ITERATIVE SUPPRESSION OF CO-CHANNEL INTERFERENCE

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ABSTRACT

Co-Channel Interference (CCI) is one of the major bottlenecks in the performance of today's densely planned wireless cellular systems. A receiver equipped with multiple receive antennas and sufficient computational budget typically exploits the spatial dimension and iterative processing between the detector and the channel decoder to suppress this interference. In this paper, we propose modifications in the standard iterative turbo processing based CCI suppression to especially boost the receiver performance in a dominant CCI scenario. Performance gains of 3-5 dB are shown to be achieved over the conventional iterative approaches for an LTE uplink system.

***Index terms* — Iterative Receiver, Turbo Equalization, Interference Suppression, Co-Channel Interference**

1. INTRODUCTION

A growing demand for better spectral efficiency in wireless cellular systems pushes for a diminishing spectrum reuse factor. This allows, on one hand, the same set of resources to be reused in a much smaller neighborhood but on the other hand leads to rising *Co-Channel Interference (CCI)* levels that ultimately limit the performance of conventional receivers.

Having multiple receive antennas allows a receiver to suppress the co-channel interference via exploitation of its spatial correlation. Additional redundancy in transmission, for instance, by channel coding further helps an interfered receiver to improve its detection performance. Among the two extremes of the optimal joint detector and the linear equalization-based detector, iterative detection [1, 2] offers a nice trade-off between performance and complexity.

In this paper, we propose a few modifications to the conventional turbo processing based iterative receiver to boost its performance especially in CCI limited scenarios. Precisely speaking, we propose improvements in the estimation of interference plus noise covariance matrix and suggest better exploitation of this information for iterative channel estimation. Furthermore, we propose to revise the estimate of interference plus noise covariance matrix (to be employed in detection) by incorporating the fresh channel estimation errors in each iteration. The effectiveness of the proposed modifications is analyzed and confirmed via *EXtrinsic Information Transfer (EXIT)* charts and block error rate curves. A performance improvement of 3-5 dB is achieved over the conventional iterative receiver for an LTE uplink system in a dominant CCI scenario.

2. SYSTEM MODEL

Although the proposed set of modifications is applicable to any iterative receiver, the description as well as the performance analysis in this paper is adapted to a *Discrete Fourier Transform Spread Orthogonal Frequency Division Multiplexing (DFT-SOFDM)* system such as the one in LTE uplink specifications [3]. Owing to the conventional OFDM processing, the system model diagonalizes in the frequency domain [4]. Incorporating the additional DFT block at transmitter (having single transmit antenna) and the corresponding *Inverse DFT (IDFT)* block at receiver (having M receive antennas), the overall system model at m-th receive antenna is depicted via the block diagram in Fig. 1. With $\boldsymbol{x} \in \mathbb{C}^{N_c}$ denoting the vector of transmitted frequency domain samples of the desired user, the received frequency domain vector at m-th antenna $\boldsymbol{y}_m \in \mathbb{C}^{N_c}$ can therefore be expressed as[1]

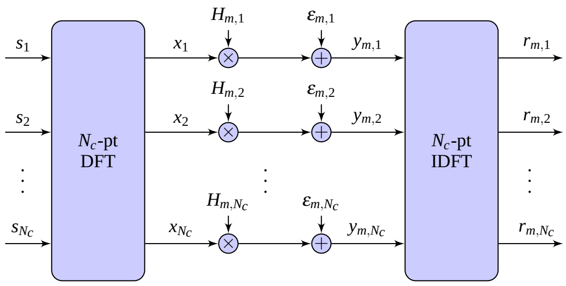


Figure 1: LTE Uplink (DFT-SOFDM) System Model. N_c denotes the number of OFDMA sub-carriers assigned to the desired user.

$$\boldsymbol{y}_m(\ell) = \boldsymbol{H}_m(\ell)\boldsymbol{x}(\ell) + \boldsymbol{\varepsilon}_m(\ell) \tag{1}$$

where $\ell \in \mathscr{L}$ is the OFDM block index and $\boldsymbol{H}_m \in \mathbb{D}^{N_c}$ is the diagonal frequency domain channel matrix containing the *Channel Frequency Response (CFR)* coefficients of the desired user along its main diagonal while $\boldsymbol{\varepsilon}_m$ denotes the impairment vector. Besides the thermal white Gaussian noise, an additional impairment that we consider in our system is the presence of strong co-channel interference. In the context of uplink cellular communications, the co-channel interference may well originate from one or more similar users located in the neighbouring cells and assigned the same portion of *Orthogonal Frequency Division Multiple Access (OFDMA)* spectrum as the desired user. Thus the residual signal[2] vector $\boldsymbol{\varepsilon}_m$ can be expressed for Q co-channel interferers as

