Proportionally Sampled Multicarrier Modulation

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Abstract—A new proportionally sampled multicarrier modulation (PSMCM) system with optimum receiver is presented in this paper. By adopting different sampling rates at transmitter and receiver, the PSMCM system can achieve multipath diversity with simple linear operations. It has the same spectral efficiency as conventional orthogonal frequency division multiplexing (OFDM) system, while doesn’t require complex precoder/decoder as used by most other multicarrier systems for the fulfillment of multipath diversity. In addition, it is insensitive to timing phase offset caused by the phase difference between transmitter clock and receiver clock, thus eliminates the need of precise sample level synchronization as required by conventional OFDM system. Exact bit error rate expression of the new system is derived. Both analytical and simulation results show that PSMCM system outperforms OFDM system with the same spectral efficiency by as much as 10 dB.

I. INTRODUCTION

Multicarrier modulation (MCM) is emerging as a leading data transmission technique for broadband digital communication systems [1], [2]. By modulating a large number of narrow band data streams over closely spaced subcarriers, MCM is less sensitive to intersymbol interference (ISI) compared to single carrier systems with the same spectral efficiency. The most widely adopted MCM scheme is orthogonal frequency division multiplexing (OFDM), which modulates parallel data streams onto orthogonal subcarriers. The orthogonality among subcarriers eliminates intercarrier interference (ICI). In addition, OFDM system can completely remove ISI by inserting a short cyclic prefix between OFDM symbols.

The ICI- and ISI-free configuration of OFDM is obtained at the cost of multipath diversity, or frequency diversity, introduced by the time dispersion of frequency selective fading [3] – [5]. Given the fact that narrow band data streams are transmitted over orthogonal subcarriers, conventional uncoded OFDM system cannot achieve multipath diversity [3]. Various linear precoders with suboptimum decoders have been proposed to introduce controlled interferences among subcarriers, thus to achieve partial or full multipath diversity [3] – [5]. Optimum maximum likelihood decoding or maximum a posteriori decoding is practically impossible for such systems due to the prohibitive complexity incurred by the precoders.

The performance of OFDM system also suffers from timing phase offset, which is caused by the phase difference between the transmitter clock and receiver clock. The presence of timing phase offset will introduce ICI [6]. It has been shown in [7] that receiver oversampling can effectively remove the negative effects caused by timing phase offset. The result in [7] was developed based on the assumption of perfect ISI cancellation in single carrier systems. Capitalizing on the robustness of MCM against ISI, we expect similar results in oversampled MCM system.

This paper presents a proportionally sampled MCM (PSMCM) scheme operating at quasi-static frequency selective fading. The PSMCM system doesn’t suffer from any aforementioned limitations of convention MCM or OFDM systems. By employing different sampling rates at transmitter and receiver, the PSMCM system can achieve multipath diversity with simple linear operation in both time domain and frequency domain. It doesn’t require complex precoding/decoding, still enjoys ICI- and ISI-free operation, and has the same spectral efficiency as conventional OFDM systems. In addition, a higher sampling rate at the receiver removes the negative effects of timing phase offset as predicted by [7]. An optimum diversity receiver for the PSMCM system is developed by investigating the time-frequency properties of frequency selective fading and noise samples. Exact bit error rate (BER) expression is derived for the PSMCM with optimum receiver. Both analytical and simulation results show that the new PSMCM system significantly outperforms conventional OFDM system without sacrificing spectral efficiency.

II. SYSTEM STRUCTURE

The time domain model and frequency domain operations of the PSMCM system is described in this section. The block diagram of the proposed PSMCM system with N subcarriers is shown in Fig. 1.

