

Adaptive Spectrally-Precoded OFDM Under Flat Fading

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Abstract—Adaptive transmission technique adopting spectrally-precoded orthogonal frequency-division multiplexing (SP-OFDM) signals is studied on the flat fading channel. By jointly adapting the precoding order for spectral precoder, transmission power, and the component modulation for OFDM, adaptive SP-OFDM is shown to outperform conventional adaptive OFDM significantly in average spectral efficiency.

Keywords—Adaptive modulation, orthogonal frequency division multiplexing, spectral precoding, average spectral efficiency.

I. INTRODUCTION

Adaptive modulation adopting orthogonal frequency division multiplexing (OFDM) [1]-[4] is an efficient technique to significantly improve the system throughput in wireless applications. Adaptive OFDM adapts the transmission parameters, such as power and constellation size, according to the estimated subband or subcarrier channel status in a way that a preassigned bit error rate (BER) performance is instantaneously sustained. Typically, adaptive OFDM systems shape the transmitted signal with the rectangular pulse which facilitates the efficient implementation by discrete Fourier transform (DFT) and the insertion of guard intervals to counteract channel dispersion, but results in large power spectral sidelobes [5]-[8]. In order to prevent excessive interference to adjacent channels, redundant guard bands [4] and transmitter filtering [9] are commonly employed in practical applications at the cost of spectral and power efficiencies.

Recently, several spectral precoding schemes [5]-[8] are proposed to improve the spectral compactness of the rectangularly-pulsed OFDM signals. These spectrally precoded OFDM (SP-OFDM) signals, can not only facilitate the standard DFT-based OFDM implementation and the insertion of cyclic prefix (CP) or zero padding (ZP) in the guard intervals to counteract channel dispersion, but also provide very small power spectral sidelobes. Among the spectral precoding schemes developed in [5]-[8], $\mathcal{G}_L$ -coded SP-OFDM with ZP or CP in [5] provides extremely high spectral compactness and thus outperforms the spectrally

uncoded OFDM signal significantly in spectral efficiency, especially when the precoding order L is large. As a tradeoff, the $\mathcal{G}_L$ -coded SP-OFDM schemes degrade, however, from spectrally uncoded OFDM on the additive white Gaussian noise (AWGN) channel, the delay-spread channels with small amplitude fluctuation, and lightly faded channels, and the degradation gets worse as L is larger [5]-[8]. Since the precoding order L can be increased to provide much higher spectral efficiency, $\mathcal{G}_L$ -coded SP-OFDM signal is suited to adaptive transmission with constant transmission power [10]-[11]. In this paper, an adaptive $\mathcal{G}_L$ -coded SP-OFDM system is proposed to jointly adapt precoding order, modulation, and transmission power according to the channel status in a way that much higher system throughput can be achieved.

The remainder of this paper is organized as follows. Section II describes the $\mathcal{G}_L$ -coded adaptive SP-OFDM (ASP-OFDM) system model. Section III presents the adaptation schemes with constant or variable transmission power for $\mathcal{G}_L$ -coded ASP-OFDM. Numerical results are presented for $\mathcal{G}_L$ -coded ASP-OFDM schemes with constant or variable power adaptation in Section IV. Section V concludes the paper.

II. SYSTEM MODEL

Figure 1 depicts the $2N$ -dimensional ($2N$ -D) ASP-OFDM system. Denote $T = T_d + T_g$ as the length of the block time interval for transmitting a block of M data symbols, where T_d and T_g are the lengths for the useful data subinterval and the guard subinterval, respectively, and $N > M$ is assumed. Specifically, in a nominal block interval, the source generates a block of M independent and memoryless K -ary quadrature amplitude modulation (QAM) symbols $\{D_m | m \in \mathcal{Z}_M\}$ with zero mean and unit variance, where $\mathcal{Z}_I \triangleq \{0, 1, \dots, I-1\}$.

