Channel Estimation and Co-Channel Interference Rejection for LTE-Advanced MIMO Uplink

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Abstract—Two techniques for channel estimation and co-channel interference rejection are presented for the multiple-input, multiple-output (MIMO) uplink in the Long Term Evolution-Advanced (LTE-Advanced) system. The first technique consists of estimating the desired users’ channels in the frequency domain (FD) before estimating interference covariance matrices for interference rejection. In the second technique, channel and interference covariance matrices are estimated jointly via a time domain (TD) approach. Interference rejection is achieved by using the channel and interference covariance matrix estimates in the MIMO equalizer. Significant interference rejection gains are observed.

I. INTRODUCTION

Modern cellular systems such as the 3rd Generation Partnership Project Long Term Evolution-Advanced (3GPP LTE-Advanced) system reuse the same frequency spectrum at each cell (frequency reuse 1). Hence, co-channel interference is a major impairment in current and future wireless systems. Channel estimation and interference mitigation are essential to maintaining quality-of-service (QoS) for the users.

Several papers discuss channel estimation for LTE-Advanced. For example, the discrete cosine transform (DCT) is used in [1] for channel estimation. Time-domain channel estimation and selection of the cyclic shifts of the training sequence are discussed in [2], [3]. As expected, cyclic shifts should be chosen to minimize the spectral leakage between users. In most of the works on LTE-Advanced channel estimation, co-channel interference from other cells is not considered.

For co-channel interference mitigation, neighboring cells can coordinate schedulers to allocate resources with reduced interference. Such inter-cell interference coordination (ICIC) requires communication between cells and is supported in the LTE-Advanced standard [4]. Another approach is estimation and cancellation (subtraction) of co-channel interfering signals. A difficulty with this approach is accurate interferer channel estimation and detection of the interferer’s signal, which may require modification to pilot sequences [5]. A third approach is interference suppression, wherein statistics of the interference plus noise are estimated to attenuate interference via noise whitening or equalization [6], [7], [8]. This technique is also known as interference rejection combining (IRC). Related work includes a multi-stage Wiener filter [9] for reduced-rank interference suppression in code-division multiple access (CDMA), a turbo equalizer [10] for single-carrier multiple-input, multiple-output (MIMO) systems with co-channel interference and IRC performance studies for the Global System for Mobile Communications (GSM) [11].

This paper presents frequency- and time-domain approaches to channel estimation and interference suppression in the LTE-Advanced MIMO uplink. Since different users participating in multi-user MIMO (MU-MIMO) communication have different propagation delays and delay spreads, MU-MIMO is more challenging for channel estimation than single-user MIMO (SU-MIMO). Hence, the remainder of the paper focuses on MU-MIMO with single-antenna users, although the techniques are directly applicable to SU-MIMO. The frequency-domain (FD) approach consists of two steps: MIMO channel estimation followed by interference covariance matrix estimation. Since the FD approach involves channel estimation in the presence of interference, a time-domain (TD) approach is also presented that estimates channel and interference covariance matrices jointly. The essential novelty of these approaches is the application to LTE-Advanced of both channel estimation and interference rejection with reasonable complexity and with minimal assumptions on channel and interference statistics. Simulations show significant performance gains using these techniques in the presence of interference.

The remainder of the paper is organized as follows. In Section II, the system model is introduced. The frequency-domain channel and interference covariance matrix estimation technique is described in Section III. In Section IV, the time-domain approach is presented. Simulation results are provided in Section V, and conclusions are given in Section VI.

II. SYSTEM MODEL

Fig. 1 is an illustration of a cellular base station (BS) with $M_r$ receiver antennas, $U \leq M_r$ desired users and $L$ co-channel interferers. The users and interferers are assumed to have single transmitter antennas. Note that in SU-MIMO, the spatial layers can be regarded as single-antenna users participating in MU-MIMO with the additional properties of...
nearly equal timing offsets and delay spreads. The channel and interference covariance matrix estimation is performed on the demodulation reference signal (DM-RS) in each slot of an LTE-Advanced uplink subframe, as illustrated in Fig. 2 for four resource blocks (RBs) [12]. The size of each tile in Fig. 2 is one single-carrier frequency-division multiple-access (SC-FDMA) symbol in the time domain and 12 subcarriers in the frequency domain.