A. Time Domain System Model

At the transmitter, N data symbols, \( s = [s(0), \ldots, s(N - 1)]^T \in \mathbb{C}^{N \times 1} \), with \( A^T \) denoting matrix transpose, are modulated onto N orthogonal subcarriers through N basis functions

\[
\phi_n(t) = \begin{cases} 
\frac{1}{\sqrt{N}} e^{j2\pi nF_0 t}, & 0 \leq t \leq T_0, \\
0, & \text{otherwise}
\end{cases}
\]

where \( F_0 = \frac{1}{F_0} \) is the subcarrier space, and \( T_0 = \frac{1}{F_0} \) is the duration of one symbol. The multicarrier modulated signal, \( x(t) = \sum_{n=0}^{N-1} s(n)\phi_n(t) \), are sampled at the transmitter at a sampling period of \( T_1 = \frac{T_0}{N} \), which yields N time domain samples, \( x = [x(0), x(1), \ldots, x(N-1)]^T \), per symbol period. The resulting time domain samples can be represented as \( x = F^H s \), where \( A^H \) represents matrix Hermitian, and \( F_N \in \mathbb{C}^{N \times N} \) is the normalized N-point discrete Fourier
transform (DFT) matrix with the \((m, n)\)-th element being
\[
\frac{1}{\sqrt{N}} \exp \left\{ -j 2\pi \frac{(m-1)(n-1)}{N} \right\}.
\]
Cyclic prefix is inserted in the time domain to avoid ISI. The prefixed time domain samples are passed through a transmit filter, \(p_1(t)\). At the receiver, the received signal is passed through a receive filter, \(p_2(t)\).

Define the composite impulse response (CIR) of the frequency selective fading channel as
\[
g(t) = p_1(t) \otimes c(t) \otimes p_2(t),
\]
where \(c(t)\) is the impulse response of the physical frequency selective fading, and \(\otimes\) represents convolution. It is assumed that the channel is quasi-static, i.e., it keeps constant during one symbol duration, and varies from symbol to symbol.

With the definition of \(g(t)\), the signal at the output of the receive filter can be represented by
\[
y(t) = \sum_n x(n)g(t - nT_1) + z(t),
\]
where \(z(t) = v(t) \otimes p_2(t)\) with \(v(t)\) being the additive white Gaussian noise (AWGN) with variance \(N_0\).

The output of the receive filter is sampled at the instant \(t = nT_2 + \tau_0\), where \(\tau_0 \in [-T_2/2, T_2/2]\) is the receiver timing phase offset, and the sampling period \(T_2\) is proportional to \(T_1\) as \(T_1/T_2 = u\), with the sampling factor \(u\) being an integer. The sampled output of the receive filter can be represented as
\[
y_T(n) = \sum_{l=0}^{uL-1} g_T(l)\hat{x}(n - l) + z_T(n),
\]
where \(g_T(n) = y(nT_2 + \tau_0)\) and \(z_T(n) = z(nT_2 + \tau_0)\) are the \(T_2\)-spaced samples of received signals and noise, respectively, \(uL\) is the channel length of the discrete-time CIR \(g_T(l) = g(lT_2 + \tau_0)\), and \(\hat{x}(n)\) is the oversampled version of the time domain data samples defined as
\[
\hat{x}(n) = \begin{cases} 
x(n/u), & \text{if } n/u \text{ is integer}, \\
0, & \text{otherwise}.
\end{cases}
\]

Eqn. (3) is an equivalent discrete-time representation of the multicarrier communication system in the time domain. The performance of the system depends on the properties of the channel vector, \(g_T = [g_T(0), g_T(1), \ldots, g_T(uL-1), 0, \ldots, 0]^T \in \mathbb{C}^{uL \times 1}\), and noise vector, \(z_T = [z_T(0), \ldots, z_T(uN-1)]^T \in \mathbb{C}^{uN \times 1}\).

The statistical properties of \(g_T\) and \(z_T\) are summarized in the following two Lemmas [8].

