A Simple Scalable Space-Frequency Coding Scheme for MIMO-OFDM

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***Abstract*— There are several space-time and space-frequency coding schemes proposed in the literature for MIMO-OFDM. In this paper, we present a linear space-frequency coding scheme based on permutation matrices, that is scalable with number of antennas and spatial-multiplexing (SM) rate, and is easy to implement. Furthermore, the design allows for low cost power amplifier (PA) design. For SM rate 1, our space-frequency coding scheme is backwards compatible with legacy receivers using only 1 receive antenna. These features of our so-called *Permuted Space Frequency Codes* enable seamless deployment of multiple antenna OFDM transceivers in existing wireless networks with legacy single antenna transceivers.**

I. Introduction

A coded MIMO-OFDM system [1] uses multiple transmit and receive antennas, and coded-OFDM to improve link reliability and data rates. Link reliability is obtained through transmit and receive diversity, and leads to improved coverage. A linear increase in data rates is obtained through spatial multiplexing, where an independant data stream is transmitted on each antenna. The number of independent streams is referred to as the multiplexing order (M).

In MIMO-OFDM wireless systems, edge-of-cell users typically operate in pure diversity mode ($M = 1$) to get improved reliability against fading and path-loss. Such a scenario require the use of space-time and space-frequency codes that maximize diversity gains. The optimal space-frequency code construction criterion that maximizes diversity gain is proposed in [5], but the codes lack closed form solution and require high complexity maximum likelihood (ML) decoding. Codes optimized for flat-fading environment such as the well-known Orthogonal Space-Time Block codes [6][7] can also be used in coded-OFDM systems. Unfortunately, the code structure is not scalable with number of antennas and has implementation bottlenecks. For example, the time domain implementation of block codes requires buffering of multiple OFDM symbols for decoding in the receiver, thereby increasing latency and implementation complexity. The frequency domain implementation of block codes results in performance loss in frequency selective channels, since the channel can vary within a space-frequency code block. The codes proposed by [2] using constellation precoding are scalable with number of antennas, but require high complexity ML decoding, and optimized for uncoded OFDM systems. It is worth mentioning that *none of the above codes are backwards compatible* with legacy systems such as IEEE 802.11a wireless local area networks (WLAN) [13] with single antenna receivers, i.e., such receivers cannot decode the signals sent from multiple-antenna transmitters, without making changes to the decoding algorithms. For example, the Alamouti code requires the receiver to use an Alamouti decoder, which is absent in legacy single antenna transceivers already deployed in the network.

In wireless environments with rich scattering and reasonably high SNR, spatial multiplexing is favorable. In practice, the multiplexing order can vary as $1 \leq M \leq min(M_R, M_T)$,where M_R and M_T are the number of receive and transmit antennas respectively. In such scenarios, codes that maximize data rates [4] while trading-off diversity [11] have been proposed, but they unfortunately lack closed form construction and require high complexity ML decoding. In [2], a scalable (with M, M_R and M_T) Layered Constellation Galois-Field (LCF) coding scheme was proposed. However, the code-construction framework does not assume a coded-OFDM system. Furthermore, since such codes require the so-called LCF [2] matrix multiplications of transmit symbol vectors, they cannot be decoded by legacy single antenna OFDM receivers, without making changes to the legacy decoding algorithms.

In this paper, we propose a simple linear scalable space-frequency coding scheme for MIMO-OFDM based on permutation matrices (also referred to as *Permuted Space Frequency (PSF) Codes* [9] in this article) that is (a) scalable with multiplexing rate and antennas and (b) backwards compatible with legacy receivers employing a single receiver antenna and coded-OFDM. Other advantages include simplicity in implementation, closed form code construction and less stringent power amplifier (PA) requirements. Simulation results are presented using the 802.11a physical layer system parameters and the recently adopted 802.11n channel models [12] to show scalability with rates and antennas, as well as backwards compatibility with legacy 802.11a receivers. Comparison with well-known schemes such as the Alamouti scheme and the circular delay diversity scheme [10] is also undertaken.

