

DISTRIBUTED MIMO SYSTEMS: RECEIVER DESIGN AND ML DETECTION

Yi Cao, Weifeng Su, and Stella N. Batalama

Department of Electrical Engineering, University at Buffalo, SUNY, NY 14260

ABSTRACT

We propose a novel receiver design for distributed MIMO systems that accounts for multiple carrier frequency offsets (CFOs) and multiple timing offsets (TOs). The proposed structure utilizes a bank of pulse matched filters (one per effective CFO) at each receive antenna, followed by an information symbol detector. Each filter in the bank is sampled at the symbol rate with sampling timing selected according to the corresponding TO. For the proposed receiver configuration, we derive the maximum likelihood (ML) detector. Our theoretical developments are illustrated through extensive simulation studies and indicate that the proposed receiver structure together with the optimal ML detection offers significant performance gains compared to the current state of the art.

Index Terms— Distributed MIMO systems, receiver designs, carrier frequency offsets, timing offsets

1. INTRODUCTION

Multiple-input multiple-output (MIMO) communication techniques have been extensively investigated and deployed, with advantages such as high data rates without extra bandwidth and diversity to combat channel fading. In most existing MIMO systems such as WLAN and 4G LTE cellular networks, multiple transmit antennas are co-located and utilize a single oscillator to generate carrier signals. Similarly, at the receiver side, multiple receive antennas share a common oscillator and thus there exhibits a single carrier frequency offset between multiple transmit antennas and multiple receive antennas. So, single-input single-output (SISO) receiver design methodologies to compensate for timing and frequency offset which have been extensively studied, e.g. [1]-[5], can be directly applied to the co-located MIMO systems. However, in distributed MIMO systems such as emerging cooperative communication systems and underwater MIMO communications systems, multiple transmit antennas are generally not co-located and cannot utilize a common oscillator to generate high-frequency carrier signals. Thus, direct application of SISO receiver design methodologies is not valid in distributed MIMO systems where multiple carrier frequency offsets (CFOs) and multiple timing offsets (TOs) are observed [6].

Some existing literature have considered receiver designs to address multiple CFOs in distributed MIMO systems, e.g.

[7]-[10]. However, most existing literature on distributed MIMO receiver design exploit only the effect of multiple CFOs to the phase of the received signal and do not consider the effect of the CFOs to the symbol energy attenuation at the receiver output. This is evident even in SISO systems [1] where the output of the pulse matched filter experiences amplitude degradation due to a single CFO. Thus, in distributed MIMO systems, it is expected that different CFOs will have different symbol-energy attenuation effects at the receiver output which should be addressed properly. In addition, most receiver designs in the literature assume perfect timing synchronization between multiple distributed transmit antennas and multiple distributed receive antennas. The assumption of perfect timing synchronization, however, appears to be incompatible with the very nature of distributed MIMO systems where transmission paths exhibit different propagation delays which implies different timing offsets at each receive antenna. Recently, a receiver design, which considers both multiple CFOs and multiple TOs, was proposed in [11] for distributed MIMO systems. In this design, a supervised compensation matrix to combat the joint effect of multiple CFOs and multiple TOs is applied on the received signals which is sampled with a uniform sampling rate, followed by an MMSE detector.

In this paper, we propose a novel receiver structure that is able to better accommodate multiple CFOs and multiple TOs in distributed MIMO systems. The receiver consists of a bank of matched filters (one per CFO) at each receive antenna, followed by an information symbol detector. The output of each matched filter is sampled at the symbol rate with sampling timing determined by the corresponding TO. For this given receiver configuration, we derive the maximum likelihood (ML) detector. Extensive simulation studies illustrate our theoretical developments and show that the proposed scheme outperforms the current state of the art in terms of receiver bit-error-rate (BER) under a variety of communication conditions and system parameters.