**Lemma 1:** For system undergoing wide sense stationary uncorrelated scattering (WSSUS) frequency selective fading, \(g_T\) is zero-mean complex Gaussian distributed with covariance matrix \(R_{g_T}\). The \((m,n)\)-th element of \(R_{g_T}\) is
\[
\left( R_{g_T} \right)_{m,n} = \begin{cases} 
\rho(m-1, n-1), & 1 \leq m, n \leq uL, \\
0, & \text{otherwise}.
\end{cases}
\]
The coefficient, \(\rho(m, n) = \mathbb{E} \left[ g_T(m) g_T^*(n) \right]\), represents correlation between channel taps, and it can be calculated by
\[
\rho(m, n) = \int_{-\infty}^{+\infty} R_{p_1p_2} (mT_2 + \tau_0 - \mu) \times \overline{R_{p_1p_2}} (nT_2 + \tau_0 - \mu) G(\mu) d\mu,
\]
where \(\mathbb{E}(\cdot)\) represents mathematical expectation, \(R_{p_1p_2}(t) = p_1(t) \otimes p_2(t)\), \(\overline{\cdot}\) denotes complex conjugate, and \(G(\mu)\) is the normalized channel power delay profile.

**Lemma 2:** The time domain noise vector, \(z_T\), is zero-mean complex Gaussian distributed with covariance matrix \(R_{z_T} = N_0 R_p\), where \(N_0\) is the variance of AWGN. The \((m,n)\)-th element of the matrix \(R_p \in \mathbb{C}^{uN \times uN}\) is \((R_p)_{m,n} = R_{p,p}(m-n)\), where \(R_{p,p}(n) = \int_{-\infty}^{+\infty} p_2(nT_2 + \tau) \overline{p_2}(\tau) d\tau\) is the autocorrelation function of the receive filter.

The results from Lemmas 1 and 2 state that the equivalent discrete-time system has correlated channel taps and operates in the presence of colored noise. The channel tap correlation and the correlation among noise samples are introduced by the time dispersion of the filters, \(p_1(t)\) and \(p_2(t)\).

**B. Frequency Domain Signal Processing**

The adoption of higher sample rate at receiver results in \(uN\) samples after the removal of cyclic prefix. Before converting the time domain samples to frequency domain, we demultiplex the \(uN\) time domain samples into \(u\) length-N streams as shown in Fig. 1. The samples in the \(v\)-th stream are \(y_T^{(v)} = [y_T^{(v)}(v), y_T^{(v)}(v+u), \ldots, y_T^{(v)}(v+(N-1)u)]^T \in \mathbb{C}^{N \times 1}\). From (3), we can write the system equation related to the \(v\)-th sample stream, \(y_T^{(v)}\), as
\[
y_T^{(v)} = G_T^{(v)} x + z_T^{(v)}, \quad \text{for } v = 0, 1, \ldots, u - 1.
\]
where \(z_T^{(v)} = [z_T^{(v)}(v), z_T^{(v)}(v+u), \ldots, z_T^{(v)}(v+(N-1)u)]^T \in \mathbb{C}^{N \times 1}\) is the noise sample vector, and \(G_T^{(v)} = [g_T^{(v)}, (g_T^{(v)})_1, \ldots, (g_T^{(v)})_{N-1}] \in \mathbb{C}^{N \times uN}\) is a Toeplitz matrix, with \(g_T^{(v)} = [g_T(v), g_T(v+u), \ldots, g_T(v+(uL-1))]^T\).
and \((g_T^{(v)})_k\) obtained by cyclic shifting \(g_T^{(v)}\) downwards by \(k\) elements.

