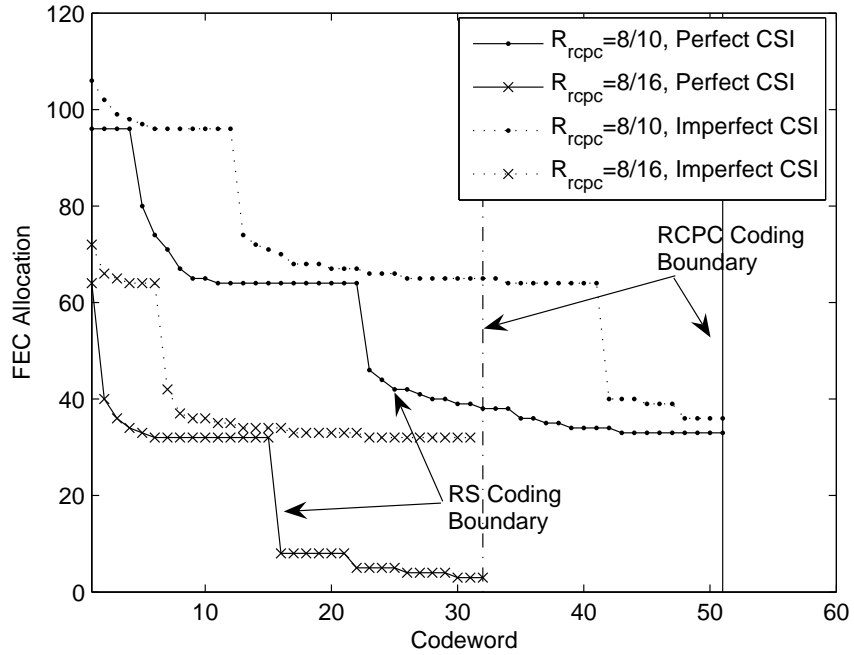


Fig. 4. Profiles showing the optimal allocation of source and channel symbols for systems with $(N, M) = (4, 32)$, $f_{nd} = 10^{-3}$ and $\text{SNR} = 16.0$ dB for different choices of RCPC coding rates and for both perfect CSI and imperfect CSI.

at $\text{SNR} = 16$ dB for systems with perfect CSI and imperfect CSI. The maximum order of diversity achieved in the frequency domain is $\mathcal{D}_f = 4$, while the maximum order of diversity in the time domain is $\mathcal{D}_t = 1$. In other words, no diversity can be exploited by using RCPC codes, although coding gain can still be obtained. In general, lower code rates in the time domain improve the packet loss performance, thus reducing the number of RS parity symbols required for minimizing the expected distortion $E[D]$, as can be noticed from the figures. Moreover, since for a fixed code rate in the time domain the perfect-CSI system outperforms the imperfect-CSI system, the latter system requires more protection in the frequency domain than does the perfect system.

As can be seen from Fig. 4, the RS code boundaries exhibit similar degree of tilting for both R_{rpc} rates, and for both the perfect CSI and imperfect CSI systems for the same diversity order. As discussed in [9], the degree of tilt of the RS boundary indicates the importance of unequal error protection (UEP) relative to equal error protection (EEP) which has a horizontal RS boundary line. Hence, the results demonstrate that although the packet loss performance of an individual subcarrier can be improved by using a lower channel coding rate, the degree of UEP, represented by the tilt of the RS boundaries, mainly depends on

the frequency diversity order of the system, and is relatively insensitive to the selection of the channel code rate in the time domain. In addition to the similar degree of tilting, the curves also show similar stepwise behavior. In particular, the RS boundaries show similar leveling behavior at approximately the same FEC value with step-size roughly equal to the coherence bandwidth. This observation agrees with the simulation results shown in [9] which is mainly due to, in addition to the same diversity order, the perfectly correlated fading within a subband in the frequency domain, which results in, with high probability, the simultaneous loss of the correlated subcarriers when a subband is under a deep fade.

In general, as the frequency diversity order increases, the variation of the number of lost packets decreases and thus reduces the need and hence the relative advantages of UEP, as shown in Fig. 5. In particular, in Fig. 5, we show the optimal allocation of source and channel symbols for imperfect CSI systems with different frequency diversity orders ($N = 1, 4, 32, 128$) in an environment with $f_{nd} = 10^{-2}$. The time domain channel code rate R_{rcpc} is fixed at $8/24$. As can be seen, in spite of significant difference in the time domain channel conditions due to the effect of fast fading and the time domain channel coding, similar behavior of the RS boundaries can still be observed when compared with the slow fading system reported in Fig.12 of [9]. Specifically, the amount RS code rate increases with the increasing frequency diversity order while the degree of UEP decreases with increasing frequency diversity order. Observe that at $N = 128$, the RS boundary is almost flat.

The tradeoff between RCPC codes and RS codes for both perfect and imperfect CSI is further illustrated in Fig. 6, where the optimal R_{rs} vs. R_{rcpc} is shown. By lowering the RCPC code rates, better packet loss performance is achieved due to the coding gain, and less protection in the frequency domain is required.

In Fig. 7, we plot the optimal peak-signal-to-noise ratio (PSNR) performances against R_{rcpc} for selected normalized Doppler spreads in systems with $(N, M) = (4, 32)$, SNR = 16 dB and imperfect CSI at the receiver. In the figure, for each selection of the time domain channel coding rate R_{rcpc} , the RS boundary is optimally constructed to maximize the delivered PSNR based on the frequency diversity order of the system. For comparison, in the plot, we also include the curves for the normalized Doppler spread with $f_{nd} = 10^{-1}$ and 10^{-4} with perfect CSI, representing the fast fading and slow fading scenarios with ideal channel estimation. As can be observed, the curve corresponding to of $f_{nd} = 10^{-4}$ with perfect CSI tracks the performance of the system with imperfect CSI closely, with minor degradation due to channel estimation errors. However, the curve corresponding to $f_{nd} = 10^{-1}$ and imperfect CSI deviates significantly from the system with perfect CSI due to the high channel estimation errors in a fast fading environment. It is worth mentioning that for the fast fading $f_{nd} = 10^{-1}$ environment and high signal-to-noise ratio, the system with imperfect CSI performs close to the perfect CSI system, indicating that

