Symbol-Aided Approach to Nonlinear Distortion Cancellation for Self-heterodyne OFDM Direct Conversion Receivers

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***Abstract*-We propose a symbol-aided approach to nonlinear distortion cancellation and data demodulation in self-heterodyne direct conversion (SHDC) receivers for discrete Fourier transform (DFT)-based orthogonal frequency division multiplexing (OFDM) systems. In order to cancel the intermodulation distortion generated at the square-law device output in the receiver, the SHDC transmitter generates an OFDM symbol followed by the next symbols having a phase difference of 90 degree. By using this approach, quadrature modulation can be supported with direct conversion. The advantages of the proposed technique are demonstrated through the computer simulation.**

I. INTRODUCTION

The self-heterodyne direct conversion (SHDC) with system predistortion technique was proposed in [1]. Baseband predistortion in the transmitter is conducted to compensate the intermodulation distortion generated at the square-law device output of a receiver. The SHDC scheme has the advantage that simple, without local oscillators, low-cost receivers can be built that are completely immune to any phase noise or frequency offset [2]. This is of importance to OFDM systems that are much sensitive to frequency and phase errors [3]. Furthermore the 17 GHz for an OFDM-based broadband wireless LAN (WLAN) system, frequencies from 10 to 66 GHz for IEEE 802.16a [4], and millimeter wave bands are explored to meet the strong demand for high transmission rate in indoor and short-range outdoor wireless applications.

However, according to the system proposed in [1], phase modulation can not be supported with self-heterodyne direct conversion systems.

This paper introduces an alternative direct conversion technique for complex-valued data demodulation and nonlinear distortion cancellation in SHDC schemes. In the proposed scheme, we first generate a DFT-based OFDM symbol, and then construct the next two symbols having phase differences of 90 degree. The received signals through the square-law device are processed to cancel out the intermodulation distortion without system predistortion and then recover the complex-valued transmit signal. One realizes that there is a reduction in throughput when compared to the SHDC with system predistortion.

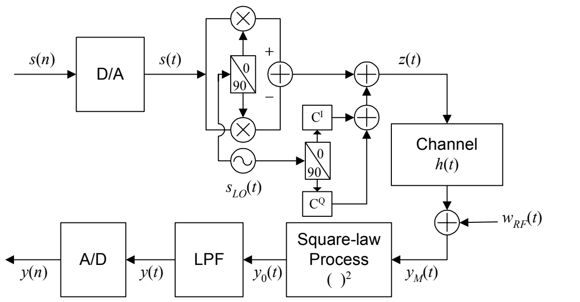

Figure 1. Self-heterodyne system employing direct conversion architecture and complex transmission.

II. SIGNAL MODEL

Shown in Fig. 1 is the SHDC system model considered for complex-valued signal. The transmitter sends out a local carrier as well as a modulated RF signal. Here, $s(t)$ is the complex signal to be transmitted; $s_{LO}(t)$ is the transmit local carrier, i.e., $exp(j\omega_c t)$, where ω_c is the carrier frequency; $z(t) = [s(t) + C]s_{LO}(t)$ is the signal expression at the output of the transmitter, where C is a local carrier gain factor; $h(t)$ is the channel impulse response; and $w_{RF}(t)$ is the zero mean additive white noise in the passband. The passband signal $y_M(t)$ entering the square-law device is represented as $y_M(t) = x(t)\ exp(j\omega_c t) + x^*(t)\ exp(-j\omega_c t) + w_{RF}(t)$, where $x(t)$ is an equivalent lowpass signal at the receiver antenna, which is given by the convolution of $h(t)$ and $s(t) + C$. If the type of fading exhibits by a flat fading channel, $y_M(t)$ can be rewritten by [5]

$$\begin{aligned} y_M(t) = & 2\{ h^I[s^I(t) + C^I] - h^Q[s^Q(t) + C^Q]\}\cos\omega_c t \\ & - 2\{ h^I[s^Q(t) + C^Q] + h^Q[s^I(t) + C^I]\}\sin\omega_c t \\ & + 2w^I(t) \cos\omega_c t - 2w^Q(t) \sin\omega_c t, \end{aligned} \tag{1}$$

This passband signal $y_M(t)$ is the input of the square-law device. Lowpass filtering the signal $y_0(t)$, the output of the square-law device, yields the baseband analog signal $y(t)$. If $y(t)$ is sampled with a symbol rate, the discrete signal $y(n)$ is obtained as

$$y(n) = |h|^2\{|s(n)|^2 + |C|^2 + 2s^I(n)\, C^I + 2s^Q(n)\, C^Q\} + v(n), \quad (2)$$

where $v(n)$ denotes the aggregate noise in terms of the additive white Gaussian noise $w(n)$. The right-hand side (RHS) of (2) represents the desired signal components interfered with second-order intermodulation product and dc offset. Note that $y(n)$ is normalized to unit gain for simplicity.

