# LOW-COMPLEXITY SUCCESSIVE INTERCELL INTERFERENCE CANCELLATION

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#### ABSTRACT

Due to the data traffic explosion, cellular networks tend to have larger cell density and smaller cell size, which will give rise to significant intercell interference. In this work, we focus on receiver side signal processing techniques to perform intercell interference suppression and successive interference cancellation, where adaptive antenna arrays are used to develop a receiver beamforming approach within an iterative receiver. Specifically, the matrix inversion for calculating the beamforming weights is avoided, which drastically reduce the computational complexity.

*Index Terms*— OFDM, turbo processing, interference rejection combining, successive interference cancellation, adaptive antenna array

## 1. INTRODUCTION

Due to the data traffic explosion, cellular networks tend to have larger cell density and smaller cell size, which will give rise to significant intercell interference. There are various interference management schemes to improve the system performance. We focus on receiver side signal processing techniques to effectively combat the performance degradation due to the presence of strong interference signals. Specifically, receiver optimum combining [1], also known as interference rejection combining (IRC), combined with turbo receivers [2] is one of the most promising approaches.

An advanced receiver for signal detection in the presence of strong intercell interference has been developed in [3], in the context of 3GPP long term evolution (LTE) downlink transmission, which is based on orthogonal frequencydivision multiplexing (OFDM). It has been shown in [3] that, per-carrier beamforming is preferred for receivers experiencing highly frequency-selective channel, where each subcarrier has its own beamforming weight vector. However, each beamforming weight calculation requires one matrix inverse operation, which gives rise to high computational complexity, due to the large number of subcarriers.

In this work, we develop a low-complexity algorithm which avoids the matrix inversion operation when calculating the beamforming weight. The computational complexity could be drastically reduced especially for receivers equipped with large number of receive antennas. The proposed algorithm initializes the beamforming weight using reference signal only, since the signal of interest should be excluded to calculate the beamforming vector [4]. In the later iterations, successive interference cancelation (SIC) is applied to remove the previously estimated signal, and the beamforming weight vectors are updated in each iteration using IRC, which take into account the soft information from the channel decoder and previously estimated interfering signals.

*Notation:* We use capital (lower case) boldface letters to denote matrices (column vectors).  $(\cdot)^T$  and  $(\cdot)^H$  represent transpose and complex conjugate transpose (Hermitian), respectively.  $E(\cdot)$  stands for the statistical expectation. The covariance matrix between **x** and **y** is defined as  $Cov(\mathbf{x}, \mathbf{y}) = E(\mathbf{x}\mathbf{y}^H) - E(\mathbf{x})E(\mathbf{y}^H)$ . Finally,  $\mathbf{I}_N$  denote the  $N \times N$  identity matrix.

## 2. SYSTEM MODEL

We consider a single user equipment (UE) receiving signal from the serving base station (BS), as well as interference signals from neighboring BSs. BSs from different cells are assumed to be synchronized in time and frequency, and each BS has only one antenna to transmit one data stream. We consider a LTE downlink transmission system based on OFDM [5], with the minimum scheduling unit being one resource block, which contains  $N_{\rm sc}^{\rm RB} = 12$  subcarriers and  $N_{\rm symb}^{\rm DL} = 14$ OFDM symbols, and one subframe consists of  $N_{RB}$  resource blocks.

The channel is assumed to be constant during one subframe in time, while channel fading is characterized by different channel models in frequency. In this work, the receiver is assumed to have perfect channel state information (CSI) between the serving BS and the UE, as well as the interfering BSs and the UE to perform SIC. In order to reduce the decoding complexity and signaling overhead, only the desired signal will be decoded, and the interference signal is detected at symbol level without further decoding.

The transceiver diagram is shown in Fig. 1. At the transmitter, a sequence of bits is encoded with error correction coding, and the coded bits are interleaved and mapped into complex data symbols. The reference signals are multiplexed with data symbols to form the time-frequency grid for

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**Fig. 1**. System model for the iterative OFDM receiver with combined interference suppression and cancellation.

transmission. For simplicity, we consider unit-energy quaternary phase-shift keying (QPSK) with the symbol alphabet  $\alpha_i, i = 1, ..., 4$ , which are used for desired signal as well as all other interference signals, extension to other constellations is straightforward [6].