Performing \(N\)-point DFT over \(y_F^{(v)}\), we have
\[
y_F^{(v)} = G_F^{(v)}s + z_F^{(v)}, \quad \text{for } v = 0, 1, \ldots, u - 1.
\]  
(8)

where \(y_F^{(v)} = F_Ny_T^{(v)}\) and \(z_F^{(v)} = F_Nz_T^{(v)}\). The frequency domain channel matrix, \(G_F^{(v)} = F_NG_T^{(v)}F_N^H\), is a diagonal matrix with the \(k\)-th diagonal element being
\[
g_F^{(v)}(k) = \sum_{l=0}^{L-1} g_T(v + lu)e^{-j2\pi (k-1)/N}.
\]  
(9)

Demultiplexing the time domain samples results in \(u\) parallel sub-systems as described in (8). We propose to perform linear combination over the \(u\) sub-systems in the frequency domain, such that the original \(N\)-subcarrier system can be converted to an equivalent system with \(uN\) subcarriers.

**Proposition 1:** Define a group of diagonal matrices, \(D_m = \text{diag}\left[e^{-j2\pi u_k/mN}, e^{-j2\pi (u_k+1)/mN}, \ldots, e^{-j2\pi (u_k+u)/mN}\right]\), for \(m = 0, \ldots, u - 1\), then the frequency domain sample vectors, \(\{y_F^{(v)}\}_{v=0}^{u-1}\), can be linearly combined in the following manner
\[
r^{(m)} = \sum_{v=0}^{u-1} D_m^{(v)}y_F^{(v)} = H^{(m)}s + w^{(m)}, \quad \text{for } m = 0, \ldots, u - 1.
\]  
(10)

such that the diagonal channel matrix, \(H^{(m)} = \sum_{v=0}^{u-1} D_m^{(v)}G_F^{(v)}\), and the noise vector, \(w^{(m)} = \sum_{v=0}^{u-1} D_m^{(v)}z_F^{(v)}\), contain elements obtained from \(uN\)-point DFT of their respective time domain samples, up to a scaling factor, as
\[
h^{(m)}(k) = \frac{1}{\sqrt{uN}} \sum_{n=0}^{uL-1} g_T(n)e^{-j2\pi (mN+k-1)/uN},
\]  
(11a)

\[
w^{(m)}(k) = \frac{1}{\sqrt{uN}} \sum_{n=0}^{uL-1} z_T(n)e^{-j2\pi (mN+k-1)/uN},
\]  
(11b)

where \(h^{(m)}(k)\) is the \(k\)-th diagonal element of \(H^{(m)}\), and \(w^{(m)}(k)\) is the \(k\)-th element of \(w^{(m)}\).

**Proof:** Since both \(D_m\) and \(G_F^{(v)}\) are diagonal matrices, \(H^{(m)}\) is also diagonal. From (10), the \(k\)-th diagonal element of \(H^{(m)}\) is
\[
h^{(m)}(k) = \sum_{v=0}^{u-1} e^{-j2\pi (mN+k-1)v/uN} g_F^{(v)}(k)
\]  
(12)

where \(g_F^{(v)}(k)\) is the \(k\)-th diagonal element of \(G_F^{(v)}\) as defined in (9). Combining (9), (12), and the identity that
\[
e^{-j2\pi (k-1)/N} = e^{-j2\pi (mN+k-1)/uN},
\]  
(13)

we have
\[
h^{(m)}(k) = \sum_{v=0}^{u-1} \sum_{l=0}^{L-1} g_T(v + lu)e^{-j2\pi (mN+k-1)(v+lu)/uN},
\]  
(14)

and (11a) immediately follows.

Similarly, from (8) and (10), the \(k\)-th element of \(w^{(m)}\) can be expressed as
\[
w^{(m)}(k) = \frac{1}{\sqrt{uN}} \sum_{v=0}^{uL-1} \sum_{l=0}^{L-1} e^{-j2\pi (mN+k-1)(v+lu)/uN} z_T(v + lu)e^{-j2\pi (k-1)/N}.
\]  
(15)

Combining (13) with (15) leads to (11b).

