Downlink Multicell MIMO-OFDM: An Architecture for Next Generation Wireless Networks

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Abstract

In this paper, we consider a multicell MIMO-OFDM TDD system where the emphasis is on the downlink for future data-intensive applications. We propose a MIMO scheme that can effectively combat co-channel interference with only local BS coordination and *retain the high peak rate achievable for point-to-point single-cell communications*. We describe several different levels of CSI availability at both the transmitter and the receiver that lead to different system architecture choices. The performance of rate-1 SFBC code developed in [1] with channel estimation is investigated via simulation; our results show that with pragmatic channel estimation schemes, multi-cell MIMO-OFDM is a good candidate for future high-rate applications.

Keywords

Multicell MIMO-OFDM, TDD, co-channel interference, local BS coordination, channel estimation

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***Abstract*— In this paper, we consider a multicell MIMO-OFDM TDD system where the emphasis is on the downlink for future data-intensive applications. We propose a MIMO scheme that can effectively combat co-channel interference with only local BS coordination and *retain the high peak rate achievable for point-to-point single-cell communications*. We describe several different levels of CSI availability at both the transmitter and the receiver that lead to different system architecture choices. The performance of rate-1 SFBC code developed in [1] with channel estimation is investigated via simulation; our results show that with pragmatic channel estimation schemes, multi-cell MIMO-OFDM is a good candidate for future high-rate applications.**

I. INTRODUCTION

Orthogonal Frequencey Division Multiplexing (OFDM) is a multi-carrier transmission scheme that is well-recognized for it's potential for attaining high rate transmission over frequency selective channels. OFDM is thus a promising candidate for mobile broadband wireless and has already been adopted in many high rate wireless communications standards such as DAB/DVB-T, 802.11 WLAN, 802.16 WMAN etc.

The main impediment to the state-of-art of WLANs or WMANs for wireless access is their limited coverage. As is well known from mobile radio, cellular layouts can provide wide-area coverage for wireless communications via frequency reuse. However, requiring each access point (AP) or base station (BS) in such networks to be individually connected to the backbone (wired) network is not a feasible (cost-effective) solution to network scaling. Accordingly, there has been considerable interest in *multi-hop* architectures where the data to/from mobile user reaches the backbone network via multiple hops. This has two clear implications: while all APs or BS need not be wired to the backbone network (only a fraction of gateway APs or BSs are so connected), it requires some (future) protocol for inter-networking the APs or BSs. Our work is thus forward looking in proposing a system concept for the next generation of multi-cellular OFDM based wireless WLANs or WMANs with spatial frequency reuse.

Further, we anticipate that both the mobile user and the access point (the two ends of a link) may potentially have multiple antennas in next generation networks. As is well known, point-to-point (i.e. single user) multi-input, multi-output (MIMO) links potentially can provide diversity and array gains for increased robustness as well as multiplexing gains for increased rate via appropriate space-time-frequency codes that map information symbols to frequency-time and antennas[13][14][15]. However, in the presence of multiple access interference (MAI) in a spatial reuse scenario[6][5], some degrees of MIMO design choice should be allocated to suppression of co-channel interference for higher link spectral efficiency [10]. Clearly, this implies a trade-off: transmit beamforming for downlink is used to mitigate the strong co-channel interferers at the reference receiver, thereby preserving a (nearly) interference-free link for the reference pair. Hence, the remaining degrees of freedom can be used for space-time-frequency (STF) coding as in single-user systems for achieving either desired diversity order or rate per user. Since data-centric applications such as web-browsing imply that traffic will be asymmetric with downlink (AP to user) throughput being the bottleneck, we focus on the design of a downlink TDD multicell MIMO-OFDM system.

As we demonstrate, exploiting transmit diversity on the downlink in the presence of MAI requires some level of channel state information (CSI) availability at the transmitter (access point) for achieving high spectral efficiency. In typical indoor environments, the multipath propagation channels undergo slow fading (at rates no more than 1-2Hz in Doppler) and can be modelled as quasi-static over an OFDM frame. However, in the TDD mode, the uplink and downlink channels can be regarded as reciprocal for contiguous transmissions. Thus channel estimates based on uplink transmission can be exploited by the access point for configuring the transmitter (namely the space-time-frequency coder) for the downlink.

This work therefore looks forward to a future multi-cell scenario based on MIMO-OFDM and specifically investigates downlink schemes with requisite BS coordination that can retain the peak rate achievable for point-to-point communications in a single cell with no co-channel interference through effective co-channel interference avoidance.