For a given slot, let $H_{k,u}$ denote the $M_r \times 1$ channel response during the DM-RS symbol from user $u$ on subcarrier $k$, where $k = k_1, \ldots, N + k_1 - 1$ and $N$ and $k_1$ are, respectively, the number of subcarriers and starting index of the allocation under consideration. Furthermore, let $H^{(l)}_k$ denote the $M_r \times 1$ channel response during the DM-RS symbol from interferer $l$ on subcarrier $k$. For ease of exposition, it is assumed that the interferer transmissions are synchronized with that of the desired users. This assumption is reasonable for strong interferers that contribute to the largest performance degradation. These interferers are likely located near the cell edges of both the desired and interfering cells; in this case, the timing advance value used to synchronize the interferer to its cell will most likely cause the interferer to synchronize to the desired cell also. Moreover, the cell-edge interferer propagation delay and delay spread are contained within the cyclic prefix; hence, strong interferers are most likely time-synchronized with the desired users. Note, however, that the algorithms in this paper are also applicable to the case of unsynchronized interferers.

The $M_r \times 1$ received vector $Y_k$ during the DM-RS symbol on subcarrier $k$ can be written as

$$Y_k = X_k \sum_{u=1}^{U} H_{k,u} + W_k$$

(1)

where $X_k$ is the (known) DM-RS training sample ($|X_k| = 1$) and $W_k$ is the $M_r \times 1$ interference-plus-noise vector on subcarrier $k$. One can write the channel vector as $H_{k,u} = H_{k,u} e^{jk\theta_u}$, where $H_{k,u}$ is the channel vector without the linear phase shift across subcarriers, $\theta_u = -\phi_u + 2\pi n_{cs,u}/N_{cs}$, $\phi_u$ is the unknown phase shift from subcarrier $k$ to subcarrier $k+1$ caused by user $u$’s timing offset, $n_{cs,u}$ is the known cyclic shift index for user $u$ and $N_{cs}$ is the total number of equally-spaced cyclic shifts.

The interference plus noise vector $W_k$ can be written as

$$W_k = \sum_{l=1}^{L} H^{(l)}_k X^{(l)}_k + V_k$$

(2)

where $X^{(l)}_k$ is the (unknown) DM-RS training sample for interferer $l$ on subcarrier $k$ and $V_k$ is an additive white Gaussian noise (AWGN) vector with covariance matrix $N_0 I_M$. Signals from different interferers and signals from the same interferer on different subcarriers are assumed to be uncorrelated, i.e., $E[X^{(l)}_k X^{(m')}_{k}] = \delta(l,m')$, where $\delta(l,m)$ is 1 or 0 depending on interferer frequency allocations and $\delta$ denotes the Kronecker delta function. In addition, the wide-sense stationary uncorrelated scattering (WSSUS) model is assumed for the interferer channels, i.e., $E[H^{(l)}_k H^{(l)H}_{k+p}] = B^{(l)}(p)$, where $B^{(l)}(p)$ is a $M_r \times M_r$ matrix function of $p$. Hence,

$$E[W_k W^{H}_{k+p}] = \sum_{l=1}^{L} \epsilon_l B^{(l)}(0) + N_0 I_{M_r} \delta_p = R_{ww,k} \delta_p$$

(3)

where $R_{ww,k} = E[W_k W^{H}_{k}]$ is the interference covariance matrix on subcarrier $k$.

Substituting (2) into (1) and multiplying by $X^*_k$, one obtains

$$Z_k = \sum_{u=1}^{U} H_{k,u}^* X_k + \sum_{l=1}^{L} H^{(l)}_k X^{(l)}_k + \hat{V}_k$$

(4)

where $\hat{X}^{(l)}_k = X^*_k X^{(l)}_k$ and $\hat{V}_k = X^*_k V_k$. As discussed in Sections III and IV, $Z_k$ is used for the FD and TD channel and interference covariance matrix estimation techniques.

III. FREQUENCY-DOMAIN ESTIMATION

In the FD approach, MIMO channel estimates are computed in the presence of interference; subsequently, the interference covariance matrix $R_{ww,k}$ is estimated. FD MIMO channel estimation relies on the narrowband approximation, wherein the channel magnitude is approximately constant over a frequency band less than the channel coherence bandwidth. Consider a set of $P$ adjacent subcarriers over which a narrowband approximation holds, i.e.,

$$\hat{H}_{k+p,u} \approx \hat{H}_{k+1} e^{jk\theta_u}, 0 \leq p \leq P - 1$$

(5)

since the phase ramp $k\theta_u$ is not included in $\hat{H}_k$. The constraint $P > U$ is required to estimate desired users’ channels and the interference covariance matrix.

