Sounding Reference Signal Design for TDD LTE-Advanced System

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Abstract—In the 3\textsuperscript{rd} generation partnership project (3GPP) long term evolution (LTE) time division duplex (TDD) systems, base stations use uplink sounding reference signals (SRs) to estimate downlink channel state information (CSI) for downlink beamforming transmissions. Because SRs sent by up to 8 users are multiplexed on the same time-frequency resources, there exist mutual interferences among these SRs over fading channels. Thus the current scheme inevitably impacts the estimation accuracy of CSI and thereby degrades the beamforming performance. Based on the characteristics of TDD systems, we propose a novel sounding scheme to improve the quality of CSI, in which according to the CQIs (channel quality indicators) periodically reported by each user and the bandwidth requirement of each user, BS dynamically schedules each user to send SRs only on the requisite bandwidth with best CQIs instead of on full bandwidth or specified sub-bandwidth as in the current LTE TDD systems. Simulations show that the proposed scheme outperforms the current scheme in estimation accuracy for various frequency-selective and/or time-selective fading channels, and is especially robust to frequency selective channels. Therefore the proposed scheme is applicable to the uplink SR design in LTE-Advanced and has been adopted as a candidate scheme for 3GPP LTE-Advanced.

Index Terms—channel estimation error, Long Term Evolution (LTE), sounding, time division duplex (TDD)

I. INTRODUCTION

The commercial deployment of cellular infrastructure based on the 3\textsuperscript{rd} generation partnership project (3GPP) long term evolution (LTE) is becoming more and more widespread due to the continuous and strong demand of broadband mobile services, especially the appearance of smart handsets in recent years. LTE can work in FDD (frequency division duplex) mode or TDD (time division duplex) mode. In this paper we focused on LTE TDD systems. According to 3GPP specification [1], there are seven kinds of transmission modes (TMs) in LTE TDD systems. For the TMs 4, 5 and 6, the base station (BS) should know the downlink channel state information (CSI) for beamforming transmission.

As well known, accurate CSI is very important for multiple-input single-output (MISO) and multiple-input multiple-output (MIMO) beamforming systems. However, in practice CSI is always imperfect due to the existence of CSI delay, channel estimation error and quantization error, which would degrade the system performance. Hence, it is significant to optimize the performance of MISO/MIMO beamforming systems in the case of imperfect CSI.

Recently there are a lot of papers which optimized the performance of beamforming systems with imperfect CSI from the different perspectives. In [2], authors proposed to weight channel sounding pilots (used to measure CSI) using the level of interferences at each user to increase beamforming system throughput. A Kalman-filter-based prediction method was proposed to overcome the feedback delay effect in [3]. In [4], ergodic capacity bounds were investigated with channel estimation error and quantization error, and then were optimized over the number of both finite training symbols and limited feedback bits under the condition of fixed frame length. The optimal bandwidth allocation between data channel and feedback channel was studied in [5] to maximize the average throughput in the data channel using MISO beamforming scheme. In [6], the authors investigated the optimal precoding algorithm and corresponding power allocation in the presence of imperfect CSI feedback, such as an imperfect channel coefficient feedback, channel mean feedback and channel covariance feedback. The optimal spatial and temporal power allocation was studied in [7] to minimize outage probability for a MISO system with delayed feedback. In [8] we optimized time-domain period of sounding to minimize the impact of CSI delay on TDD MISO beamforming systems. However, all the above papers focused on either FDD systems or TDD narrowband systems and few papers concern beamforming performance optimization of TDD wideband systems, especially for LTE TDD communication systems.

In the current LTE TDD beamforming systems, a BS estimates downlink CSI via channel reciprocity based on uplink sounding reference signal (SRS) [9] [10] sent by users, and then uses it to generate beamforming vectors / matrices for downlink transmission. There are two kinds of sounding schemes as described in LTE Release 8 [9] and 9 [10]: one is full bandwidth scheme that each user sends SRs on all physical resource blocks (PRBs, PRB is minimum resource allocation unit and one PRB is equivalent to one subband); the other is sub-bandwidth scheme that each user sends SRs on the specified partial continuous PRBs. In each scheme SRs of up to 8 users are multiplexed on the same time-frequency resource. Each user applies a unique time-domain cyclic shift, which allows the signals to be separated at the receiver, to own SRS sequence. However, the cyclic shifts do not perfectly orthogonalize these SRs over fading channel at the BS. Thus the residual interferences would cause serious channel estimation errors, especially where most of the available cyclic shift values are utilized and/or the multi-path delay is large. So the current sounding schemes impact the accuracy of CSI as well as degrade the performance of beamforming systems.