To provide a concrete motivation, consider the impact of un-cancelled cochannel interference on a point-to-point MIMO system design with known good STF codes as shown in Fig. 1. Here we used the rate-1 space-frequency block code (SFBC) developed in [1][2] that achieves maximum diversity to encode both the desired signal and all interferers.

We note from Fig.1 that in the absence of CSI at the transmitter, even having number of receive antennas larger ($M_r = 3$) than the number of independent transmitted streams

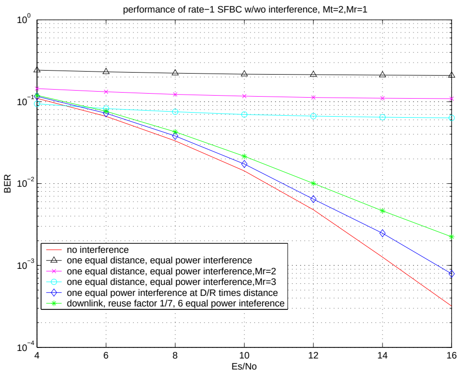

Fig. 1. Performance degradation of rate-1 SFBC in interference-limited scenario

(2) does not prevent loss of diversity. The receiver is a single user sphere decoder that assumes no knowledge of the co-channel interference. For one equal power interferer at equal distance, the number of receive antennas M_r is varied to observe its effect. In all other cases, the number of receive antennas was kept at $M_r = 1$. Significant performance degradation vis-a-vis interference-free point-to-point MIMO STF coded link is observed with only one (sufficiently strong) cochannel interferer. The key lesson is that to combat co-channel interference, MIMO schemes need to exploit any *available CSI at the transmitter*.

Achieving such transmit diversity on the downlink presupposes some form of BS coordination; in that case, the BSs may share not only CSI but also the information streams themselves as a possibility. In the latter case, the participating BS-set may engage in *full joint encoding* [7][8]. However, cooperation among BSs incurs communication overhead that scales exponentially with the number of coordinating BSs. Accordingly, such methods are only feasible if adapted for *local* BS coordination in a wide-area network. Alternately, a mobile user station may implement a suitable multiuser detector (MUD) to combat intercell interference[4][3]. However, the desire to keep the mobile user terminal simple and cost-effective suggests an asymmetric design where the processing complexity is shifted to the BS.

Multicell system architectures based on more limited information exchange among BS are also desirable. If the coordinated BS-set exchange multi-cell CSI, this may be exploited by the reference BS to null it's interference to other co-channel cells, thereby improving link performance significantly.

II. MULTI-CELL MIMO-OFDM ARCHITECTURE

In this work, we consider a cellular wireless time-division duplex (TDD) MIMO-OFDM with TDMA for uplink and time division multiplexing (TDM) for downlink. Therefore, at any time instant, only one user or transmission occupies the whole bandwidth assigned to this cell. In other words, there is only point-to-point communication within a cell.

We consider a conventional FDMA type of cellular overlay with co-channel reuse factor $D/R = \sqrt{3N}$, where N is the cluster size. Fig. 2 shows the traditional $N = 7$ case that equally partition the available system bandwidth. We concentrate on the downlink (i.e. from the BS to user stations) in this work since future data intensive networks are expected to be asymmetric with greater demand on aggregate downlink capacity to support applications such as web browsing.

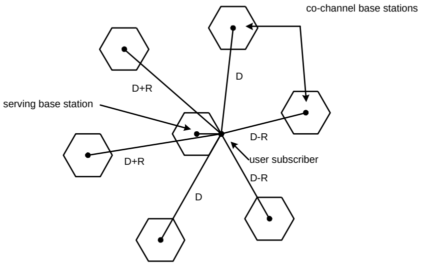

Fig. 2. Multicell Downlink Scenario

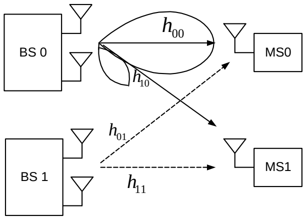

Fig. 3. Channel Estimation in Multicell scenario (BS 0 estimates h_{10}, whereas BS 1 estimates h_{01})

Fig.3 shows a simplified downlink scenario for only 2 cells; each user station receives signals from both the serving BS and the interfering BS. Let BS i denote the serving BS of user station i. h_{ij} stands for the channel from BS j to user station i. If each BS can estimate the channels from itself to the victim user stations in other cells, it can form nulls in those directions by exploiting the differences in spatial signatures. Therefore, inter-cell interference in the downlink can be avoided.

