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MIMO-UFMC in the Presence of Antenna Mutual Coupling and Phase Noise

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Abstract

In this work, we study the performances of universal filtered multicarrier (UFMC) based multiple-input multiple-output (MIMO) systems in the presence of phase noise and antenna mutual coupling. A phase noise mitigation scheme for MIMO-UFMC systems is presented. The scheme does not require detailed knowledge of the phase noise statistics and can effectively mitigate the phase noise within each UFMC symbol. Moreover, it is shown that, at small antenna separations, the performance of the MIMO-UFMC system taking the (Electromagnitic) mutual coupling effect into account can be better than that when the mutual coupling effect is overlooked.

1. Introduction

Many multicarrier waveforms have been proposed recently for the fifth generation (5G) communications (and beyond 5G) in addition to the conventional orthogonal frequency division multiplexing (OFDM) [1], and will probably be reconsidered for beyond 5G. The universal filtered multicarrier (UFMC) [2, 3] is one of them. By grouping subcarriers into blocks of consecutive subcarriers and filtering each block separately, UFMC can reduce the out-of-band (OOB) leakage (at the cost of increased complexity), while still enjoying the compatibility with the multiple-input multiple-output (MIMO) technique, e.g., [4, 5] (where transmission impairments [6] were omitted). Like OFDM, the subcarrier spacing of the UFMC needs to be sufficiently small in order to use the one-tap channel equalization at each subcarrier; and high modulation order is preferable for high throughput. However, with denser subcarriers or modulation constellation, the phase noise results in not only common phase error (CPE) but also inter-carrier interference (ICI), which can severely degrade the UFMC performance.

Phase noise compensations for OFDM systems have been studied intensively in the literature [7-13], just to name a few. The authors in [7] proposed an iterative ICI correction method which eliminates the phase noise effect by estimating discrete spectral components of the phase noise. A non-iterative phase noise mitigation method based on maximum likelihood estimation was proposed in [8], which was further improved by exploiting the spectral geometry of the phase noise [9]. The phase noise mitigation schemes [7-9] are for single-antenna OFDM systems, whereas phase noise effects on MIMO-OFDM systems have been studied in, e.g., [10-13]. Generally speaking, the phase noise effect is more severe on MIMO-OFDM than that on single-antenna OFDM.

The above mentioned work [7-12] is exclusively limited to the conventional OFDM systems. There is only limited work on phase noise compensation for UFMC systems. A phase noise mitigation scheme for single-antenna UFMC systems was proposed in [14]. The scheme uses scattered pilots to estimate the time-domain phase noise and compensate for it. Nevertheless, to the best knowledge of the authors, no phase noise mitigation methods have been proposed for MIMO-UFMC to date. In this work, we extend the phase noise mitigation scheme for single-antenna UFMC to MIMO-UFMC. Moreover, unlike the previous work on phase noise effects on MIMO-OFDM systems [10-13], where the mutual coupling effect is overlooked, we also investigate the (Electromagnetic) mutual coupling effect on MIMO-UFMC (with/without phase noise mitigation) in this work. On one hand, the mutual coupling (or mutual scattering) tends to make the antenna patterns of different antenna elements more orthogonal and, therefore, reduce the antenna correlation, as compared with the correlation of isolated antenna patterns (i.e., open-circuit correlation) [15-19]. On the other hand, the mutual coupling reduces the total embedded antenna efficiency (including impedance mismatch) [15, 18]. Previously, the mutual coupling effect was usually studied using diversity gain or MIMO capacity [15-18]. In this paper, we investigate the effects of mutual coupling and phase noises on MIMO-UFMC systems (with/without phase noise mitigation) in terms of error rate performance.

Notations: Throughout this paper, boldface lower and upper case letters **a** and **A** represents column vector and matrix, respectively; diag(**a**) denotes a diagonal matrix whose diagonal elements are given by **a**, $\mathrm{Re}\{\cdot\}$ and $\mathrm{Im}\{\cdot\}$ denote the real and imaginary parts of their arguments, respectively; ^{T}, ^{H} and † denote transpose, transpose conjugate (Hermitian), and Moore-Penrose pseudoinvse, respectively.

2. System Model

In order to focus on the effects of mutual coupling and phase noise on MIMO-UFMC, we assume perfect time synchronization and block fading channel that is known at the receiver. For simplicity and without loss of generality, we assume spatial multiplexing with zero-forcing (ZF) MIMO decoder.