$$\boldsymbol{\varepsilon}_m(\ell) = \sum_{q=1}^{Q} \boldsymbol{H}^q_{\mathrm{Int},m}(\ell)\boldsymbol{x}^q_{\mathrm{Int}}(\ell) + \boldsymbol{\eta}_m(\ell) \tag{2}$$

where $\boldsymbol{H}^q_{\mathrm{Int},m} \in \mathbb{D}^{N_c}$ and $\boldsymbol{x}^q_{\mathrm{Int}}$ denote the unknown interferer channels and transmit vectors (similar to those of desired user) respectively while $\boldsymbol{\eta}_m$ denotes the white Gaussian noise. For notational

[1]Notation: Small bold faced symbols denote vectors while capital bold faced symbols denote matrices. The set $\mathbb{C}, \mathbb{P}, \mathbb{D}$ respectively denote the set of complex, positive semidefinite and diagonal matrices. The set $\mathbb{M}$ denotes the modulation alphabet. The operators $\mathrm{E}[\bullet], |\bullet|^2, (\bullet)^*, (\bullet)^{\mathrm{H}}$ stand for expectation, absolute value square, complex conjugate and hermitian respectively. The notation $|\bullet|$ used with a set as its argument, denotes the cardinality of set. Overhead $\hat{\bullet}$ and $\bar{\bullet}$ are used to denote the estimate and expected (mean) value of a variable, while sans serif small case letters are used to represent realizations of random variables.

[2]The term residual signal would be used for the undesired interference plus noise signal.

convenience, we may vertically stack the received signal at all the M antenna elements to get the following compact system model,

$$\boldsymbol{y}(\ell) = \boldsymbol{H}(\ell)\boldsymbol{x}(\ell) + \boldsymbol{\varepsilon}(\ell) \tag{3}$$

with $\boldsymbol{y} = \begin{bmatrix}\boldsymbol{y}_1^{\mathrm{T}} & \boldsymbol{y}_2^{\mathrm{T}} & \dots & \boldsymbol{y}_M^{\mathrm{T}}\end{bmatrix}^{\mathrm{T}}$ and $\boldsymbol{\varepsilon} = \begin{bmatrix}\boldsymbol{\varepsilon}_1^{\mathrm{T}} & \boldsymbol{\varepsilon}_2^{\mathrm{T}} & \dots & \boldsymbol{\varepsilon}_M^{\mathrm{T}}\end{bmatrix}^{\mathrm{T}}$ being MN_c dimensional vectors, while $\boldsymbol{H} \in \mathbb{C}^{MN_c \times N_c}$ reads as $\boldsymbol{H} = \begin{bmatrix}\boldsymbol{H}_1^{\mathrm{T}} & \boldsymbol{H}_2^{\mathrm{T}} & \dots & \boldsymbol{H}_M^{\mathrm{T}}\end{bmatrix}^{\mathrm{T}}$. The overall interference plus noise covariance matrix

$$\boldsymbol{R}_{\varepsilon} = \begin{bmatrix} \boldsymbol{R}_{\varepsilon_{1,1}} & \boldsymbol{R}_{\varepsilon_{1,2}} & \dots & \boldsymbol{R}_{\varepsilon_{1,M}} \\ \boldsymbol{R}_{\varepsilon_{2,1}} & \boldsymbol{R}_{\varepsilon_{2,2}} & \dots & \boldsymbol{R}_{\varepsilon_{2,M}} \\ \vdots & \vdots & \ddots & \vdots \\ \boldsymbol{R}_{\varepsilon_{M,1}} & \boldsymbol{R}_{\varepsilon_{M,2}} & \dots & \boldsymbol{R}_{\varepsilon_{M,M}} \end{bmatrix} \in \mathbb{P}^{MN_c} \tag{4}$$

with $\boldsymbol{R}_{\varepsilon_{m,n}} = \sum_{q=1}^{Q} \mathrm{E}[\boldsymbol{H}_{\mathrm{Int},m}^{q} \;\; \boldsymbol{H}_{\mathrm{Int},n}^{q}{}^{\mathrm{H}}] + \boldsymbol{R}_{\eta_{m,n}}$, can be seen to be composed of diagonal blocks and being spatially coloured.