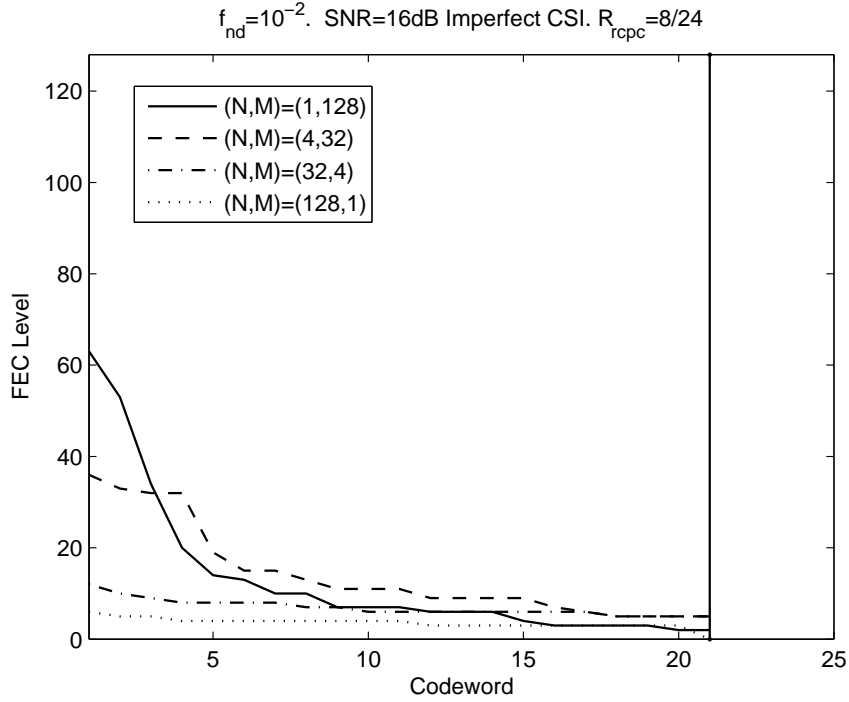


Fig. 5. Profiles showing the optimal allocation of source and channel symbols for systems with $R_{rcpc} = 8/24$ and $\text{SNR} = 16.0$ dB and imperfect CSI for systems with frequency diversity orders $N = 1, 4, 32, 128$, respectively.

extra redundancy can effectively compensate for channel estimation errors in a highly mobile scenario. Note that if the signal-to-noise ratio is low, even the lowest channel code rate in the time domain cannot sufficiently compensate for the effects of channel estimation errors.

Perhaps the more interesting observation is the crossovers among the curves with different fade rates. To explain crossovers, we look at the two extremes of the plot, i.e., the highest and lowest time domain channel code rate $R_{rcpc} = 1$ and $R_{rcpc} = 0.333$. For the selected fade rates, at $R_{rcpc} = 0.333$, the PSNR performance increases monotonically with the fade rate, while the PSNR performance at $R_{rcpc} = 1$ shows a monotonic decreasing behavior with increasing fade rate. The different behaviors are due to the two countering effects on the system performance as a result of increasing fade rate. As stated previously, on the one hand, the increase in fade rate increases the diversity order in the time domain and hence the efficiency of the RCPC channel coding. However, on the other hand, due to the higher level crossing rate in a fast fading system, errors are scattered across multiple packets rather than being bursty. For systems with little or no channel coding in the time domain, this scattered nature of the error pattern can

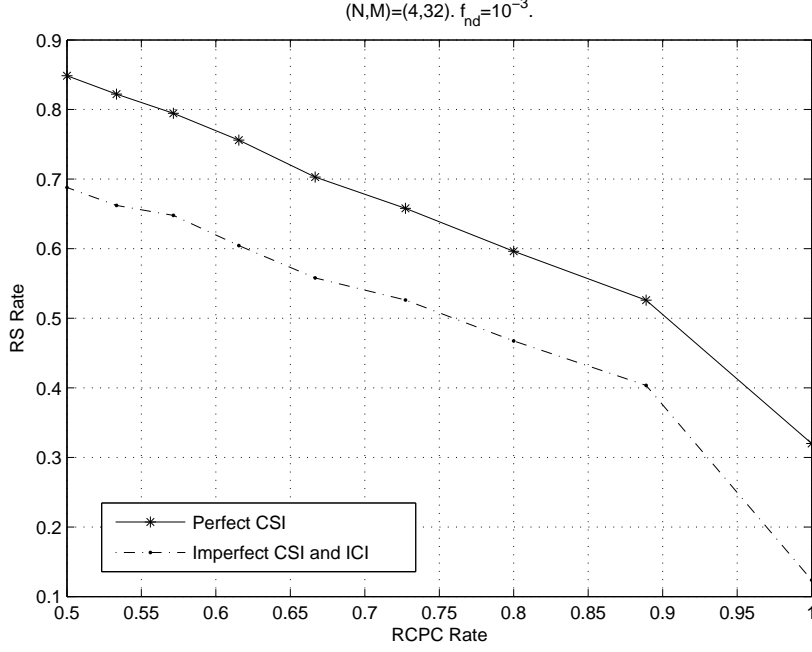


Fig. 6. Optimal R_{rs} vs. R_{rpc} for systems with $(N, M) = (4, 32)$, $f_{nd} = 10^{-3}$ SNR = 16.0 and both perfect and imperfect CSI.

significantly increase the packet loss rate and reduce the application layer throughput. Consequently, the PSNR performance drops drastically, as can be noticed by the significant performance degradation for $f_{nd} = 10^{-1}$ and 10^{-2} . As we shall see below, due to the higher level crossing rate associated with fast fading environments, the correct selection of an RCPC rate is more important for a fast fading system than for a slow fading system.