2.1 UFMC Modulation and Demodulation

The UFMC modulation (at each transmit antenna) is achieved by first grouping the active subcarriers into B blocks of consecutive subcarriers, and then applying inverse discrete Fourier transform (IDFT) and filtering to each of the blocks. For the sake of implementation simplicity, it is assumed that each block contains the same number (N_0) of subcarriers. The filter bank is obtained by exponential modulation of a prototype filter of length N_v. Mathematically, the UFMC modulation at the jth transmit antenna can be expressed as

$$\mathbf{x}_j = \mathbf{VDs} = \mathbf{Ms}_j, \tag{1}$$

where $\mathbf{M} = \mathbf{VD}$, $\mathbf{V} = [\mathbf{V}_0, \ldots, \mathbf{V}_{B-1}]$, $\mathbf{D} = \mathrm{diag}(\mathbf{D}_0, \ldots, \mathbf{D}_{B-1})$, and $\mathbf{s}_j = \begin{bmatrix} \mathbf{s}_{j,0}^{\mathrm{T}} & \cdots & \mathbf{s}_{j,B-1}^{\mathrm{T}} \end{bmatrix}^{\mathrm{T}}$. $\mathbf{s}_{j,i} = [s_{j,i}(0), \ldots, s_{j,i}(N_0-1)]^{\mathrm{T}}$ ($i = 0, \ldots, B-1$) consist of subcarrier symbols and scattered pilots to be modulated on the N_0 subcarriers in the ith block, $\mathbf{D}_i$ is an $N_f \times N_0$ IDFT matrix corresponding to the subcarriers in the ith block, and $\mathbf{V}_i$ is an $N \times N_f$ Toeplitz matrix whose first column vector is given as $[v_i(0), \ldots, v_i(N_v), 0, \ldots, 0]^{\mathrm{T}}$ where $v_i(n) = v(n)\exp(-j2\pi nk_i/N_f)$ with k_i as the center frequency of the ith block and $v(n)$ being the nth tap of the prototype filter, and the time-domain UFMC symbol length is $N = N_f + N_v - 1$.

Assuming the filter length is smaller than the number of subcarriers of one UFMC symbol (i.e., $N < 2N_f$), the UFMC symbol (at each receive antenna) is demodulated by first performing $2N_f$-point discrete Fourier transform (DFT) of the zero-padded time-domain UFMC symbol and then downsampling the frequency-domain signal by a factor of two. Note that a time-domain preprocessing has to be performed at the receiver prior to the UFMC demodulation (or at the transmitter after UFMC modulation) in order to remove the distortion introduced by the filter bank [2].

2.2 Mutual Coupling

Assuming the MIMO antenna consists of identical and lossless antenna elements, the MIMO channel including the antenna mutual coupling effect is given as [15]

$$\mathbf{H} = 2R_{11}\mathbf{R}_{\mathrm{L}}^{1/2}[\mathbf{Z}_{\mathrm{r}} + \mathbf{Z}_{\mathrm{L}}]^{-1}\mathbf{K}_{\mathrm{r}}^{\mathrm{oc}}\mathbf{H}_{\mathrm{w}}\mathbf{K}_{\mathrm{t}}^{\mathrm{oc}}\mathbf{R}_{\mathrm{t}}^{-1/2} \tag{2}$$

where R_{11} and $\mathbf{R}_{\mathrm{L}}$ are the real parts of the self-impedance of the antenna element Z_{11} and the diagonal load matrix $\mathbf{Z}_{\mathrm{L}}$ respectively, $\mathbf{Z}_{\mathrm{r}}$ is the impedance matrix of the receive antennas, $\mathbf{R}_{\mathrm{t}}$ is the real part of the impedance matrix of the transmit antennas, $\mathbf{H}_{\mathrm{w}}$ is uncorrelated MIMO channel, and $\mathbf{K}_{\mathrm{t}}^{\mathrm{oc}}$ and $\mathbf{K}_{\mathrm{r}}^{\mathrm{oc}}$ are Hermitian square root of the open-circuit correlation matrices at the transmitter and the receiver, respectively. For small (single-mode and lossless) two-port antennas in three-dimensional (3-D) uniform multipath environments, the open-circuit correlation can be approximated by $\mathrm{Re}\{Z_{12}\}/\mathrm{Re}\{Z_{11}\}$ [16], where Z_{12} is the mutual impedance.

2.3 Phase Noise

The phase noise of a free-running oscillator can be modeled by the Wiener process, whose discrete-time expression is

$$\phi(n+1) = \phi(n) + \eta(n) \tag{3}$$

where ϕ is the phase noise and η is a zero-mean Gaussian random variable whose variance is $4\pi\beta/f_s$ with f_s denoting the sampling frequency and β representing the 3-dB bandwidth of the phase noise. The phase noise model (3) is widely used in the literature [7-14]. In addition, we assume the MIMO receiver (transmitter) is equipped with a single oscillator [10-13].

