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Performance evaluation of CO-OFDM systems based on electrical constant-envelope signals

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ABSTRACT

The influence of the electrical phase modulation index h in the performance of constant-envelope orthogonal frequency division multiplexing (CE-OFDM) in coherent detection optical systems is treated analytically and its range of validity examined by simulations. A compromise between h and subcarrier mapping is identified according to differences in sensitivity related to non-linearities inserted by the optical modulator. It is shown that the proposed scheme outperforms conventional coherent detection OFDM systems, which is strongly dependent on both phase and optical modulation indexes.

1. Introduction

The resilience towards fiber dispersion is the main attractive feature of orthogonal frequency division multiplexing (OFDM) signal processing in optical communications systems [1–3]. However, large peak-to-average power ratio (PAPR) is one of the main drawbacks in both direct and coherent detection optical orthogonal frequency division multiplexing (CO-OFDM) systems, as it not only limits the resolution of digital-to-analog converters and RF amplifiers, but also reduces the tolerance to the non-linearities introduced by optical modulators and optical fibers [1,4].

To reduce PAPR, an effective technique based on the transmission of constant-envelope (CE-OFDM) signals is a suitable solution to the aforementioned problems [5]. Proposed in [6] and experimentally demonstrated in [7,8] to combat nonlinear degradations in direct-detection optical OFDM systems, this electrical phase modulation based technique is, for the first time, introduced in coherent detection systems.

Unlike the approaches evaluated in [9,10], the intermediate electrical CE-OFDM signals of this solution are used to modulate the continuous wave (CW) laser source, employing a conventional one-branch Mach–Zehnder modulator (MZM).

Section 2 of this letter develops an analytical expression from which the influence of the electrical phase modulation index h may be derived. Its validity is examined by means of numerical simulations in Section 3. A tradeoff between the index h and the subcarrier mapping is clarified from the simulation results described in this Section. It is shown that, in CO-OFDM systems that employ such multicarrier constant-envelope signals, high subcarrier modulation levels requires reduction in the aforementioned index due to non-linearities inserted by the MZM. Section 4 concludes the paper.

2. Theoretical model

Fig. 1 shows the CE-OFDM considered in this analysis. It is a modulation format where an electrical carrier is phase modulated by OFDM waveforms, which results in envelope signals with PAPR =3 dB [6]. Therefore, a single OFDM signal with mean power given by σ_s^2 is written as

$$x(t) = C\sum_{k=0}^{N_s-1}\Re[X(k)]\cos\left(2\pi\Delta_f kt\right) - \Im[X(k)]\sin\left(2\pi\Delta_f kt\right) \quad (1)$$

with $\{X(k)\}_{k=1}^{N_s-1}$ the $M-$QAM data symbols, $T=\frac{N}{F_s}$ the OFDM symbol time, $N=2N_s+2$ the fast-Fourier transform length, F_s the sampling rate and C a constant, modulates the phase of a carrier resulting in a bandpass signal given by

$$c(t) = A\cos[2\pi f_0 t + \phi(t)] = A\cos[2\pi f_0 t + \theta_n + 2\pi h C_N x(t)], \quad (2)$$

where A is the signal amplitude, f_0 is the carrier frequency, $\phi(t)$is the phase signal during the n-th signal interval $nT \leqslant t < (n+1)T$,

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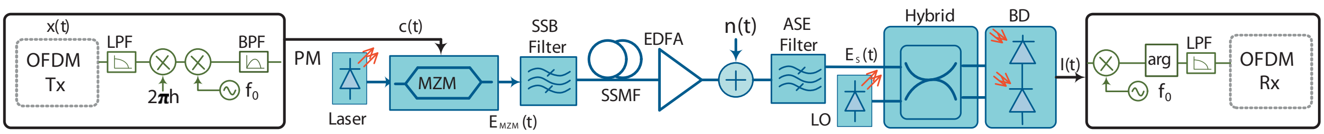

Fig. 1. CO-CE-OFDM system model. LPF: Low Pass Filter; BPF: Band Pass Filter; PM: Phase modulation; MZM: Mach–Zehnder Modulator; SSB: Single-Sideband; EDFA: Erbium-doped Fiber Amplifiers; SSMF: Standard Single-Mode Fiber; ASE: Amplified Spontaneous Emission; LO: Local Oscillator; BD: Balanced Photoreceiver.