Let $Z_k = [Z_k, \ldots, Z_{k+P-1}]$, $\hat{H}_{k+[\frac{p-1}{P}],1} = [\hat{H}_{k+[\frac{p-1}{P}],1}, \hat{H}_{k+[\frac{p-1}{P}],2}, \ldots, \hat{H}_{k+[\frac{p-1}{P}],U}]$, $\hat{W}_k = X^*_k W_k$ and $\hat{V}_k = [W_k, \ldots, W_{k+P-1}]$. From (2), (4) and (5), concatenation of signals on $P$ adjacent subcarriers yields

$$Z_k \approx \hat{H}_{k+1} D_k A_{P} + \hat{V}_k$$

(6)

where $D_k = \text{diag}(e^{jk\theta_1}, \ldots, e^{jk\theta_U})$ and

$$A_{P} = \begin{bmatrix} 1 & \cdots & e^{j(P-1)\theta_1} \\ \vdots & \ddots & \vdots \\ 1 & \cdots & e^{j(P-1)\theta_U} \end{bmatrix}$$

(7)

Since $\hat{H}_k$ and $\theta_u$ (and, hence, $A_{P}$ and $D_k$) are unknown, an iterative procedure is adopted for channel estimation. The basic idea of the algorithm is to iterate between least-squares channel estimation given the timing offsets and estimation of
Algorithm 1: Frequency-Domain Channel Estimation

Data: $Z_k$, max. iteration count $i_{\text{max}}$, $U, N, k_1, P, K \leq P$, subcarrier skip value $Q$ for timing offset estimate, frequency-domain filter impulse response $g_k$

Result: Channel and timing offset estimates, $\hat{\mathcal{H}}_k$ and $\hat{\phi}_u$

Initialization:
\[
i \leftarrow 1, \quad \hat{\theta}_u^{(i)} \leftarrow 0, \quad u = 1, \ldots, U
\]
Compute estimates $\hat{A}_p^{(i)}$ and $\hat{D}_k^{(i)}$ from $\hat{\theta}_u^{(i)}$

while $i \leq i_{\text{max}}$

for $k = k_1 : K : K([N - 1] / K) + k_1$

Compute least squares channel estimate from (6), given estimates $\hat{A}_p^{(i)}$ and $\hat{D}_k^{(i)}$:
\[
\hat{\mathcal{H}}_k^{(i)} = Z_k \hat{A}_p^{(i)} H (\hat{A}_p^{(i)} \hat{A}_p^{H})^{-1} \hat{D}_k^{(i)} H
\]

end

Interpolate channel estimates for each subcarrier $k = k_1, k_1 + 1, \ldots, N + k_1 - 1$:
\[
\hat{\mathcal{H}}_k^{(i)} \leftarrow \sum_{m = 0}^{[(N - Q - 1)]} g_{k - Km} \hat{\mathcal{H}}_k^{(i)}_{Km + k + 1} \quad (k_{\text{max}})
\]

Update timing offset estimate for user $u$:
\[
\hat{\phi}_u^{(i+1)} = \frac{1}{Q} \arg \left\{ \sum_{k = 0}^{N - Q - 1} \hat{F}_{[k + 1, u]} \hat{\mathcal{H}}_k^{(i)}_{Km + k + 1} \right\}
\]

From $\hat{\phi}_u^{(i+1)}$, obtain updated estimates of $\hat{\theta}_u^{(i+1)}$, $\hat{A}_p^{(i+1)}$ and $\hat{D}_k^{(i+1)}$

\[
i \leftarrow i + 1
\]

end

Algorithm 2: Frequency-Domain Interference Covariance Matrix Estimation

Data: $Z_k$, $\mathcal{H}_k$, $U, N, k_1, P, K \leq P$, averaging coefficients and interval $c_m$ and $m_1$, shrinkage coefficients $\alpha_k (0 \leq \alpha_k \leq 1)$, frequency-domain filter impulse response $g_k$

Result: Interference covariance matrix estimates $\hat{R}_{ww,k}^{fd}$

for $k = k_1 : K : K([N - 1] / K) + k_1$

Compute $S_{k + [\frac{p - 1}{P - U}]}$ from (11)

Compute $\hat{R}_{ww,k + [\frac{p - 1}{P - U}]}$ from (16)