The proposed iterative receiver is implemented with SIC, where the desired signal as well as a certain number of dominant interference signals are detected. For a discrete-time baseband model, after cyclic prefix (CP) removal and FFT at the receiver, the system input-output relationship can be written as

$$\mathbf{y}_{k,l} = \sum_{j=0}^{X'} \mathbf{h}(j) s_{k,l}(j) + \mathbf{n}_{k,l}$$
(1)

where X' is the number of interfering signals,  $s_{k,l}(j)$  is the transmitted signal from the *j*th BS on subcarrier *k* of the *l*th OFDM symbol;  $\mathbf{y}_{k,l} = [y_{k,l}^1, \ldots, y_{k,l}^{N_R}]^T$ ,  $N_R$  is the number of receive antennas, and  $y_{k,l}^m$  is the received signal at the *m*th received antenna;  $\mathbf{h}(j) = [h_k^1(j), \ldots, h_k^{N_R}(j)]^T$  and  $h_k^m(j)$  is the channel coefficient between the *j*th BS's transmit antenna and the *m*th receive antenna on the *k*th subcarrier;  $\mathbf{n}_{k,l} = [n_{k,l}^1, \ldots, n_{k,l}^{N_R}]^T$ , and  $n_{k,l}^m$  stands for the additive temporally and spatially white Gaussian noise at the *m*th receive antenna, which is uncorrelated with the data symbols from serving and interfering BSs, and with zero mean and variance  $\sigma_n^2$ . The index is ordered according to their signal power at the receiver, and  $\mathbf{h}(0)s_{k,l}(0)$  is assumed to have the largest received signal-to-noise ratio (SNR).

### 3. ITERATIVE INTERFERENCE SUPPRESSION AND CANCELLATION

In [3], per-carrier and block beamforming approaches have been considered for iterative receiver based on OFDM systems. In this section, we will briefly describe the per-carrier beamforming algorithm which has high complexity due to the matrix inversion operation, and then we develop a lowcomplexity receiver beamforming approach which avoids the matrix inversion.

The received signal is processed by a set of beamforming weight vectors in each iteration, where the transmitted symbols are estimated and then subtracted from the received signal successively. The beamforming weight is calculated based on the received reference signals only in the first iteration, and then updated using previously detected data symbols.

At the *q*th iteration, the updated unbiased estimate of  $s_{k,l}(x)$  can be expressed as [1]

$$\hat{s}_{k,l}(x) = \mathbf{w}_{k,l}^H \bar{\mathbf{y}}_{k,l} = \frac{\mathbf{h}^H(x) \mathbf{R}_{k,l}^{-1} \bar{\mathbf{y}}_{k,l}}{\mathbf{h}^H(x) \mathbf{R}_{k,l}^{-1} \mathbf{h}(x)}$$
(2)

where  $\mathbf{w}_{k,l}$  is the beamforming weight vector for estimating symbol  $s_{k,l}(x)$  at the *q*th iteration.  $\bar{\mathbf{y}}_{k,l}$  is the processed receive signal for estimating symbol  $s_{k,l}(x)$  at the *q*th iteration, which is obtained by subtracting the preciously detected signals, and can be written as

$$\bar{\mathbf{y}}_{k,l} = \mathbf{y}_{k,l} - \sum_{j=0}^{X} \mathbf{h}(j) m_{k,l}(j), j \neq x$$
(3)

where X is the number of signals that have been detected at that stage, and  $m_{k,l}(j)$  is the mean of the symbol  $s_{k,l}(j)$ at that stage, which is calculated from soft information. The vector  $\bar{\mathbf{y}}_{k,l}$  can be regarded as the sum of noise, other interference signals which have not been detected at the *q*th iteration when estimating symbol  $s_{k,l}(j)$ , and residual interference after SIC. It can be further written as

$$\bar{\mathbf{y}}_{k,l} = \sum_{j=0, j \neq x}^{X} \mathbf{h}(j) [s_{k,l}(j) - m_{k,l}(j)] + \sum_{j=X+1}^{X'} \mathbf{h}(j) s_{k,l}(j) + \mathbf{n}_{k,l}$$
(4)

 $\bar{\mathbf{y}}_{k,l}$  is assumed to be a temporally white and spatially correlated vector with complex normal distribution  $\mathcal{CN}(\mathbf{0}, \mathbf{R}_{k,l})$ .