The rest of the paper is organized as follows. Section II provides the system model. Section III provides the space-frequency code construction, illustration of backwards compatibility with legacy systems and comparison with delay diversity code. Sections IV and V show simulation results

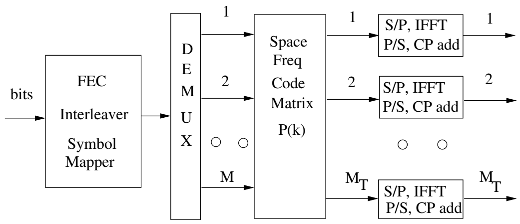

Fig. 1. MIMO-OFDM transmitter with space-frequency coding

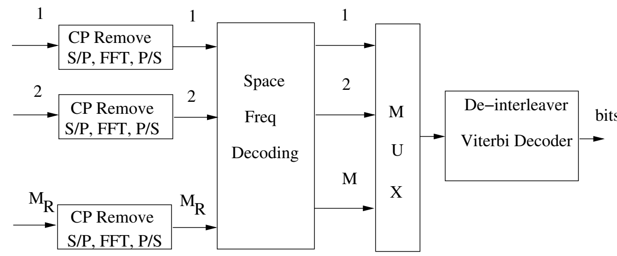

Fig. 2. MIMO-OFDM receiver with space-frequency decoding

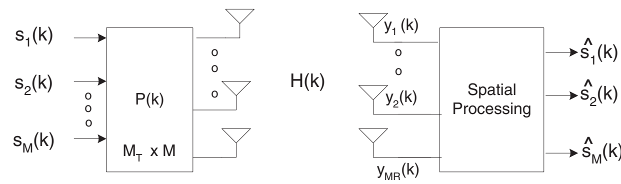

Fig. 3. System block diagram for an OFDM tone, k

using 802.11a parameters, and conclusions respectively.

II. System Model

Figure 1 shows a MIMO-OFDM transmitter with M_T transmit antennas and N OFDM tones, employing our space-frequency coding scheme. As in a typical coded-OFDM system (for example, the 802.11a transmitter), the input bit stream is first subject to scrambling, forward error correction coding (FEC), interleaving and symbol mapping to generate transmit symbols. The transmit symbol stream is then de-multiplexed so that M parallel symbols $s_1(k)\ s_2(k)\ ...\ s_M(k)$ are generated for each OFDM tone k, $0 \leq k \leq N-1$. We denote the $M \times 1$ vector $\mathbf{s}(k) = vec\,([s_1(k)\, s_2(k) \ldots s_M(k)\,])$. The symbol vector $\mathbf{s}(k)$ at each OFDM tone is multiplied by an $M_T \times M$ space frequency code matrix $\mathbf{P}(k)$ to generate an $M_T \times 1$ transmit vector $\mathbf{x}(k) = vec([x_1(k)\, x_2(k) \ldots x_{M_T}(k)])$ at each tone. The collection of matrices $\mathbf{P}(k)$ form a space-frequency code map. For each transmit antenna, j, the symbols $x_j(0)$, $x_j(1)$, ... $x_j(N-1)$ are subject to typical OFDM transmit processing such as serial-to-parallel conversion (S/P), IFFT processing, parallel-to-serial conversion (P/S), cyclic prefix (CP) addition and frequency up-conversion before launching the signal into the wireless channel.

Figure 2 shows the MIMO-OFDM receiver with M_R receive antennas. The signal received on each receive antenna is subject to typical OFDM processing such as frequency down-conversion, CP removal, S/P, FFT processing and P/S conversion to get the $M_R \times 1$ received vector, $\mathbf{y}(k) = vec([y_1(k)\, y_2(k) \ldots y_{M_R}(k)])$ at each tone. The system equation for each tone k (see Figure 3) can be written as:

$$\mathbf{y}(k) = \mathbf{H}(k)\mathbf{P}(k)\mathbf{s}(k) + \mathbf{n}(k) \qquad (1)$$

where $\mathbf{H}(k)$ is the $M_R \times M_T$ MIMO channel matrix at tone k, and $\mathbf{n}(k)$ is the $M_R \times 1$ noise vector at tone k, with the noise covariance matrix given as $\mathbf{R}_{nn}(k) = \sigma^2 \mathbf{I}_{M_R \times M_R}$. We assume without any loss of generality that $E_k\,[\,\mathbf{s}(k)\mathbf{s}(k)^*] = \mathbf{I}_{M \times M}$, i.e., unity constellation energy.