Besides the information bearing blocks, we also consider the transmission of intermittent pilot blocks that are typically provided in cellular standards to allow for pilot-aided channel estimation at the receiver. We use the OFDM block index $d \in \mathscr{D}$ and $p \in \mathscr{P}$ in place of $\ell \in \mathscr{L}$ (c.f. (3) for instance) to distinguish respectively between the data and pilot blocks wherever necessary. The sets $\mathscr{D}$ and $\mathscr{P}$ are naturally mutually exclusive and $\mathscr{D} \cup \mathscr{P} = \mathscr{L}$.

In the sequel, we focus on the suppression of co-channel and inter-symbol interference from the perspective of a realistic receiver, whereby we handle the tasks of parameter estimation and incorporate the resulting estimation errors into our analysis.

3. PREVIOUS WORK

With the target of improving the detection performance of the desired user in an interference limited scenario, a number of approaches have been investigated in the past. Broadly speaking, these can be categorized into iterative and non-iterative schemes, but only iterative schemes fall into the scope of this analysis.

The pioneering work on the iterative turbo processing between channel decoder and *Maximum Aposteriori Probability (MAP)* symbol detector was presented in [1]. A low complexity variant, replacing the optimal exponential complexity MAP detector with a combination of soft *Inter-Symbol Interference (ISI)* canceler, linear *Minimum Mean Square Error (MMSE)* equalizer and an apriori information aware demodulator was proposed in [2] and reduced rank version thereof in [5]. Adapted to our system, it implies that after pilot-aided channel estimation, we first estimate the interference plus noise covariance matrix and pursue the *Spatial Whitening (SW)* and *Maximum Ratio Combining (MRC)* for spatial suppression of interference, and then employ the apriori info based MMSE equalizer and demodulator for residual interference suppression. Finally the channel decoder is configured to produce besides the aposteriori *Log Likelihood Ratios (LLRs)* of information bits, also the extrinsic LLRs of coded bits. These extrinsic LLRs of coded bits serve as the apriori information for the detector in next iteration. The process is repeated until convergence is achieved.

In order to allow the spatial interference suppression (SW & MRC) to benefit from the apriori information generated by the channel decoder, [6] proposed to revise the estimate of interference plus noise covariance matrix in each iteration by exploiting the fresh apriori information of the transmitted data symbols. This is achieved by estimating the residual signal from data blocks as well, by using the apriori estimates of the transmitted sub-carriers. The residual signal estimates of data blocks, along with those of pilot blocks, are then used to obtain an improved estimate of the spatial interference plus noise covariance matrix in each iteration [7, 8].

The channel estimation block can also be configured to exploit the apriori information generated by the channel decoder as advocated in [9, 10] under the name of iterative (turbo) channel estimation. For an co-channel interference dominated scenario, [7] proposed the conventional soft feedback based *Decision Directed Channel Estimation (DDCE)*, while [11] proposed the *Recursive Least Squares (RLS)* based adaptive channel estimation techniques.

Since channel estimation is pursued independently for each antenna, a quick intuitive analysis reveals that it is unable to benefit from the spatial suppression of co-channel interference and as such forms the bottleneck for the overall detection performance in a strong CCI environment. Although schemes such as [7, 10, 11] above, bring in the channel estimation procedure into the iterative framework, they help primarily to track channel variations better by effectively creating virtual pilot symbols (Soft DDCE). Thus these methods are of great help in high mobility or highly frequency selectivity scenarios, but once the impairment comes from strong co-channel interference, iterative channel estimation as above can not promise significant gains.

4. PROPOSED ITERATIVE FRAMEWORK

Thus we need a strategy that exploits the available apriori information of the coded bits in a way that the co-channel interference suppression aspect of the iterative receiver can be improved. To this end, we propose three modifications. First, to update the estimate of interference plus noise covariance matrix along the iterations, by taking into account the relative reliabilities of the residual signal estimates from the data and pilot blocks. Second, to improve the channel estimation performance in each iteration by incorporating the feedback on the spectral variation of interference power. And finally, to incorporate the impact of channel estimation errors in the estimation of interference plus noise covariance matrix to be employed for SW and MRC. These modifications improve the spatial interference suppression aspect, while the conventional apriori information based MMSE equalization [2, 5] help in temporal interference suppression. The performance of the proposed iterative framework is analyzed via EXIT charts and low number of iterations is found to be sufficient to attain convergence.