It has been shown in [3] that per-carrier beamforming approach is preferred for receivers experiencing highly frequency-selective channel, where each subcarrier has its own beamforming weight vector. However, each beamforming weight calculation requires one matrix inverse operation as in (2), which gives rise to high computational complexity, due to the large number of subcarriers. In the following, we develop a low-complexity algorithm to calculate the beamforming weight without matrix inversion, which will reduce the complexity drastically especially for receivers equipped with large number of receive antennas.

#### 4. LOW COMPLEXITY ALGORITHM

We rewrite the system input-output relationship in (3) as

$$\bar{\mathbf{y}}_{k,l} = \mathbf{h}(x)s_{k,l}(x) + \mathbf{h}(u)[s_{k,l}(u) - m_{k,l}(u)] + \eta_{k,l} \quad (5)$$

where we only consider one dominant interferer u after SIC at that stage, the other interference signals are treated as additional temporally and spatially white Gaussian noise. The effective noise term  $\eta_{k,l}$  is assumed to have a complex normal distribution  $\mathcal{CN}(\mathbf{0}, \sigma_{\eta}^{2}\mathbf{I}_{N_{R}})$ . In this way, the covariance matrix  $\mathbf{R}_{k,l}$  in (2) can be written as

$$\mathbf{R}_{k,l} = \mathbf{h}(u)|s_{k,l}(u) - m_{k,l}(u)|^{2}\mathbf{h}^{H}(u) + \sigma_{\eta}^{2}\mathbf{I}_{N_{R}}$$
$$= \mathbf{h}(u)v_{k,l}(u)\mathbf{h}^{H}(u) + \sigma_{\eta}^{2}\mathbf{I}_{N_{R}}$$
(6)

where  $v_{k,l}(u)$  is the variance of the estimated symbol  $\hat{s}_{k,l}(u)$  at the *q*th iteration.

The one dominant interferer u can be chosen according to the order of SIC, which depends on the received signal power, or according to a certain metric, for example by comparing the residue interference power at that stage. The variance of the effective noise  $\eta_{k,l}$  can be calculated as

$$\sigma_{\eta}^{2} = \sigma_{n}^{2} + \sum_{j=0}^{X} v_{k,l}(j) |\mathbf{h}(j)|^{2}, j \neq x, j \neq u$$
(7)

which takes into account the residue interference after SIC.

By applying the matrix inversion lemma, the inverse of the covariance matrix  $\mathbf{R}_{k,l}$  can be expressed as

$$\mathbf{R}_{k,l}^{-1} = \left[\mathbf{h}(u)v_{k,l}(u)\mathbf{h}^{H}(u) + \sigma_{\eta}^{2}\mathbf{I}_{N_{R}}\right]^{-1}$$

$$= \frac{1}{\sigma_{\eta}^{2}}\mathbf{I}_{N_{R}} - \frac{1}{\sigma_{\eta}^{4}}\mathbf{h}(u)\left[v_{k,l}^{-1}(u) + \frac{\mathbf{h}^{H}(u)\mathbf{h}(u)}{\sigma_{\eta}^{2}}\right]^{-1}\mathbf{h}^{H}(u)$$

$$= \frac{1}{\sigma_{\eta}^{2}}\left(\mathbf{I}_{N_{R}} - \frac{\mathbf{h}(u)\mathbf{h}^{H}(u)}{\sigma_{\eta}^{2}/v_{k,l}(u) + |\mathbf{h}(u)|^{2}}\right)$$
(8)