θ_n is a memory term designed to make the modulation phase continuous (when $\theta_n = 0$ the modulation is memoryless), h is referred to as the electrical phase modulation index, and C_N a constant used to normalize the variance of the message signal $x(t)$ [6]. Continuous phase modulation (CPM) can be achieved by introducing memory in Eq. (2) making $\theta_n \neq 0$ [11]. Although the constant-envelope property and the compact signal spectrum of this class of CPM signaling, the memory increases the receiver complexity [5,12,13].

The bandwidth of the electrical signal $c(t)$ is usefully expressed as $B = \max(2\pi h, 1)B_W$, which is an RMS (root-mean-square) bandwidth lower bounded by the conventional OFDM signal bandwidth B_W [5]. A conservative bandwidth is estimated with this expression for small values of $2\pi h$. For $2\pi h > 1$ it measures at most 92% of the total bandwidth [14].

Assuming $\theta_n = 0$ and $C_N = 1$, for the sake of simplicity, the CE-OFDM signal becomes $c(t) = A\cos[w_0 t + s(t)]$, for $w_0 = 2\pi f_0$ and $s(t) = 2\pi h x(t)$. Therefore, the optical field at the output of a conventional single arm Mach-Zender modulator (MZM) characterized by its switching voltage V_π and biased by V_{bias} is a bandpass signal centered at frequency $w_c = 2\pi f_c$ given by

$$E_{MZM}(t) = \cos\left[\frac{\pi c(t)}{2V_\pi} - \frac{\pi V_{bias}}{2V_\pi}\right] \cdot \sqrt{2P}\cos(w_c t), \tag{3}$$

where P is the power of the CW laser signal fed into the MZM optical input. Biased at its null point ($V_{bias} = -V_\pi$) the MZM output signal can be rewritten as

$$\begin{aligned} E_{MZM}(t) &= \sin\left[\frac{\pi c(t)}{2V_\pi}\right] \cdot \sqrt{2P}\cos(w_c t) \\ &= \sqrt{2P_c} \cdot \sin[F(t)] \cdot \cos(w_c t), \end{aligned} \tag{4}$$

for $F(t) = \frac{\pi c(t)}{2V_\pi}$. After optical pre-amplification with gain G, the input signal into the balanced receiver can be written as

$$\begin{aligned} E_s(t) = {}& \sqrt{2GP} \cdot \sin[F(t)] \cdot \cos(w_c t) + n_i(t)\cos(w_c t) + n_q(t) \\ & \times \sin(w_c t), \end{aligned} \tag{5}$$

for $n_i(t)$ and $n_q(t)$ the in-phase and quadrature component of the amplifier noise with power spectrum density given by $N_{ASE}/2$ and variance $\sigma_n^2 = N_{ASE} \cdot B_o$, with B_o the bandwidth of the optical filter at its output.

If we describe the optical field of the local oscillator (LO) light source as $E_{LO} = A_{LO} e^{jw_{LO}t + j\theta_{LO}}$, then $\mathbb{R}\{E_{LO}\} = A_{LO}\cos(w_{LO}t)$ for E_{LO} its amplitude, w_{LO} its frequency and θ_{LO} the laser phase noise, which is zeroed for the sake of simplicity. Because of the intrinsic heterodyne detection of the CO-CE-OFDM system model, we assume that electrical synchronous demodulation is used to estimate the phase noise in order to benefit the phase demodulation [15].

