JOINT SOURCE SEPARATION AND OFFSET ESTIMATION FOR ASYNCHRONOUS OFDM SYSTEMS USING SUBSPACE FITTING

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Abstract- An asynchronous multi-user cyclic prefix (CP) transmission system, such as OFDM is considered where the receiver is offset in time by an unknown value with respect to the desired user. The user signals are transmitted with superimposed training and an antenna array is used in the receiver to compute the offset and perform space-time (ST) source separation using subspace fitting. Smoothing is used to improve on the offset estimate and alternating projections are used to improve the beamformer weights to cancel the interference. Simulation results indicate that the blind source separation scheme converges to a reference scheme, where the receiver is synchronized to the desired user and has complete knowledge of its signal.

Index Terms— Ad hoc networks, asynchronous OFDM, superimposed training, subspace fitting.

1. INTRODUCTION

WLAN systems are increasingly popular in household situations. A user buys transceiver equipment and expects it to work, however his neighbors might cause interference with their uncoordinated WLAN systems. He might also own other equipment (e.g. bluetooth) that operates in the same frequency band. It is interesting to consider small additions to existing WLAN standards (based on orthogonal frequency division multiplexing (OFDM)) that can help in reducing interference, while remaining compatible with the legacy equipment. Joint estimation techniques using superimposed training of asynchronous OFDM systems is an attractive option to achieve interference cancellation and maximize spectral efficiency. The data rate is maximized since no bandwidth is lost due to training and the receiver can retrain at any time instant, independent of the MAC layer. In our scenario, the interfering users can enter and exit the wireless channel on an *ad-hoc* basis. The receiver is not required to be synchronized to the user of interest, or to have any coordination between interfering users. In such systems, the CP of the desired user does not align with that of interfering users. Thus subcarrier frequency domain weights are not applicable to separate the sources [1].

Consider a multiuser-OFDM system as shown in Fig. 1 where each user transmits data continuously and the desired user signal is received using an antenna array corrupted by interference from other users, with unknown offset, say τ, corrupted by interference from other users and noise. Our objective is to use the signals from the antenna array to separate the interfering signals at the receiver and estimate the desired user signal and its offset.

Superimposed training algorithms (STA) [2-3] consider a non random (not necessarily periodic) sequence being arithmetically added to the transmitted sequence. These sequences introduce a time-varying mean in the received symbols at the antenna array. The time varying mean property can be exploited to separate the required user and to estimate the channel. In [2] the desired user is assumed to be synchronized for a multi-user channel with memory and a beamformer is designed using the correlation between the receive antenna array and superimposed sequence. In [3] the problem is extended for OFDM systems and a multiple signal classification (MUSIC) type search gives offset of the desired user. However this scheme performs relatively poor for high SNR's and longer channel lengths.

We improve on this scheme in 3 ways: (a) replace MUSIC with a more systematic nonlinear least squares (NLLS) problem, (b) apply smoothing to eliminate spurious local maxima (c) further improve estimation using alternating projections. Subspace fitting or eigenstructure methods [4] are known to have high resolution capabilities and yield accurate estimates. The estimates from subspace fitting are improved using a moving average filter to smooth out local maxima and followed by alternating least squares (LS).

Notation: $(.)^T$, $(.)^H$ and $(.)^\dagger$ represent transpose, hermitian and pseudo inverse. $\odot$, $\oslash$ and $\|.\|$ represent pointwise multiplication, division and Frobenius norm. $\mathcal{F}_D$ is a $D \times D$ DFT matrix.

2. DATA MODEL

In a CP based OFDM transmission system, the data is transmitted in blocks with a redundant CP of length L_{CP} inserted between successive blocks of length N, N being

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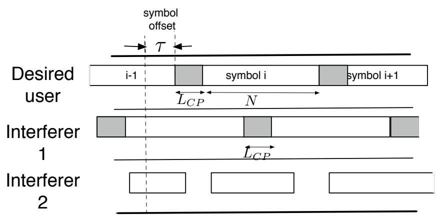

Fig. 1: *transmission with the desired user signals offset by* τ

the number of subcarriers. In such transmission systems a known sequence with period P can be arithmetically superimposed to the time domain OFDM signal including CP, $x_k^{(1)} = c_k^{(1)} + s_k^{(1)}$, as shown in Fig. 1. $s_k^{(1)}$ and $c_k^{(1)}$ are the desired user signal (PSK) and superimposed sequence as a function of time index k. The observation window is a continuous transmission of successive OFDM symbols (such that there is no edge effect), of length $L_{obs}$where the start of the CP is offset from the receiver by an unknown value τ. For the sake of convenience P is an integer fraction of L_{CP}. For simplicity the length of OFDM symbol and its cyclic prefix $(N + L_{CP})$ is a multiple of P and the start of the OFDM symbol coincides with the start of superimposed sequence[1].