We emphasize two pertinent points that impact the channel estimation approaches within this architecture. (1) The assumption of TDD implies reciprocity in the uplink and downlink channels, hence no feedback between the link end-points is needed. However, both the user stations and BS need to estimate appropriate channels.

(2) Because the channel h_{01} (from BS1 to user station 0) is different from the channel h_{10} (from BS0 to user station 1), to avoid interference to user 0, BS1 need to estimate this channel and can't obtain h_{01} even if BS0 and BS1 can share CSI.

Because of the above, a multi-cell MIMO scenario requires both uplink and downlink channel estimation. During the first phase, cochannel user stations send training sequences (uplink) to their respective BSs. In the second phase, cochannel BSs send training sequences (downlink) to their user stations. Just as the frequency bands are reused in the cellular network, training sequence reuse is also needed, which will be discussed in detail in the subsection IV.A.

III. SYSTEM MODEL

We consider a broadband wireless MIMO-OFDM/TDMA system with time-division duplex (TDD) mode. We presume that the 1st-tier co-channel BSs are coordinated (i.e. are synchronized and share knowledge of their respective training sequences) and each estimates not only the channel from the desired user but also those from the 6 interfering user stations on the uplink. Due to the reciprocity of the TDD mode, the estimated channels on the uplink may be used for downlink beamforming at the BS. We assume M_t transmit antennas at each BS and one receive antenna at each user station. Assuming a single user detector at the mobile users, the receiver only acquires CSI for the link to it's serving BS. The channel responses between each transmit antenna to a user are assumed to be uncorrelated. We consider only the first tier cochannel BSs as interferers and the effect of all other (more distant) cochannel BSs is modelled as white Gaussian noise.

In addition, we assume that all channels undergo quasi-static frequency-selective fading. With reference to Fig. 2, the general received frequency domain signal(indexed by OFDM tone index n) at the desired user in reference cell 0 in the presence of J co-channel BSs is given by

$$\mathbf{r}_0^n \quad = \quad \sum_{j=0}^{J} \mathbf{H}_{0,j}^n \mathbf{x}_j^n + \mathbf{v}_0^n \tag{1}$$

where length-M_r vectors (with M_r the number of receive antennas) $\mathbf{r}_0^n$ and $\mathbf{v}_0^n$ denote the received signal and noise in the reference cell at the n-th tone respectively and $\mathbf{x}_j^n$ is the length-M_t transmitted signal vector at the n-th tone from the j-th BS to desired user. The size $M_r \times M_t$ channel matrix $\mathbf{H}_{0,j}^n$ denotes the *scaled* channel from j-th BS to the user station in reference cell at the n-th tone. The scaling reflects the additional pathloss from interfering BS j to the desired user and is implemented using the scale factor $\rho_{0,j} < 1$ based on a pathloss exponent of 4 to determine $\mathbf{H}_{0,j}^n$ relative to $\mathbf{H}_{0,0}^n$, where $\mathbf{H}_{0,0}^n$ is a random matrix composed of i.i.d. zero-mean unit-variance complex Gaussian random variables. Then, the worst case co-channel interference on the forward channel (see Fig.2) occurs when the desired user is at the vertex of the reference hexagon, with distances from the six co-channel BSs given by D, D, D+R, D+R, D-R, D-R, respectively. The corresponding receive amplitude scaling $\rho_{0,j}$ for $j = 0, 1, \cdots, 6$ takes values of $(R/D)^2$,$(R/D)^2$,$(R/(D+R))^2$,$(R/(D+R))^2$,$(R/(D-R))^2$,$(R/(D-R))^2$, respectively.

IV. COMBINING NULL-BEAMFORMER WITH EXISTING MIMO SCHEME WHEN LOCAL MULTICELL CSI IS AVAILABLE

The main idea of the scheme is as follows: since each BS can estimate the channels to the six interfering user in the first-tier co-channel cells, a beamformer can form nulls in the directions of all cochannel interferers on the downlink and improve link quality at the desired user. Assuming perfect cancellation of all the co-channel 1st tier interfering BS transmissions, existing schemes such as rate-1 SFBC [1][2] can be used for the desired AWGN link. A block diagram of the system with the processing details is shown in Fig. 4.