Based on the characteristics of TDD systems, we propose a novel sounding scheme for TDD LTE-Advanced systems to improve the quality of CSI, in which according to the CQIs (channel quality indications) periodically reported by each user and the bandwidth requirement of each user, BS dynamically schedules each user to send SRs only on the requisite bandwidth with best CQIs instead of on full bandwidth or specified sub-bandwidth as in the current LTE TDD systems.
In our proposed scheme, the characteristic of “distribution according to need” leads that the interferences among SRSs of different users are eliminated completely and the characteristic of “scheduling based on CQI” leads that SRSs are sent always on the best channel. Simulation results verify that the proposed scheme can obtain higher channel estimation accuracy than the current scheme for various frequency-selective and/or time-selective fading channels, and especially is robust to frequency selective channels. Hence, the proposed scheme can be applicable to the uplink SRS design in TDD LTE-Advanced and has been adopted as a candidate scheme for 3GPP LTE-Advanced.

Notations: Upper and lower boldface letters denote matrices and (column) vectors, respectively. ( ) and [ ] denote complex conjugation and Frobenius norm, respectively. I denotes the identity matrix. \( \emptyset \) denotes empty set. \( \mathcal{CN}(\mu, \Sigma) \) denotes the complex Gaussian distribution with mean vector \( \mu \) and variance matrix \( \Sigma \). \( \text{diag}(\mathbf{x}) \) denotes a diagonal matrix with elements of vector \( \mathbf{x} \) on its main diagonal. \( \mathcal{F} \) and \( \mathcal{F}^{-1} \) denote FFT (fast Fourier transform) and IFFT (inverse fast Fourier transform), respectively.

II. SYSTEM MODEL

Consider an LTE TDD system, where a BS with \( M \) receive antennas communicates with \( K \) single-antenna users. These users send SRSs on the same set of \( N_c \) subcarriers at the same time. The channels between all users and BS are modeled as time-varying, spatially uncorrelated, frequency selective and independently and identically distributed (i.i.d.) Rayleigh fading channel. The channel from user \( k \) to BS on the \( i \)-th subcarrier is denoted by an \( M \times 1 \) vector \( \mathbf{h}_k(i) \), where \( \mathbf{h}_k(i) \sim \mathcal{CN}(0, \mathbf{I}) \).

In LTE TDD systems, SRSs are located in alternate subcarriers (referred as \( \text{comb} \) in terms of LTE terminology [9]) in frequency domain and are transmitted in the last two OFDM (orthogonal frequency division multiplexing) symbols of the special subframe (and/or the last OFDM symbol of uplink subframe) in time domain. It is used to estimate downlink CSI at BS for beamforming transmission besides measuring CQIs of uplink PRBs for uplink PRB scheduling. In order to let more users be accessed in one transmission time interval (TTI, one TTI equals one subframe duration), the SRSs of up to \( K \) users, as a set, are multiplexed on the same comb. The transmitted SRSs for each user are derived from a constant amplitude zero auto correlation (CAZAC) base sequence, which may be cyclically shifted in time domain by a unique number of samples, causing a time-domain circular shift of \( \tau = (\alpha/K)T \), where \( \alpha \) is a cyclic shift value uniquely assigned to each user in the set, \( T \) is the duration of one OFDM symbol.

Let \( \mathbf{s}_k = [s_k(1), s_k(2), \ldots, s_k(N_c)] \) denote the full-bandwidth SRS sequence transmitted by user \( k \), the element \( s_k(i) \) of which is the SRS symbol on the \( i \)-th subcarrier. At the user side, \( \mathbf{s}_k \) is mapped onto corresponding time-frequency resources located in the special subframe and/or uplink subframe, and transformed via IFFT into time-domain signals. Finally the time-domain signals with added cyclic prefix (CP) are sent out.