For per-carrier beamforming approach, we use one covariance matrix to approximate all  $\mathbf{R}_{k,l}$  for all OFDM symbols corresponding to subcarrier k, which can be obtained by averaging the symbol variance  $v_{k,l}(u)$  in the time domain, and it can be expressed as

$$\mathbf{R}_{k}^{-1} = \frac{1}{\sigma_{\eta}^{2}} \left( \mathbf{I}_{N_{R}} - \frac{\mathbf{h}(u)\mathbf{h}^{H}(u)}{\sigma_{\eta}^{2}/v_{k}(u) + |\mathbf{h}(u)|^{2}} \right) \tag{9}$$

where  $v_k(u) = \sum_{l=1}^{N_{\text{symb}}^{\text{DL}}} v_{k,l}(u) / N_{\text{symb}}^{\text{DL}}$ . The beamforming weight for the *k*th subcarrier can be subsequently calculated according to (2), where matrix inversion is avoided by applying the new noise plus interference signals' covariance matrix in (9).

For the desired signal, the soft information of the estimated symbol from symbol detector will be sent to the channel decoder. For simplicity, we drop the corresponding user index x, and denote the desired signal as  $s_{k,l}$ , with  $(s_{k,l,1}, s_{k,l,2})$  as the related bits. After each iteration of the detector and decoder, we update the mean using the soft estimated symbols. Specifically, we need to calculate the extrinsic log-likelihood ratio (LLR),  $L_e(s_{k,l,g}) = L(s_{k,l,g}|\hat{s}_{k,l}) - L(s_{k,l,g})$ , where  $L(s_{k,l,g})$  is the *a priori* LLR and  $L(s_{k,l,g}|\hat{s}_{k,l})$  is the *a posteriori* LLR [2], which are defined as

$$L(s_{k,l,g}) = \ln \frac{P(s_{k,l,g} = 0)}{P(s_{k,l,g} = 1)}$$
$$L(s_{k,l,g}|\hat{s}_{k,l}) = \ln \frac{P(s_{k,l,g} = 0|\hat{s}_{k,l})}{P(s_{k,l,g} = 1|\hat{s}_{k,l})}$$

where g = 1, 2,  $\hat{s}_{k,l}$  is the estimated symbol at the current iteration,  $P(s_{k,l,g} = 0)$  and  $P(s_{k,l,g} = 1)$  are the a prior probabilities, and  $P(s_{k,l,g} = 0|\hat{s}_{k,l})$  and  $P(s_{k,l,g} = 1|\hat{s}_{k,l})$  are the a posteriori probabilities.

In order to calculate  $L_e(s_{k,l,g})$ , the probability density function (PDF)  $p(\hat{s}_{k,l}|s_{k,l} = \alpha_i)$  can be approximated as Gaussian:

$$p(\hat{s}_{k,l}|s_{k,l} = \alpha_i) = \frac{1}{\pi \sigma_{k,l,i}^2} \cdot e^{-|\hat{s}_{k,l} - \mu_{k,l,i}|^2 / \sigma_{k,l,i}^2}$$
(10)

with mean  $\mu_{k,l,i} = E(\hat{s}_{k,l}|s_{k,l} = \alpha_i)$  and variance  $\sigma_{k,l,i}^2 = Cov(\hat{s}_{k,l}, \hat{s}_{k,l}|s_{k,l} = \alpha_i)$  [2, 6]. The mean  $\mu_{k,l,i}$  and variance  $\sigma_{k,l,i}^2$  can be derived from (2) as

$$\mu_{k,l,i} = \alpha_i$$
  
$$\sigma_{k,l,i}^2 = 1/(\mathbf{h}^H \mathbf{R}_k^{-1} \mathbf{h})$$
(11)

where  $\mathbf{R}_k$  is the covariance matrix of  $\mathbf{u}_{k,l}$  at the *q*th iteration for  $\mathbf{u}_{k,l}$  corresponding to the *k*th subcarrier in (9). The extrinsic LLR can be subsequently calculated as [3]

$$L_e(s_{k,l,1}) = \sqrt{8} \operatorname{Re}(\hat{s}_{k,l})(\mathbf{h}^H \mathbf{R}_k^{-1} \mathbf{h})$$
  

$$L_e(s_{k,l,2}) = \sqrt{8} \operatorname{Im}(\hat{s}_{k,l})(\mathbf{h}^H \mathbf{R}_k^{-1} \mathbf{h})$$
(12)

