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Low-Complexity Single-Channel Blind Separation of Co-Frequency Coded Signals

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Abstract—Single-channel blind source separation (SCBSS) of uncoordinated, non-spread, co-frequency interfering signals with non-zero carrier frequency offset (CFO) and timing offset, and without training sequence for channel estimation, is a challenging task. Iterative SCBSS of coded signals leads to good performance but is computationally expensive as it involves turbo processing of soft per-survivor processing and soft-input soft-output (SISO) channel decoding. In this paper, we propose a low-complexity SCBSS (LC-SCBSS) scheme, which reduces the computational complexity of the conventional iterative SCBSS by using interference-cancellation (IC) from the second iteration onwards, and adaptive channel truncation (CT) for certain users. Simulation results show that by using the proposed LC-SCBSS, the complexity of the iterative SCBSS can be reduced by more than 99% of the conventional one, with only a marginal degradation in performance.

I. INTRODUCTION

Single-channel blind source separation (SCBSS) is a problem that can be found in many practical applications, such as co-frequency overlapped signal separation [1], [2], paired carrier multiple access (PCMA) signal monitoring [3], and space-based automatic identification system (AIS) signal detection [4]. Since the number of observations is less than the number of sources in SCBSS, it is an ill-conditioned mathematical problem which is very difficult to be solved. Nevertheless, SCBSS has been shown to be possible by taking advantage of certain properties of the signals, such as differences in symbol rates [5], pulse shapes [6] or acoustic time-frequency occupancy [7], etc.

However, when both the overlapped (mutually interfering) signals are digitally coded and modulated, have the same data rate, similar pulse shapes, and are transmitted in the same time-frequency plane without pilot symbols (for channel estimation), many techniques in [5]–[7] no longer work. In such situations, the per-survivor processing (PSP) algorithm presents a feasible solution to perform joint channel and data detection for each individual user [8]. When error correction codes are used, the error rate performance of the SCBSS can be further improved by iteratively exchanging extrinsic information between the PSP detector and soft-input soft-output

(SISO) channel decoder [9], [10]. However, such SCBSS processing has very high computational complexity, especially when the number of users is large and/or the modulation order is high. In this paper, a low-complexity SCBSS (LC-SCBSS) scheme is proposed to reduce the complexity of the iterative SCBSS algorithm. The proposed LC-SCBSS includes two “tricks”. Firstly, it replaces the joint multi-user PSP detector in the iterative receiver by a number of separate interference-cancellation (IC) detectors *from the second iteration onwards*. Next, it uses channel truncation (CT) to reduce the length of the estimated channel impulse response (CIR) *of certain users*, and hence reduces the number of states in the PSP algorithm significantly. Simulation results show that the proposed schemes leads to substantial complexity reduction while largely maintaining the bit error probability.

This paper is organized as follows. In Section II, the basic system model for single-channel blind separation of two co-frequency phase-shift keying (PSK) signals is introduced. In Section III, the PSP algorithm and the conventional iterative SCBSS receiver are described. In Section IV, the proposed low-complexity iterative SCBSS is presented. Simulation results are shown in Section V and conclusions are drawn in Section VI.

II. BASIC SYSTEM MODEL

To make the problem tractable, consider a two-user SCBSS system in which the received signal at time index k can be expressed as

$$y_k = h_1 e^{j(2\pi\Delta f_1 kT + \theta_1)} x_{1,k} + h_2 e^{j(2\pi\Delta f_2 kT + \theta_2)} x_{2,k} + w_k, \quad (1)$$

where h_i ($i = 1, 2$) are the amplitude of the received signals, Δf_i ($i = 1, 2$) are their carrier frequency offsets (CFOs), θ_i ($i = 1, 2$) are the initial phases, and w_k is the additive white Gaussian noise (AWGN) with zero mean and per dimension variance σ^2 . Each source signals $x_{i,k}$ is the output of a pulse shaping filter which can be expressed as

$$x_{i,k} = \sum_{m=-L}^L s_{i,m+k} g_i(-mT - \tau_i), \quad i = 1, 2 \quad (2)$$

where $s_{i,m}$ is the m th symbol transmitted by source i , $g_i(t)$ is the filter impulse response which has a raised-cosine spectrum and finite duration from $-LT$ to LT (L is an integer constant). $0 \leq \tau_i < T$ is the time invariant relative delay between the i th received signal and the local clock reference, and T is the symbol period.