To further illustrate the effect of the error pattern on the PSNR performance, in Fig. 8, we plot the optimized PSNR performance vs. the normalized fading speed (f_{nd}) for $R_{rpc} = 1$ for a system with a frequency diversity order $N = 16$ and SNR = 16 dB. We include the PSNR performance curves for both perfect CSI and imperfect CSI. As expected, the performance difference between perfect CSI and imperfect CSI widens as the fade rate increases due to the increasing channel estimation errors. However, more importantly, both curves show a monotonic decreasing behavior with an increasing fade rate due to the increasingly scattered error pattern. In the Appendix, we provide some further analysis for the packet error rates due to the effects of error patterns resulting from different fading rates on the application layer throughput. In particular, by combining the threshold model [33] and the analysis on fade duration

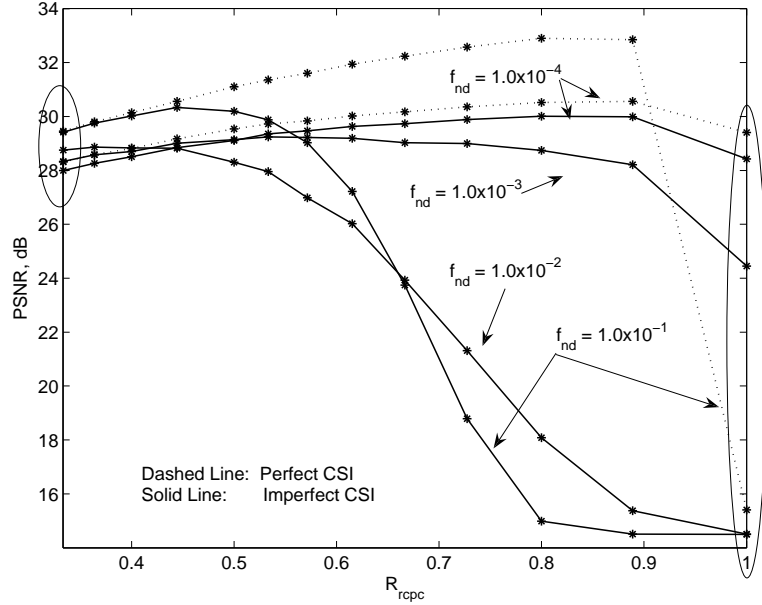


Fig. 7. Optimized PSNR vs R_{rpc} for different Doppler spreads in systems with $(N, M) = (4, 32)$, SNR = 16 dB and imperfect CSI.

distribution [34], [35], we provide a simple analytic solution showing that for an uncoded system, the application layer throughput decreases exponentially with increasing fade rates due to the fact that deep fading events are shorter but occur more frequently.

In Fig. 9(a) and Fig. 9(b), we illustrate the effects of the frequency and time diversity orders on the selection of optimal coding schemes. In Fig. 9(a), we show the optimal PSNR performance vs. R_{rpc} for different diversity orders ($N = 1, 2, \dots, 128$) in a system with $f_{nd} = 10^{-3}$, SNR = 16 dB and imperfect CSI. In the figure, we also mark with a circle (o) the optimal R_{rpc} . As the system experiences low Doppler spread with $\mathcal{D}_t = 1$ and channel estimation becomes more accurate, the selection of optimal coding schemes is dominated by the frequency diversity order of the system. As can be observed, generally a better performance can be achieved with a higher diversity order. More importantly, as the diversity order N increases, the optimal R_{rpc} increases and the delivered image quality improves accordingly. Notice also that, except for the case $N = 1$, the PSNR performance curves are relatively flat around the optimal R_{rpc} . To give a specific example, consider the PSNR performance curve for $N = 8$. Although $R_{rpc} = 0.62$ gives the optimal performance, if $R_{rpc} = 0.5$ or 0.87 is selected instead, only minor degradation is suffered. This is because, in a slow fading environment, the performance loss due to the

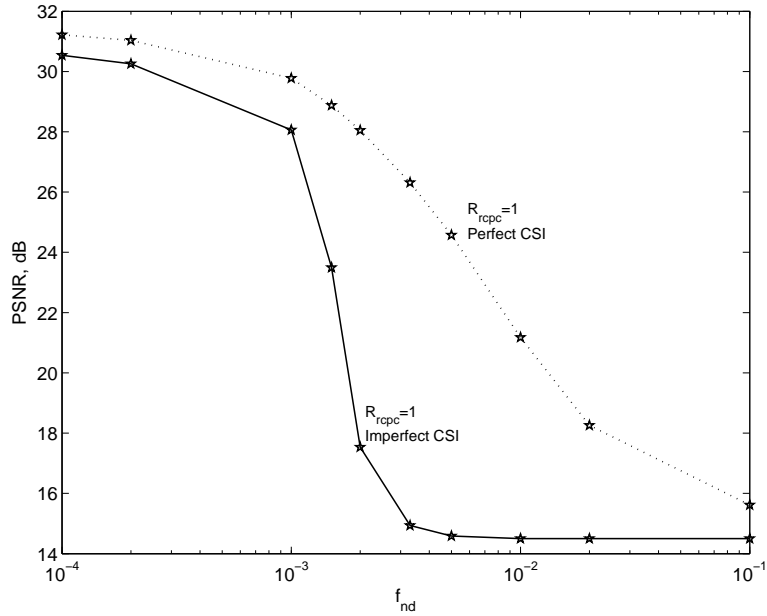


Fig. 8. Optimized PSNR vs. f_{nd} for systems with $(N, M) = (16, 8)$, SNR = 16 dB and $R_{rpc} = 1$ for perfect and imperfect CSI systems .

non-optimal selection of R_{rpc} is partly compensated by the RS coding in the frequency domain. The results indicate that in a slow fading environment, by using the proposed 2D coding scheme, the results are relatively insensitive to the selection of R_{rpc} , which can be selected on a broad range. The sub-optimal approach only sacrifices marginal performance degradation. The case $N = 1$ represents a flat fading environment, in which RS coding across the subcarriers becomes ineffective.

In Fig. 9(b), instead of a slow fading environment, we study the performance of under fast fading conditions. Specifically, we plot the optimal PSNR performance vs. R_{rpc} for different frequency diversity orders for a fast fading system with $f_{nd} = 10^{-1}$, SNR = 16 dB and imperfect CSI. As can be easily noticed, by comparing Fig. 9(b) with Fig. 9(a), the performance of a fast fading system is drastically different from that of a slow fading system due to a combination of higher diversity, more scattered errors and poorer channel estimation accuracy associated with fast fading environments. In particular, the system experiences a relatively flat region at low R_{rpc} rates and a drastic drop in PSNR as it moves towards high R_{rpc} . Observe that, although the system with a higher frequency diversity order generally provides a better performance, unlike the slow fading systems, the optimal R_{rpc} 's are relatively insensitive to the frequency diversity order. This is because in the time domain, the performance is dominated by the high

time diversity gain, and thus PSNR depends only slightly on the frequency diversity order. Moreover, it is worth noting that, due to the high time diversity order, the time domain channel coding is very effective and the optimal channel code rate in the time domain is dominated by the channel estimation errors and ICI.