Apply regularization and shrinkage as follows:
\[
\hat{R}_{ww,k + [\frac{p - 1}{P - U}]} \leftarrow (1 - \alpha_k) \hat{R}_{ww,k + [\frac{p - 1}{P - U}]} + \frac{\alpha_k}{M_r} \text{tr} (\hat{R}_{ww,k + [\frac{p - 1}{P - U}]} I_{M_r})
\]

end

Interpolate covariance matrix estimates for each subcarrier $k = k_1, k_1 + 1, \ldots, N + k_1 - 1$:
\[
\hat{R}_{ww,k}^{fd} = \sum_{m = 0}^{[(N - 1) / K]} g_{k - Km} \hat{R}_{ww,Km + k + 1} + [\frac{p - 1}{P - U}]
\]

\[
\hat{A}_p^H (\hat{A}_p \hat{A}_p^H)^{-1} \hat{A}_p
\]

Now,
\[
E[S_{k + [\frac{p - 1}{P - U}]}^H S_{k + [\frac{p - 1}{P - U}]}] = \sum_{p = 0}^{P - 1} m_p^H m_p R_{ww,k + p}
\]

Under a narrowband assumption for the interference covariance matrix, i.e., $R_{ww,k + p} \approx R_{ww,k + [\frac{p - 1}{P - U}]}$, $p = 0, 1, \ldots, P - 1$, (14) can be approximated by
\[
E[S_{k + [\frac{p - 1}{P - U}]}^H S_{k + [\frac{p - 1}{P - U}]}] \approx ||M_p||_F^2 R_{ww,k + [\frac{p - 1}{P - U}]} = (P - U) R_{ww,k + [\frac{p - 1}{P - U}]}.
\]

Hence, an estimate for the interference covariance matrix on subcarrier $k + [\frac{p - 1}{P - U}]$ is given by
\[
\hat{R}_{ww,k + [\frac{p - 1}{P - U}]} = \frac{1}{P - U} \sum_{m = m_1}^{m_1} c_m S_{k + [\frac{p - 1}{P - U}] + mK}^H S_{k + [\frac{p - 1}{P - U}] + mK}^H
\]

where $K$ is defined in Algorithm 1 and $m_1$ and $c_m$ are selected to approximate $E[S_{k + [\frac{p - 1}{P - U}]}^H S_{k + [\frac{p - 1}{P - U}]}]$. Regularization and shrinkage are used to improve interference rejection performance. Finally, interpolation to all subcarriers yields the final covariance estimate $R_{ww,k}^{fd}$. The algorithmic complexity of the FD approach is $O(M_r^2 N P / K)$. The FD interference covariance matrix estimation procedure is summarized in Algorithm 2.

IV. TIME-DOMAIN ESTIMATION

In the TD approach, the channel and interference covariance matrices are estimated jointly. The basic idea is to convert the signal $Z_k$ in (4) to the time domain, estimate the channel impulse response taps of each user, convert the channel estimates...
to the frequency domain and subtract the channel estimates to obtain a residual signal. The interference covariance matrix is estimated from the residual signal. In order to exploit the localization of the channel impulse responses near the cyclic shifts, spectral leakage caused by finite resource allocations needs to be minimized.

Given a frequency-domain allocation of \( N \) subcarriers, a suitable window function \( \Omega_k \) is applied to (4) to reduce spectral leakage. An \( N_f \)-point inverse discrete Fourier transform (IDFT) is computed on \( \tilde{Z}_k = \Omega_k Z_k \), where zero-padding is used if \( N_f > N \). The resultant time-domain vector is

\[
\tilde{z}_n = \sum_{u=1}^{U} h_{n,u} \otimes \omega_n + \sum_{l=1}^{L} h_{n,l}^{(l)} \otimes \tilde{x}_n^{(l)} \otimes \omega_n + \tilde{v}_n \otimes \omega_n
\]  