In the well known pilot tone assisted modulation (PTAM) scheme for OFDM systems, a number of dedicated training pilots are placed in specific frequency bins to acquire the channel state information, resulting in reduced bandwidth. The PTAM scheme can be seen as a special case of superimposed training where a pilot sequence is inserted into the frequency bins which is equivalent to being added onto the time domain OFDM signal.

Let N_t be the number of users occupying the common wireless channel with user 1 being the desired user and the other users being interferers. Similar to [2-3] we make the following assumptions about the transmitted signals from desired and interfering users:

A1: The information sequence of the desired user $s_k^{(1)}$ is zero mean and white with $E\{|s_k^{(1)}|^2\} = \sigma_s^2$.

A2: The superimposed sequence of the desired user $c_k^{(1)}$ is a non-random periodic sequence with period P such that $c_k^{(1)} = c_{k+mP}^{(1)}$ for all m, k.

A3: The information sequence of the asynchronous interfering users is zero mean. The interfering users may or may not use superimposed training.

Each user is associated with one antenna and the receiver is supplied with N_r receive antennas. Consider a

[1] Although this paper is written in the context of an OFDM system, most of the techniques are in fact applicable to any signal with zero mean and time domain samples.

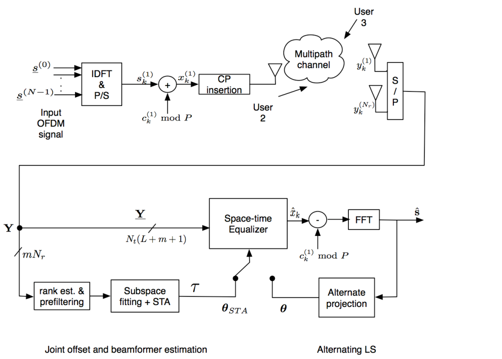

Fig. 2: *Source separation for asynchronous OFDM*

$N_r \times N_t$ frequency selective fading channel modeled by a time invariant FIR filter H_l of order L; $H_l = [\mathbf{h}_l^{(1)} \cdots \mathbf{h}_l^{(N_t)}]$ ($L \leq L_{CP}$ cyclic prefix length). $\mathbf{h}_l^{(1)}$ is the channel response from user 1 to the receiver and the received $N_r \times 1$ signal is given by

$$\mathbf{y}_k = \sum_{l=0}^{L} H_l \mathbf{x}_{k-l} + \mathbf{w}_k \tag{1}$$

where $\mathbf{x}_k = [x_k^{(1)} \cdots x_k^{(N_t)}]^T$ is a $N_t \times 1$ vector of the transmitted signals from the desired and interfering users and $\mathbf{w}_k$ is the $N_r \times 1$ noise vector at the receive antennas at time instant k. For simplicity of expression, we will assume that all $N_r \times N_t$ MIMO channels have the same length.

3. EQUALIZER DESIGN

3.1. Setup

Consider a situation where the desired and the interfering users transmit data at random time instants unknown to the receiver. Fig. 2 shows a multi-user discrete-time baseband schematic of such an OFDM transmission system. The transmitter is a standard OFDM system. The receiver has two blocks: A ST beamformer followed by a single user OFDM receiver. The dimensionality offered by the multiple antennas and the data stacking is used to cancel all interfering signals in the ST beamformer. It has been shown in [1] that when the desired user is synchronized to the receiver (but corrupted with asynchronous interfering users) and OFDM demodulation is performed directly after the receive antenna array, beamformer coefficients of degree L are required for each subcarrier to cancel the interfering users. If the conventional OFDM demodulation scheme were to be used in our scenario the synchronization offset has to be estimated prior to the OFDM demodulation followed by L degree beamforming on each subcarrier, which requires at least $N_r = L$ receiver antennas.