These M symbols $\{D_m | m \in \mathcal{Z}_M\}$ are then spectrally precoded to yield a block of N transmitted complex symbols $\{B_n | n \in \mathcal{Z}_N\}$ as

$$B_n = \sum_{m=0}^{M-1} G_{n,m} D_m, \quad n \in \mathcal{Z}_N. \quad (1)$$

With $L \triangleq N - M$ denoting the precoding order, the complex valued precoding coefficient $G_{n,m}$ [5] is defined by

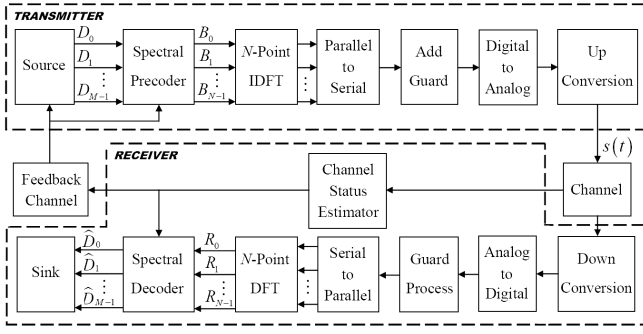


Figure 1. The ASP-OFDM system model

$$G_{n,m} = \begin{cases} \left(\frac{2L}{L}\right)^{-\frac{1}{2}} \binom{L}{n-m} \zeta_n, & 0 \leq n-m \leq L \\ 0, & \text{otherwise} \end{cases} \quad (2)$$

with $\zeta_n = \exp\{-j\omega_d T_g n/2\}$ for SP-OFDM with CP and $\zeta_n = 1$ for SP-OFDM with ZP, and $\omega_d \triangleq 2\pi/T_d$.

By monitoring the channel status, i.e., the instantaneously received signal-to-noise power ratio (SNR) γ fed back from the receiver, both source and spectral precoder operate adaptively so that a $\mathcal{G}_L$ -coded SP-OFDM component modulation with appropriately chosen parameters L (the precoding order) and K (the modulation size) is used for transmission. Note that the SP-OFDM component modulation signal with larger L or K can provide higher spectral efficiency when the required inband power fraction is large. This property is to be explored in the design of ASP-OFDM systems.

After the precoder, $\{B_n | n \in \mathcal{Z}_N\}$ are modulated in parallel with N subcarriers, uniformly spaced by ω_d , and multiplexed to form the transmitted SP-OFDM signal [5]–[8] as

$$s(t) = \sqrt{S} \operatorname{Re} \left\{ \sum_{n=0}^{N-1} B_n \exp \{j(\omega_0 + n\omega_d)t\} \right\} \quad (3)$$

in the useful data subinterval $-Td/2 \leq t < Td/2$, where S represents the instantaneous power and ω_0 is the reference frequency with $\omega_0 \gg \omega_d$. In the guard subinterval $-Tg - Td/2 \leq t < -Td/2$, the transmitted signal is given by $s(t) = s(t+Td)$ for SP-OFDM with CP and $s(t) = 0$ for SP-OFDM with ZP. Note that $s(t)$ can be generated from $\{B_n | n \in \mathcal{Z}_N\}$ by a standard inverse discrete Fourier transform (IDFT)-based OFDM process as in Figure 1.

In this paper, ASP-OFDM with a small N (the number of subcarriers) is considered for design and performance in flat fading channels. Although not treated explicitly, the results can also be applied to ASP-OFDM with a large number of subcarriers in the frequency-selective fading channel wherein the total band is modeled as a partition of subbands in a way that all subcarriers in a subband suffer the same fade. This subband flat fading model is plausible when the coherence bandwidth of the frequency-selective fading channel exceeds the bandwidth of a subband and commonly assumed for the study in adaptive OFDM systems [1]–[3].

Assume that the received signal is perfectly synchronized in amplitude, phase, symbol timing, and frequency by the DFT based receiver in Figure 1. The received waveform is down converted, digitally sampled, and then manipulated block by block by the guard process. The guard process extracts useful samples for data detection by overlapping and adding the guard samples to the useful data samples for SP-OFDM with ZP and by simply removing the guard samples for SP-OFDM with CP [8]. After the guard process, the output samples are transformed by DFT to yield $\{R_n | n \in \mathcal{Z}_N\}$ where R_n represents the received complex symbol on the n -th subcarrier. Assuming that the guard subinterval is long enough to reject the inter-symbol interference, R_n can be modeled by $R_n = \sqrt{SH}B_n + W_n$ where H is the flat fading amplitude that is assumed to remain constant within the nominal block interval and $\{W_n | n \in \mathcal{Z}_N\}$ are the identically-distributed circularly symmetric complex AWGN samples with mean zero and variance $E\{|W_n|^2\} = N_0$. For SP-OFDM with CP, W_n 's are mutually independent. For SP-OFDM with ZP and $T_g = J T_d/N$, W_n 's are correlated with $E\{W_n W_m^*\} = N_0 \sigma_{n-m}$ for $n \neq m$ where $\sigma_k = \frac{1}{N+J}$