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For compactness, the discrete received mixture signal y_k can also be expressed in a matrix form as

$$y_k = \mathbf{s}_k \mathbf{f}_k + w_k, \quad (3)$$

where $\mathbf{s}_k = [\mathbf{s}_{1,k} \ \mathbf{s}_{2,k}]$ and $\mathbf{f}_k = [\mathbf{f}_{1,k} \ \mathbf{f}_{2,k}]^T$, with $(\cdot)^T$ denotes matrix transpose. $\mathbf{s}_{i,k} = [s_{i,k-L}, s_{i,k-L+1}, \dots, s_{i,k+L}]$ is the i th source symbol vector at time index k , and $\mathbf{f}_{i,k}$ is the CIR vector which is defined as

$$\mathbf{f}_{i,k} = h_i e^{j(2\pi \Delta f_i k T + \theta_i)} [g_i(LT - \tau_i), \dots, g_i(-LT - \tau_i)]^T. \quad (4)$$

The goal of the receiver is to accurately detect the source symbol sequences $\mathbf{s}_{i,1:N} = [s_{i,1}, s_{i,2}, \dots, s_{i,N}]$ by using the received sequence $\mathbf{y}_{1:N} = [y_1, y_2, \dots, y_N]$. More specifically, the detector for the source symbol sequences is given by

$$\{\hat{\mathbf{s}}_{1,1:N}, \hat{\mathbf{s}}_{2,1:N}\} = \arg \max_{\substack{\tilde{\mathbf{s}}_{1,1:N} \\ \tilde{\mathbf{s}}_{2,1:N}}} \Pr(\tilde{\mathbf{s}}_{1,1:N}, \tilde{\mathbf{s}}_{2,1:N} | \mathbf{y}_{1:N}) \quad (5)$$

where $\tilde{\mathbf{s}}_{i,1:N}$ is a trial of $\mathbf{s}_{i,1:N}$. At the receiver side, if the knowledge of the CIR $\mathbf{f}_k$ is available, the conventional Viterbi algorithm [11] can be directly employed for multi-user detection. However, in this paper, we are focused on *blind* receiver design which assumes that the CIR $\mathbf{f}_k$ is unknown. To this end, a joint detection of $\mathbf{s}_k$ and $\mathbf{f}_k$ is necessary and will be carried out by using the PSP algorithm.

III. PSP ALGORITHM AND ITERATIVE SCBSS RECEIVER

In the PSP algorithm, the symbol detection and channel estimation are performed alternately. In every stage of the PSP algorithm, the Viterbi algorithm is used for the symbol detection, based on the estimated CIR in the previous stage. Define the state at time index k in the PSP algorithm as $\phi_k = [\bar{\mathbf{s}}_{k-L+1}, \bar{\mathbf{s}}_{k-L+2}, \dots, \bar{\mathbf{s}}_{k+L}]$, where $\bar{\mathbf{s}}_k = [s_{1,k}, s_{2,k}]$ is the pair of symbols transmitted at time index k from sources 1 and 2 respectively. At time index k , the path metric at state ϕ_k is defined as $P_k(\phi_k)$ and the branch metric associated with the state transition from ϕ_{k-1} to ϕ_k is denoted by $m(\phi_{k-1} \rightarrow \phi_k)$. The PSP algorithm includes the following steps:

- 1) Initialization of the CIR ($\mathbf{f}_0$). For each state ϕ_0 , initialize the metric as $P_0(\phi_0) = 0$.
- 2) Calculate the branch metrics at time index k by

$$m(\phi_{k-1} \rightarrow \phi_k) = |e(\phi_{k-1} \rightarrow \phi_k)|^2, \quad (6)$$

where

$$e(\phi_{k-1} \rightarrow \phi_k) = y_k - \mathbf{s}_k(\phi_{k-1} \rightarrow \phi_k) \mathbf{f}_k. \quad (7)$$