To avoid this situation, the setup in Fig. 2 will estimate the delay using subspace fitting and cancel the asynchronous interference with a ST beamformer followed by OFDM demodulation.

We assume without loss of a generality that the maximum offset between the receiver and the desired user is P (the period of superimposed sequence). The receiver collects L_{obs} samples to jointly estimates the offset of the desired user and beamformer taps to cancel the interference. L_{obs} is a multiple of $N + L_{CP}$. The transmitted packet falls completely within the observation window.

3.2. Subspace fitting technique to jointly estimate offset and design beamformer

We start by following the approach described earlier by us in [3]. The objective is to estimate the desired user data using a space-time beamformer ($mN_r \times 1$ vector) $\boldsymbol{\theta} = [\boldsymbol{\theta}_1^T \cdots \boldsymbol{\theta}_m^T]^T$, where m is the length of the beamformer which combines $N_r \times 1$ receive antenna array signals at successive time instants. Stacking the successive antenna array signals from time k to $k-m+1$ gives the usual data model

$$\begin{aligned}\mathbf{y}_{k:k-m+1} &= [\mathbf{y}_k^T \ \cdots \ \mathbf{y}_{k-m+1}^T]^T \\ &= \mathcal{H}\mathbf{x}_{k:k-m-L+1} + \mathbf{w}_{k:k-m+1}, \qquad (2)\\ \mathcal{H} &= \begin{bmatrix} H_0 H_1 \ldots H_L & & \\ & \ddots & \ddots \\ & & H_0 H_1 \ldots H_L \end{bmatrix}.\end{aligned}$$

$\mathbf{x}_{k:k-m-L+1}$ and $\mathbf{w}_{k:k-m+1}$ are defined in a similar way as $\mathbf{y}_{k:k-m+1}$ and $\mathcal{H}$ is a $mN_r \times (m+L+1)N_t$ block toeplitz matrix. Interference cancellation relies on the existence of a filter coefficient vector $\boldsymbol{\theta}$ such that $\mathbf{y}_{k:k-m+1}^T\boldsymbol{\theta}$ is an estimate of the transmitted signal $x_k^{(1)}$. For a time domain symbol transmitted with delay τ, $x_k^{(1)} = s_{k-\tau}^{(1)} + c_{k-\tau}^{(1)}$. A necessary condition for space-time equalization is that $\mathcal{H}$ should be a tall matrix [6]. The instantaneous error of the ST beamformer $\boldsymbol{\theta}$ is

$$\begin{aligned}\varepsilon_k &= \mathbf{y}_{k:k-m+1}^T\boldsymbol{\theta} - x_k^{(1)} \\ &= \left[\mathbf{y}_k^T \cdots \mathbf{y}_{k-m+1}^T\right] \begin{bmatrix}\boldsymbol{\theta}_1 \\ \vdots \\ \boldsymbol{\theta}_m\end{bmatrix} - (s_{k-\tau}^{(1)} + c_{k-\tau}^{(1)})\end{aligned}$$

Extending the error signal ε_k for P successive time instants to a $P \times 1$ vector by right shifting and stacking P times, we obtain the

$$\begin{bmatrix}\varepsilon_k \\ \vdots \\ \varepsilon_{k+P-1}\end{bmatrix} = \begin{bmatrix}\mathbf{y}_k^T & \mathbf{y}_{k-1}^T & \cdots \mathbf{y}_{k-m+1}^T \\ \vdots & \ddots & \vdots \\ \mathbf{y}_{k+P-1}^T & \mathbf{y}_{k+P-2}^T & \cdots \mathbf{y}_{k+P-m}^T\end{bmatrix} \begin{bmatrix}\boldsymbol{\theta}_1 \\ \vdots \\ \boldsymbol{\theta}_m\end{bmatrix} - \begin{bmatrix} s_{k-\tau}^{(1)} + c_{k-\tau}^{(1)} \\ \vdots \\ s_{k-\tau+P-1}^{(1)} + c_{k-\tau+P-1}^{(1)}\end{bmatrix} \qquad (3)$$