This can be illustrated by comparing Fig. 9(b) with Fig. 9(c). In Fig. 9(c), we plot the corresponding system with perfect CSI as opposed to the system with imperfect CSI shown in Fig. 9(b). Observe that, generally, high $R_{r_{cpc}}$'s are preferred for better system performance. However, the performance is relatively insensitive to the frequency diversity order. Moreover, both systems exhibit precipitous drops in PSNR performance due to a more dispersed error pattern, leading to poor application layer throughput if the system is under-protected.

In Fig. 10(a) and Fig. 10(b), we show the optimized PSNR vs. both normalized Doppler spread f_{nd} and the number of independent subbands (N) using the proposed coding scheme for a 2D time-frequency OFDM resource block with perfect and imperfect CSI, respectively. The SNR is set to 16 dB. As can be observed from Fig. 10(a), without channel estimation errors, systems with greater diversity opportunities in time and/or frequency domains generally give better performance. However, more importantly, observe the relatively stable performance under different physical environments. Only for both low Doppler and flat fading environments does the system perform poorly. For other values of frequency and time diversity, the PSNR provided is always more than 30dB, even in the case of systems with low time or low frequency diversity order.

In Fig. 10(b), we plot the corresponding system with imperfect CSI. Complicated by the effects of channel estimation errors, the optimal performance becomes more irregular. While in general systems with higher frequency diversity orders outperform systems with lower frequency diversity orders, some irregularities are observed in the time domain. In particular, the PSNR drops with decreasing fade rate and starts to rise again at fade rates around 10^{-2} and 10^{-3} . To have a better understanding of the optimal behavior, in Fig. 11 we show the optimal PSNR performance vs. f_{nd} for systems with two different frequency diversity orders, $\mathcal{D}_f = 4$ and $\mathcal{D}_f = 32$. The SNR is set to 16 dB. The OFDM resource block is constructed with optimal RS profiles and RCPC rates based on the proposed scheme. For comparison, systems with both perfect and imperfect CSI are considered. By first looking at the slow fading section, i.e., the region with $f_{nd} < 10^{-3}$, it can be observed that the optimal PSNR performances are relatively flat, with some degradation in the systems with imperfect CSI due to channel estimation errors. The performance gap between the perfect CSI and imperfect CSI in this region is relatively small due to better channel estimation accuracy in a slow fading environment. At the middle section, i.e.,

($10^{-3} < f_{nd} < 10^{-2}$), the drops in PSNR performance are steepened for systems with imperfect CSI, due to the combined effects of low time diversity gain, increasing channel estimations errors, and the impact of the more scattered nature of the error pattern. For $f_{nd} > 10^{-2}$, the large channel variations provide significant time diversity gain which improves the efficiency of the RCPC codes and partly compensate for the performance loss due to the channel estimation errors.

In Table II and Table III, by defining the total/combined channel code rate as $R_{tot} = R_{rcpc} \cdot R_{rs}$ and the total system diversity order available as $\mathcal{D}_{tot} = \mathcal{D}_f \cdot \mathcal{D}_t$, we provide further analysis regarding the proposed 2D channel coding scheme. In particular, in Table II, the optimal PSNR, R_{rcpc} , R_{rs} , and R_{tot} are presented for a system with SNR = 16dB³, perfect CSI and different diversity orders in both the time and frequency domains. In each row, we keep \mathcal{D}_{tot} fixed and investigate the corresponding performance for different combinations of \mathcal{D}_f and \mathcal{D}_t . To provide a specific comparison, consider systems with \mathcal{D}_{tot} equal to 4. Obviously, this can be achieved with three different combinations: ($\mathcal{D}_f = 4, \mathcal{D}_t = 1$), ($\mathcal{D}_f = 2, \mathcal{D}_t = 2$), and ($\mathcal{D}_f = 1, \mathcal{D}_t = 4$). Generally, as expected, the higher the total diversity, the better is the quality of the received image, and the less is the required redundancy, which is reflected by the increase in combined channel coding rate R_{tot} . Comparing a system with no diversity to a system with $\mathcal{D} = 2$, a substantial gain in terms of PSNR can be observed (at least 2dB), although the gain diminishes with increasing diversity orders. This is because, although the error rate of a wireless communication system is generally a strictly decreasing function of the order of diversity, the gain diminishes with increasing order of diversity [36]. What is worth noting from the table is the behavior of optimal PSNR and R_{tot} for a constant total diversity order. Specifically, it can be seen that for a given \mathcal{D}_{tot} , with perfect CSI, both the optimal PSNR and the R_{tot} are roughly constant for all the possible combinations of \mathcal{D}_f and \mathcal{D}_t , independent of whether the diversity gain comes from the frequency domain or time domain.

In Table III, we provide a similar study for a system with imperfect CSI. Similarly to the observation above, we see an enhancement of the performance with an increase of the total diversity. However, for a fixed \mathcal{D}_{tot} , moving from slow fading to fast fading results in a decrease of the optimal PSNR, because of the channel estimation errors. Thus, for a fixed total order of diversity, the system with maximum frequency diversity order performs better than the system with maximum time diversity order.

³Although only SNR = 16dB is reported, we also considered other SNR values and behavior similar to the one for SNR = 16dB were observed.

TABLE II

OPTIMAL PSNR, R_{rcpc} , R_{rs} , AND R_{tot} FOR A SYSTEM WITH PERFECT CSI, SNR = 16dB AND DIFFERENT DIVERSITY ORDER IN BOTH TIME AND FREQUENCY DOMAINS.