2. SYSTEM MODEL

We consider a distributed MIMO system with M_t transmit antennas and M_r receive antennas where the transmit antennas are not co-located and thus cannot utilize a common oscillator to generate high-frequency carrier signals. In this scenario, each transmit antenna uses its own oscillator and different transmit antennas may experience different carrier fre-

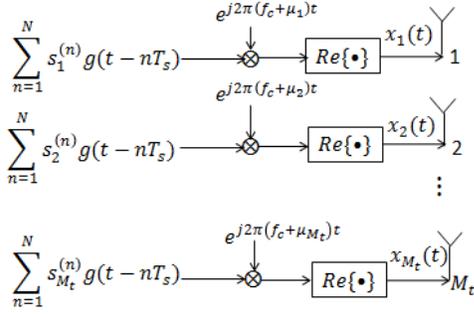


Fig. 1: Transmitter processing diagram.

quency drifts. The transmitter diagram is shown in Fig. 1. In general, if each transmit antenna i , $i = 1, 2, \dots, M_t$, sends a symbol sequence of length N , we may model the bandpass signal sent from the i th transmit antenna as

$$x_i(t) = \sum_{n=1}^N \text{Re} \left\{ s_i^{(n)} g(t - nT_s) e^{j2\pi(f_c + \mu_i)t} \right\} \quad (1)$$

where $s_i^{(n)}$ is the symbol sent by the i th transmit antenna at the n th time slot, $g(t)$ is the pulse shaping signal, T_s is the symbol duration, f_c is the assumed target carrier frequency, and μ_i denotes the carrier frequency drift of the i th transmit antenna.

At the receiver side, we assume each receive antenna may have its own carrier frequency drift which may or may not be the same for all receive antennas. Thus, the lowpass equivalent signal at the j th receive antenna $j = 1, 2, \dots, M_r$ as

$$r_j(t) = \frac{1}{2} \sum_{i=1}^{M_t} \sum_{n=1}^N h_{i,j} e^{-j2\pi(f_c + \mu_i)\tau_{i,j}} e^{j2\pi\Delta f_{i,j}t} \times s_i^{(n)} g(t - nT_s - \tau_{i,j}) + n_j(t) \quad (2)$$

where $h_{i,j}$ represents the channel coefficient from the i th transmit antenna to the j th receive antenna which is assumed to be quasi-static during the transmission of a symbol sequence, $\tau_{i,j}$ denotes the propagation delay from the i th transmit antenna to the j th receive antenna, ν_j denotes the carrier frequency drift of the j th receive antenna, $\Delta f_{i,j}$ denotes the carrier frequency offset between the i th transmit antenna and the j th receive antenna which is given by $\Delta f_{i,j} = (f_c + \mu_i) - (f_c + \nu_j) = \mu_i - \nu_j$, and $n_j(t)$ is the lowpass noise which has the following power spectrum:

$$S_n(f) = \begin{cases} N_0, & |f| \leq \frac{B}{2} \\ 0, & \text{elsewhere} \end{cases} \quad \text{where } B \text{ is the bandwidth of the bandpass filter at each receive antenna. We can estimate the propagation delay } \tau_{i,j} \text{ and the carrier frequency offset } \Delta f_{i,j} \text{ at each receive antenna based on the received lowpass signal.}$$

3. PROPOSED RECEIVER STRUCTURE AND OPTIMAL DETECTION

In this section, we first present the receiver design which is able to better accommodate multiple CFOs and multiple TOs in distributed MIMO systems. Then, we derive the optimal

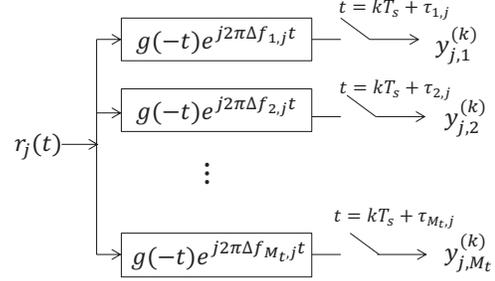


Fig. 2: Proposed receiver structure at each receive antenna.

ML detection algorithm based on our proposed receiver structure.

3.1. Proposed Receiver Structure

We would like to propose a receiver design with M_t parallel matched filters at each receive antenna, as shown in Fig. 2. The parallel matched filters operate on the lowpass received signals, and each of the filters targets a specific CFO with output sampled at certain timing to accommodate specific TO. More specifically, at each receive antenna j , $j = 1, 2, \dots, M_r$, the matched filter on the m th, $m = 1, 2, \dots, M_t$, branch is adjusted to $g(-t)e^{j2\pi\Delta f_{m,j}t}$ to accommodate CFO $\Delta f_{m,j}$ and sampled at $t = kT_s + \tau_{m,j}$ to accommodate TO $\tau_{m,j}$. With the lowpass received signal $r_j(t)$ in (2), the discrete sample output from the m th branch matched filter with sampling at $t = kT_s + \tau_{m,j}$ can be given by