To mitigate the effect of the unknown symbols $s_{k-\tau}^{(1)}$, we propose to use the assumptions A1, A2 and A3 and average them out over M periods of length P, $M = \frac{L_{obs}}{P}$, where M is an integer. When M is sufficiently large the cost function can be averaged for $\bar{\varepsilon}_l = \sum_{i=0}^{M-1} \varepsilon_{iP+l}$ for all $l = 0 : P-1$ as

$$\begin{bmatrix}\bar{\varepsilon}_0 \\ \vdots \\ \bar{\varepsilon}_{P-1}\end{bmatrix} \approx \begin{bmatrix}\mathbf{z}_0^T & \cdots \mathbf{z}_{-m+1}^T \\ \vdots & \ddots & \vdots \\ \mathbf{z}_{P-1}^T & \cdots \mathbf{z}_{P-m}^T\end{bmatrix} \begin{bmatrix}\boldsymbol{\theta}_1 \\ \vdots \\ \boldsymbol{\theta}_m\end{bmatrix} - \begin{bmatrix} c_{-\tau}^{(1)} \\ \vdots \\ c_{P-1-\tau}^{(1)}\end{bmatrix}$$
$$\Leftrightarrow \bar{\boldsymbol{\varepsilon}} \approx \mathcal{Z}\boldsymbol{\theta} - \mathbf{c}_\tau^{(1)} \qquad (4)$$

where $\mathbf{z}_l = \frac{1}{M}\sum_{i=0}^{M-1} \mathbf{y}_{iP+l}$. The negative indices of the receiver antenna array averages $\mathbf{z}_l$ have some samples from cyclic prefix of the received signal and $-\tau$ in the expression $c_{-\tau}^{(1)}$ is the superimposed sequence $c_0^{(1)}$, shifted by τ times to the left. Since $\mathbf{c}$ is periodic in our case, $c_{-\tau}^{(1)} = c_{P-\tau}^{(1)}$.

If τ is known, a necessary condition for a unique solution $\boldsymbol{\theta}$ (4) is that $\mathcal{Z}$ is tall. This requires $P > mN_r$. To estimate τ we exploit the shift invariance property [3, 5] that a delay in time domain corresponds to phase progression in frequency domain.

$$\mathcal{F}_P \mathbf{c}_\tau^{(1)} = \mathcal{F}_P \mathbf{c}^{(1)} \odot \boldsymbol{\phi}_\tau \qquad (5)$$

where $\boldsymbol{\phi}_\tau = [1 \ \ \phi \ \ \cdots \ \ \phi^{P-1}]^T$, $\phi = \exp\frac{-j2\pi\tau}{P}$. Applying the DFT matrix $\mathcal{F}_P$ to (4) gives the LS problem

$$\mathcal{F}\mathcal{Z}\boldsymbol{\theta} \approx \mathcal{F}_P \mathbf{c}^{(1)} \odot \boldsymbol{\phi}_\tau \qquad (6)$$

In [3] this lead to MUSIC type algorithm, here we take a systematic approach and the LS problem can be expressed as subspace fitting:

$$\begin{aligned}\{\hat{\boldsymbol{\theta}}, \hat{\tau}\} &= \underset{\boldsymbol{\theta},\tau}{\operatorname{argmin}} \|\mathcal{F}\mathcal{Z}\boldsymbol{\theta} - \mathcal{F}_P\mathbf{c}^{(1)} \odot \boldsymbol{\phi}_\tau\|^2 \\ &= \underset{\boldsymbol{\theta},\tau}{\operatorname{argmin}} \|\mathcal{Z}_f\boldsymbol{\theta} - C_f^{(1)}\boldsymbol{\phi}_\tau\|^2 \qquad (7)\end{aligned}$$

where $\mathcal{Z}_f = \mathcal{F}\mathcal{Z}$ and $C_f^{(1)} = \text{diag}(\mathcal{F}_P\mathbf{c}^{(1)})$. For a fixed $C_f^{(1)}$, the minimum of (7) with respect to τ and $\boldsymbol{\theta}$ is a measure of how well the range space of $\mathcal{Z}_f\boldsymbol{\theta}$ and $C_f^{(1)}\boldsymbol{\phi}_\tau$ match. This problem can be viewed as a subspace fitting problem [4] but for a 1-dim subspace so that the technique becomes a quite feasible NLLS problem. The subspace fitting estimate selects the τ such that these subspaces are as close as possible. By substituting the pseudoinverse solution $\boldsymbol{\theta} = \mathcal{Z}_f^\dagger C_f^{(1)}\boldsymbol{\phi}_\tau$ solution back in (7)