The received vector $\mathbf{y}(k)$ is then processed with a spatial decoder to recover an estimate of the $M \times 1$ transmitted symbol vector, $\hat{\mathbf{s}}(k) = vec([\hat{s}_1(k)\ \hat{s}_2(k)\ \ldots \hat{s}_M(k)])$. Assuming a linear zero-forcing (ZF) spatial decoder[1], the estimated symbol vector can be written as $\hat{\mathbf{s}}(k) = \mathbf{R}(k)\mathbf{y}(k)$, where $\mathbf{R}(k)$ is the $M \times M_R$ ZF spatial decoder given as:

$$\mathbf{R}(k) = [\mathbf{P}^*(k)\mathbf{H}^*(k)\mathbf{H}(k)\mathbf{P}(k)]^{-1}\mathbf{P}^*(k)\mathbf{H}^*(k) \qquad (2)$$

The M estimated symbol streams $\hat{s}_1(k)\ \hat{s}_2(k)\ \ldots \hat{s}_M(k)$ across all OFDM tones, are then multiplexed and subject to typical coded-OFDM decoding process, such as de-interleaving and Viterbi decoding to recover the transmitted bits (see Figure 2).

[1]Note that other linear decoders such as the minimum mean square error (MMSE) receiver, or non-linear decoders such as maximum-likelihood (ML) receiver, lattice-decoders or the well-known BLAST decoders can also be used instead.

III. Code Construction

There are 2 desirable power constraints on $\mathbf{P}(f)$ from an implementation point of view.

1) $||\,\mathbf{P}(k)\,||_F^2 = 1$, where $||.||_F$ is the Frobenius norm. This constraint ensures the total transmit power across all antennas on each OFDM tone to be unity without any loss of generality, thereby guaranteeing spectral flatness of transmitted power. A similar constraint is also enforced in standards such as 802.11a [13].
2) $\sum_{k=1}^{N} ||\,p_r(k)\,||^2 = \frac{N}{M_T}$, where $p_r(k)$ is the r^{th} row of $\mathbf{P}(k)$ and $1 \leq r \leq M_T$. This constraint ensures that the transmit power for each antenna is N/M_T, or better yet, the total transmit power across all antennas is N, regardless of the number of transmit antennas. Hence, as M_T increases, the power transmitted per transmit power amplifier (PA) decreases, *enabling low cost PA design*. It is worth mentioning that the transmit power per (transmit) antenna is identical regardless of the spatial multiplexing rate. This allows the transmitter to vary the spatial multiplexing rate (depending on the user location and channel conditions) on a OFDM symbol-by-symbol basis and avoid the task of changing the output power at the PA on a symbol-by-symbol basis.

Let us define an $M_R \times M$ equivalent channel matrix seen by the receiver at tone k to be:

$$\mathbf{H}_{eq}(k) = \mathbf{H}(k)\mathbf{P}(k) \qquad (3)$$

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Our code construction is motivated by the fact that the set of matrices $\mathbf{P}(k),\ 0 \le k \le N-1$, can be chosen so that the equivalent channel matrix $\mathbf{H}_{eq}(k)$ seen by the receiver can have improved frequency selectivity compared to the channel $\mathbf{H}(k)$, for any spatial multiplexing rate. The set of matrices $\mathbf{P}(k)$ can transform the available spatial selectivity across the transmit antennas into frequency selectivity. The improved frequency selectivity can then be exploited by the Viterbi decoder (across tones) in the coded-OFDM system to obtain improved frequency diversity gains.