$$y_{j,m}^{(k)} = \frac{1}{2} \sum_{i=1}^{M_t} \sum_{n=1}^N h_{i,j} e^{-j2\pi(f_c + \mu_i)\tau_{i,j}} e^{j2\pi\Delta f_{i,j}(kT_s + \tau_{m,j})} \times s_i^{(n)} \int_{-\infty}^{+\infty} g([k-n]T_s + \tau_{m,j} - \tau_{i,j} - \tau) \times g(-\tau) e^{j2\pi(\Delta f_{m,j} - \Delta f_{i,j})\tau} d\tau + N_{j,m}^{(k)} \quad (3)$$

where $N_{j,m}^{(k)} = \{n_j(t)\} * \{g(-t)e^{j2\pi\Delta f_{m,j}t}\} |_{t=kT_s + \tau_{m,j}}$ is the sampled noise at the m th branch matched filter at receive antenna j . For simplicity of notation, let us denote $\tilde{h}_{i,j} \triangleq h_{i,j} e^{-j2\pi(f_c + \mu_i)\tau_{i,j}}$, and

$$G_{i,j,m,p} \triangleq \int_{-\infty}^{+\infty} g(pT_s + \tau_{m,j} - \tau_{i,j}) g(-\tau) e^{j2\pi(\Delta f_{m,j} - \Delta f_{i,j})\tau} d\tau, \quad (4)$$

then the discrete sample output in (3) can be represented as

$$y_{j,m}^{(k)} = \frac{1}{2} \sum_{i=1}^{M_t} \sum_{n=1}^N \tilde{h}_{i,j} e^{j2\pi\Delta f_{i,j}(kT_s + \tau_{m,j})} s_i^{(n)} G_{i,j,m,k-n} + N_{j,m}^{(k)}. \quad (5)$$

We can see that if there is no multiple TOs, signals from all transmit antennas are aligned, i.e. $\tau_{i,j} = \tau_{m,j}$, $1 \leq i \neq m \leq M_t$, then $G_{i,j,m,k-n}$ is non-zero only when $k-n = 0$. Hence, the discrete sample output $y_{j,m}^{(k)}$ without multiple TOs can be reduced to

$$y_{j,m}^{(k)} = \frac{1}{2} \sum_{i=1}^{M_t} \tilde{h}_{i,j} e^{j2\pi\Delta f_{i,j}kT_s} s_i^{(k)} G_{i,j,m,0} + N_{j,m}^{(k)} \quad (6)$$

where

$$G_{i,j,m,0} = \int_{-\infty}^{+\infty} g^2(-\tau) e^{j2\pi(\Delta f_{m,j} - \Delta f_{i,j})\tau} d\tau. \quad (7)$$

Furthermore, if both multiple TOs and multiple CFOs do not present, i.e. $\tau_{i,j} = \tau_{m,j}$ and $\Delta f_{m,j} = \Delta f_{i,j}$, $1 \leq i \neq m \leq M_t$, then $G_{i,j,m,0}$ in (7) can be further reduced to

$$G_{i,j,m,0} = \int_{-\infty}^{+\infty} g^2(-\tau) d\tau. \quad (8)$$

In co-located MIMO systems, since all transmit antennas can share one oscillator and all receive antennas can share one oscillator, there exists only one CFO and one TO. Thus, the M_t parallel matched filters can be reduced to one matched filter to accommodate the single CFO and sample at one timing to accommodate the single TO which is the case of (8). For simplicity of the model and discussion, we consider the

following pulse shaping: $g(t) = \begin{cases} g_T(t), & -\frac{T_s}{2} \leq t \leq \frac{T_s}{2} \\ 0, & \text{elsewhere} \end{cases}$,

where $g_T(t)$ is the pulse shape within one symbol duration. We also assume that the difference of the TOs from different transmit antennas is within one symbol duration, i.e. $0 \leq |\tau_{m,j} - \tau_{i,j}| \leq T_s$. In this scenario, the parameter $k - n$ in (3) can have only three possible values, i.e. 1, 0, -1, which means the discrete sample output of the matched filter involves only the symbols from the $(k - 1)$ th, k th and $(k + 1)$ th time slots. In addition, $G_{i,j,m,k-n}$ in (4) and (5) has the following properties: i) When $k - n = 1$, we have $G_{i,j,m,1} = 0$ if $\tau_{m,j} - \tau_{i,j} > 0$; ii) When $k - n = -1$, we have $G_{i,j,m,-1} = 0$ if $\tau_{m,j} - \tau_{i,j} < 0$; iii) When $k - n = 0$ and $i = m$, we have $G_{i,j,i,0} = \int_{-\infty}^{+\infty} g^2(-\tau) d\tau$, which is the ideal output amplitude without any CFO and TO. It means that with the proposed receiver, symbol sent from each transmit antenna has the maximum amplitude on its corresponding branch. We note that for general pulse shaping and arbitrary TOs, the proposed receiver structure and the optimal ML detection can be similarly developed with larger window (i.e. involving more time slots).