At the BS side, after removing CP, implementing FFT and de-mapping subcarrier, the received signal on the \( i \)-th subcarrier (summing over all SRSs from all users in the set), denoted by \( \mathbf{y}(i) \) of size \( M \times 1 \), is given by

\[
\mathbf{y}(i) = \sum_{k=1}^{K} \mathbf{h}_k(i)s_k(i) + \mathbf{n}(i), \quad 0 < i \leq N_c, \tag{1}
\]

where \( \mathbf{n}(i) \sim \mathcal{CN}(0, \sigma_n^2 \mathbf{I}) \) is an additive white Gaussian noise (AWGN) vector of size \( M \times 1 \) on the \( i \)-th subcarrier at the BS. For simplicity, the discrete time index is ignored in the rest of paper. We rewrite (1) in a full-band form as

\[
\mathbf{Y} = \sum_{k=1}^{K} \mathbf{H}_k \text{diag}(\mathbf{s}_k) + \mathbf{N}, \tag{2}
\]

where \( \mathbf{Y} = \begin{bmatrix} \mathbf{y}(1), \mathbf{y}(2), \ldots, \mathbf{y}(N_c) \end{bmatrix}, \mathbf{H}_k = \begin{bmatrix} \mathbf{h}_k(1), \mathbf{h}_k(2), \ldots, \mathbf{h}_k(N_c) \end{bmatrix}, \mathbf{N} = \begin{bmatrix} \mathbf{n}(1), \mathbf{n}(2), \ldots, \mathbf{n}(N_c) \end{bmatrix}. \)

In the current sounding scheme, \( K \) equals to 8. Let \( \alpha_k \in \{0, 1, 2, \ldots, K-1\} \) denote the cyclic shift value assigned to user \( k \). The SRS symbol \( s_k(i) \) on the \( i \)-th subcarrier of user \( k \) can be generated in the frequency domain as [9] [10]

\[
s_k(i) = r_{bs}(i) \exp(-j2\pi \alpha_k i/K), \quad 0 < i \leq N_c, \quad 0 < k \leq K, \tag{3}
\]

where \( i \) is the subcarrier index in the natural frequency order, and \( r_{bs} \) is the frequency-domain CAZAC base sequence with \( |r_{bs}(i)|^2 = 1 \) for all \( i \). The cyclic shift (in time-domain samples at baseband) resulting from the phase term is given by \( (\alpha_k/K)N_c \).

Substituting (3) into (1), the received signal at the BS is as follows:

\[
\mathbf{y}(i) = \sum_{k=1}^{K} \mathbf{h}_k(i)r_{bs}(i) \exp(-j2\pi \alpha_k i/K) + \mathbf{n}(i), \quad 0 < i \leq N_c. \tag{4}
\]

Prior to channel estimation, the received signal is circularly cross-correlated with the CAZAC base sequence, which can be implemented in the frequency domain as a conjugate multiplication

\[
\mathbf{z}(i) = \mathbf{y}(i) r^*_\text{eq}(i) = \sum_{k=1}^{K} \mathbf{h}_k(i) \exp(-j2\pi \alpha_k i/K) + \mathbf{n}(i)r^*_\text{eq}(i), \quad 0 < i \leq N_c. \tag{5}
\]

The (5) can be rewritten in full-band form as

\[
\mathbf{Z} = \sum_{k=1}^{K} \mathbf{H}_k \theta_k + \mathbf{N}_\text{eq}, \tag{6}
\]

where \( \mathbf{Z} = \begin{bmatrix} \mathbf{z}(1), \mathbf{z}(2), \ldots, \mathbf{z}(N_c) \end{bmatrix}, \theta_k \) is an \( N_c \)-dim phase shift diagonal matrix with elements \( \exp(-j2\pi \alpha_k i/K) \) on its main diagonal, \( \mathbf{N}_\text{eq} = \begin{bmatrix} \mathbf{n}_\text{eq}(1), \mathbf{n}_\text{eq}(2), \ldots, \mathbf{n}_\text{eq}(N_c) \end{bmatrix} \)
with $n_{eq}(i) = n(i) e^{j\theta_i}$, $0 < i \leq N_w$.

To separate the frequency-multiplexed SRSs, we perform time-domain cyclic shift with

$$
F^{-1}(Z) = \sum_{k=1}^{K} \hat{H}_k (n-m_k) + F^{-1}(N_{eq}), \quad 0 < n \leq N_w, \quad (7)
$$

where $\hat{H}_k (n-m_k) = [\hat{h}_1 (1-m_k), \hat{h}_2 (2-m_k), \ldots, \hat{h}_n (N_w-m_k)]$ is a discrete time-domain channel matrix, $\hat{h}_n (n-m_k)$ is a discrete time-domain channel vector, $n$ is the discrete time index, $m_k = (\alpha_k/K) N_w$ is cyclic shift in time-domain sample as described before.