III. Proposed Technique

A. Derivation

In the proposed scheme, we first generate a DFT-based OFDM symbol, and then construct the next symbol having a phase difference of 90 degree, and finally construct the last symbol out of phase to the second symbol. The received signal with respect to those transmit OFDM symbols are given by

$$y_1(n) = |h|^2\{|s(n)|^2 + |C|^2 + 2s^I(n)\, C^I - 2s^Q(n)\, C^Q\} + v_1(n). \quad (3)$$

$$y_2(n) = |h|^2\{|s(n)|^2 + |C|^2 - 2s^I(n)\, C^I + 2s^Q(n)\, C^Q\} + v_2(n). \quad (4)$$

The received signal of the second symbol through the self-mixer is subtracted from that of the first symbol, which removes the nonlinear distortion and produces the quadrature (Q) symbol components. Next, the received signal of the last symbol through the self-mixer is subtracted from that of the first symbol, which removes the nonlinear distortion and produces the in-phase (I) symbol components. A direct calculation using (1), (3), and (4) yields

$$y(n) - y_1(n) = 4\,|h|^2 s^Q(n)\, C^Q + v(n) - v_1(n). \quad (5)$$

$$y(n) - y_2(n) = 4\,|h|^2 s^I(n)\, C^I + v(n) - v_2(n). \quad (6)$$

Base on this process, we are able to cancel out the intermodulation distortion without a system predistortion and demodulate the quadrature-modulated data.

B. Data Demodulation and Interference Cancellation

The M-ary PSK-modulated data stream is partitioned into blocks of length N symbols, and the generic uth block is used to generate a corresponding block of N transmitted data symbols $\mathbf{S}^u = [S_1^u, S_2^u, \ldots, S_{N-1}^u]$, referred to as frequency-domain symbols. As the next step, the IDFT of the sequence $\mathbf{S}^u$ is calculated, yielding the N dimensional block of $\mathbf{s}^u = [s(0)^u, s(1)^u, \ldots, s(N-1)^u]$ transmitted symbols, referred to as time-domain symbols. If one omit the notation u, the samples of OFDM signal on the uth block, $s(n)$ is given by

$$s(n) = \frac{1}{\sqrt{N}} \sum_{k=0}^{N-1} S_k \exp\left(j\frac{2\pi kn}{N}\right), \quad 0 \le n < N, \quad (7)$$

where S_k is a quadrature PSK symbol, k refers to a sub-carrier index and N is the length of IDFT. Note that no data is assigned to S_0, which is used for transmitting the local oscillator (LO) carrier.

Suppose that one OFDM symbol, $s(n)$, constructed by data symbols $\{S_k | k = 1, 2, \ldots, N-1\}$ and other OFDM symbols, $s^*(n)$, $-s^*(n)$, constructed by data symbols $\{S_k^* | k = 1, 2, \ldots, N-1\}$, $\{-S_k^* | k = 1, 2, \ldots, N-1\}$, respectively, are consecutively transmitted. Here, $(\cdot)^*$ denotes complex conjugation. Three consecutive OFDM symbols are finally transmitted.

The right-hand side (RHS) of (5) represents the channel parameters multiplied by the quadrature components of OFDM symbols. And the right-hand side (RHS) of (6) represents the channel parameters multiplied by the in-phase components of OFDM symbols. Accordingly, the distortion-free-signal can be obtained by

$$q_1(n) = \frac{y(n) - y_1(n)}{4C^Q} = |h|^2 s^Q(n) + \frac{v(n) - v_1(n)}{4C^Q}. \quad (8)$$

$$q_2(n) = \frac{y(n) - y_2(n)}{4C^I} = |h|^2 s^I(n) + \frac{v(n) - v_2(n)}{4C^I}. \quad (9)$$