3. Phase Noise Mitigation

The received time-domain signal (in the presence of phase noise) at the nth time sample can be expressed as

$$\mathbf{y}(n) = \exp\left(j\phi(n)\right)\sum_{l=0}^{L-1}\mathbf{H}_l\mathbf{x}(n-l) + \mathbf{w}(n)\,. \tag{4}$$

where $\mathbf{x}(n)$ and $\mathbf{y}(n)$ are $N_t \times 1$ and $N_r \times 1$ transmit and receive vectors, $\mathbf{H}_l$ is an $N_r \times N_t$ MIMO channel impulse response (CIR) matrix of the lth tap that is modeled by (2), and $\mathbf{w}(n)$ is a $N_r \times 1$ vector consisting the additive white Gaussian noises (AWGNs) at the receive antennas. The frequency-domain expression of (4) at the kth subcarrier is given as

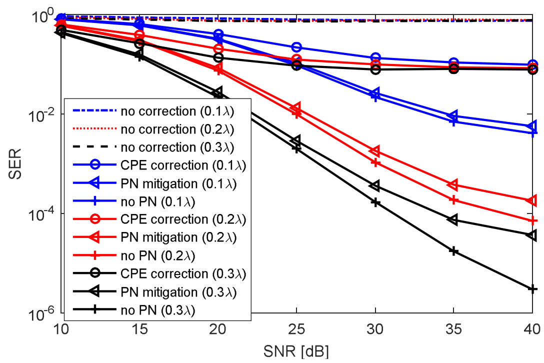
$$\mathbf{r}(k) = J(0)\mathbf{C}(k)\mathbf{s}(k) + \sum_{l=0, l\neq k}^{N-1} J(k-l)\mathbf{C}(l)\mathbf{s}(l) + \mathbf{z}(k) \tag{5}$$

where $\mathbf{s}(k)$ and $\mathbf{r}(k)$ are the frequency-domain $N_t \times 1$ transmit and $N_r \times 1$ received signal vectors, $\mathbf{C}(k)$ is the $N_r \times N_t$ MIMO channel transfer function (CTF) matrix, $\mathbf{z}(k)$ is the frequency-domain $N_r \times 1$ AWGN vector, and $J(k)$ is the spectral component of the phase noise

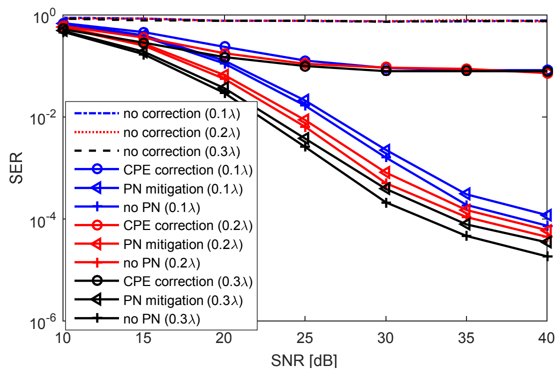
$$J(k) = \frac{1}{\sqrt{N}}\sum_{n=0}^{N-1}\exp\left(j\phi(n)\right)\exp\left(\frac{-j2\pi kn}{N}\right). \tag{6}$$

$J(0)$ in (5) represents the CPE, whereas the second term in the right hand side of (5) represents the ICI. The CPE can be estimated as [10-12]

$$\hat{J}(0) = \frac{\sum_{k\in S_{\mathrm{p}}}\left(\mathbf{C}(k)\mathbf{s}(k)\right)^H \mathbf{r}(k)}{\sum_{k\in S_{\mathrm{p}}}\left(\mathbf{C}(k)\mathbf{s}(k)\right)^H\left(\mathbf{C}(k)\mathbf{s}(k)\right)} \tag{7}$$



(a)



(b)

Figure 1. SER performances of 4×2 MIMO-UFMC systems in multipath fading channel: (a) without mutual coupling; (b) with mutual coupling.

where S_{p} denotes the set of the scattered pilot subcarriers. Once the CPE is estimated, the CPE effect can be readily corrected. By comparison, the ICI effect is much more difficult to mitigate. In order to mitigate both CPE and ICI, we extend the phase noise mitigation scheme for single-antenna UFMC [14] to MIMO-UFMC in this work.

The phase noise compensation at the mth ($m = 1, \ldots, N_{\mathrm{r}}$) receive antenna is given by $\mathbf{\Phi r}_m$, where $\mathbf{r}_m$ is the N×1 time-domain signal vector of one received UFMC symbol and $\mathbf{\Phi} = \mathrm{diag}\left(\left[\exp(-j\phi(0)) \quad \cdots \quad \exp(-j\phi(N-1))\right]^{\mathrm{T}}\right)$. Let $\mathbf{s}_j$ ($j = 1, \ldots, N_{\mathrm{t}}$) be an N_{p}×1 vector consisting of N_{p} scattered pilots from the jth transmit antenna and $\mathbf{c}_{mj}$ be an N_{p}×1 vector consisting of the CTFs at the N_{p} pilot subcarriers from the jth transmit antenna to the mth receive antenna. The relation between $\mathbf{\Phi}$ and scattered pilots is given as