(19)

where \( h_{n,u}, \omega_n, \tilde{x}_n^{(l)} \) and \( \tilde{v}_n \) are the IDFT’s of \( \mathbf{H}_{k,u}, \Omega_k, \tilde{X}_k^{(l)} \) and \( \tilde{V}_k \), respectively, and \( \otimes \) denotes circular convolution. The IDFT size \( N_f \geq N \) is selected based on the desired sampling period. The channel taps of user \( u \) are estimated within a time interval \( I_u = \{ \tilde{n}_{cs,u} - n_l, \ldots, \tilde{n}_{cs,u} + n_r \} \) where \( \tilde{n}_{cs,u} = (n_{cs,u}/N_{cs})N_f \) is the sample index associated with the cyclic shift for user \( u \). The interval values \( n_l \) and \( n_r \) are chosen such that the intervals \( I_1, \ldots, I_U \) are disjoint. The spectral leakage from dominant channel taps of other users is removed before estimating the channel of the user under consideration. After the channel taps for all users are estimated, \( N_f \)-point discrete Fourier transforms (DFTs) are taken to obtain frequency-domain channel estimates. The channel estimates are subtracted from \( Z_k \) in (4) to form a vector \( \mathbf{E}_k \), whose covariance is the interference covariance matrix. Regularization and shrinkage are applied to provide the final covariance matrix estimate \( \mathbf{R}_{w,u,k}^{td} \).

Regularization and shrinkage are applied to provide the final covariance matrix estimate \( \mathbf{R}_{w,u,k}^{td} \). Since

\[
\mathbf{E}_k = \mathbf{Z}_k - \sum_{u=1}^{U} \tilde{\mathbf{H}}_{k,u}
\]  

(20)

\[
= \sum_{u=1}^{U} (\mathbf{H}_{k,u} - \tilde{\mathbf{H}}_{k,u}) + \sum_{l=1}^{L} \mathbf{H}_k^{(l)} \tilde{X}_k^{(l)} + \tilde{\mathbf{V}}_k
\]  

(21)

the covariance matrix of \( \mathbf{E}_k \) automatically includes the channel estimation error and interference covariance matrices. The complexity for the DFTs is \( O(M_r Un \log N) \). The remaining operations have complexity \( O(M_r^2 N) \). The TD estimation procedure is summarized in Algorithm 3.

**Algorithm 3: Time-Domain Channel and Interference Covariance Matrix Estimation**

**Data:** \( Z_k, \Omega_k, U, N, N_f \geq N, \{ I_u \}_{u=1}^{U} \), max. number of pulses per user \( C_{p,\text{max}} \), \( c_m, m_1, \alpha_k, \beta (0 \leq \beta) \)

**Result:** Channel and interference covariance matrix estimates, \( \hat{\mathbf{H}}_{k,u} \) and \( \mathbf{R}_{w,u,k}^{td} \)

**Initialization:**

\( i \leftarrow 1 \)

\[
\hat{Z}_k = \Omega_k Z_k, \tilde{z}_n = \text{IDFT}_{N_f}(\hat{Z}_k), \tilde{z}_n^{(i)} \leftarrow \tilde{z}_n
\]

(22)

\[
\omega_n = \text{IDFT}_{N_f}(\Omega_k)
\]

(23)

\[
C_{p,u} \leftarrow 0, N_u \leftarrow 0, u = 1, \ldots, U
\]

(24)

\[
n_{\text{max}} = \arg \max_n \| \tilde{z}_n \|^2
\]

(25)

while \( n_{\text{max}} \in \bigcup_{u=1}^{U} I_u \) and max. \( C_{p,u} \leq C_{p,\text{max}} \) do

\( u_{\text{max}} \leftarrow u \)

\[
\mathcal{N}_{(i)}_{\text{max}} \leftarrow \mathcal{N}_{(i)_{\text{max}}} \cup \{ n_{\text{max}}, \tilde{z}_n^{(i)} \}, \quad C_{p,u_{\text{max}}} \leftarrow C_{p,u_{\text{max}}} + 1
\]

(26)

end

Cancel spectral leakage:

\[
\tilde{z}_n^{(i+1)} = \tilde{z}_n^{(i)} - \tilde{z}_n^{(i)} \left( \frac{\omega_n - n_{\text{max}}}{\omega_0} \right)
\]

(27)

Compute \( n_{\text{max}} = \arg \max_n \| \tilde{z}_n^{(i+1)} \|^2 \)

(28)

\( i \leftarrow i + 1 \)

end

Set \( \tilde{z}_n^{\text{res}} \leftarrow \tilde{z}_n^{(i)} \)

Estimate energy of \( \tilde{z}_n^{\text{res}} \) outside the intervals \( I_u \):

\[
\varepsilon_{\tilde{z}_n^{\text{res}}} \approx \frac{1}{N_f - U(n_r + n_l + 1)} \sum_{n \notin \bigcup_{u=1}^{U} I_u} \| \tilde{z}_n^{\text{res}} \|^2
\]