Based on the above idea, we choose our $M_T \times M$ matrices $\mathbf{P}(k)$ that also satisfies the power constraints 1) and 2), as follows:

Lemma 1:

$$\mathbf{P}(k) = \frac{1}{\sqrt{M}}\mathcal{P}\{\ \lfloor \frac{k}{B} \rfloor\ mod\ M_T\} \qquad (4)$$

where $\mathcal{P}\{0\}, \mathcal{P}\{1\}, ..., \mathcal{P}\{M_T-1\}$ are the size $M_T \times M$ sub-permutation matrices derived from the M columns of the identity matrix $\mathbf{I}_{M_T \times M_T}$, and B is a parameter that is used to control the frequency selectivity of the equivalent channel.

As an example, the sub-permutation matrices for $M_T = 3, M = \{1, 2, 3\}$ are given below. For $M = 1$ we have:

$$\mathcal{P}\{0\} = \begin{pmatrix} 1 \\ 0 \\ 0 \end{pmatrix}\ \mathcal{P}\{1\} = \begin{pmatrix} 0 \\ 1 \\ 0 \end{pmatrix}\ \mathcal{P}\{2\} = \begin{pmatrix} 0 \\ 0 \\ 1 \end{pmatrix}$$

For $M = 2$ we have:

$$\mathcal{P}\{0\} = \begin{pmatrix} 1 & 0 \\ 0 & 1 \\ 0 & 0 \end{pmatrix}\ \mathcal{P}\{1\} = \begin{pmatrix} 0 & 0 \\ 1 & 0 \\ 0 & 1 \end{pmatrix}\ \mathcal{P}\{2\} = \begin{pmatrix} 0 & 1 \\ 0 & 0 \\ 1 & 0 \end{pmatrix}$$

For $M = 3$ we have:

$$\mathcal{P}\{0\} = \begin{pmatrix} 1 & 0 & 0 \\ 0 & 1 & 0 \\ 0 & 0 & 1 \end{pmatrix}\ \mathcal{P}\{1\} = \begin{pmatrix} 0 & 0 & 1 \\ 1 & 0 & 0 \\ 0 & 1 & 0 \end{pmatrix}$$

$$\mathcal{P}\{2\} = \begin{pmatrix} 0 & 1 & 0 \\ 0 & 0 & 1 \\ 1 & 0 & 0 \end{pmatrix}$$

From Lemma 1, we see that the matrices $\mathbf{P}(k)$ select antenna-tone pairs such that the transmit symbols are launched on an equivalent channel with increased frequency selectivity, compared to the channel responses from each transmit antenna. The increase in frequency selectivity is inversely proportional to B, as will be illustrated in next section. The optimum choice of B is dependant on implementation issues. A smaller value of B is desirable to obtain increased frequency selectivity that can be exploited by a powerful outer FEC code to obtain frequency diversity gains. On the other hand, if a weaker FEC is used, it is desirable to lower the frequency selectivity by using a larger value of B. Finally, a higher value of B may be desirable to curb excessive frequency selectivity and thereby improve accurate channel tracking across OFDM tones.

A. Backwards Compatibility with Legacy Receivers

The coding scheme proposed in Lemma 1 with $M = 1$ is backwards compatible with legacy receivers using $M_R = 1$ receiver antenna. This is explained as follows. When $M_R = 1$, the channel matrix at tone k is an $1 \times M_T$ vector, $\mathbf{h}(k)$. The equivalent channel matrix as seen by the legacy receiver is a complex scalar, $h_{eq}(k) = \mathbf{h}(k)\mathbf{P}(k)$, since $\mathbf{P}(k)$ is a $M_T \times 1$ vector (assuming $M = 1$). Since the receiver only sees a complex scalar equivalent channel $h_{eq}(k)$ regardless of the number of the number of transmit antennas, it can decode the multiple-antenna transmitted signal assuming $M = 1$. This enables seamless introduction of multiple antenna coded-OFDM transceivers in a wireless network with legacy receivers employing coded-OFDM as well. For communication between MIMO-OFDM transceivers, we can have $1 \le M \le min(M_R, M_T)$, whereas for communication between a legacy transceiver and multiple antenna OFDM transceiver, we restrict $M = 1$. It is worth mentioning that several purely transmit diversity schemes such as the Alamouti scheme [7], block codes proposed by Tarokh et at [6], and the scalable codes proposed by [2] are not backwards compatible with legacy receivers. Furthermore, they are not always scalable (in non-backwards compatible mode) with multiplexing rates and transmit antennas.