$$\sum_{j=1}^{N_{\mathrm{t}}} \mathrm{diag}(\mathbf{h}_{mj})\mathbf{s}_j \approx \mathbf{W\Phi r}_m \tag{8}$$

where $\mathbf{W}$ consists of N_{p} rows (that correspond to the scattered pilots) of $\mathbf{M}^{\dagger}$. Since $\mathbf{\Phi}$ is a diagonal matrix, $\mathbf{\Phi r}_m = \mathrm{diag}(\mathbf{r}_m)\boldsymbol{\alpha}$, where $\boldsymbol{\alpha}$ denotes an N×1 column vector consisting of the diagonal elements of $\mathbf{\Phi}$. We model the time-varying phase noise as a piecewise linear function, i.e., $\boldsymbol{\alpha}=\mathbf{Tz}$ where $\mathbf{z}$ is a q×1 column vector consisting of q unknowns to be estimated, and $\mathbf{T}$ is the linear interpolation matrix. The elements in $\mathbf{z}$ are PN estimates at the ends of the piecewise linear segments. In order to be robust against the model noise, q should be strictly smaller than N_{p} ($q < N_{\mathrm{p}}$) so that the problem of estimating $\mathbf{z}$ becomes overdetermined. The phase noise compensation matrix $\mathbf{\Phi}$ can be estimated as

$$\hat{\mathbf{\Phi}} = \frac{1}{N_{\mathrm{r}}}\sum_{m=1}^{N_{\mathrm{r}}} \mathrm{diag}\left(\left(\mathbf{W}\,\mathrm{diag}(\mathbf{r}_m)\mathbf{T}\right)^{\dagger}\left(\sum_{j=1}^{N_{\mathrm{t}}}\mathrm{diag}(\mathbf{c}_{mj})\mathbf{s}_j\right)\right). \tag{9}$$

4. Simulation

For simulations, we assume 1024 subcarriers (including 832 data subcarriers, 32 scattered pilots, and 160 guard band subcarriers). The active subcarriers are loaded with 64 quadrature amplitude modulation (64-QAM) symbols. The free-running oscillator model with β = 100 Hz is used. q = 7 unknowns (cf. Section 3) are used for phase noise mitigation. The sampling frequency is assumed to be 20 MHz. The UFMC prototype filter is chosen to be the Dolph-Chebyshev filter with 64 taps and 40-dB stopband suppression. The multipath channel is a 3-tap Rayleigh fading channel, whose taps occurs at the 0, 4, and 8th time samples with equal variance of 1/3, and that the channel stays constant within 20 UFMC symbols after which an independent channel realization is drawn. In total 100 channel realizations are generated. In order to focus on the effects mutual coupling and phase noise, we assume 3-D uniform multipath environment, a pair of parallel half-wavelength (λ/2) dipole antennas at the transmitter, and four ideal antennas (i.e., no correlation or mutual coupling) at the receiver.

Fig. 1 shows the symbol error rate (SER) of a 4×2 MIMO-UFMC system in multipath fading channel under different dipole separations (i.e., 0.1, 0.2, and 0.3 λ). For each dipole separation, the SERs of the MIMO-UFMC system without phase noise (PN) correction, with CPE correction, with PN mitigation are plotted, respectively. The corresponding PN-free SERs (no PN) are also plotted in the same graph as references. Fig. 1a corresponds to the case when the mutual coupling is not considered, whereas Fig. 1b corresponds to the case when the mutual coupling is taken into account. As can be seen, 1) without any compensation, the phase noise severely degrades the system performance (the corresponding SER curves of the three dipole separations overlap); 2) the CPE correction only results in marginal improvement; 3) the proposed PN mitigation scheme can effectively remove the phase noise effect. The above observations hold with/without considering the mutual coupling effect. By comparing Fig. 1a and Fig. 1b, it is found that, at very small dipole separations (< 0.3 λ), the SER performance of the MIMO-UFMC system when taking the mutual coupling effect into account is better than that when the mutual coupling effect is ignored. This is because that, at such small antenna separations, the correlation reduction due to the (Electromagnetic) mutual coupling (as compared with the open-circuit correlation) is more prominent than the degradation of the total embedded antenna efficiency (including mismatch) caused by the mutual coupling. Nevertheless, at 0.3-λ separation, mutual coupling effect on correlation reduction becomes less influential than that on efficiency reduction (cf. Fig. 2 in [18]), the SER performance when taking the mutual coupling effect into account becomes slightly worse than that without considering the mutual coupling effect. As the separation increases beyond 0.4 λ, the mutual coupling has little impact on the system. Therefore, we focus on the small separation of 0.1-0.3 λ in this paper.

5. References

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