The beating of the optical signal and the LO at the 180° optical hybrid results in two photocurrents outputs from the balanced photodiodes (PD) defined as

$$\begin{aligned} I_1(t) &= R|\mathbb{R}\{\frac{1}{\sqrt{2}}(E_S + E_{LO})\}|^2 \\ I_2(t) &= R|\mathbb{R}\{\frac{1}{\sqrt{2}}(E_S - E_{LO})\}|^2, \end{aligned} \tag{6}$$

with R denoting the PD responsivity and $\mathbb{R}$ the real number set. Hence, the $I_2(t)$ current can be expressed as

$$\begin{aligned} I_2(t) = R\Big\{& 2GP\sin^2(F(t))\cos^2(w_c t) + \\ & + n_i^2\cos^2(w_c t) + n_q^2\sin^2(w_c t) + \\ & + A_{LO}\cos^2(w_{LO}t) + \\ & + 2[\sqrt{2GP}\sin(F(t))\cos^2(w_c t)n_i + \\ & + \sqrt{2GP}\sin(F(t))\cos(w_c t)\sin(w_c t) + \\ & + n_i n_q \sin(w_c t)\cos(w_c t) - \\ & - A_{LO}\cos(w_{LO}t)(\sqrt{2GP}\sin(F(t))\cos(w_c t) + \\ & + n_i\cos(w_c t) + n_q\sin(w_c t))]\Big\}. \end{aligned} \tag{7}$$

Using the same mathematical description in $I_1(t)$ current, the resulting photon electrical current can be expressed as follows

$$\begin{aligned} I(t) &= I_1 - I_2 \\ &= 4RA_{LO}\cos(w_{LO}t)[\sqrt{2GP}\sin(F(t))\cos(w_c t) + \\ &\quad + n_i\cos(w_c t) + n_q\sin(w_c)]. \end{aligned} \tag{8}$$

Assuming that the LO is frequency locked to the received signal ($w_{LO} = w_c$), the current at the balanced receiver output can be written as

$$\begin{aligned} I(t) &= 4RA_{LO}[\sqrt{2GP}\sin(F(t))\cos^2(w_c t) + n_i\cos^2(w_c t) \\ &\quad + n_q\sin(w_c t)\cos(w_c t)] \\ &= 2RA_{LO}[\sqrt{2GP}\sin(F(t))(1 + \cos(2w_c t)) + n_i(1 + \cos(2w_c t)) \\ &\quad + n_q\sin(2w_c t)]. \end{aligned} \tag{9}$$

After low-pass filtering to filter out the high-order items, this signal is approximated to

$$\begin{aligned} I(t) &= 2RA_{LO}\left[\sqrt{2GP}\sin(F(t)) + n_i\right] \\ &\approx 2RA_{LO}\left[\sqrt{2GP}\left(\frac{\pi c(t)}{2V_\pi}\right) + n_i\right], \end{aligned} \tag{10}$$

considering the first-order Taylor's expansion that allows $\sin[F(t)] = \sin\left(\frac{\pi c(t)}{2V_\pi}\right) \approx \frac{\pi c(t)}{2V_\pi}$. Substituting $F(t)$ and Eq. (2) into (10), the received bandpass signal becomes

$$\begin{aligned} I(t) &= \frac{RA_{LO}\sqrt{2GP}\pi}{V_\pi} \cdot A\cos[w_0 t + s(t)] + 2RA_{LO} \cdot n_i \\ &= K_1\cos[w_0 t + s(t)] + K_2 n_i, \end{aligned} \tag{11}$$

for $K_1 = \frac{RA_{LO}\sqrt{2GP}\pi A}{V_\pi}$ and $K_2 = 2RA_{LO}$. After multiplying this passaband signal by the electrical carrier $\cos\left(w_0 t + \frac{\pi}{2}\right)$, for downconversion, the expression

$$\begin{aligned} I(t) \approx {}& \frac{K_1}{2}\left[\cos\left(2w_0 t + \frac{\pi}{2} + s(t)\right) + \cos\left(\frac{\pi}{2} - s(t)\right)\right] + K_2 n_i \\ & \times \cos\left(w_0 t + \frac{\pi}{2}\right) \end{aligned} \tag{12}$$

is obtained. Neglecting the high frequency contributions and considering $\sin(s(t)) \approx s(t) = 2\pi h x(t)$ after first-order Taylor expansion this photocurrent becomes