B. Comparison with Circular Delay Diversity Scheme

The well-known delay-diversity scheme (see for example [10]) is backwards compatible with legacy transceivers as well. A natural question to ask is how does our PSF encoding scheme with $M = 1$ compare with the delay diversity scheme. In the latter scheme, the signal sent on each transmit antenna (other than the first antenna) is a circularly delayed copy of the signal sent on the first transmit antenna. In the frequency domain, we can write the $M_T \times 1$ vector $\mathbf{P}(k) = vec\left([\,\phi_0\ \phi_1\ \phi_2\ ...\,\phi_{M_T-1}]\right)$, where $\phi_m(k) = e^{\frac{j2\pi\tau_m k}{N}}$ and τ_m is the circular delay (in samples) of the OFDM symbol sent on the $m+1\ ^{th}$ antenna, and assuming $\tau_0 = 0$. Substituting the expression for $\mathbf{P}(k)$ in (3), the $M_R \times 1$ equivalent channel becomes $\mathbf{h}_{eq}(k) = \sum_{m=0}^{M_T-1} \mathbf{h_m}(k)\phi_m(k)$, where $\mathbf{h_m}(k)$ is the $m+1\ ^{th}$ column of $\mathbf{H}(k)$.

We note that the equivalent channel response is a summation of the phase-rotated channel responses from different transmit antennas and hence can destructively combine at some tones to create additional frequency domain nulls. Hence, in line-of-sight channel conditions with low delay-spread, the delay diversity scheme converts an AWGN-like channel into a channel with frequency domain nulls. As a result, the performance can be worse than a single antenna legacy transceiver ! In contrast, the PSF coding scheme does not constructively or destructively add the channel responses from different antennas. As a result, no additional frequency domain nulls are created and the performance can only improve relative to the single antenna legacy transceivers. We will illustrate this via simulation results in the next section.

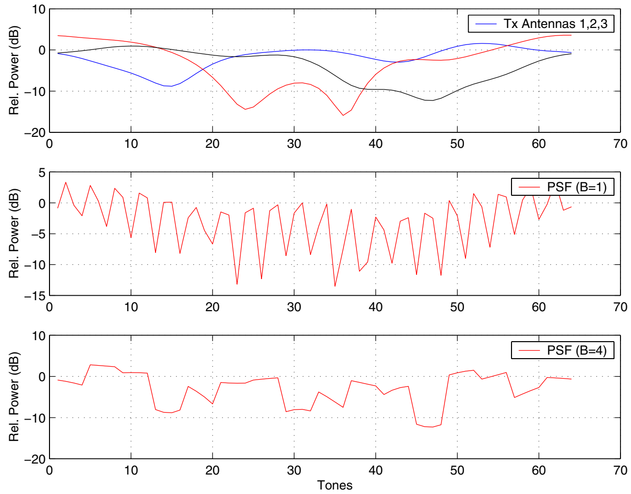

Fig. 4. Equivalent channel response for PSF code with $B = \{1, 4\}$

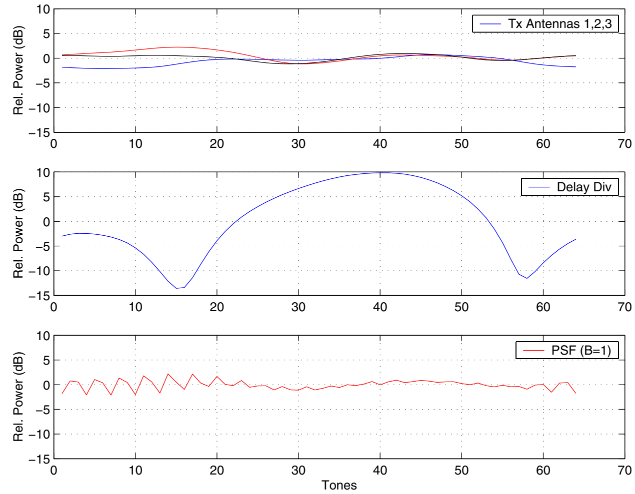

Fig. 5. Equivalent channel response in LOS conditions: Delay diversity code vs. PSF code