$\cdot \sum_{l=0}^{J-1} \exp\{-j\frac{2\pi}{N}lk\}$ with J being the number of zero-padded samples in each guard subinterval. By monitoring $\{R_n | n \in \mathcal{Z}_N\}$, the channel status estimator estimates the fading amplitude H and the received SNR γ , and passes the estimate for H to the spectral decoder and the estimate for γ to the transmitter through a feedback channel, respectively. It is assumed that H and γ are perfectly measured and the estimate for γ is fed back to the transmitter without error nor delay. Using the estimate for H , the symbol decisions $\{\hat{D}_m = \hat{D}_m^{(I)} + j\hat{D}_m^{(Q)} | m \in \mathcal{Z}_M\}$ can be obtained by the spectral decoder through applying the block decoding rule [5] on $\{R_n = R_n^{(I)} + jR_n^{(Q)} | n \in \mathcal{Z}_N\}$ for $\mathcal{G}_L$ -coded SP-OFDM as

$$\begin{aligned} \{\hat{D}_m^{(x)} | m \in \mathcal{Z}_M\} &= \arg \left\{ \min_{\{D_m^{(x)} | m \in \mathcal{Z}_M\}} \sum_{n=0}^{N-1} |R_n^{(x)}| \right. \\ &\quad \left. - \sqrt{SH} \sum_{m=\max\{0, n-L\}}^{\min\{M-1, n\}} \binom{2L}{L}^{-\frac{1}{2}} \right. \\ &\quad \left. \cdot \binom{L}{n-m} D_m^{(x)} \right\} \end{aligned} \quad (4)$$

with $D_m = D_m^{(I)} + jD_m^{(Q)}$ and $x = I$ and Q .

An invertible expression of the BER in AWGN is required for the design and analysis of adaptive modulation systems over flat fading channels [12]. Motivated by the approximate BER expression in [12], the following invertible exponential function is adopted here to approximate the BER of $\mathcal{G}_L$ -coded SP-OFDM in AWGN as

$$BER(\gamma) \approx c_0 \exp \left\{ \frac{-c_1 \gamma}{2c_2 L K^{c_3 + c_4 L}} \right\} \quad (5)$$

where the coefficients $\{c_i | i \in \mathcal{Z}_5\}$ are acquired by curvefitting the simulated BER of the block decoding rule in (4). Note that the expression in (5) with $c_2 = c_4 = 0$ reduces to the BER approximate expression for K -ary QAM as in [12]. As an

example, Table 1 lists $\{c_i | i \in \mathcal{Z}_5\}$ of the SP-OFDM schemes when $N = 8$, where the approximate BER's are tight to the simulation results within 1 dB for $10^{-5} < BER(\gamma) < 10^{-2}$, $L \in \{1, 2, 3\}$, and $K \in \{4, 16, 64\}$.

Table 1. The Parameters for the Approximate Invertible BER of SP-OFDM with $N=8$.

Modulation Scheme	c_0	c_1	c_2	c_3	c_4
$\mathcal{G}_L$ -coded SP-OFDM with CP	0.3	3.1	0.27	0.65	0.07
$\mathcal{G}_L$ -coded SP-OFDM with ZP	0.29	3.0	0.33	0.65	0.05

III. ADAPTATION SCHEME

We consider both constant and variable power adaptation schemes with the instantaneous BER constrained by a target BER, i.e., BER_T , as developed in [12]. The constant-power adaptation scheme fixes transmission power while the variable power adaptation scheme adopts the waterfilling approach to optimize power allocation according to the received SNR. In general, the constant-power adaptation scheme is simpler to implement than the variable-power adaptation scheme, at the cost of reduced average spectral efficiency [12].