In (11), \(h^{(m)}(k)\) is represented as the \((mN+k)\)-th subcarrier of the \(uN\)-point DFT of the oversampled discrete-time CIR \(g_T\), for \(m = 0, \ldots, u\) and \(k = 0, \ldots, N - 1\). The \(uN\)-point DFT is achieved through \(u\)-times oversampling in the time domain and the combination of the oversampled data streams in the frequency domain. From (10), each data symbol, \(s(k)\), is equivalently transmitted over \(u\) subcarriers, \(\{h^{(m)}(k)\}_{m=0}^{u-1}\). Stacking all the \(u\) subcarriers related to \(s(k)\) leads to an alternative system model
\[
r_k = h_k \cdot s(k) + w_k, \quad \text{for } k = 0, \ldots, N - 1,
\]  
(16)

where \(r_k = [r^{(0)}(k), r^{(1)}(k), \ldots, r^{(u-1)}(k)]^T\), \(h_k = [h^{(0)}(k), h^{(1)}(k), \ldots, h^{(u-1)}(k)]^T\), and \(w_k = [w^{(0)}(k), w^{(1)}(k), \ldots, w^{(u-1)}(k)]^T\).

In (16), the multicarrier system is equivalently represented as a single input multiple output (SIMO) system with each data symbol, \(s(k)\), transmitted over \(u\) subcarriers. Therefore, frequency diversity is achieved by means of oversampling and simple linear operations performed at the receiver.

### III. OPTIMUM DIVERSITY RECEIVER

An optimum diversity receiver for the equivalent SIMO system described in (16) is presented in this section to collect the multipath diversity enabled by the new PSMCM structure.

**A. Design of Optimum Diversity Receiver**

The optimum diversity receiver is designed by investigating the statistical properties of the channel vector, \(h_k\), and the noise vector, \(w_k\). From (11) and (16), we have
\[
h_k = \sqrt{uN} \cdot F_{uN}g_T, \quad \text{(17a)}
\]
\[
w_k = \sqrt{u} \cdot F_{uN}^Hz_T, \quad \text{(17b)}
\]

where \(F_{uN}^H \in \mathbb{C}^{u \times uN}\) is a sub-matrix of the normalized \(uN\)-point DFT matrix, and the \((m, n)\)-th element of \(F_{uN}^{(k)}\) is
\[
\frac{1}{\sqrt{uN}} \exp(-j2\pi ((m-1)/u)(n-1)/N)\cdot \text{Exp}(\cdot).
\]

Based on (17) and Lemmas 1 and 2, both \(h_k\) and \(w_k\) are zero mean complex Gaussian distributed with their respective covariance matrices, \(R_h = \mathbb{E}(w_kw_k^H)\), and \(R_w = \mathbb{E}(w_kw_k^H)\), given by
\[
R_h^{(k)} = uN \cdot F_{uN}R_{g_T}F_{uN}^H, \quad \text{(18a)}
\]
\[
R_w^{(k)} = uN \cdot F_{uN}R_{p}F_{uN}^H, \quad \text{(18b)}
\]

The results in (18) indicate that the SIMO system of (16) has correlated channel taps and operates in colored noise. Since the frequency domain noise samples are mutually correlated, the covariance matrix, \(R_w^{(k)}\), might be rank deficient. To facilitate analysis, define the pseudo-inverse of \(R_w^{(k)}\) as
\[
\Phi_k = V_k \Omega_k^{-1}V_k^H \in \mathbb{C}^{u \times u},
\]  
(19)
with
\[
V_k = [v_{k1}, v_{k2}, \ldots, v_{knw}] \in C^{n \times u_w},
\]
\[
\Omega_k = \text{diag} [\omega_{k1}, \omega_{k2}, \ldots, \omega_{knw}] \in C^{u_w \times u_w},
\]
where \( u_w \) is the number of non-zero eigenvalues of \( R_h^{(k)} \), \( \Omega_k \) is a diagonal matrix with its diagonal elements being the non-zero eigenvalues of \( R_h^{(k)} \), and the corresponding eigenvectors, \( \{v_i\}_{i=1}^{u_w} \), formulate the reduced eigenvector matrix \( V_k \).