We may organize the discrete sample output $y_{j,m}^{(k)}$ compactly in vector form as

$$y_{j,m}^{(k)} = \boldsymbol{\psi}_{j,m}^{(k)} \mathbf{S}^T + N_{j,m}^{(k)} \quad (9)$$

where $\mathbf{S} = [\mathbf{s}^{(k-1)T} \mathbf{s}^{kT} \mathbf{s}^{(k+1)T}]^T$, $\mathbf{s}^{(k)} = [s_1^{(k)}, \dots, s_{M_t}^{(k)}]^T$ is the symbol vector sent at the k th time slot from the M_t transmit antennas, and $\boldsymbol{\psi}_{j,m}^{(k)}$ denotes the vector of the coefficients of symbols from the $(k - 1)$ th, k th and $(k + 1)$ th time slots with size $3M_t$ which is given by

$$\boldsymbol{\psi}_{j,m}^{(k)} \triangleq \frac{1}{2} \begin{bmatrix} \tilde{h}_{1,j} e^{j2\pi\Delta f_{1,j}(kT_s + \tau_{m,j})} G_{1,j,m,1} \\ \vdots \\ \tilde{h}_{M_t,j} e^{j2\pi\Delta f_{M_t,j}(kT_s + \tau_{m,j})} G_{M_t,j,m,1} \\ \tilde{h}_{1,j} e^{j2\pi\Delta f_{1,j}(kT_s + \tau_{m,j})} G_{1,j,m,0} \\ \vdots \\ \tilde{h}_{M_t,j} e^{j2\pi\Delta f_{M_t,j}(kT_s + \tau_{m,j})} G_{M_t,j,m,0} \\ \tilde{h}_{1,j} e^{j2\pi\Delta f_{1,j}(kT_s + \tau_{m,j})} G_{1,j,m,-1} \\ \vdots \\ \tilde{h}_{M_t,j} e^{j2\pi\Delta f_{M_t,j}(kT_s + \tau_{m,j})} G_{M_t,j,m,-1} \end{bmatrix}. \quad (10)$$

Collecting the discrete sample output $y_{j,m}^{(k)}$, $m = 1, \dots, M_t$, from all M_t branches at the j th receiver, we can form a received signal vector of size M_t as

$$\mathbf{Y}_j^{(k)} = [y_{j,1}^{(k)}, \dots, y_{j,M_t}^{(k)}]^T = \boldsymbol{\psi}_j^{(k)} \mathbf{S}^T + \mathbf{N}_j^{(k)} \quad (11)$$

where $\boldsymbol{\psi}_j^{(k)} = [\boldsymbol{\psi}_{j,1}^{(k)T} \dots \boldsymbol{\psi}_{j,M_t}^{(k)T}]^T$ is the coefficient matrix at the j th receiver of size M_t by $3M_t$, and $\mathbf{N}_j^{(k)} = [N_{j,1}^{(k)}, \dots, N_{j,M_t}^{(k)}]^T$ is the vector of sampled noise at the j th receiver of size M_t . Finally, we stack $\mathbf{Y}_j^{(k)}$, $j = 1, \dots, M_r$, from all M_r receivers, and denote a received signal vector of size $M_t M_r$ as

$$\mathbf{Y}^{(k)} = [\mathbf{Y}_1^{(k)T} \dots \mathbf{Y}_{M_r}^{(k)T}]^T = \boldsymbol{\Psi}^{(k)} \mathbf{S}^T + \mathbf{N}^{(k)} \quad (12)$$

where $\boldsymbol{\Psi}^{(k)} = [\boldsymbol{\psi}_1^{(k)T} \dots \boldsymbol{\psi}_{M_r}^{(k)T}]^T$ is the corresponding coefficient matrix of size $M_t M_r$ by $3M_t$, and $\mathbf{N}^{(k)} = [N_1^{(k)T} \dots N_{M_r}^{(k)T}]^T$ is the corresponding sampled noise vector of size $M_t M_r$.