IV. SIMULATIONS

For our simulations, we assume a $M_T \times M_R$ MIMO system overlaid on a 802.11a coded-OFDM system with $N = 64$ tones spanning 20 MHz bandwidth. The 802.11a FEC rates are $1/2, 2/3$ and $3/4$ and possible modulation orders are BPSK, 4 QAM, 16 QAM and 64 QAM. We use the 802.11n channel models [12] to quantify the performance.

Figure 4 shows an example equivalent channel response as seen by a receiver antenna for PSF code with $B = \{1, 4\}$ and $M_T = 3$ transmit antennas. The individual channel responses from each of the transmit antennas is shown in top sub-plot, and the equivalent channel responses for $B = \{1, 4\}$ are shown in the middle and bottom sub-plots. The figure was generated assuming a the D-NLOS channel model [12] with 50nsec RMS delay spread. We see that the PSF coding scheme increases frequency selectivity of the equivalent channel, which can then be exploited by the 802.11a FEC to get improved frequency diversity gains. Indeed, for $B = 4$, the frequency selectivity is diminished, compared to the case when $B = 1$.

Figure 5 shows an example equivalent channel response for the circular delay diversity scheme with $\tau_0 = 0, \tau_1 = 1, \tau_2 = 2$ and a PSF code with $M = 1, B = 1$, for a system with $M_T = 3$ transmit antennas. As before, the individual channel responses from each of the transmit antennas is shown in top sub-plot, and the equivalent channel responses for circular delay diversity scheme and the $B = 1$ PSF scheme are shown in the middle and bottom sub-plots, respectively. The plot was generated assuming a LOS channel model with K-factor of 20 dB and RMS delay spread of 50nsec. We see that the circular delay diversity scheme introduces additional frequency domain nulls whereas the PSF code does not introduce additional frequency domain nulls. As a result, the PSF code performs better in LOS channel conditions.

Figure 6 shows the Packet Error Rate (PER) vs SNR performance of our PSF code with $B = 4$ and spatial multiplexing rates of $M = \{1, 2, 3\}$, for a system with $M_T = 3, M_R = 4$ antennas. We assume the transmitted packets are 1000 bytes long, and channel model is D-NLOS model with 50nsec RMS delay spread. For this simulation, we assume the base 802.11a code rate of 54 Mbps with 64QAM modulation and rate 3/4 FEC. For $M = \{1, 2, 3\}$, the effective data rates are $54 \times 1 = 54$ Mbps, $54 \times 2 = 108$ Mbps and $54 \times 3 = 162$ Mbps respectively. For comparison, we also plot the legacy 802.11a 54 Mbps performance for $M_R = M_T = 1$. We see that our PSF code scales with spatial multiplexing rate and offer performance and throughput improvement, compared to the 1×1 legacy system. For example, at the 1% PER, a 3×4 MIMO-OFDM system employing the PSF code can achieve 162 Mbps while the legacy system can only achieve 54 Mbps, given the same SNR. Also, at the 1% PER, the 3×4 MIMO-OFDM system in the 54 Mbps mode gives about 12 dB improvement over the legacy 54 Mbps mode. Another advantage of our PSF coding scheme is that it can support intermediate multiplexing rates ($1 < M < min(M_R, M_T)$), as evidenced by the 108 Mbps performance plot. This enables the modem to vary data rate with finer granularity when compared to a modem that cannot support intermediate spatial multiplexing rates.