Finally, by applying time-domain filter $w_k$ with window length $N_w/K$, inverse time-domain circular shift with $m_k$, delay samples and FFT to (7), we can get the frequency-domain channel estimate $\hat{H}_k$ for user $k$.

$$
\hat{H}_k = H_k + \sum_{l=1; l \neq k}^{K} H_l \theta_l + N_{eq}, \quad (8)
$$

where $H_k$ is the expected frequency-domain channel corresponding to the windowed time-domain signal $H_k \omega_k$, $H_k$ is the frequency-domain interference channel corresponding to residual interference signal $H_k \omega_k$, $\theta_k$ is a phase shift diagonal matrix with element $e^{-jM_{(l,j)}/K}$, $N_{eq}$ is caused by AWGN.

From (8), we can see the estimated channel $\hat{H}_k$ is severely impacted by interferences from other users in current sounding scheme. Furthermore the interferences increases as the number of multiplexed users increases and the length of multi-path delay, which is reflected by $\hat{H}_k \omega_k$, increases.

### III. PROPOSED SOUNDING REFERENCE SIGNAL DESIGN

The sounding schemes above are shared by LTE FDD and LTE TDD, however, the own characteristics of TDD systems can be used to improve the current sounding schemes. In TDD systems: according to the CQIs periodically reported by each user and the bandwidth requirement of each user, BS dynamically schedules each user to send SRSs only on the allocated PRBs with best CQIs instead of on all PRBs or specified PRBs as in the current schemes. Moreover, each user periodically reports the CQIs of all PRBs to BS, so BS can allocate the $N_{eq}$ PRBs with best CQIs to each user and thus can schedule that each user sends SRSs on the allocated PRBs with best CQIs.

In single-user MIMO (SU-MIMO) systems, the PRBs allocated to each user don’t overlap in frequency domain, so the SRSs sent by each user never overlap in frequency domain according to the characteristic of “distribution according to

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**Fig. 1. Snapshot of SRS bandwidth on one comb.** Assume only 3 users are multiplexed on the same comb, full-bandwidth of system includes only 3 PRBs and bandwidth requirement is one PRB for each user. These three users periodically report their CQIs. At this moment, the CQI of User 1 is the biggest on PRB1; the CQI of User 2 is the biggest on PRB2; the CQI of User 3 is the biggest on PRB3, so in proposed scheme, SRS bandwidth of user 1 is PRB1, SRS bandwidth of user 2 is PRB2, SRS bandwidth of user 3 is PRB3. Note: at next moment, BS can schedule that each user sends SRSs on other PRB due to the changing CQIs.

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**Fig. 2. Operation of the proposed sounding scheme**

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need” in the proposed sounding scheme, thus the interferences among these SRSs multiplexed on the same comb are completely eliminated in the proposed scheme by separating these SRSs in frequency domain. The proposed scheme can be also easily extended to multiple-user MIMO (MU-MIMO) systems, which will be discussed in the last paragraph of this section.

Assume $\mathcal{A}_k$ denoting the set of alternate subcarriers of PRBs allocated to user $k$, $\mathcal{A}_1$, $\mathcal{A}_2$, …, and $\mathcal{A}_K$ belong to the same comb and meet the following constraints in the proposed scheme

$$\mathcal{A}_i \cap \mathcal{A}_j = \emptyset, \quad 0 < k, l \leq K, \quad l \neq k,$$

(9)

where $|\mathcal{A}_i|$ denotes the cardinality of set $\mathcal{A}_i$. So the SRS symbol $s_k(i)$ on the $i$-th subcarrier of user $k$ can be generated in the frequency domain as

$$s_k(i) = \begin{cases} \beta_k r_n(i) \exp(-j2\pi\alpha_k i/K) & i \in \mathcal{A}_i, \\ 0 & i \not\in \mathcal{A}_i, \end{cases},$$

(10)

$$0 < k \leq K,$$

where $\beta_k$ is the amplitude of the transmitted signal to meet the transmit power constraint. Thus the SRS sequence of user $k$ can be expressed in full-band form as

$$s_k = \{s_k(i)\}, \quad 0 < i \leq N_w.$$

(11)