(29)

for \( u=1:U \) do

Add pulses associated with user \( u \) to residual signal:

\[
\tilde{z}_n^{\text{res}} = \tilde{z}_n^{\text{res}} + \sum_{n \notin I_u} \tilde{z}_n^{(i)} \left( \frac{\omega_n-n_{\text{max}}}{\omega_0} \right)
\]

(30)

Estimate channel impulse response for user \( u \) from samples with energy greater than \( \beta \varepsilon_{\tilde{z}_n^{\text{res}}} \):

\[
\hat{h}_{n,u} = \begin{cases} \tilde{z}_n^{\text{res}}, & n \in I_u, \| \tilde{z}_n^{\text{res}} \|^2 > \beta \varepsilon_{\tilde{z}_n^{\text{res}}} \\ 0, & \text{otherwise} \end{cases}
\]

(31)

Compute \( \hat{\mathbf{H}}_{k,u} = \text{DFT}_{N_f}(\hat{h}_{n,u}) \)

end

Form the vector \( \mathbf{E}_k \) according to (20)

Obtain estimate of interference covariance matrix:

\[
\mathbf{R}_{w,u,k}^{td} = \sum_{m=-m_1}^{m_1} c_m \mathbf{E}_k+m \mathbf{E}_k^H \]

(32)

Apply regularization and shrinkage:

\[
\mathbf{R}_{w,u,k}^{td} = (1 - \alpha_k) \mathbf{R}_{w,u,k}^{td} + \frac{\alpha_k}{M_r} \text{tr}(\mathbf{R}_{w,u,k}^{td}) \mathbf{I}_{M_r}
\]

(33)
channel and interference covariance matrix estimates: \( G_k = H_k^H (H_k H_k^H + R_{ww,k})^{-1} \).

Fig. 3 is a plot of the block error rate (BLER) versus carrier-to-interference ratio (CIR) for a single user transmitting quadrature phase shift keying modulation (QPSK) with rate-1/3 turbo coding on 15 RBs \((N = 180)\) to a base station with \( M_r \) = 4 receive antennas. A single co-channel interferer is also present. The signal-to-noise ratio (SNR) is 1 dB, and the channel model assumed for both user and interferer is the Extended Typical Urban (ETU) channel with 5 Hz Doppler frequency [13]. Various regularization factors \( \alpha_k \) are considered, including \( \alpha_k = 1/\kappa_k \), where \( \kappa_k \) is the condition number (ratio of the largest to smallest singular value) of \( R_{ww,k} \). The number of cyclic shifts is \( N_{cs} = 12 \). For FD, \( \max_{I} = Q = 2, K = 1. \) For the TD approach, \( N_t = 240, n_t = 0.4N_t/N_{cs}, n_r = 1.4N_t/N_{cs}, C_{p,max} = 4, \beta = 1. \) The filter \( \hat{c}_{\alpha} \) is a moving average with 11 taps \((n_t = 5)\).

Fig. 4 is similar to Fig. 3, except the desired user transmits 64-quadrature amplitude modulation (64-QAM) with rate-5/6 turbo coding, SNR = 21 dB and the desired user’s channel is the Extended Pedestrian A (EPA) model [13]. Here, for TD, \( n_t = 0.8N_t/N_{cs}, n_r = 1.8N_t/N_{cs}. \) FD performs better than TD at low CIR since the strong interference affects the channel tap selection in the TD case. At higher CIR, TD performs better than FD because of the averaging gain from selecting a few dominant taps. Furthermore, higher regularization factors \( \alpha_k \) improve performance as the CIR increases, while low regularization factors perform well at low CIR. A reasonable choice for the regularization is the inverse condition number of the instantaneous interference covariance estimate. At BLER = \( 10^{-3} \), interference rejection gains compared to a white covariance matrix are 6–9 dB and 10–12 dB for QPSK and 64-QAM, respectively. Simulations with greater than one interferer also show significant gains.

VI. CONCLUSION

Channel estimation and interference mitigation are important receiver functions in cellular systems. This paper presents two schemes for channel estimation and co-channel interference rejection in the LTE-Advanced MIMO uplink. The FD approach is sequential (channel estimation followed by interference suppression) and relies on a narrowband approximation. The TD approach involves joint estimation of the channel and interference covariance matrices by relying on the localization of the channel taps near the cyclic shifts. Both approaches provide significant gains in the presence of interference; a good regularization factor is the inverse condition number of the instantaneous covariance estimate.

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