After performing a N-point DFT operation of $\mathbf{q}_{N(m)} = [q_m(0), q_m(1), \ldots, q_m(N-1)]$, $m = 1, 2$, the received signal can be written in vector form as

$$\mathbf{r}^{(2)} = \mathbf{S}^I \mathbf{D}\mathbf{h}_{eff} + \boldsymbol{\xi}^{(2)}, \quad (10)$$

$$\mathbf{r}^{(1)} = \mathbf{S}^Q \mathbf{D}\mathbf{h}_{eff} + \boldsymbol{\xi}^{(1)}, \quad (11)$$

where $\mathbf{r} = [r_0, r_1, \ldots, r_{N-1}]^T$; $\mathbf{h}_{eff} = [h_{eff0}, h_{eff1}, \ldots, h_{eff(L-1)}]^T$, where $h_{eff0} = |h|^2$ and $h_{effl} = 0$; $\mathbf{S} = diag\,\{S_1, S_2, \ldots, S_{N-1}\}$ is a diagonal M-ary PSK symbol matrix; an $(N-1)$-by-L matrix $\mathbf{D}$ with entries $[\mathbf{D}]_{k,l} = \exp(-j2\pi kl/N)$, $0 \le k < N$, $0 \le l \le L-1$; and $\boldsymbol{\xi}^{(\cdot)}$ is an N-dimensional column noise vector, $[\xi_0^{(\cdot)}, \xi_1^{(\cdot)}, \ldots, \xi_{N-1}^{(\cdot)}]^T$, given by

$$\xi_k^{(\cdot)} = \frac{1}{\sqrt{N}} \sum_{n=0}^{N-1} v'(n) \exp\left(-j\frac{2\pi nk}{N}\right), \quad (12)$$

where $v'(n) = [v(n) - v_2(n)]/(2C^I)$ for $\boldsymbol{\xi}^{(2)}$, $[v(n) - v_1(n)]/(2\,C^Q)$ for $\boldsymbol{\xi}^{(1)}$ and $0 \le k < N$. The both in-phase and quadrature components of the desired data symbol can also be acquired by using (8) through (11).

In a preamble mode, the channels are to be estimated by a least-squares (LS) method [6]. If the equation (10) or (11) is rewritten as $\mathbf{r}_p = \mathbf{U}_p \mathbf{h}_{eff} + \boldsymbol{\xi}$, where $\mathbf{U}_p = \mathbf{S}_p \mathbf{D}$, then with known data symbols, the LS estimate of $\mathbf{h}_{eff}$ is given by

$$\hat{\mathbf{h}}_{eff} = \left(\mathbf{U}_p^H \mathbf{U}_p\right)^{-1} \mathbf{U}_p^H \mathbf{r}_p, \quad (13)$$

where $(\cdot)^H$ denotes complex conjugate transpose.

IV. Simulation Results

In the self-heterodyne direct conversion systems, the signal to noise ratio is defined by

$$\mathrm{SNR}_{SHDC} = \frac{E\left[d^2(n)\right]}{E\left[v^2(n)\right]}, \tag{14}$$

where $E[d^2(n)]$ denotes the expected value of the power of $d(n) = 2|h|^2\{s^I(n)C^I + s^Q(n)C^Q\}$ from (2). A computer simulation is conducted to prove the effectiveness of the proposed technique for the cancellation of the receiver nonlinearity and the recovery of the original data symbols in AWGN channels. The modulation formats for the DFT-OFDM subcarriers are QPSK, which is assuming data blocks of N = 128 symbols. The channel impulse response (CIR) estimate is obtained in a preamble mode with ideal timing synchronization. Carrier frequency synchronization is not required due to the typical characteristics of self-heterodyne systems [2].

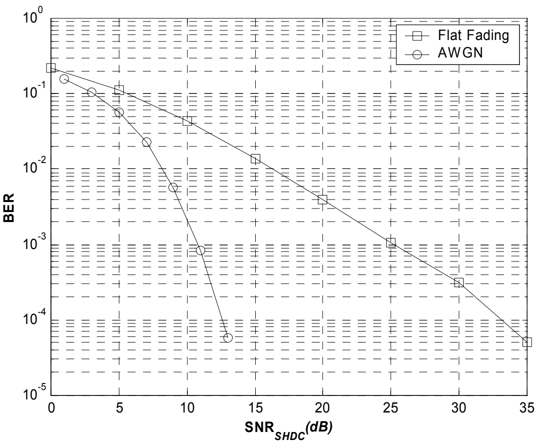

Figure 2. Comparison of BER performance for the case of QPSK self-heterodyne transmission system.

Note that this proposed self heterodyne direct conversion scheme should be considered as an alternative to the conventional transceiver system for wireless channel environments where the retransmission justifies its application. The self heterodyne direct conversion system with proposed scheme could be developed for millimeter-wave application where the transmission link is assumed to be based on a line-of-sight (LOS) path and be operated with a comparatively high-gain antenna.

V. Conclusion

A novel complex data transmission and interference cancellation technique is proposed for self-heterodyne direct conversion systems. Simulation results indicated that the proposed method performed well assuming the I/Q perfect match.

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