DIVERSITY ORDER=1: ($N = 1, f_{nd} = 10^{-4}$)			
psnr = 27.28 dB $R_{rcpc} = 0.36, R_{rs} = 0.91$ $R_{tot} = 0.33$			
DIVERSITY ORDER = 2			
$\mathcal{D}_f = 2, \mathcal{D}_t = 1$ ($N = 2, f_{nd} = 10^{-4}$) psnr = 29.49 dB $R_{rcpc} = 0.73, R_{rs} = 0.69$ $R_{tot} = 0.50$		$\mathcal{D}_f = 1, \mathcal{D}_t \sim 2$ ($N = 1, f_{nd} = 3, 3 \cdot 10^{-3}$) psnr = 30.71 dB $R_{rcpc} = 0.53, R_{rs} = 0.95$ $R_{tot} = 0.51$	
DIVERSITY ORDER = 4			
$\mathcal{D}_f = 4, \mathcal{D}_t = 1$ ($N = 4, f_{nd} = 10^{-4}$) psnr = 30.55 dB $R_{rcpc} = 0.89, R_{rs} = 0.60$ $R_{tot} = 0.54$	$\mathcal{D}_f = 2, \mathcal{D}_t \sim 2$ ($N = 2, f_{nd} = 3, 3 \cdot 10^{-3}$) psnr = 30.96 dB $R_{rcpc} = 0.61, R_{rs} = 0.89$ $R_{tot} = 0.54$	$\mathcal{D}_f = 1, \mathcal{D}_t \sim 4$ ($N = 1, f_{nd} = 8 \cdot 10^{-3}$) psnr = 31.38 dB $R_{rcpc} = 0.67, R_{rs} = 0.92$ $R_{tot} = 0.61$	
DIVERSITY ORDER = 8			
$\mathcal{D}_f = 8, \mathcal{D}_t = 1$ ($N = 8, f_{nd} = 10^{-4}$) psnr = 31.23 dB $R_{rcpc} = 0.90, R_{rs} = 0.65$ $R_{tot} = 0.58$	$\mathcal{D}_f = 4, \mathcal{D}_t \sim 2$ ($N = 4, f_{nd} = 3.3 \cdot 10^{-3}$) psnr = 31.22 dB $R_{rcpc} = 0.73, R_{rs} = 0.78$ $R_{tot} = 0.57$	$\mathcal{D}_f = 2, \mathcal{D}_t \sim 4$ ($N = 2, f_{nd} = 8 \cdot 10^{-3}$) psnr = 31.59 dB $R_{rcpc} = 0.73, R_{rs} = 0.83$ $R_{tot} = 0.60$	$\mathcal{D}_f = 1, \mathcal{D}_t \sim 8$ ($N = 1, f_{nd} = 1.5 \cdot 10^{-2}$) psnr = 32.01 dB $R_{rcpc} = 0.73, R_{rs} = 0.92$ $R_{tot} = 0.67$
DIVERSITY ORDER = 16			
$\mathcal{D}_f = 16, \mathcal{D}_t = 1$ ($N = 16, f_{nd} = 10^{-4}$) psnr = 31.75 dB $R_{rcpc} = 0.89, R_{rs} = 0.70$ $R_{tot} = 0.62$	$\mathcal{D}_f = 8, \mathcal{D}_t \sim 2$ ($N = 8, f_{nd} = 3.3 \cdot 10^{-3}$) psnr = 31.64 dB $R_{rcpc} = 0.73, R_{rs} = 0.82$ $R_{tot} = 0.60$	$\mathcal{D}_f = 4, \mathcal{D}_t \sim 4$ ($N = 4, f_{nd} = 8 \cdot 10^{-3}$) psnr = 31.82 dB $R_{rcpc} = 0.73, R_{rs} = 0.88$ $R_{tot} = 0.64$	$\mathcal{D}_f = 1, \mathcal{D}_t \sim 16$ ($N = 1, f_{nd} = 3 \cdot 10^{-2}$) psnr = 32.37 dB $R_{rcpc} = 0.73, R_{rs} = 0.95$ $R_{tot} = 0.69$
DIVERSITY ORDER = 32			
$\mathcal{D}_f = 32, \mathcal{D}_t = 1$ ($N = 32, f_{nd} = 10^{-4}$) psnr = 32.13 dB $R_{rcpc} = 0.89, R_{rs} = 0.75$ $R_{tot} = 0.67$	$\mathcal{D}_f = 16, \mathcal{D}_t \sim 2$ ($N = 16, f_{nd} = 3.3 \cdot 10^{-3}$) psnr = 32.03 dB $R_{rcpc} = 0.80, R_{rs} = 0.81$ $R_{tot} = 0.64$	$\mathcal{D}_f = 8, \mathcal{D}_t \sim 4$ ($N = 8, f_{nd} = 8 \cdot 10^{-3}$) psnr = 32.16 dB $R_{rcpc} = 0.80, R_{rs} = 0.83$ $R_{tot} = 0.66$	$\mathcal{D}_f = 4, \mathcal{D}_t \sim 8$ ($N = 4, f_{nd} = 1.5 \cdot 10^{-2}$) psnr = 32.23 dB $R_{rcpc} = 0.73, R_{rs} = 0.93$ $R_{tot} = 0.68$

TABLE III

OPTIMAL PSNR, R_{rcpc} , R_{rs} , AND R_{tot} FOR A SYSTEM WITH IMPERFECT CSI, SNR = 16dB AND DIFFERENT DIVERSITY ORDER IN BOTH TIME AND FREQUENCY DOMAINS.