Substituting (10) into (1), the received signal on the $i$-th subcarrier at BS can be obtained as follows

$$y(i) = \sum_{k=1}^{K} h_i(i) \beta_k r_n(i) \exp(-j2\pi\alpha_k i/K) + n(i), \quad 0 < i \leq N_w.$$

(12)

Especially a user set consisting of less than $K$ users can be supported in this model simply by setting some $\beta_k$ to zero. Similar to (2), the full-bandwidth received signal at BS is given by

$$Y = \sum_{k=1}^{K} H_k \text{diag}\{s_k\} + N.$$  

(13)

Define SRSs frequency-domain filter for user $k$ as

$$\text{diag}\{w_k\} \equiv \text{diag}\{w_k(i)\}, \quad w_k(i) \text{ is defined as}$$

$$w_k(i) = \begin{cases} 1, & i \in \mathcal{A}_k, \\ 0, & i \not\in \mathcal{A}_k. \end{cases}$$

(14)

Based on (9), (10) and (14), the following relationship can be deduced

$$\text{diag}\{s_k\} \text{diag}\{w_k\} = \begin{cases} \text{diag}\{s_k\} & k = l, \\ \mathbf{0} & k \neq l, \end{cases}$$

(15)

where $\mathbf{0}$ means the zero matrix of size $N_w \times N_w$. Now apply SRSs frequency-domain filter $\text{diag}\{w_k\}$ to (13), the received signal, denoted by $Y_k$, of user $k$ can be obtained

$$Y_k = \sum_{l=1}^{K} H_k \text{diag}\{s_l\} \text{diag}\{w_k\} + N \text{diag}\{w_k\} \equiv H_k \text{diag}\{s_k\} \text{diag}\{w_k\} + N \text{diag}\{w_k\} + N_{eq,k},$$

(16)

where $N_{eq,k} \triangleq N \text{diag}\{w_k\}$, $Y_k \triangleq \{y_k(1), y_k(2), \ldots, y_k(N_w)\}$ with $y_k(i)$ ($M \times 1$) which is the received signal vector of user $k$ at the $i$-th subcarrier at BS. (16) indicates clearly that the interferences caused by other users multiplexed on the same comb, which always exist in fading channel in current scheme, are completely eliminated in the proposed scheme.

Then multiply (16) with the conjugated CAZAC sequence of user $k$, we have

$$R_k = Y_k \text{diag}\{s_k^*\} = H_k \text{diag}\{s_k\} \text{diag}\{s_k^*\} + N_{eq,k} \text{diag}\{s_k^*\}$$

(17)

$$= \beta_k^2 H_k + N_{eq,k},$$

where $R_k \triangleq \{\bar{r}_k(1), \bar{r}_k(2), \ldots, \bar{r}_k(N_w)\}$, $N_{eq,k} \triangleq N \text{diag}\{s_k^*\}$ and $-CN(0, \beta_k^2 \sigma_n^2 \mathbf{I})$. From (17), we can know that channel estimation is still impacted by noise. In the following, we add a noise filter to effectively reduce noise.

Because the energy of the signal above is more concentrated in time domain compared with that in frequency domain, the noise filter is implemented in time domain. Apply IFFT to (17), we have

$$\tilde{R}_k = \beta_k^2 \hat{H}_k + \mathcal{F}^{-1}(N_{eq,k}),$$

(18)

where $\tilde{R}_k \triangleq \{\tilde{r}_k(1), \tilde{r}_k(2), \ldots, \tilde{r}_k(N_w)\}$, $\tilde{r}_k(n)$ is an $M \times 1$ vector with element $\bar{r}_k(n,a)$, where $n$ is the discrete time index and $0 < n \leq N_w$, and $a$ is antenna index and $1 \leq a \leq M$. $\hat{H}_k \triangleq \{\hat{h}_k(1), \hat{h}_k(2), \ldots, \hat{h}_k(N_w)\}$ is a discrete time-domain channel matrix, $\hat{h}_k(n)$ is a discrete time-domain channel vector of size $M \times 1$.

Now we define the following noise filter, named as threshold filter, meaning that only those samples emerging from noise will be kept.

$$\text{if } |\tilde{r}_k(n,a)|^2 \leq \beta_k^2 \sigma_n^2, \quad \tilde{r}_k(n,a) = 0,$$

(19)

where $\sigma_n^2$ is noise power. In practical communications systems, e.g., LTE systems, $\sigma_n^2$ can be estimated via demodulation reference signal.