For adapting an SP-OFDM component modulation, the received SNR region is partitioned into $V + 1$ subregions in a way that the v -th SP-OFDM component modulation λ_v is chosen for transmission when $\gamma \in [\Upsilon_{\lambda_v}, \Upsilon_{\lambda_{v+1}})$ with $v \in \mathcal{Z}_V$ where Υ_{λ_v} denotes the v -th modulation-switching threshold, and that no signal is transmitted when $\gamma < \Upsilon_{\lambda_0}$. By default, $\Upsilon_{\lambda_V} \triangleq \infty$. Specifically, the adaptation criterion maximizes the average spectral efficiency η as

$$\eta = \sum_{v=0}^{V-1} \Phi_{\lambda_v} [F(\Upsilon_{\lambda_{v+1}}) - F(\Upsilon_{\lambda_v})] \text{ bits/sec/Hz} \quad (6)$$

under the constraints that the average transmission power satisfies

$$\int_{\Upsilon_{\lambda_0}}^{\infty} S f(\gamma) d\gamma = 1 \quad (7)$$

for the constant-power adaptation scheme and

$$\sum_{v=0}^{V-1} \int_{\Upsilon_{\lambda_v}}^{\Upsilon_{\lambda_{v+1}}} S(\gamma) f(\gamma) d\gamma \leq 1 \quad (8)$$

for the variable-power adaptation scheme with $S(\gamma)$ representing the variable power function, and that the instantaneous BER is not larger than BER_T as

$$BER_{\lambda_v}(\gamma) \leq BER_T, v \in \mathcal{Z}_V. \quad (9)$$

Here, $BER_{\lambda_v}(\gamma)$ is the BER function when adopting λ_v . Φ_{λ_v} is the spectral efficiency defined by the inverse of the normalized bandwidth $(CT_b)^{-1}$ required to capture a preassigned fraction $\wp$ of total power within a bandwidth C , with T_b being the bit time. $f(\gamma)$ and $F(\gamma)$ are the probability density function and the cumulative distribution function of γ , respectively. The optimum solutions of $\{\Upsilon_{\lambda_v} | v \in \mathcal{Z}_V\}$ and the instantaneous transmission power to

the above adaptation criterions have been obtained in [12] as

$$\Upsilon_{\lambda_v} = \Gamma_{\lambda_v} [1 - F(\Gamma_{\lambda_0})] \text{ for } v \in \mathcal{Z}_V \quad (10)$$

and

$$S = \int_{\Upsilon_{\lambda_0}}^{\infty} f(\gamma) d\gamma \quad (11)$$

for the constant-power adaptation scheme, where Υ_{λ_v} is the power efficiency defined by the SNR value required to achieve BER_T when adopting λ_v , and

$$\Upsilon_{\lambda_0} = \frac{\Gamma_{\lambda_0}}{\Phi_{\lambda_0}} \epsilon, \Upsilon_{\lambda_v} = \frac{\Gamma_{\lambda_v} - \Gamma_{\lambda_{v-1}}}{\Phi_{\lambda_v} - \Phi_{\lambda_{v-1}}} \epsilon, v \in \{1, 2, \dots, V-1\} \quad (12)$$

and

$$S(\gamma) = \frac{\Gamma_{\lambda_v}}{\gamma} \text{ for } \Upsilon_{\lambda_v} \leq \gamma < \Upsilon_{\lambda_{v+1}} \text{ and } v \in \mathcal{Z}_V \quad (13)$$

for the variable-power adaptation scheme, where the Lagrangian ϵ can be solved by satisfying the equality in the power constraint in (8). Note that Υ_{λ_v} can be approximated through inverting (5) as $\Gamma_{\lambda_v} \approx 2^{c_2 L} K^{c_3 + c_4 L} \ln(c_0 / BER_T) / c_1$.