A block diagram of the proposed iterative receiver framework is shown in Fig. 2. A comparison with previous approaches shows that the lower portion of the diagram involves primarily three aforemen-

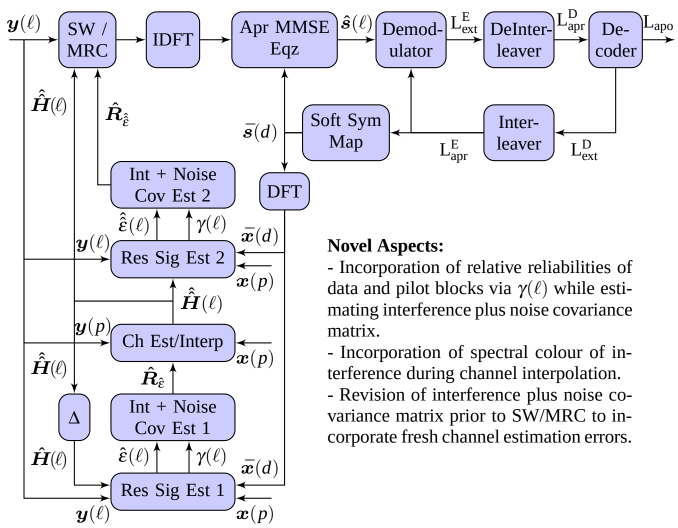


Figure 2: Block diagram of the proposed iterative receiver framework adapted to a DFT-Spread OFDM system.

tioned novel aspects which are discussed more elaborately in Sections 4.2, 4.3 and 4.4 respectively. We present the discussion in the reverse order — elaborating how the feedback from the channel decoder is exploited by various blocks in the iterative receiver chain.

4.1 Transformation of extrinsic LLRs to soft symbols

Besides the aposteriori LLRs of information bits, the channel decoder is configured to produce the extrinsic LLRs of coded bits as well. These extrinsic LLRs represent the new information generated in each iteration by the decoder via exploitation of the code structure. Adherence to the turbo principle [1] requires that the feedback is constituted of this new information only lest small cycles should limit the gains of the iterative procedure.

The extrinsic LLRs of the coded bits L_{ext}^{D} are interleaved back to generate the apriori information for next iteration $L_{apr}^{E} = \Pi(L_{ext}^{D})$.[3] The *Soft Sym Map* block in Fig. 2 transforms these LLRs to soft symbols. First, the bit probabilities are obtained as [2],

$$\Pr(c_{q,n} = \mathsf{c_q}) = \frac{1}{2}\left(1 + (1 - 2\mathsf{c_q})\tanh\left(\frac{L_{apr}^{E}(c_{q,n})}{2}\right)\right), \quad \mathsf{c_q} = \pm 1$$

which under the assumption of independent consecutive LLRs (owing to interleaver), lead to symbol probabilities $\Pr(s_n = \mathsf{s}) = \prod_q \Pr(c_{q,n} = \mathsf{c_q})$ where $\{\mathsf{c}_1, \mathsf{c}_2, \ldots, \mathsf{c}_Q\}$ are the $Q = \log_2(|\mathbb{M}|)$ bits that map to the symbol $\mathsf{s} \in \mathbb{M}$. These are then used to obtain the expected (soft) value of the symbols as $\bar{s}_n = \sum_{\mathsf{s}\in\mathbb{M}} \mathsf{s}\Pr(s_n = \mathsf{s})$ and the corresponding reliability measured by the variance

$$\sigma_{s_n}^2 = \left(\sum_{\mathsf{s}\in\mathbb{M}} |\mathsf{s}|^2 \Pr(s_n = \mathsf{s})\right) - |\bar{s}_n|^2. \tag{5}$$

The linear transformation via DFT can now be used to arrive at the apriori information of the transmitted frequency domain samples, so that if $\bar{\boldsymbol{s}}(d) \in \mathbb{C}^{N_c}$ denotes the vector containing data symbol expected values, the vector $\bar{\boldsymbol{x}}(d) = \boldsymbol{F}\bar{\boldsymbol{s}}(d) \in \mathbb{C}^{N_c}$ (with $\boldsymbol{F}$ being the Fourier matrix) contains the apriori information of the frequency domain samples.

4.2 Estimation of interference plus noise covariance matrix

Given the apriori information of the transmitted data blocks as $\bar{\boldsymbol{x}}(d) \in \mathbb{C}^{N_c}$, we obtain (c.f. *Res Sig Est 1* block in Fig. 2) the residual signal estimates by subtracting the expected contribution of the desired user's sub-carriers from the overall received signal to get

$$\hat{\varepsilon}(d) = \boldsymbol{y}(d) - \hat{\boldsymbol{H}}(d)\bar{\boldsymbol{x}}(d) \qquad d \in \mathscr{D} \tag{6}$$

$$\hat{\varepsilon}(p) = \boldsymbol{y}(p) - \hat{\boldsymbol{H}}(p)\boldsymbol{x}(p) \qquad p \in \mathscr{P} \tag{7}$$