With the pseudo-inverse matrix defined in (19), the optimum diversity receiver for the equivalent SIMO system is derived in this subsection.

**Proposition 2:** For an SIMO system described in (16) with colored Gaussian noise, the optimum decision rule is
\[
\hat{s}(k) = \arg \max_{s(k) \in S} |\beta_k - q_k s(k)|^2
\]
where \( S \) is the modulation alphabet set, \( q_k = h_k^H \Phi_k h_k \), and \( \beta_k \) is the decision variable defined as \( \beta_k = h_k^H \Phi_k r_k \).

**Proof:** From (19), define the noise whitening matrix, \( B_k = \Omega_k^{\frac{1}{2}} V_k^H \). Applying \( B_k \) to both sides of (16) leads to
\[
\tilde{r}_k = \tilde{h}_k \cdot s(k) + \tilde{w}_k,
\]
where \( \tilde{r}_k = B_k r_k \), \( \tilde{h}_k = B_k h_k \), and \( \tilde{w}_k = B_k w_k \). The covariance matrix of the noise vector, \( \tilde{w}_k \), is \( R_{\tilde{w}}^{(k)} = B_k R_w B_k^H = I_{u_w} \) where \( I_{u_w} \) is a \( u_w \times u_w \) identity matrix. Therefore, (23) is an equivalent system with white noise. Applying the optimum maximal ratio combining (MRC) to (23), i.e., \( \beta_k = h_k^H \tilde{r}_k \), leads to the optimum decision rule in (22).

**B. Performance Analysis**

The theoretical BER of the PSMCM system with the optimum diversity receiver is derived in this subsection.

Combining (16) and Proposition 2 yields an alternative representation of the decision variable as
\[
\beta_k = q_k \cdot s(k) + h_k^H \Phi_k w_k,
\]
Thus, the signal to noise ratio (SNR) at the output of the optimum receiver is \( \gamma_k = q_k \gamma_0 \), where \( \gamma_0 = E_s / N_0 \) is the SNR without fading, with \( E_s \) being the energy of one symbol. Based on the SNR \( \gamma_k \), the conditional error probability (CEP) of the \( k \)-th data stream for system with binary-phase-shift-keying (BPSK) can be written as [7]
\[
P(E_k|q_k) = \frac{1}{\pi} \int_0^{\pi/2} \exp \left( -\frac{\gamma_0 \cdot q_k}{\sin^2 \theta} \right) d\theta.
\]

The unconditional error probability can be obtained by evaluating the characteristic function (CHF) of \( q_k \), which is a quadratic form of the zero mean complex Gaussian vector \( h_k \). The CHF of \( q_k \) is [9]
\[
\Psi_q(j\omega) = \mathbb{E} \left( e^{j\omega q_k} \right) = \left[ \det \left( I_{u_w} - j\omega R_h^{(k)} \Phi_k \right) \right]^{-1}.
\]

Combining (25) and (26), we have the unconditional error probability, \( P(E_k) = \mathbb{E} [P(E_k|q_k)] \), as
\[
P(E_k) = \frac{1}{\pi} \int_0^{\pi/2} \prod_{i=1}^{u_k} \left( 1 + \frac{\gamma_0 \cdot \lambda_i}{\sin^2 \theta} \right)^{-1} d\theta,
\]
where \( u_k \) is the rank of the product matrix, \( R_h^{(k)} \Phi_k \), with \( \lambda_i \) being the corresponding non-zero eigenvalues. The closed-form expression can be obtained by partial fraction expansion of the integrand of (27) as in [7], and the details are omitted here for brevity. With \( P(E_k) \) given in (27), the bit error rate (BER) of the PSMCM system can then be calculated as
\[
P(E) = \frac{1}{N} \sum_{k=0}^{N-1} P(E_k).
\]

It’s apparent from (27) that the multipath diversity order of the PSMCM is equal to the rank of the product matrix, \( R_h^{(k)} \Phi_k \), which is in turn determined by the time domain covariance matrices, \( R_{\gamma_T} \) and \( R_p \).