Denote $\Lambda \triangleq \{\lambda_v | v \in \mathcal{Z}_V\}$ as the set of 2N-D SP-OFDM component modulations which is obtained from the modulation set Θ containing all 2N-D SP-OFDM modulations with $L \in \{1, 2, \dots, L_{\max}\}$ and $K \in \{4, 16, \dots, K_{\max}\}$. Given the preassigned system design values BER_T and $\wp$, the following rule is adopted to choose the component modulations in Λ from Θ : First, the most spectrally efficient modulation in Θ is chosen as λ_{V-1} . Second, when $\lambda_{V-1}, \lambda_{V-2}, \dots, \lambda_{V-m+1}$ are determined, λ_{V-m} is chosen as the modulation giving the highest spectral efficiency among all the modulations in Θ which are more power-efficient than λ_{V-m+1} . The second step is recursively repeated for $m \in \{2, 3, \dots, V\}$ or until Θ is exhausted. For the variable-power adaptation schemes, the set of component modulations chosen from this rule should be further pruned to satisfy the logical threshold relation $\Upsilon_{\lambda_v} < \Upsilon_{\lambda_{v+1}}$ for $v \in \mathcal{Z}_V$ [13]. As an example, Table 2 lists Λ 's obtained from Θ with $L_{\max} = 3$ and $K_{\max} = 64$ for both constant and variable - power adaptation schemes when $N = 8$, $T_g = T_d/8$, $BER_T = 10^{-3}$, and $\wp = 0.99$ and 0.9999. When $\wp = 0.99$, all chosen component modulations are shown to have $L = 1$ in that $\mathcal{G}_1$ -coded SP-OFDM scheme provides sufficiently high spectral efficiency while yielding the largest possible power efficiency among all $\mathcal{G}_L$ -coded SP-OFDM schemes. However, such a performance phenomenon vanishes when a higher value $\wp = 0.9999$ is required.

IV. NUMERICAL RESULTS

Fig. 2 compares the average spectral efficiency characteristics among ASP-OFDM and adaptive OFDM under flat Rayleigh fading when $N = 8$, $T_g = T_d/8$, $BER_T = 10^{-3}$, and $\wp = 0.99$ and 0.9999, where the modulation sets Λ 's are chosen from Θ with $L_{\max} = 3$ and $K_{\max} = 64$ for ASP-OFDM as listed in Table 2, and from the set of OFDM

Table 2. The Chosen Λ out of Θ with $L_{\text{MAX}} = 3$ and $K_{\text{MAX}} = 64$ When $N = 8$, $T_g = T_d/8$, and $BER_T = 10^{-3}$. Here, λ_v is Represented by Corresponding Modulation Parameter Pair (K, L) .

$\wp$		0.99						0.9999					
Adaptation Scheme	v	$\mathcal{G}_L$ -coded ASP-OFDM with ZP			$\mathcal{G}_L$ -coded ASP-OFDM with CP			$\mathcal{G}_L$ -coded ASP-OFDM with ZP			$\mathcal{G}_L$ -coded ASP-OFDM with CP		
		λ_v	Υ_{λ_v}	Φ_{λ_v}	λ_v	Υ_{λ_v}	Φ_{λ_v}	λ_v	Υ_{λ_v}	Φ_{λ_v}	λ_v	Υ_{λ_v}	Φ_{λ_v}
Constant Power Adaptation	0	(4, 1)	7.24	1.39	(4, 1)	6.94	1.35	(4, 1)	6.94	0.78	(4, 1)	6.94	0.05
	1	(16, 1)	11.46	2.77	(16, 1)	11.09	2.71	(4, 2)	8.17	0.89	(4, 2)	8.17	0.55
	2	(64, 1)	15.67	4.16	(64, 1)	15.25	4.06	(16, 1)	11.09	1.56	(4, 3)	9.41	0.67
	3	-	-	-	-	-	-	(16, 2)	12.75	1.78	(16, 2)	12.75	1.10
	4	-	-	-	-	-	-	(64, 1)	15.25	2.33	(16, 3)	14.40	1.33
	5	-	-	-	-	-	-	(64, 2)	17.32	2.66	(64, 2)	17.32	1.66
Variable Power Adaptation	0	(4, 1)	6.75	1.39	(4, 1)	6.25	1.35	(4, 1)	6.75	0.78	(4, 2)	8.00	0.55
	1	(16, 1)	8.89	2.77	(16, 1)	8.30	2.71	(16, 1)	8.80	1.56	(4, 3)	10.01	0.67
	2	(64, 1)	13.11	4.16	(64, 1)	12.45	4.06	(64, 1)	12.95	2.33	(16, 2)	10.88	1.10
	3	-	-	-	-	-	-	(64, 2)	16.67	2.66	(16, 3)	13.07	1.33
	4	-	-	-	-	-	-	-	-	-	(64, 2)	16.37	1.66
	5	-	-	-	-	-	-	-	-	-	(64, 3)	17.01	2.00