After applying the noise filter in (19) and FFT to (18) together with being divided by $\beta_k^2$, we can obtain the frequency-domain estimation channel for user $k$ as follows

$$\hat{H}_k = H_k + N_{eq,k},$$

(20)

where the covariance of the residual noise $N_{eq,k}$ is far less than that of the original noise $N_{eq,k}$ due to the noise filter.

Figs. 3.a and 3.b summarize the sounding processing at the user and BS sides respectively for the proposed sounding
scheme.

![Fig. 3.a. Sounding transmit processing at user side](image)

![Fig. 3.b. Sounding receive processing at BS side](image)

As mentioned in Section II, SRSs are also used to estimate CQIs of all uplink PRBs for the purpose of uplink PRB scheduling. Although the proposed scheme requests each user send uplink SRSs only on the subcarriers of \( K \) downlink PRBs with the best CQIs, the proposed scheme has almost little impact on uplink PRB scheduling. In LTE TDD systems, due to the channel reciprocity, the downlink/uplink PRBs with the same time-frequency indices have the same CQIs under the condition of no inter-cell interferences. Even though inter-cell interferences are considered for the users at the centre of cell, the inter-cell interferences can almost be ignored because the cell-centre users are closed to the desired BS. For the users at the edge of cell, the inter-cell interferences can also be ignored if ICIC (Inter-Cell Interference Coordination) technique is taken at the edge of cell. Therefore the proposed scheme can make sure that each user is allocated the best uplink PRBs and thus has almost little impact on uplink PRB scheduling.

The proposed scheme could be extended from SU-MIMO to MU-MIMO. In SU-MIMO, the allocated PRBs of each user are orthogonal in frequency domain, so these users can use different part of the same CAZAC sequence to send sounding reference signals. If a certain user has a paired user in MU-MIMO mode, the SRSs of the paired user can be obtained via cyclic shift of the CAZAC sequence of this user. A BS can first separate those users in SU-MIMO mode in frequency domain (users in MU-MIMO mode are considered as an equivalent user in SU-MIMO mode), and then separate those users in MU-MIMO mode in time domain via cyclic shift. This method can increase the number of users which are accessed in one TTI, which is especially valuable for the LTE-Advanced, e.g., enhanced MU-MIMO, CoMP (Coordinated Multi-Point) techniques etc.

IV. SIMULATION RESULTS

Consider an LTE TDD system, where the BS with 4 antennas communicates with 8 users, each user is equipped with 1 antenna. The channels are assumed to be time-varying, spatially uncorrelated, frequency selective and Rayleigh fading. Jakes model [11] is used to simulate the time-varying channels. BS uses the channel estimator in (20) for the proposed scheme and the channel estimator in (8) for the current LTE TDD scheme to estimate CSI at transmitter side. Each user sends SRSs on the allocated subcarriers in the proposed scheme (equally subcarrier allocation for each user) while on the all subcarriers in the current LTE TDD scheme. TDD frame configuration 1 is used [9]. We run 3 radio frames (10ms per radio frame) for each simulation case. Simulation parameters, most of which comes from [9] [12], are listed in Table 1 and Table 2.

![Fig. 4](image)

Fig. 4 shows the NMSE in the cases with small, mid and large multi-path but with the same Doppler spread (5Hz) for the proposed scheme and current scheme. It can be observed that the NMSE performance of the proposed scheme is significantly better than that of the current scheme in the case of mid and large multi-path delay. The reason is that in the case of mid and large multi-path delay, the interferences from SRSs of all other users are strong and are the primary factors dominating estimation accuracy. In the proposed scheme the interferences among SRSs don’t exist due to the frequency-domain orthogonalization and separation among SRSs of each user. However, in the current scheme the interferences inevitably exist because the current scheme separates SRSs in time domain and the mid and large multi-path delay damages severely the time-domain orthogonalization of SRSs.