3.2. ML Detection

In the following, we would like to develop an optimal ML detector based on the proposed receiver structure. With the receiver signal model in (12), the probability density function of $\mathbf{Y}^{(k)}$ can be derived as

$$P(\mathbf{Y}^{(k)}) = \frac{1}{(2\pi)^{\frac{M_t M_r}{2}} |\Sigma|^{\frac{1}{2}}} \exp \left\{ -\frac{1}{2} (\mathbf{Y}^{(k)} - \boldsymbol{\Psi}^{(k)} \mathbf{S}^T)^H \Sigma^{-1} (\mathbf{Y}^{(k)} - \boldsymbol{\Psi}^{(k)} \mathbf{S}^T) \right\} \quad (13)$$

where Σ denotes the covariance matrix of the overall noise vector $\mathbf{N}^{(k)}$ of size $M_t M_r$ by $M_t M_r$ which is given by

$$\Sigma = \begin{pmatrix} \text{Var}(N_{1,1}^{(k)}) & \text{Cov}(N_{1,1}^{(k)}, N_{1,2}^{(k)}) & \dots & \text{Cov}(N_{1,1}^{(k)}, N_{M_r, M_t}^{(k)}) \\ \text{Cov}(N_{1,2}^{(k)}, N_{1,1}^{(k)}) & \text{Var}(N_{1,2}^{(k)}) & \dots & \text{Cov}(N_{1,2}^{(k)}, N_{M_r, M_t}^{(k)}) \\ \vdots & \vdots & \ddots & \vdots \\ \text{Cov}(N_{M_r, M_t}^{(k)}, N_{1,1}^{(k)}) & \text{Cov}(N_{M_r, M_t}^{(k)}, N_{1,2}^{(k)}) & \dots & \text{Var}(N_{M_r, M_t}^{(k)}) \end{pmatrix}. \quad (14)$$

With the power spectrum of $n_j(t)$, its autocorrelation function can be calculated as $R_n(t) = N_0 B \text{sinc}(Bt)$ [12]. Then the entries of the covariance matrix in (14) can be specified as

$$\begin{aligned} \text{Var}(N_{j,m}^{(k)}) &= \int_{-\infty}^{+\infty} \int_{-\infty}^{+\infty} R_n(\tau - \lambda) g(-\tau) g(-\lambda) \\ &\quad \times e^{j2\pi\Delta f_{m,j}(\tau - \lambda)} d\tau d\lambda, \\ \text{Cov}(N_{j,m}^{(k)}, N_{j',m'}^{(k)}) &= \int_{-\infty}^{+\infty} \int_{-\infty}^{+\infty} R_n(\tau_{m,j} - \tau_{m',j'} + \tau - \lambda) \\ &\quad \times g(-\tau) g(-\lambda) e^{j2\pi(\Delta f_{m,j}\tau - \Delta f_{m',j'}\lambda)} d\tau d\lambda. \end{aligned}$$

Thus, the optimal ML detector to decode the symbols $\mathbf{s}^{(k)}$ is given by

$$\hat{\mathbf{s}}^{(k)} = \underset{\mathbf{s}^{(k-1)}, \mathbf{s}^{(k)}, \mathbf{s}^{(k+1)}}{\arg \max} P(\mathbf{Y}^{(k)}). \quad (15)$$

Note that the above ML detector is based on the exhaustive search among the symbols from the previous $(k - 1)$ th time slot to the next $(k + 1)$ th time slot in order to find the symbols $\mathbf{s}^{(k)}$ at the current k th time slot, and the detector output is only for the symbols $\mathbf{s}^{(k)}$ at the k th time slot.

4. SIMULATION RESULTS

We carry out some simulations to show the performances of our proposed receiver, and compare it with the receiver proposed in [11]. In the simulation setup, we apply a root-raised cosine filter with a roll-off factor of 0.5 and the symbol duration is set to be $T_s = 10$ ms. We use a binary phase-shift keying (BPSK) constellation and channel coefficients are assumed to be quasi-static during the transmission of a symbol sequence.