modulations using K -ary QAM with $K \in \{4, 16, 64\}$ as subcarrier modulation for adaptive OFDM, respectively. In this case, H is Rayleigh with $F(\Upsilon_{\lambda_v}) = 1 - \exp(-\Upsilon_{\lambda_v}/\bar{\gamma})$ where $\bar{\gamma}$ denotes the average SNR. Note that the spectral efficiency in Table 2 can be numerically computed through the power spectral density expressions of the SP-OFDM signals provided in [5]. Several remarks can be drawn from Fig. 2. First, all ASP-OFDM signals significantly outperform the adaptive OFDM signals in average spectral efficiency, especially when a higher $\wp$ (inband power fraction) is required. This is primarily due to the fact that the adopted $\mathcal{G}_L$ -coded SP-OFDM signals exhibit much higher spectral compactness than the uncoded OFDM signals. Second, the variable-power adaptation schemes give uniformly better average spectral efficiency than the corresponding constant-power adaptation schemes for both ASP-OFDM [10] and adaptive OFDM, as was analytically justified in [12]. Finally, the adaptive OFDM schemes with CP perform better in average spectral efficiency than the corresponding adaptive OFDM schemes with ZP. However, a reverse trend is observed for ASP-OFDM. This phenomenon results from the fact that the $\mathcal{G}_L$ -coded SP-OFDM signals with ZP can provide the power spectral sidelobes decaying asymptotically as f^{-2L-2} with f being the frequency in Hz, and thus outperform in spectral efficiency the $\mathcal{G}_L$ -coded SP-OFDM signals with CP which yield the power spectral sidelobes decaying asymptotically as f^{-2} [5].

V. CONCLUSIONS

The $\mathcal{G}_L$ -coded ASP-OFDM system which adapts the spectral precoder order L and the component modulation size K in conjunction with constant or variable power adaptation are studied for the flat fading channels. Because the $\mathcal{G}_L$ -coded SP-OFDM signals exhibit extremely high spectral compactness, $\mathcal{G}_L$ -coded ASP-OFDM is found to significantly outperform the conventional adaptive OFDM in average spectral efficiency

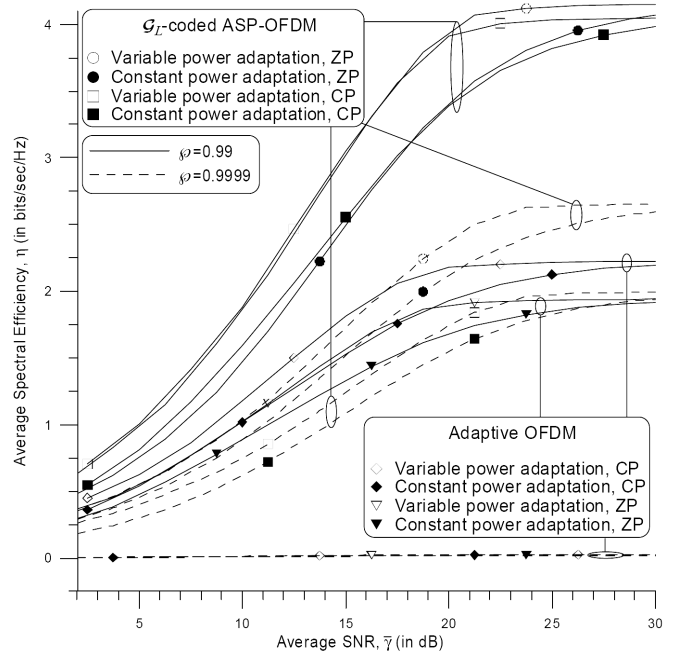


Figure 2. Average spectral efficiency characteristics for ASP-OFDM and adaptive OFDM

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