for the data blocks and pilot blocks respectively. Collectively labeled as $\hat{\varepsilon}(\ell)$ with $\ell \in \mathscr{L} = \mathscr{D} \cup \mathscr{P}$, these can now be used for updating the estimate of interference plus noise covariance matrix. However, the relative reliabilities of the residual signal estimates obtained from pilot and data blocks must be taken into account. Unlike the transmitted vectors for the pilot blocks which are known accurately in (7), the transmitted vectors for the data blocks in (6) could only be estimated based on the feedback from the decoder and therefore may be incorrect. This introduces an additional impairment arising from possible errors between the transmitted samples and their estimates, i.e. if we let $\bar{\boldsymbol{x}}(d) = \boldsymbol{x}(d) + \boldsymbol{e}(d)$, with $\boldsymbol{e}(d)$ being the error vector for block d, then the residual signal estimates for the data blocks can be expressed as

$$\hat{\varepsilon}(d) = \boldsymbol{\varepsilon}(d) + \left(\boldsymbol{H}(d) - \hat{\boldsymbol{H}}(d)\right)\boldsymbol{x}(d) - \hat{\boldsymbol{H}}(d)\boldsymbol{e}(d) \tag{8}$$

which in contrast to the similar expression for pilot blocks

$$\hat{\varepsilon}(p) = \boldsymbol{\varepsilon}(p) + \left(\boldsymbol{H}(p) - \hat{\boldsymbol{H}}(p)\right)\boldsymbol{x}(p) \tag{9}$$

has an extra impairment term $\hat{\boldsymbol{H}}(d)\boldsymbol{e}(d)$ that directly depends on the quality of feedback from the decoder. In order to take the relative reliabilities of pilot and data blocks into account, we introduce block dependent scalar weighting factors $\gamma(\ell)$ while obtaining the interference plus noise covariance matrix estimate (c.f. *Int + Noise Cov Est 1* block), i.e.

$$\hat{\boldsymbol{R}}_{\hat{\varepsilon}}^{SS} = \frac{1}{\sum\gamma(\ell)} \sum_{\ell\in\mathscr{L}} \gamma(\ell)\hat{\varepsilon}(\ell)\hat{\varepsilon}^{H}(\ell) \tag{10}$$

It should be noted that all prior iterative CCI suppression schemes such as [6, 7, 8, 11] consider $\gamma(\ell) = 1$ for all $\ell \in \mathscr{L}$, i.e. disregard the block reliability while estimating $\boldsymbol{R}_{\hat{\varepsilon}}^{SS}$. The superscript $(\bullet)^{SS}$ emphasizes that $\hat{\boldsymbol{R}}_{\hat{\varepsilon}}^{SS}$ is just a sufficient statistic. The structural constraints, such as block diagonal nature, are imposed later to obtain the improved estimate as $\hat{\boldsymbol{R}}_{\hat{\varepsilon}}$ [12, Ch. 3].

The value of $\gamma(\ell)$ needs to be chosen keeping in view the relative reliability of block ℓ. Precisely speaking, it should lie in the interval [0,1] and be monotonically decreasing in $\sigma_s^2(\ell)$, the mean variance of the expected symbols (c.f. (5)) of the block ℓ. A simple heuristic choice that fulfills these requirements is,

$$\gamma(\ell) = 1 - (\sigma_s^2(\ell))^{1/r} \tag{11}$$

with r being a positive integer controlling the behaviour of $\gamma(d)$ as $\sigma_s^2(d)$ i.e. the uncertainty increases[4]. We choose the value of r such that the contribution in estimating $\boldsymbol{R}_{\varepsilon}$ from data blocks are trusted only if their block mean error variance is sufficiently low. This implies the following condition to hold approximately,

$$\sum_{d\in\mathscr{D}} \left(1 - (\sigma_s^2(d))^{1/r}\right) \approx \sum_{p\in\mathscr{P}} \left(1 - (\sigma_s^2(p))^{1/r}\right) \tag{12}$$

whereby putting $\sigma_s^2(p) = 0$ and assuming that $\sigma_s^2(d) \approx \sigma^2$ (reliability threshold), we finally express r as a function of the threshold and the number of pilot and data blocks as follows,

$$r \approx \frac{\log(\sigma^2)}{\log(1 - |\mathscr{P}|/|\mathscr{D}|)} \tag{13}$$

For LTE uplink specifications with $|\mathscr{P}| = 2$ pilot blocks and $|\mathscr{D}| = 12$ data blocks in each subframe, we get $r \approx 25$ and $r \approx 12$ for $\sigma^2 = 10^{-2}$ and 10^{-1} respectively.