In Fig. 3 (a) and (b), we show the performances of the proposed receiver structure with the optimal ML detection in a system with different levels of multiple TOs and multiple CFOs. We also compare the performances of the proposed receiver with the existing receiver in [11]. Simulations are carried out in a system with $M_t = 2$ distributed transmit antennas and $M_r = 1$ receive antenna, and the Alamouti code is applied at the transmitter side. For fair comparison, the oversampling rate of the receiver in [11] is set as 2 in this case. In Fig. 3 (a), we consider CFOs $\Delta f_{1,1} = \Delta f_{2,1} = 0$ Hz and difference between TOs $\Delta\tau = \tau_{1,1} - \tau_{2,1} = 1, 2, 3$ and 4 ms, respectively. In this figure, we observe that the BER performance of the proposed receiver is much better than that of the receiver in [11]. For example, with difference between TOs $\Delta\tau = 1$ ms, the proposed receiver yields almost 4 dB gain over the receiver in [11] at a BER of 10^{-4} . In Fig. 3 (b), we consider difference between TOs $\Delta\tau = \tau_{1,1} - \tau_{2,1} = 0$ ms and CFOs $\Delta f_{1,1} = -\Delta f_{2,1} = 5, 10, 15$ and 20 Hz, respectively. In this figure, we have similar observation that the BER performance of the proposed receiver is significantly better than that of the receiver in [11], for example, with CFOs $\Delta f_{1,1} = -\Delta f_{2,1} = 15$ Hz, the proposed receiver yields about 10 dB gain over the receiver in [11] at a BER of 10^{-4} .

Fig. 4 shows the overall BER performances of the proposed receiver with both multiple CFOs and multiple TOs in systems with $M_t = 2$ distributed transmit antennas and $M_r = 1$ receive antenna. In this figure, we consider CFOs $\Delta f_{1,1} = -\Delta f_{2,1} = 15$ Hz and difference between TOs $\Delta\tau = \tau_{1,1} - \tau_{2,1} = 4$ ms. We observe that the proposed receiver achieves about 3 dB improvement compared to the receiver in [11] at a BER of 10^{-4} .

5. CONCLUSION

In this paper, we considered the problem of effective receiver designs for distributed MIMO systems in the presence of multiple CFOs and multiple TOs. We proposed a novel receiver structure which consists of a bank of matched filters at each receive antenna. Each matched filter in the bank is sampled at the symbol rate with sampling timing determined by the corresponding TO. We derived the optimal ML detector based on this proposed receiver structure. Simulation results show that the proposed receiver structure together with the optimal ML detection achieves significant performance gains compared to the current state of the art. For example, the proposed receiver achieves about 3 dB improvement compared to the receiver in [11] at a BER of 10^{-4} .

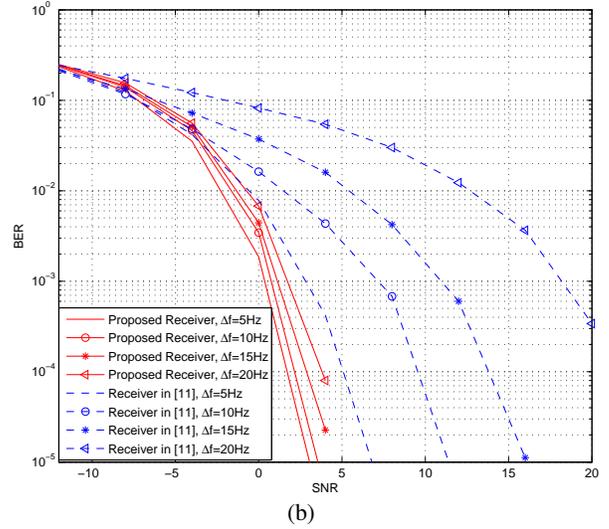
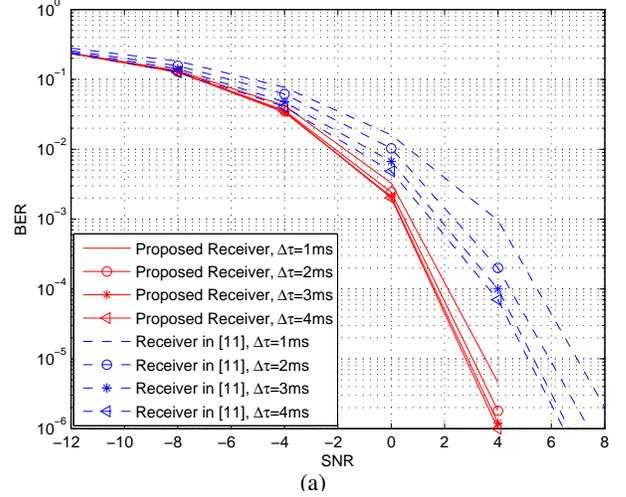


Fig. 3: Performances of the proposed receiver and the receiver in [11]: (a) with multiple TOs but without CFOs; (b) with multiple CFOs but without TOs.