Figure 7 demonstrates the backwards compatibility of the $M = 1$ PSF code employed on a transmitter with $M_T = 2$ antennas, with the legacy 802.11a receiver. We assume the base 802.11a code rate of 24 Mbps assuming 16 QAM modulation and rate 1/2 FEC, and a D-NLOS channel model with 50nsec RMS delay spread. We see that the legacy receiver with no modifications to it's decoding algorithms can decode the $M = 1$ PSF code from a transmitter with $M_T = 2$ transmit antennas. Furthermore, we see that the PSF-coded system has improved diversity order when compared to the legacy 1×1 system. This is because the PSF code transforms

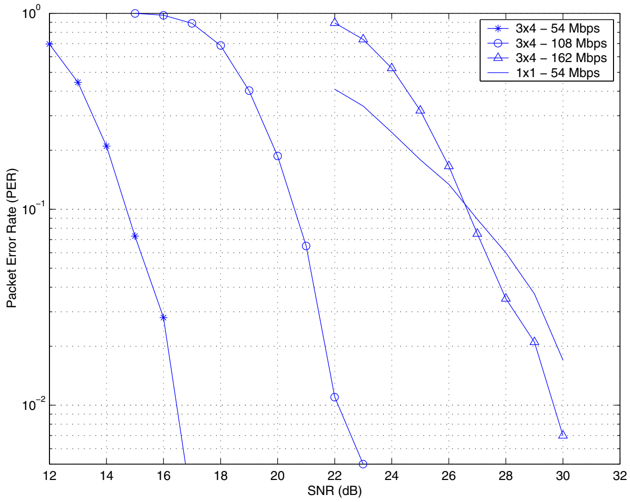

Fig. 6. Performance of PSF code for $M = \{1, 2, 3\}, M_T = 3, M_R = 4$

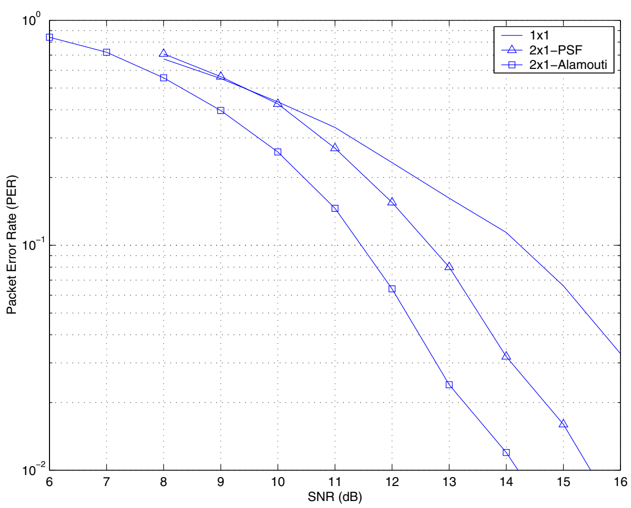

Fig. 7. Backwards compatibility of the PSF code and performance comparison with Alamouti code

the spatial selectivity across the transmit antennas to improved frequency selectivity, that can be exploited by the rate 1/2 FEC code to obtain frequency diversity gains. We also compare the performance of the PSF code with the Alamouti scheme with $M_T = 2, M_R = 1$. The Alamouti code (implemented in the time-domain) provides the best performance since it is the optimal code for $M_T = 2$. However, the drawback is that the Alamouti scheme is not backwards compatible with legacy receivers, unlike the PSF scheme. Finally, we note that the PSF encoding scheme in general may lose performance compared to the optimal encoding scheme for a given number of antennas and spatial multiplexing rate. However, the PSF scheme has several advantages from an implementation point of view that make it attractive for MIMO-OFDM.

V. Conclusion

In this paper, we presented a linear space-frequency coding scheme based on permutation matrices, that is scalable with number of antennas and spatial-multiplexing (SM) rate, as well as easy to implement. Furthermore, the design allows for low cost power amplifier (PA) design. For SM rate 1, our space-frequency coding scheme is backwards compatible with legacy receivers using only 1 receive antenna. These features of our so-called *Permuted Space Frequency Codes* enable seamless deployment of multiple antenna OFDM transceivers in existing wireless networks with legacy single antenna transceivers. Simulation results were presented for MIMO-OFDM wireless LAN system assuming the base architecture of 802.11a system, and 802.11n channel model.

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