As shown in Fig. 1, the extrinsic LLR  $L_e(s_{k,l,g})$  is passed to the decoder to generate a new extrinsic LLR  $L_e^d(s_{k,l,g})$ , and it is added to the a priori LLR to form the a posteriori LLR, which is used to update the mean of the estimated symbol as in [2,7]:

$$L_{\text{new}}(s_{k,l,g}) = L(s_{k,l,g}) + L_e^d(s_{k,l,g})$$

$$m_{k,l,\text{new}} = \frac{\tanh(\frac{L_{\text{new}}(s_{k,l,1})}{2}) + i \cdot \tanh(\frac{L_{\text{new}}(s_{k,l,2})}{2})}{\sqrt{2}}$$

$$v_{k,l,\text{new}} = 1 - |m_{k,l,\text{new}}|^2$$
(13)

The above calculation is for the desired signal, whose soft information is passed to the channel decoder for decoding the information bits. For interference signals, only the extrinsic information  $L_e(s_{k,l,g})$  is used to update the mean value in (13), which is calculated in the same way as (12).

#### 5. SIMULATION RESULTS

In this section, the proposed algorithms are examined and compared by simulations. The receiver is assumed to have two receive antennas. The Extended Vehicular A (EVecA) channel models [8] are used for simulation, which has a large number of channel taps and long delay spread. The signalto-interference ratio (SIR) for the xth interferer is defined as SIR<sub>x</sub> =  $1/E\{|h_t^x|^2\}$ , where  $|h_t^x|^2$  is the energy of all the channel taps in time domain, and the SNR is defined as  $SNR = 1/E\{\sigma_n^2\}$ . For all the serving and interfering BSs, the channel coefficients between the BS transmit antenna and the receive antennas are i.i.d.. A rate 1/3 turbo code is used at the transmitter, and the decoder at the receiver employs a constant log-MAP decoding algorithm. For each detector and decoder iteration, the turbo decoder performs two iterations within the decoder. The signal which has larger received power will be detected first and then canceled.

In Fig. 2, we show the frame error rate (FER) where two interferers are present with  $SIR_1 = -5 \, dB$  and  $SIR_2 =$  $+5 \,\mathrm{dB}$  respectively. The FER performance for three iterations is plotted, where both algorithms have the same performance in the first iteration. It is shown that the developed low-complexity per-carrier beamforming algorithm has similar FER performance as the original one, but with reduced complexity. After the third iteration, the low-complexity algorithm even achieves better FER performance than the exact solution. Compare to the per-carrier beamforming approach, the low-complexity algorithm only takes into account single dominant interference signal. This might give rise to performance degradation in the early iteration when the residue interference still has large power, as the interference signals are spatially colored but being approximated as additive spatially white Gaussian noise. However, there is only a few dominant interferers in cellular systems, and the simulation results show that the interference signals can be largely removed after the first iteration. On the other hand, the covariance matrix  $\mathbf{R}_{k,l}$  might be a singular matrix, which causes large error when performing matrix inversion operation. Thus, the developed low-complexity per-carrier beamforming algorithm is expected to have similar performance as the original one, as validated in the simulation results.

In the simulation results we don't show here due to space limitation, we observe that if both interferers have SINR close to 0dB, the receiver will not be able to decode the desired signal. As there is not enough degrees of freedom in spatial domain to reject the interference at first iteration, and the subsequent SIC cannot remove the interference signal. In this case, the MIMO theory requires that the number of receive antennas should be at least equal to the number of dominant interferers minus one.



Fig. 2. FER performance comparison for system with two dominant interferers,  $SIR_1 = -5 dB$  and  $SIR_2 = +5 dB$ .

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