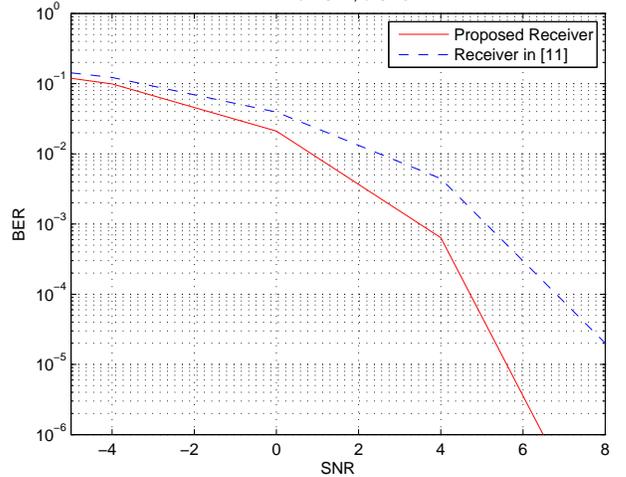


Fig. 4: Performances of the proposed receiver and the receiver in [11] with both multiple CFOs and multiple TOs in systems with $M_t = 2$ distributed transmit antennas.

6. REFERENCES

- [1] Wen-Yi Kuo and Michael P. Fitz, "Frequency offset compensation of pilot symbol assisted modulation in frequency flat fading," *IEEE Trans. Communications*, vol. 45, pp. 1412–1416, Aug. 2002.
- [2] Chia-Liang Li, Djen W.S. and Feher Kamilo, "A low-cost dual-mode noncoherent receiver with robust frequency-offset compensation," *VTC, Secaucus, NJ*, 1993, pp. 412–415
- [3] Young-Jae Ryu and Dong-Seong Han, "Frequency synchronization algorithm robust to pilot distortion in ATSC system," *IEEE Trans. Consumer Electronics*, vol. 48, pp. 610–616, Aug. 2002
- [4] T. Wiegand and J. Rust, S. Paul, "Reconfigurable architecture of a hybrid synchronisation algorithm for LTE," *2012 IEEE Int. Symposium on PIMRC*, Sydney, pp. 2291–2296, Sept. 9-12, 2012
- [5] B. Ramakrishnan, "Matched nonlinearities for frame synchronisation in presence of frequency offsets," *Electronics Letters*, vol. 45, pp. 236–237, Feb. 2009
- [6] Kai Deng, "Frequency synchronization in MIMO systems," *CECNet, Yichang*, pp. 1832–1835, 2012.
- [7] J. Lee, H. Lin, and T. Sang, "An SFBC-OFDM Receiver with MLSE Equalizer to Combat Multiple Carrier Frequency Offsets," *Proc. IEEE ISCAS*, pp. 1524–1527, May 2012.
- [8] Hsin-Yi Shen, Haiming Yang, B. Sikdar, and S. Kalyanaraman., "A distributed system for cooperative mimo transmissions," *IEEE In Global Telecommunications Conference, 2008. IEEE GLOBECOM 2008. IEEE*, pp. 1-5, 2008.
- [9] Z. Li, D. Qu, and G. Zhu, "An equalization technique for distributed stbc-ofdm system with multiple carrier frequency offsets," *Proc. IEEE WCNC*, pp. 839-843, 2006.
- [10] Huiming Wang, Xia, Xiang-Gen, Qinye Yin and Lin Bai, "A distributed linear convolutive space-frequency coding for cooperative communication systems with multiple frequency offsets," *Communications. ICC. IEEE Int., Dresden*, pp. 1–5, 2009.
- [11] A. A. Nasir, H. Mehrpouyan, S. Durrani, S. D. Blostein, R. A. Kennedy and B. Ottersten, "Transceiver design for distributed STBC based AF cooperative networks in the presence of timing and frequency offsets," *IEEE Trans. Signal Process*, vol. 61, pp. 3143–3158, Jun. 2013.
- [12] J. G. Proakis and M. Salehi, *Digital Communications*, 5th ed., McGraw-Hill, NY, NY, 2008, pp. 67–69.