In (7), $\mathbf{s}_k(\phi_{k-1} \rightarrow \phi_k) = [s_{1,k}, s_{2,k}]$ is the symbol vector associated with the state transition $\phi_{k-1} \rightarrow \phi_k$. The accumulated metrics at time index k are given by

$$P_k(\phi_k) = \min_{\phi_{k-1}} (P_{k-1}(\phi_{k-1}) + m(\phi_{k-1} \rightarrow \phi_k)). \quad (8)$$

- 3) Update the CIR vector $\mathbf{f}_k$ by the least mean square (LMS) algorithm [8], which is given by

$$\mathbf{f}_k = \mathbf{f}_{k-1} + \eta \cdot e(\phi_{k-1} \rightarrow \phi_k) \cdot \mathbf{s}_k(\phi_{k-1} \rightarrow \phi_k)^H, \quad (9)$$

where η is the step size and $(\cdot)^H$ denotes Hermitian transposition.

- 4) Set $k \leftarrow k + 1$ and go back to step 2 or stop if $k = N$.

When error correction codes are used, the performance of the SCBSS can be further improved by iteratively exchanging extrinsic information between the PSP detector and the SISO channel decoder. A block diagram of the conventional iterative SCBSS receiver is shown in Fig. 1(a). It includes a joint PSP detector and two SISO decoders. To enable the PSP detector to provide soft outputs to the SISO decoder, soft-output Viterbi algorithm (SOVA) [11] can be used in the PSP algorithm. The soft outputs, i.e., the extrinsic information $L_e(s_{1,k})$ and $L_e(s_{2,k})$, are deinterleaved and used as the *a priori* information by the SISO decoders to decode bits from sources 1 and 2 respectively. The outputs of the SISO decoder are the decoded bits $\hat{s}_{1,k'}$ and $\hat{s}_{2,k'}$, and the extrinsic information $L_e(s_{1,k'})$ and $L_e(s_{2,k'})$. They are interleaved and fed back to the PSP detector as the *a priori* information. The above iteration continues until a stopping criterion is reached.

IV. PROPOSED SIMPLIFIED ITERATIVE SCBSS

As shown in Fig. 1(a), a joint PSP detector is used by the receiver to separate and demodulate signals from uncoordinated interfering users. Suppose each user uses the same PSK modulation scheme, then the number of states in the PSP algorithm is M^{2LU} , where U is the number of users and M is the PSK alphabet size. It is clear that the complexity of the entire receiver is dominated by the joint PSP detector complexity which is prohibitively high for large M , U , or L . When iterative SCBSS is performed, the complexity will be increased by N_{iter} times, where N_{iter} is the iteration number. To reduce the complexity of the iterative receiver, we propose to replace the joint PSP detector by U separate interference-cancellation PSP (IC-PSP) detectors *after the first iteration*, as shown in Fig. 1(b) for $U = 2$ users (which can be easily extended to $U > 2$).

Referring to Fig. 1(b), in the detection of each individual signal, the interference from the other user is removed by using its decoded bits from the previous iteration. Suppose that after the first iteration, the estimated CIR of user i is $\hat{\mathbf{f}}_{i,k} = [\hat{f}_{i,k-L}, \hat{f}_{i,k-L+1}, \dots, \hat{f}_{i,k+L}]^T$. The received signals after IC are

$$\tilde{y}_{1,k} = y_k - \hat{\mathbf{s}}_{2,k} \hat{\mathbf{f}}_{2,k} \quad (10)$$

and

$$\tilde{y}_{2,k} = y_k - \hat{\mathbf{s}}_{1,k} \hat{\mathbf{f}}_{1,k} \quad (11)$$

respectively, where $\hat{\mathbf{s}}_{i,k} = [\hat{s}_{i,k-L}, \hat{s}_{i,k-L+1}, \dots, \hat{s}_{i,k+L}]$ is the decoded source symbol vector of the i th user at time index k . (10) and (11) can be easily extended to $U > 2$ users and the signals after IC at the i th detector can be expressed as

$$\tilde{y}_{i,k} = y_k - \sum_{u=1, u \neq i}^U \hat{\mathbf{s}}_{u,k} \hat{\mathbf{f}}_{u,k}. \quad (12)$$