$$\begin{aligned}\hat{\tau} &= \underset{\tau}{\operatorname{argmin}} \|\mathcal{Z}_f\mathcal{Z}_f^\dagger C_f^{(1)}\boldsymbol{\phi}_\tau - C_f^{(1)}\boldsymbol{\phi}_\tau\|^2 \\ &= \underset{\tau}{\operatorname{argmin}} \|(\Pi_z - I)C_f^{(1)}\boldsymbol{\phi}_\tau\|^2 \\ &= \underset{\tau}{\operatorname{argmax}} \|\Pi_z C_f^{(1)}\boldsymbol{\phi}_\tau\|^2 \\ &= \underset{\tau}{\operatorname{argmax}} \{\boldsymbol{\phi}_\tau^H C_f^{(1)H}\Pi_z C_f^{(1)}\boldsymbol{\phi}_\tau\} \qquad (8)\end{aligned}$$

where $\Pi_z = \mathcal{Z}_f \mathcal{Z}_f^{\dagger}$ is the projection on the column span of $\mathcal{Z}_f$.

3.3. Smoothing

Due to the time dispersive nature of the channel, the received signals also contain several shifts of the superimposed sequence resulting in multiple solutions ($L+m-1$ local maxima) that satisfy (8) at $\tau_0 \cdots \tau_{0+L+m-2}$, where τ_0 is the true offset. This can be explained from the data model. Stack the antenna array output $\mathbf{y}_{k:k-m+1}$ for P successive time instants to compute the periodic mean similar to (4) gives the model for $\tau = \tau_0$:

$$\begin{bmatrix} \mathbf{z}_0 & \cdots \mathbf{z}_{P-1} \\ \vdots & \ddots \quad \vdots \\ \mathbf{z}_{-m+1} & \cdots \mathbf{z}_{P-m} \end{bmatrix} \approx \begin{bmatrix} \mathbf{h}_0^{(1)} \ldots \mathbf{h}_L^{(1)} & & \\ & \ddots & \ddots \\ & & \mathbf{h}_0^{(1)} \ldots \mathbf{h}_L^{(1)} \end{bmatrix} \begin{bmatrix} c_\tau^{(1)} \cdots & c_{\tau+P-1}^{(1)} \\ \vdots & \vdots \\ c_{\tau-m-L+2}^{(1)} & c_{\tau+P-m-L+1}^{(1)} \end{bmatrix}$$

$$\Leftrightarrow \mathcal{Z}^T \approx \mathcal{H}^{(1)} \mathcal{C}_\tau^{(1)T}$$

where $\mathcal{C}_\tau^{(1)}$ is a $P \times L + m - 1$ matrix and $\mathcal{H}^{(1)}$ is a $mN_r \times L + m - 1$ channel matrix containing taps only from user 1 (the channel taps from other users cancel out since they do not have periodic mean property). Assuming that $\mathcal{H}^{(1)}$ is tall and has full rank, rowspan($\mathcal{Z}$) = rowspan($\mathcal{C}_\tau^{(1)}$). Thus each row of ($\mathcal{C}_\tau^{(1)}$) of dimension $1 \times P$ is in the row span($\mathcal{Z}$). Similar to [7],

$$\begin{aligned} \left[c_\tau^{(1)} \cdots c_{\tau+P-1}^{(1)}\right] &\in \text{rowspan}(\mathcal{Z}) \\ \left[c_{\tau-1}^{(1)} \cdots c_{\tau+P-2}^{(1)}\right] &\in \text{rowspan}(\mathcal{Z}) \\ \left[c_{\tau-m-L+2}^{(1)} \cdots c_{\tau+P-m-L+1}^{(1)}\right] &\in \text{rowspan}(\mathcal{Z}) \end{aligned}$$

This results in $L + m - 1$ consecutive values of τ satisfying rowspan($\mathcal{Z}$). We can use a moving average filter of order L to smooth out the local maxima essentially, if $f(\tau) = \text{argmax}_\tau \{\boldsymbol{\phi}_\tau^H C_f^{(1)H} \Pi_z C_f^{(1)} \boldsymbol{\phi}_\tau\}$ is the cost function in (8) then we form $\mathcal{J}(\tau) = f(\tau) + \cdots + f(\tau - L+m+2)$ and search for $\text{argmax}_\tau\, \mathcal{J}(\tau)$. Simulation results show that moving average operation results in global convergence.