The impact of the proposed block dependent $\gamma(\ell)$ on the receiver performance is analyzed by the following EXIT curves [13] for the different reliability assignments to data blocks while updating the covariance matrix estimate. We consider SNR levels of 12 dB and 18 dB and plot in Fig. 3 the curves for the input vs. output *Mutual Information (MI)* of the detector (the combined block of SW/MRC, MMSE Equalizer and Demodulator). For the case where we do not include data block's residual signal estimates ($\gamma(d) = 0$), we observe only a small but monotonic increase in the output MI as the input MI increases. Incorporating data block's residual signal estimates ($\gamma(d) \neq 0$), we note a significant gain in the output MI, for the high input MI i.e. once the apriori information is more reliable. But at low values of input MI, we observe a decrease in the output MI owing to the heavily erroneous residual signal estimates being used for the covariance matrix estimation. This explains the introduction of the parameter r. Higher values of r, for instance $r = 18$, help us not only to retain the good performance at the low feed back quality but also to achieve significant gains at the high reliability region of feedback symbols.

[3]The superscript D and E on LLRs signifies their association with Decoder and Equalizer respectively.

[4]Note that owing to no uncertainty in pilot blocks i.e. $\sigma_s^2(p) = 0$, we have $\gamma(p) = 1$ regardless of r

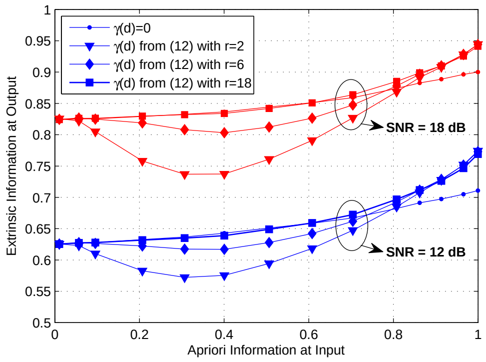


Figure 3: Equalizer EXIT characteristics obtained for different values of the estimation parameter r to be employed for computing the reliability values $\gamma(\ell)$ while estimating $\boldsymbol{R}_{\varepsilon}$ from the pilot and data block residual signals.

4.3 Exploitation of feedback during channel estimation

Given the *Least Squares (LS)* estimates of channel at pilot positions, we discuss now the strategy for interpolation to data positions (c.f. *Ch Est/Interp* block). To this end, we employ the conventional 2-D MMSE filter. However, unlike the earlier proposals [7, 11] for turbo channel estimation in presence of strong interferer, we propose to exploit the information fedback about the spectral variation of interference power. This additional information can be extracted from the diagonal values of the updated interference plus noise covariance matrix estimate in each iteration and could be used to assign different interpolation weights to the pilot CFRs at different locations depending upon their local interference plus noise power.

Let $\tilde{\boldsymbol{h}}_{\mathrm{p}} = \boldsymbol{h}_{\mathrm{p}} + \acute{\varepsilon}$ being the N_{p} dimensional vector containing the LS pilot estimates to be employed for interpolation at a particular data position.[5] The 2-D MMSE filter for estimating the data CFR $H(f,d)$ at f-th sub-carrier of the d-th block can be given as

$$\boldsymbol{w}_{\mathrm{MMSE}} = (\boldsymbol{R}_h + \boldsymbol{R}_{\acute{\varepsilon}})^{-1}\boldsymbol{r}_h, \tag{14}$$

where $\boldsymbol{R}_h = \mathrm{E}[\boldsymbol{h}_{\mathrm{p}}\boldsymbol{h}_{\mathrm{p}}^{\mathrm{H}}]$ and $\boldsymbol{r}_h = \mathrm{E}[\boldsymbol{h}_{\mathrm{p}}H(f,d)^*]$ are determined by the 2-D channel correlation sequence for which we propose to employ the Least-Favorable 2-D channel correlation sequence for robust performance [14]. $\boldsymbol{R}_{\acute{\varepsilon}} = \mathrm{E}[\acute{\varepsilon}\acute{\varepsilon}^{\mathrm{H}}] \in \mathbb{D}^{N_{\mathrm{p}}}$ denotes the covariance matrix of noise plus interference after LS estimation, and can be readily constructed from the original covariance matrix estimate $\hat{\boldsymbol{R}}_{\hat{\varepsilon}_{m,m}}$ at the given antenna.

Unlike the prior art approaches in [7] and [11], we propose to exploit the information about the spectral colour of interference in each iteration. To this end, the diagonal covariance matrix $\boldsymbol{R}_{\acute{\varepsilon}}$ is constructed by taking into account the varying interference plus noise power along frequency. The time variance of the interference plus noise power may also be incorporated in a similar fashion, but for the motion scenarios of interest it is deemed not necessary.