DIVERSITY ORDER=1: ($N = 1, f_{nd} = 10^{-4}$)			
psnr = 26.15 dB $R_{rcpc} = 0.33, R_{rs} = 0.90$ $R_{tot} = 0.30$			
DIVERSITY ORDER = 2			
$\mathcal{D}_f = 2, \mathcal{D}_t = 1$ ($N = 2, f_{nd} = 10^{-4}$) psnr = 28.75 dB $R_{rcpc} = 0.61, R_{rs} = 0.67$ $R_{tot} = 0.41$		$\mathcal{D}_f = 1, \mathcal{D}_t \sim 2$ ($N = 1, f_{nd} = 3, 3 \cdot 10^{-3}$) psnr = 25.47 dB $R_{rcpc} = 0.33, R_{rs} = 0.72$ $R_{tot} = 0.24$	
DIVERSITY ORDER = 4			
$\mathcal{D}_f = 4, \mathcal{D}_t = 1$ ($N = 4, f_{nd} = 10^{-4}$) psnr = 30.00 dB $R_{rcpc} = 0.80, R_{rs} = 0.58$ $R_{tot} = 0.47$	$\mathcal{D}_f = 2, \mathcal{D}_t \sim 2$ ($N = 2, f_{nd} = 3, 3 \cdot 10^{-3}$) psnr = 27.06 dB $R_{rcpc} = 0.33, R_{rs} = 0.70$ $R_{tot} = 0.23$		$\mathcal{D}_f = 1, \mathcal{D}_t \sim 4$ ($N = 1, f_{nd} = 8 \cdot 10^{-3}$) psnr = 28.13 dB $R_{rcpc} = 0.33, R_{rs} = 0.87$ $R_{tot} = 0.28$
DIVERSITY ORDER = 8			
$\mathcal{D}_f = 8, \mathcal{D}_t = 1$ ($N = 8, f_{nd} = 10^{-4}$) psnr = 30.74 dB $R_{rcpc} = 0.89, R_{rs} = 0.59$ $R_{tot} = 0.52$	$\mathcal{D}_f = 4, \mathcal{D}_t \sim 2$ ($N = 4, f_{nd} = 3.3 \cdot 10^{-3}$) psnr = 27.72 dB $R_{rcpc} = 0.36, R_{rs} = 0.66$ $R_{tot} = 0.24$	$\mathcal{D}_f = 2, \mathcal{D}_t \sim 4$ ($N = 2, f_{nd} = 8 \cdot 10^{-3}$) psnr = 28.48 dB $R_{rcpc} = 0.36, R_{rs} = 0.82$ $R_{tot} = 0.29$	$\mathcal{D}_f = 1, \mathcal{D}_t \sim 8$ ($N = 1, f_{nd} = 1.5 \cdot 10^{-2}$) psnr = 28.94 dB $R_{rcpc} = 0.36, R_{rs} = 0.90$ $R_{tot} = 0.33$
DIVERSITY ORDER = 16			
$\mathcal{D}_f = 16, \mathcal{D}_t = 1$ ($N = 16, f_{nd} = 10^{-4}$) psnr = 31.33 dB $R_{rcpc} = 0.89, R_{rs} = 0.65$ $R_{tot} = 0.58$	$\mathcal{D}_f = 8, \mathcal{D}_t \sim 2$ ($N = 8, f_{nd} = 3.3 \cdot 10^{-3}$) psnr = 28.22 dB $R_{rcpc} = 0.36, R_{rs} = 0.71$ $R_{tot} = 0.26$	$\mathcal{D}_f = 4, \mathcal{D}_t \sim 4$ ($N = 4, f_{nd} = 8 \cdot 10^{-3}$) psnr = 28.75 dB $R_{rcpc} = 0.36, R_{rs} = 0.83$ $R_{tot} = 0.30$	$\mathcal{D}_f = 1, \mathcal{D}_t \sim 16$ ($N = 1, f_{nd} = 3 \cdot 10^{-2}$) psnr = 29.79 dB $R_{rcpc} = 0.44, R_{rs} = 0.88$ $R_{tot} = 0.39$
DIVERSITY ORDER = 32			
$\mathcal{D}_f = 32, \mathcal{D}_t = 1$ ($N = 32, f_{nd} = 10^{-4}$) psnr = 31.77 dB $R_{rcpc} = 0.89, R_{rs} = 0.70$ $R_{tot} = 0.62$	$\mathcal{D}_f = 16, \mathcal{D}_t \sim 2$ ($N = 16, f_{nd} = 3.3 \cdot 10^{-3}$) psnr = 28.70 dB $R_{rcpc} = 0.44, R_{rs} = 0.66$ $R_{tot} = 0.29$	$\mathcal{D}_f = 8, \mathcal{D}_t \sim 4$ ($N = 8, f_{nd} = 8 \cdot 10^{-3}$) psnr = 29.23 dB $R_{rcpc} = 0.44, R_{rs} = 0.75$ $R_{tot} = 0.33$	$\mathcal{D}_f = 4, \mathcal{D}_t \sim 8$ ($N = 4, f_{nd} = 1.5 \cdot 10^{-2}$) psnr = 29.26 dB $R_{rcpc} = 0.44, R_{rs} = 0.77$ $R_{tot} = 0.34$

VI. CONCLUSION

We studied channel coding in a 2D time-frequency resource block of an OFDM system. In particular, we used symmetric n -channel FEC-based multiple descriptions based on the diversity order in the frequency domain. In the time domain, a concatenation of RCPC codes and CRC codes was employed to protect individual descriptions. We studied the performance of the proposed system in a doubly-selective channel with channel estimation errors. In a slow-fading environment, it was shown as the frequency diversity order increases, the optimal R_{rcpc} increases and the delivered image quality improves accordingly. On the other hand, in a fast-fading environment, the optimal R_{rcpc} is relatively insensitive to the frequency diversity order while the performance is limited by the channel estimation errors and ICI. It was also illustrated that the advantages of UEP protection diminishes as the frequency diversity order increases in both slow and fast fading environment. Thus, since both the optimal R_{rcpc} and R_{rs} vary depending on the channel conditions, a system can be robust only employing a 2D channel coding adaptable to both time and frequency diversity orders. Lastly, we illustrated that the bursty nature of a slow fading environment can lead to a higher application layer throughput and thereby deliver a better image quality while the scattered error pattern in a fast fading environment may lead to poor image quality.

APPENDIX

For a Rayleigh fading process $r(t)$, consider a simple two-state threshold model [33] with R_{th} being the threshold for the Rayleigh fading signal. If the signal level is above R_{th} (strong fade), the channel is considered to be in the good state, in which the probability of receiving the information is equal to 1, while if the signal level is below R_{th} (deep fade), the probably of receiving the particular information bit is equal to 0. Let us further assume that τ_f , τ_{nf} and τ_s are the deep fade duration, the strong fade duration and the deep fade inter-arrival intervals, so that

$$\tau_s = \tau_f + \tau_{nf}. \quad (9)$$

The average deep fade inter-arrival interval is the inverse of the level crossing rate (the expected rate at which the signal crosses the R_{th}), defined as [35]

$$\overline{\tau_s} = \frac{1}{\sqrt{2\pi} f_D \rho e^{-\rho^2}}, \quad (10)$$

where f_D is the maximum Doppler frequency, $\rho^2 = (R_{th}/R_{rms})^2$ is the inverse of the fade margin and R_{rms} is the root mean square of the fading signal. Denoting by N_r the level crossing rate, the average

deep fade duration is defined as [35]

$$\overline{\tau}_f = \frac{P(r(t) \leq R_{th})}{N_r} = \frac{e^{\rho^2} - 1}{\sqrt{2\pi} f_D \rho}, \quad (11)$$

so that the average strong fade duration can be expressed as

$$\overline{\tau}_{nf} = \overline{\tau}_s - \overline{\tau}_f = \frac{1}{\sqrt{2\pi} \rho f_D}. \quad (12)$$