Once the τ is computed, the beamformer coefficients are obtained through $\boldsymbol{\theta} = \mathcal{Z}_f^{\dagger} C_f^{(1)} \boldsymbol{\phi}_\tau$. This value is used as initial point $\boldsymbol{\theta} = \boldsymbol{\theta}_{STA}$ for the alternating LS as explained in section 3.5.

3.4. Prefiltering

If $\mathbf{Y} = [\mathbf{y}_{l:l-m+1} \cdots \mathbf{y}_{l+N-1:l+N-m}]$ (received antenna matrix) is rank deficient, additional nullspace solutions exist, where $l = \tau + L_{CP}$ and denotes the start of OFDM symbol. In this case there exist beamformers $\boldsymbol{\theta}_0$ such that $\mathbf{Y}^T \boldsymbol{\theta}_0 = \mathbf{0}$. These nullspace solutions lead to two independent ST beamformers $\boldsymbol{\theta}$ and $\boldsymbol{\theta} + \boldsymbol{\theta}_0$ reconstructing the same signal, and can be avoided by using a prefilter $\mathbf{F}$ (obtained from SVD($\mathbf{Y}$)), which reduces the number of rows from mN_r to $N_t(L+m-1)$. The prefilter whitens the antenna array signal $\underline{\mathbf{Y}} = \mathbf{F}^H \mathbf{Y}$. This prewhitening improves the conditioning of the received matrix, since the ST beamformers in the whitened domain are approximately orthogonal and sometimes result in the convergence of the iterative algorithms to independent solutions.

3.5. Post processing: Alternating LS

The OFDM signal in frequency domain is constant modulus (CM), this can be exploited in a block iterative algorithm. In frequency domain the source separation can be modeled as a LS problem with cost function (CP removal follows definition of $\mathbf{Y}$)

$$\underset{\hat{\mathbf{s}} \in \mathcal{CM};\, \boldsymbol{\theta}}{\text{argmin}} \|\mathcal{F}_N(\underline{\mathbf{Y}}^T \boldsymbol{\theta} - \mathbf{c}_{mod\, P}^{(1)}) - \hat{\mathbf{s}}\|$$

where $\mathcal{CM} = \{\hat{\mathbf{s}} \,\big|\, |\hat{s}_k| = 1; \text{for } k = 0 \text{ to } N-1\}$, $\hat{\mathbf{s}} = [\hat{s}^{(0)} \cdots \hat{s}^{(N-1)}]^T$ is the estimate of transmitted OFDM symbol in frequency domain, $\mathbf{c}_{mod\, P}^{(1)}$ is a repeating sequence of length N of $\mathbf{c}^{(1)}$. With an initial value of $\boldsymbol{\theta} = \boldsymbol{\theta}_{STA}$ from the STA, an alternating LS algorithm operates as follows:

(a) restricting $\mathbf{s}$ to $\mathcal{CM}$ signals estimate $\mathbf{s}$ given τ and $\boldsymbol{\theta}$:

$$\begin{aligned} \hat{\mathbf{s}} &= \mathcal{F}_N(\underline{\mathbf{Y}}^T \boldsymbol{\theta} - \mathbf{c}_{mod\, P}^{(1)}) \\ \hat{\mathbf{s}} &= \hat{\mathbf{s}} \oslash |\hat{\mathbf{s}}| \end{aligned} \quad (9)$$

(b) estimate $\boldsymbol{\theta}$ given $\hat{\mathbf{s}}$ and τ:

$$\boldsymbol{\theta} = (\underline{\mathbf{Y}}^{\dagger})^T (\mathcal{F}_N^H \hat{\mathbf{s}} + \mathbf{c}^{(1)}) \quad (10)$$