Then, $\tilde{\mathbf{y}}_{i,1:N} = [\tilde{y}_{i,1}, \tilde{y}_{i,2}, \dots, \tilde{y}_{i,N}]$, ($i = 1, 2, \dots, U$) will be used by the IC-PSP detectors to perform detection of symbols from users 1 to U respectively. Since in each separate detector, only one user is considered, the total number of states in the

Fig. 1. Structures of the conventional and proposed iterative SCBSS receivers

PSP algorithm is reduced to $U \cdot M^{2L}$, which is much lower than that of the joint PSP detector.

It should be noted that in our proposed LC-SCBSS, we only replace the joint PSP detector with parallel single-user PSP detectors *from the second iterations onwards*. To further reduce complexity of LC-SCBSS, especially for the first iteration, we propose an adaptive trellis reduction scheme based on channel truncation (CT). Our idea is based on the observation in [2], which showed that even when $g_i(t)$ and the channel condition of each user are the same, when there is a relative delay between the users, the estimated CIRs will be different for different users. This implies that we do not need to use the same value of L for all the users. For a certain user with a specific sampling instance, it is possible that the energy at some channel taps are very small (say, smaller than a certain threshold Tr) and can be ignored. Suppose at time index k , most of the energy of $s_{i,k}$ are concentrated on the $k - L_{i,1}$ to $k + L_{i,2}$ ($L_{i,1}, L_{i,2} \leq L$) channel taps. If $L_{i,1} < L$ or $L_{i,2} < L$, the length of the CIR can be truncated and consequently, the number of states in the PSP algorithm can be reduced.

In practical situation, since the CIR is unknown when detection starts, we first use the same value of L for all the users. After the CIRs of all the users are estimated (this usually could be achieved after processing a few hundred symbols), we can start to perform CT and reduce the number of states in the trellis. To make the trellis state reduction smooth, we propose to perform CT by keeping a subset of the original trellis states, which is determined by the values of $L_{i,1}$ and $L_{i,2}$. In the following, we use a simple example to illustrate how this can be achieved. In the example, we assume two-user case and in the start of the detection, we use $L = 1$ for all the users. After CT, we have $L_{1,1} = L_{2,1} = 1$ and $L_{1,2} = L_{2,2} = 0$. Before CT, the CIR is $\mathbf{f}_{i,k} = [f_{i,k-1}, f_{i,k}, f_{i,k+1}]^T$ and the

states at time index k in the PSP algorithm are defined as $\phi_k = [s_{1,k}, s_{2,k}, s_{1,k+1}, s_{2,k+1}]$. After CT, the CIR is shortened to $\mathbf{f}'_{i,k} = [f_{i,k-1}, f_{i,k}]^T$ and the states in the PSP algorithm reduce to $\phi'_k = [s_{1,k}, s_{2,k}]$. It means that the states in the original trellis with the same $s_{1,k}$, $s_{2,k}$ and different $s_{1,k+1}$, $s_{2,k+1}$ can be merged into one state, defined just by the values of $s_{1,k}$ and $s_{2,k}$. For example, the states [0000], [0001], [0010] and [0011] are merged into one state [00], as illustrated in Fig. 2. It should be noted that after state reduction, the state transitions between neighbouring time states will change. For example, before state reduction, there is no state transition from state [1111] to state [0000]. But after state reduction, there is a transition from state [11] to state [00].

Fig. 2. Trellis in PSP algorithm before and after state reduction.

In Table I, the complexity of conventional SCBSS and the proposed LC-SCBSS schemes are compared.

TABLE I
THE COMPLEXITY OF THE CONVENTIONAL AND PROPOSED ALGORITHMS.