Note that the ratio between the average strong fade duration and the average deep fade duration is not a function of the Doppler spread. Thus, increasing f_D causes a reduction of the average deep fade inter-arrival interval ($\overline{\tau}_s$), but also a shorter deep fade duration, as can be observed from (10) and (11), respectively. Hence, faster fading produces deep fade events that are shorter in duration but occur more frequently.

Since $r(t)$ is a Rayleigh random process, $r^2(t)$ is a χ^2 process and thus its asymptotic⁴ level down-crossing rate forms a Poisson process [37]. From the properties of Poisson random variables [38], it follows that

$$P_{\tau_s}(\tau) = \text{Prob}(\tau_s \leq \tau) = 1 - \exp\left(-\frac{\tau}{\overline{\tau}_s}\right), \quad (13)$$

and we can define the probability of having k deep fade arrivals within an interval of T_{PL} seconds as

$$\text{Prob}\{\mathcal{K}(T_{PL}) = k\} = \frac{(T_{PL}/\overline{\tau}_s)^k e^{-T_{PL}/\overline{\tau}_s}}{k!} \quad (14)$$

where T_{PL} corresponds to the duration of a packet, and $\mathcal{K}(T_{PL})$ is a random variable representing the number of deep fade arrivals in T_{PL} seconds.

Thus, in an uncoded system, the probability of having a packet correctly received (P_{succ}) is

$$P_{succ} = P_{good} \cdot \text{Prob}(\mathcal{K}(T_{PL}) = 0) = \exp\left(-\rho^2 - \frac{T_{PL}}{\overline{\tau}_s}\right) \quad (15)$$

where:

$$P_{good} = \text{Prob}(\text{packet starts in good fade}) = P(r(t) > R_{th}) = \exp(-\rho^2) \quad (16)$$

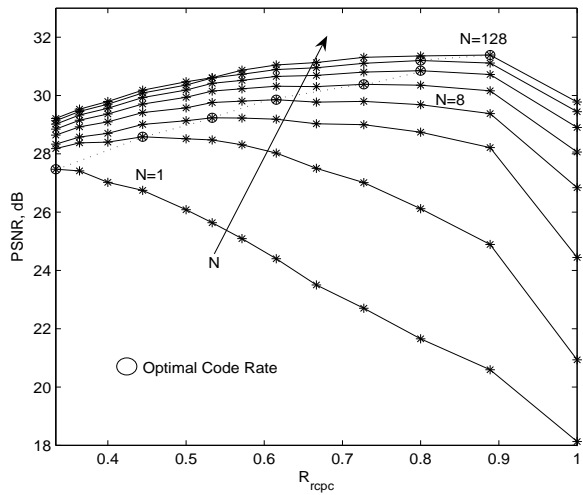
is the probability that a packet starts in the good state. From (15), it can be seen that the packet success rate probability decreases with decreasing inter-arrival time of the deep fades due to the fact that deep fading arrival are shorter but occur more frequently. In Fig. 12, we show both the simulation results and analytical results with different fade rates. As can be seen from the figure, the simulation and analytical results closely track one another. In particular, the packet success rate decreases with increasing fade rate. This explains the performance of Fig. 8, in which the PSNR of systems with either perfect CSI or imperfect CSI decreases monotonically with increasing fade rates.

⁴We use "asymptotic level down-crossing" to mean the level crossing of a very low threshold ($R_{th} \rightarrow 0$) [37].

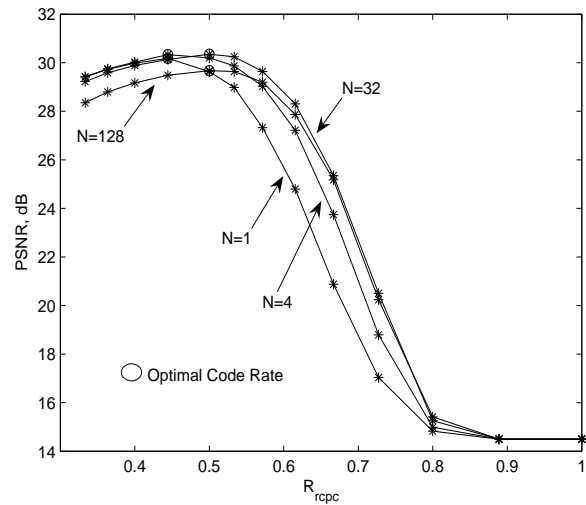
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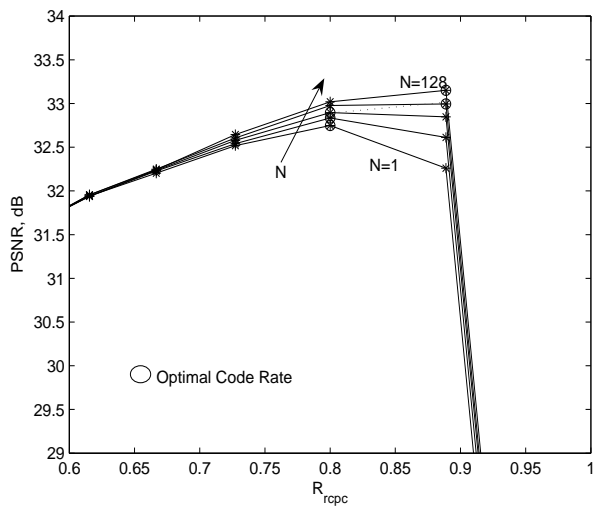
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(a) $f_{nd} = 10^{-3}$, SNR = 16 dB and imperfect CSI.



(b) $f_{nd} = 10^{-1}$, SNR = 16 dB and imperfect CSI.