It can be also observed that in the case of small multi-path delay the NMSE performance of the proposed scheme outperforms that of the current scheme in low SNR region and is almost the same as that of the current scheme in Mid-high SNR region. The reason is that in the case of small multi-path delay, the interferences among SRSs are weak and the communications system becomes a noise-limited system at

<table>
<thead>
<tr>
<th>Table 1</th>
<th>Simulation parameters</th>
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<tbody>
<tr>
<td>Parameter</td>
<td>value</td>
</tr>
<tr>
<td>Bandwidth (MHz)</td>
<td>20</td>
</tr>
<tr>
<td>SNR(dB)</td>
<td>0--30</td>
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<tr>
<td>SRS bandwidth</td>
<td>96 PRBs</td>
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<tr>
<td>PRB size</td>
<td>Frequency domain 180kHz</td>
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<tr>
<td></td>
<td>Time domain 1ms</td>
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<tr>
<td>Carrier frequency(GHz)</td>
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<tr>
<td>Doppler spread(Hz)</td>
<td>5(low), 70(Mid), 300(High)</td>
</tr>
<tr>
<td>Number of users ( (K) )</td>
<td>8</td>
</tr>
<tr>
<td>Number of BS antenna ( (M) )</td>
<td>4</td>
</tr>
<tr>
<td>Number of SRS subcarriers per user ( (N_k) )</td>
<td>72 for the proposed scheme 576 for the current scheme</td>
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<table>
<thead>
<tr>
<th>Table 2</th>
<th>Multi-path delay parameters</th>
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<tr>
<td>Delay type</td>
<td>Value (ns)</td>
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<tr>
<td>Small delay</td>
<td>[0 30 70 90 110 190 410]</td>
</tr>
<tr>
<td>Mid delay</td>
<td>[0 30 150 310 710 1090 1730 2510]</td>
</tr>
<tr>
<td>Large delay</td>
<td>[0 50 120 200 230 500 1600 2300 4600]</td>
</tr>
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</table>

We use the average normalized mean square error (NMSE) of channel estimation to evaluate the estimation accuracy for the proposed scheme and the current scheme. It is defined as

\[
NMSE = \frac{1}{MN_k^2} \sum_{k=1}^{K} \sum_{i=1}^{N_k} \left\| h_k(i) - \hat{h}_k(i) \right\|^2.
\]
this moment, the virtue of overcoming SRS interferences of
the proposed scheme is buried. Moreover, the noise power is
small in Mid-high SNR region, the advantage of the noise
filter in the proposed scheme does not come out to play, so the
performance of the proposed scheme is almost the same as
that of the current scheme in Mid-high SNR region. However,
the noise power is big in low SNR region, and the noise filter
applied in the proposed scheme effectively reduces noise, so
the NMSE performance of the proposed scheme is better than
that of the current scheme in low SNR region.

Furthermore, by comparing NMSE curves of different
multi-path delays, we find the proposed scheme is robust to
the multi-path delay (or frequency selectivity) while the
performance of the current scheme degrades as the multi-path
delay increases. The reason is that multi-path delay doesn’t
damage on frequency-domain orthogonalization of SRSs but
damages the time-domain orthogonalization of SRSs.

Fig. 5 shows the NMSE in the cases with low, mid and high
Doppler spread but with the same Mid multi-path delay for
the proposed scheme and the current scheme. It can be seen
that the NMSE performance of the proposed scheme degrades
as the Doppler spread increases while the performance of
current scheme doesn’t change as the Doppler spread changes.
The reason is that the Doppler spread damages the
frequency-domain orthogonalization of SRSs in the proposed
scheme but doesn’t damage the time-domain orthogonalization of SRSs in the current scheme. So as the
Doppler spread increases, the interferences among SRSs
increase in the proposed scheme but remain unchanged in the
current scheme. However, the NMSE performance of the
proposed scheme is still greatly better than that of the current
scheme in the case of low, mid and high Doppler spread at the
whole SNR region. The reason is that the interferences caused
by Doppler spread are less than one caused by mid multi-path
delay. Moreover, the noise filter in the proposed scheme can
well reduce noise in low SNR region.

V. CONCLUSION

In this paper, we have proposed a novel sounding design for
TDD LTE-Advanced systems. The characteristic of “distribution according to need” in our proposed scheme leads
that the interferences among SRSs of different intra-cell users
are eliminated completely. Simulation results indicate that the
proposed scheme outperforms the current scheme in estimation accuracy for various frequency-selective and time-
selective fading channels. Furthermore the proposed scheme is robust to the channel frequency selectivity. Especially our
research is applicable to the uplink sounding reference signal
design in LTE-Advanced and has been adopted as a candidate
scheme for 3GPP LTE-Advanced.

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Fig. 5. Channel estimate NMSE for low, mid and high Doppler spread.