4. SIMULATION RESULTS

The performance of the subspace fitting and the alternating LS scheme is verified for a 2-user multi-user OFDM system. All users transmit constant modulus (QPSK) information sequences, superimposed with training. The training power overhead introduced by superimposed sequences is β given by $\beta = \frac{\sigma_c^2}{\sigma_c^2 + \sigma_s^2}$, where σ_c^2 is the variance of superimposed sequence $\mathbf{c}^{(1)}$. The desired and interfering users are neither synchronized, nor any form of coordination exist between them and the receiver. The information sequences, corrupted by a multi-user channel and noise is received using $N_r = 2$ antennas. We consider a L=3 Rayleigh fading channel with signal to interference ratio 0 dB. To satisfy $\mathcal{H}$ to be a tall matrix, an oversampling ratio of 2 is added [8]. We choose $N = 64$ subcarriers and $P = 16$. This period of superimposed training is used to ensure that $\mathcal{Z}$ is a tall matrix. The time varying mean is averaged over 128 OFDM symbols, which is approximately equal to the packet length of a

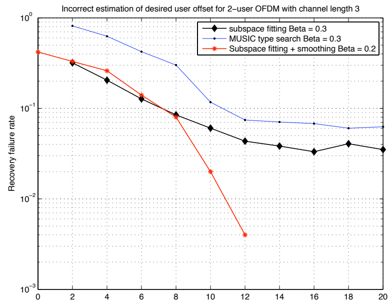

Fig. 3: *Incorrectly estimated delays for subspace fitting*

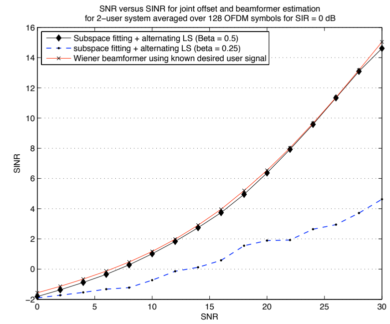

Fig. 4: *Comparison of SNR with output SINR for joint delay and source separation with SIR = 0dB*

wireless LAN system. All users transmit information sequences superimposed with periodic sequences known to the receiver. We consider a Raleigh fading channel for $N_t = 2$ users, $N_r = 2$ and signal to interference ratio (SIR) = 0 dB. An oversampling ratio of 2 is used to increase the dimensions of $\mathcal{H}$. Fig. 3 shows the recovery failure rate performance for $L = 3$ channel, where the training power overhead is $\beta = 0.3$ for subspace fitting schemes. A recovery failure occurs when the estimated τ rounded to an integer is different from the correct (integer) τ. This performance is compared with a similar algorithm but based on MUSIC type search [3] for the same channel and increased power overhead ($\beta = 0.3$). The recovery failures can be improved by performing smoothing as shown in Fig. 3.

Fig. 4 compares the SINR values of the received OFDM symbol, for different values of power overhead when the subspace fitting scheme gives correct offset estimates followed by alternating LS. The Wiener beamformer is designed assuming that signal and offset of the desired user is known. We observe that the subspace fitting scheme with $\beta = 0.5$ converges to the Wiener beamformer

Fig. 5 shows the BER performance joint offset and beamformer estimation using subspace fitting followed by alternating LS (curve 2), where the transmission starts at beginning of observation window. Curve 3 considers a continuous transmission of data. Curve 2 performs around 6 dB better than the superimposed training scheme as explained in [3].The performance of curve 3 is better than curve 2 for high SNR's, since the edge effect is removed and converges with Wiener beamformer. In the simulations, the BER computations are based on trials where the offset is correctly estimated.

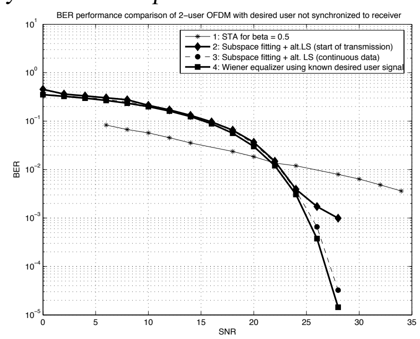

Fig. 5: *BER performance of joint offset estimation and source separation $L = 3$ and $\beta = 0.5$*

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