(c) $f_{nd} = 10^{-1}$, SNR = 16 dB and perfect CSI.

Fig. 9. Optimized PSNR vs R_{rpc} for different coherence bandwidths.

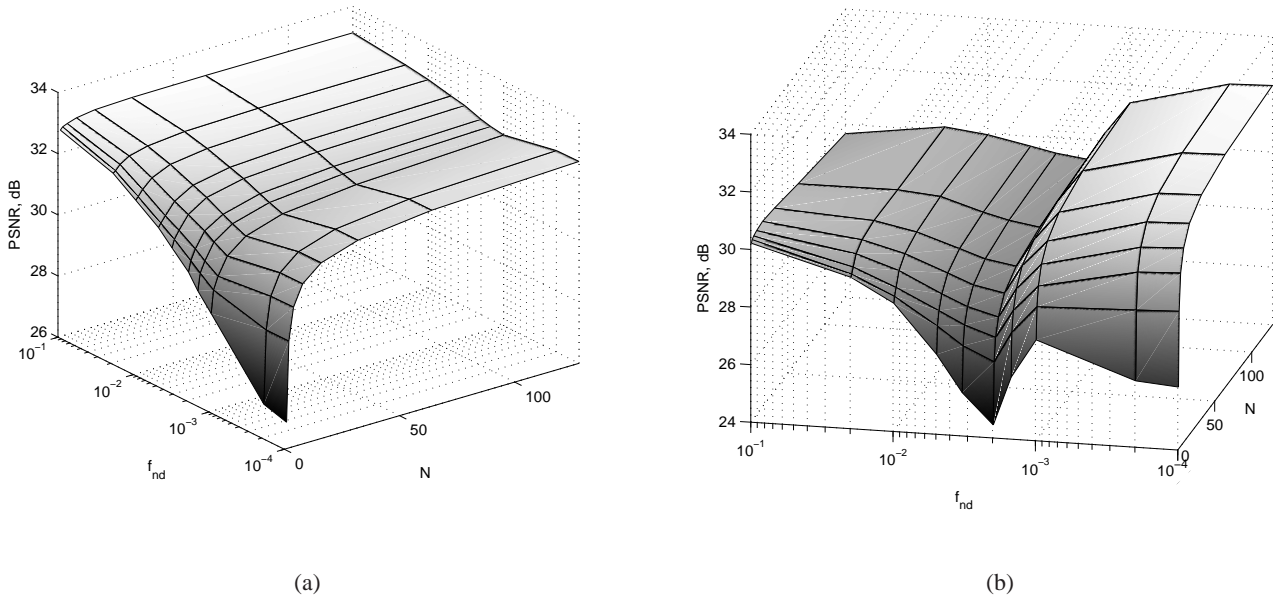


Fig. 10. Optimal PSNR performances vs. both R_{rec} and f_{nd} in systems with SNR = 16 dB for both perfect CSI systems (a) and imperfect CSI systems (b).

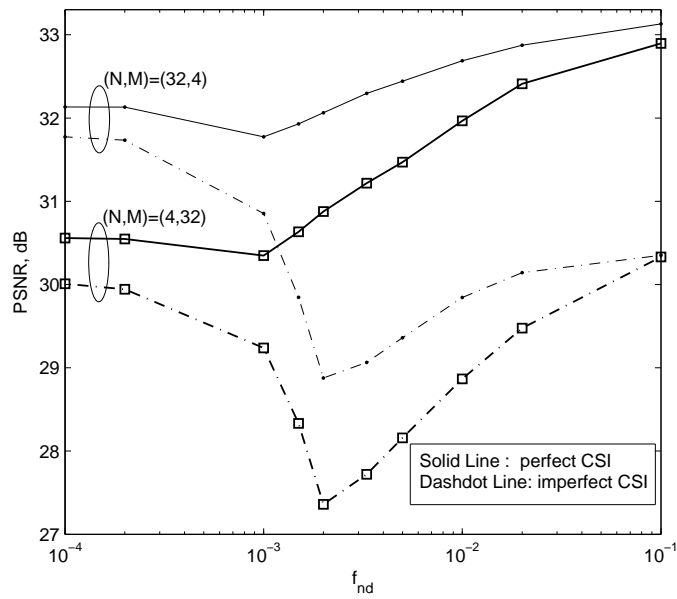


Fig. 11. Optimized PSNR vs. f_{nd} for systems with SNR = 16 dB and both perfect and imperfect CSI. Two different frequency diversity order are considered: $(N, M) = (32, 4)$ and $(N, M) = (4, 32)$.

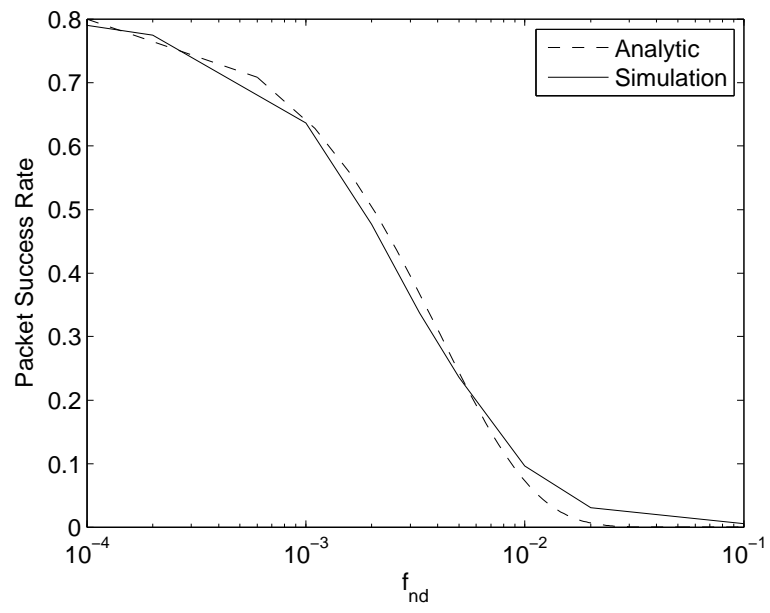


Fig. 12. Simulation and analytic results for the packet success rate for an uncoded system with different f_{nd} with $\text{SNR}(R_{th}) = 9$ dB.