Algorithm	Complexity
Conventional SCBSS	$O(N_{iter} \cdot M^{2LU})$
Proposed LC-SCBSS	$O(M \sum_{i=1}^U (L_{i,1} + L_{i,2}) + (N_{iter} - 1) \sum_{i=1}^U M^{L_{i,1} + L_{i,2}})$

V. SIMULATION RESULTS

In our simulations, we consider two users, both employing QPSK modulation ($M = 4$). The channel gains of the two modulated signals are set to be the same, i.e., $h_1 = h_2 = 1$, to represent the worst-case situation when the two users interfere with each other to the largest extent. The signal-to-noise ratio (SNR) is defined by $10\log(2/\sigma^2)$. $g_1(t)$ and $g_2(t)$ are raised-cosine pulses with a roll-off factor 0.33. Before CT, we use $L = 2$. The initial phases of the two signals are randomly chosen to be 0.8 and 0.5, respectively. The CFOs Δf_1 and Δf_2 are assumed to be 1kHz [12]. The relative delay between the two signals is $\Delta\tau = \tau_2 - \tau_1 = 0.4T$. The well known rate 1/2 convolutional (133,171) codes are used by both the users. The modulated symbols are transmitted in blocks and each block contain 5000 symbols. h_i, θ_i, g_i and τ_i are assumed to be unknown at the receiver side. For each block, we perform CT after 500 symbols. Suppose the receiver sample at time instance when $\tau_2 = 0$. If not considering the CFOs, the CIR for user 1 and user 2 can be found to be $\mathbf{f}_{1,k} = [0.05 + 0.05i, -0.12 - 0.13i, 0.52 + 0.53i, 0.34 + 0.35i, -0.10 - 0.11i]$ and $\mathbf{f}_{2,k} = [-0.0008 - 0.0005i, 0.0009 + 0.0005i, 0.88 + 0.48i, 0.0009 + 0.0005i, -0.0008 - 0.0005i]$, respectively. Suppose the threshold for CT is $\text{Tr} = 0.001$. After CT, we have

Fig. 3. Comparison of data detection performance.

$L_{1,1} = L_{1,2} = 2$ and $L_{2,1} = L_{2,2} = 0$. The iteration number is set to $N_{iter} = 4$, as more iterations only lead to marginal bit error rate (BER) improvement (shown as converged BER for both SCBSS and LC-SCBSS in Fig. 3). According to Table I, the complexity of the conventional SCBSS and proposed LC-SCBSS are $O(2.6 \times 10^5)$ and $O(1.0 \times 10^3)$ respectively, which shows that the complexity of the conventional SCBSS is reduced by more than 99%.

In Fig. 3, the average BERs of the 2 users obtained by the conventional SCBSS and the proposed LC-SCBSS are compared. It can be observed that the BER obtained by using the proposed LC-SCBSS is almost the same as that obtained by using the conventional SCBSS scheme.

In Fig. 4, the CIR estimation accuracy measured in terms of mean square error (MSE) of the estimation is shown for SNR=17dB. The measurement settings are the same as those used in [8]. It can be observed that after one iteration, the CIR estimation accuracy obtained by both our conventional SCBSS and proposed LC-SCBSS scheme is similar to that shown in [8]. After two iterations, CIR estimation accuracy obtained by us is better than that in [8], because we use joint channel estimation and decoding, where the channel decoder and the channel estimator perform iterative (turbo) exchange of soft information to improve each other. Paper [8] did not consider channel coding. Beyond two iterations, there is almost no more incremental gain on the CIR accuracy.

VI. CONCLUSION

In this paper, an iterative low-complexity SCBSS (LC-SCBSS) scheme is proposed to reduce the complexity of blindly separating overlapped, uncoordinated, misaligned, coded digital signals without training sequence for channel estimation. The LC-SCBSS firstly replaces the joint PSP detector in the iterative receiver by U separate IC-PSP detectors from the *second iteration onwards*, then uses adaptive channel truncation to further reduce the number of states in the PSP algorithm for users *with weak estimated channel taps*. The simulation results, in conjunction with Table I, show that by

Fig. 4. Comparison of channel estimation accuracy.

using the proposed LC-SCBSS scheme, the complexity of the iterative SCBSS can be drastically reduced with almost no degradation of BER performance and CIR estimation accuracy. In our future work, we will make further investigation on LC-SCBSS schemes for larger ($U > 2$) number of users with higher order modulation orders ($M > 4$).

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