With the estimate of $\boldsymbol{R}_{\acute{\varepsilon}}$ improving along iterations, both in terms of mean interference plus noise power and its variation along the sub-carriers, the overall channel estimation/interpolation performance is boosted. The updated set of channel estimates (c.f. Fig. 2) in each iteration are denoted by $\hat{\hat{\boldsymbol{H}}}(\ell)$ to distinguish them from the previous estimates $\hat{\boldsymbol{H}}(\ell)$ which are used for the estimation of interference plus noise covariance matrix for channel estimation.

The impact of exploitation of spectral colour of interference is analyzed and presented later via simulation results in Fig. 4. The results confirm that channel estimation performance in a dominant CCI scenario remained a bottleneck for the overall receiver performance. Hence, improving channel estimation performance along iterations by the incorporation of feedback from decoder leads to a significant improvement in the overall receiver performance.

[5]Note that in this sub-section we drop the antenna index m because the channel estimation is pursued independently for each antenna. Thus the vector $\tilde{\boldsymbol{h}}_{\mathrm{p}}$ contains pilot estimates $\hat{H}(f,p)$ for various frequency and block indexes f and p respectively of the given receive antenna.

4.4 SW and MRC via revised covariance matrix estimate

Typically it is considered appropriate to estimate the interference plus noise (plus channel estimation error) covariance matrix only once in each iteration [7, 11]. However, given the updated channel estimates we propose to revise the estimate the interference plus noise plus channel estimation error covariance matrix in each iteration taking into account the fresh channel estimation errors. The justification can be sought by expressing the system model, after the incorporation of updated channel estimates as

$$\boldsymbol{y}(\ell) = \boldsymbol{H}(\ell)\boldsymbol{x}(\ell) + \varepsilon(\ell) = \hat{\hat{\boldsymbol{H}}}(\ell)\boldsymbol{x}(\ell) + \hat{\hat{\varepsilon}}(\ell) \tag{15}$$

with $\hat{\hat{\varepsilon}}(\ell) = \varepsilon(\ell) + (\boldsymbol{H}(\ell) - \hat{\hat{\boldsymbol{H}}}(\ell))\boldsymbol{x}(\ell)$. Thus we note that the optimal spatial whitening plus MRC processing for the system (after incorporation of channel estimation errors) is $\hat{\hat{\boldsymbol{H}}}^{\mathrm{H}}\boldsymbol{R}_{\hat{\hat{\varepsilon}}}^{-1}$. To this end, we recompute (c.f. *Res Sig Est 2* block in Fig. 2) the residual signals from data and pilot blocks via (6) and (7) respectively (with $\hat{\boldsymbol{H}}(\ell)$ being replaced by $\hat{\hat{\boldsymbol{H}}}(\ell)$) to yield $\hat{\hat{\varepsilon}}(\ell)$ rather than $\hat{\varepsilon}(\ell)$, and then employ (10) to estimate (c.f. *Int + Noise Cov Est 2* block) the new interference plus noise plus channel estimation error covariance matrix as $\hat{\boldsymbol{R}}_{\hat{\hat{\varepsilon}}}$ to be used for SW and MRC.

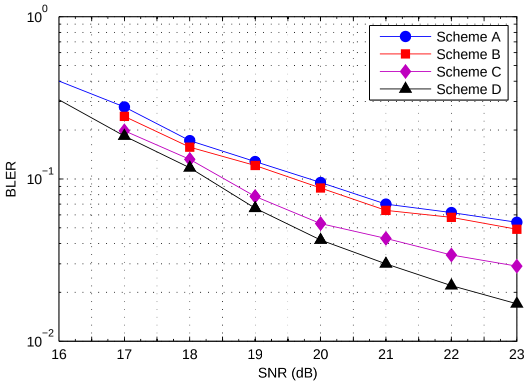
The impact of revising the estimate of interference plus noise covariance matrix twice in each iteration is also presented in the simulation results (Fig. 4), and this promises additional gains in the iterative suppression of CCI.