In addition, it has been shown in [7] that the matrices \( R_{\gamma_T}^{(v)} \) and \( R_p \) are independent of timing phase offset, provided that there is no spectrum aliasing at the receiver. Since oversampling can effectively remove spectrum aliasing at the receiver [7], the performance of the PSMCM system, which is completely determined by \( R_{\gamma_T}^{(v)} \) and \( R_p \), is independent of timing phase offset.

**IV. NUMERICAL RESULTS**

Numerical results are provided in this section to demonstrate the performance of the PSMCM system with optimum diversity receiver. Square root raised cosine (SRRC) filter with roll-off factor \( \alpha = 1 \) is used as both transmit filter and receive filter. The typical urban (TU) channel profile is used to model the frequency selective fading. The number of subcarriers is \( N = 64 \). The sample period at the transmitter is \( T_1 = 3.09 \mu s \).

Fig. 2 compares the BER of the PSMCM system with conventional OFDM system. Two times oversampling \( (u = 2) \) is employed by the PSMCM system. The receiver timing phase offset is set to \( \tau_0 = 0 \). Considerable performance improvement of the PSMCM system over OFDM system is observed from both simulation and analytical results. At the BER level of \( 3 \times 10^{-4} \), the PSMCM system with optimum receiver outperforms OFDM system by 10 dB. The performance improvement is mainly contributed by the introduction of multipath diversity. The performance of a PSMCM system with conventional MRC receiver is also shown in the figure. The MRC receiver directly performs diversity combining over the SIMO system of (16) without considering the correlation among noise samples. The employment of non-optimum MRC receiver results in a performance loss of approximately 2.5 dB compared to the PSMCM with the new optimum receiver. In addition, perfect agreement are observed between the results obtained from simulation and theoretical analysis.

The effects of timing phase offset on the performance of OFDM and PSMCM systems are illustrated Fig. 3. The performance of conventional OFDM system degrades with the introduction of timing phase offset. On the other hand, the performance of PSMCM system with two times oversampling is independent of timing phase offset. Two times oversampling completely removes spectrum aliasing of a system with at most 100% excessive bandwidth. Therefore, for PSMCM system,
only coarse synchronization on symbol level is required, as against the precise sample level synchronization required by conventional OFDM system.

Fig. 4 shows the performance of PSMCM systems with \( u = 1 \) (conventional OFDM), \( u = 2 \), and \( u = 8 \), respectively. The timing phase offset is set to \( \tau_0 = 0 \). As expected, increasing \( u \) leads to better performance due to the improvement in multipath diversity. As \( u \) becomes large, the performance improvement gradually diminishes. At the BER level of \( 3 \times 10^{-4} \), increasing \( u \) from 1 to 2 leads to a performance improvement of 10 dB. The performance improvement becomes 2 dB when \( u \) is increased from 2 to 8.

V. CONCLUSIONS

A new PSMCM structure with optimum receiver was presented in this paper. Performing oversampling in time domain and linear signal processing in frequency domain lead to an equivalent system, where each data stream is transmitted over multiple subcarriers. As a result, multipath diversity is achieved without sacrificing spectral efficiency. To collect the multipath diversity enabled by the new structure, an optimum receiver was developed by investigating the statistical properties of frequency selective fading and noise samples in the frequency domain. Theoretical analysis leads to exact BER expression of the PSMCM structure with optimum receiver. Both theoretical analysis and simulation results demonstrated that the PSMCM system outperforms conventional OFDM system by as much as 10 dB. In addition, the PSMCM system is insensitive to timing phase offset.

REFERENCES