4.5 Apriori MMSE Equalization, Demodulation and Decoding

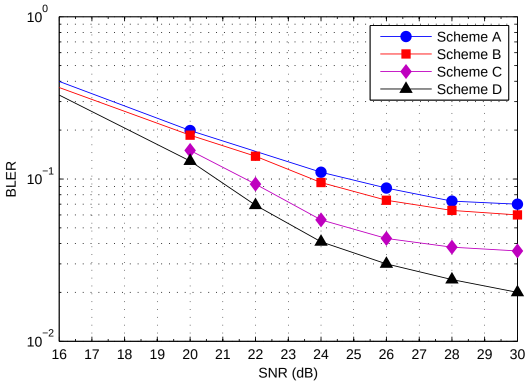
Instead of employing the optimal exponential complexity MAP detector, we employ (c.f. the block diagram in Fig. 2), the apriori information based MMSE equalization [5], followed by [1, 2] the improved demodulation taking into account apriori LLRs of neighbouring bits and the standard channel decoding. Since all these steps are standard in iterative receivers, for the sake of brevity we refrain from an in-depth discussion and refer the interested reader to the aforementioned original papers.

5. SIMULATION RESULTS

We emphasize that although the proposed modifications are applicable and beneficial to any interference limited iterative receiver equipped with multiple receive antennas, but we present here the results for an LTE uplink receiver [3] as an example. We consider two receive antennas, and a Carrier-to-Interference (C/I) level of 0 dB. The users are assumed to have a ETU power delay profile, and Jakes Doppler spectrum. We consider a turbo code with BCJR based decoding. For the iterative receiver we run 5 iterations although the convergence is mostly achieved earlier than this especially at high SNR. The two scenarios for which we present the analysis of prior art and proposed interference suppression performance are 16-QAM modulation combined with rate 2/3 turbo code at a velocity of 30 kmph and 64-QAM modulation combined with rate 1/2 turbo code at a velocity of 10 kmph. It may be pointed out here, that owing to the complexity constraints and the relatively slow channel variations across both frequency and time, we omit the Soft DDCE in our simulations for all the schemes, and note that since the proposed modifications can co-exist with Soft DDCE as well, the relative behaviour and gains of all schemes is expected to stay the same.



(a) 16-QAM, Code rate 2/3, 30 kmph



(b) 64-QAM, Code rate 1/2, 10 kmph

Figure 4: Performance comparison of proposed and prior interference suppression variants in iterative (turbo) receiver framework. Scheme A: Standard Turbo Equalization with non-iterative CCI suppression [2]. Scheme B: Turbo Equalization with iterative CCI suppression but without incorporation of spectral colour of interferer in channel estimation. Scheme C: Turbo Equalization with iterative CCI suppression with incorporation of spectral colour of interferer in channel estimation, but without revision of interference plus noise covariance matrix estimate. Scheme D: Turbo Equalization with iterative CCI suppression with incorporation of spectral colour of interferer in channel estimation and revision of interference plus noise covariance matrix estimate (c.f. Fig. 2). Both Scheme C and Scheme D include novel aspects.

Specifically, for the 16-QAM transmission scenario with 8% benchmark BLER (Block Error Rate), we note in Fig. 4(a) that inclusion of channel estimation into iterative framework but without exploitation of spectral colour of interference (Scheme B) leads to only a minor gain of around 0.5 dB over Scheme A, the standard Turbo equalization scheme (where CCI suppression is not the part of feedback loop). Incorporation of spectral colour of interference during channel estimation and interpolation (Scheme C) leads to a significant boost in iterative performance, and we gain more than 2.5 dB with respect to the reference scheme (Scheme B). Revision of the estimate of interference plus noise covariance matrix to reflect the effect of fresh channel estimation errors (Scheme D) leads to additional gain, so that the overall gain offered by the proposed iterative framework over the prior art (Scheme B) is about 3.2 dB at the benchmark level of 8% BLER.

Similarly for the 64-QAM transmission scenario, we note that in Fig. 4(b) the Scheme B leads to 1.5 dB gain at the benchmark level of 8% BLER. But once we go for the proposed incorporation of spectral colour of interferer during channel estimation we gain 3 dB over the reference Scheme B. Additionally, including the update of interference plus noise covariance matrix estimate into the framework leads to further gain, making the overall gain over the prior art around 4 dB.

6. CONCLUSION

Within the framework of conventional iterative receiver for interference suppression, we proposed the following improvements. (a). Incorporation of the relative reliabilities of various blocks while estimating the interference plus noise covariance matrix. This helps avoid giving undue weightage to the less reliable residual signal estimates from erroneous data blocks, and hence improves the covariance matrix estimate. (b). Incorporation of spectral colour of interference during channel estimation and interpolation, thereby assigning different weights to heavily and marginally interfered subcarriers' pilot estimates. (c). Revision of the interference plus noise covariance matrix estimate based on updated channel estimates taking into account the fresh channel estimation errors.

The proposed modifications in the estimation and exploitation of the interference plus noise covariance matrix, are shown to exhibit significant performance gains of 3-5 dB over the conventional iterative approaches for a CCI limited LTE uplink iterative receiver.

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