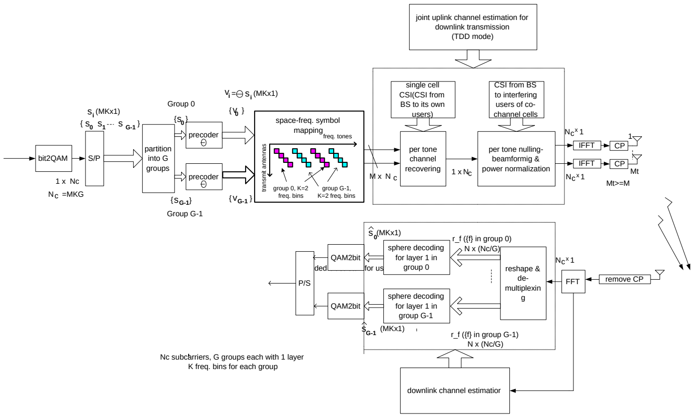

Fig. 4. combining null beamformer with rate-1 SFBC (only block diagrams at one BS and one user station are shown)

Since we only consider one receive antenna in this paper, size $M_r \times M_t$ channel matrix $\mathbf{H}_{i,j}^n$ becomes row vector of length M_t, which we denote as $\mathbf{h}_{i,j}^n$ in the following. Let $\mathbf{x}_0^n$ be the length M_t-vector of transmitted signals from reference BS 0 at the n-th tone intended for desired user; z_i^n be the scalar *received signal component* of the user station in the i-th cell due to $\mathbf{x}_0^n$ at the n-th tone. Then we can define a length-$(J+1)$ (with $J=6$) column vector $\mathbf{z}^n = [z_0^n \cdots z_J^n]^T$ which combines the $J+1$ received signal components in the cochannel cells due to $\mathbf{x}_0^n$.

$$\underbrace{\begin{bmatrix} z_0^n \\ z_1^n \\ \vdots \\ z_J^n \end{bmatrix}}_{\mathbf{z}^n} = \underbrace{\begin{bmatrix} \mathbf{h}_{0,0}^n \\ \mathbf{h}_{1,0}^n \\ \vdots \\ \mathbf{h}_{J,0}^n \end{bmatrix}}_{\mathbf{H}_0^n} \mathbf{x}_0^n \tag{2}$$

where $\mathbf{h}_{i,0}^n$ denotes the length M_t channel response (row) vector from BS 0 to the user station in i-th cell at the n-th OFDM tone and the size $(J+1) \times M_t (M_t > (J+1))$ matrix $\mathbf{H}_0^n$ is the combined frequency channel matrix at the n-th tone from BS 0 to reference user and the nearest $J=6$ interfering cochannel users.

Eqn. (2) suggests that if the reference BS transmitter is to avoid interfering with 1st-tier cochannel cells, $\mathbf{x}_0^n$ must be chosen such that the interfering components $z_i^n, i = 1, \cdots, J$ are zero, while keeping the desired received signal component z_0^n as large as possible. If each BS is able to *locally* avoid intercell interference in this manner, the link between the reference BS and it's intra-cell user reduces to a point-to-point channel. In order to form nulls in the directions of the user stations in neighboring cochannel cells, the $\mathbf{x}_0^n$ is constructed as follows:

$$\mathbf{x}_0^n \quad = \quad \frac{1}{\sqrt{\gamma_0^n}} (\mathbf{H}_0^n)^\dagger \ \mathbf{e}_1 \ \mathbf{h}_{0,0}^n[:, selected\ M\ columns] \ \mathbf{s}_0^n \tag{3}$$

where $(\cdot)^\dagger$ denotes the pseudo-inverse, i.e., $(\mathbf{H}_0^n)^\dagger = (\mathbf{H}_0^n)^H \left(\mathbf{H}_0^n (\mathbf{H}_0^n)^H\right)^{-1}$ with $(\cdot)^H$ the Hermitian transpose of the matrix argument, and the vector $\mathbf{e}_i$ is the i-th column of the identity matrix. For the pseudo-inverse to exist, $M_t \geq (J+1)$ is a necessary condition. Left-multiplying $\mathbf{x}_0^n$ by $\mathbf{H}_0^n$ yields a vector proportional to $\mathbf{e}_1$ and hence $(\mathbf{H}_0^n)^\dagger \mathbf{e}_1$ functions as a null-beamformer. In addition, $\mathbf{s}_0^n$ denotes the column of the rate-1 SFBC code matrix transmitted from the BS 0 at the n-th tone.

The length-M row vector $\mathbf{h}_{0,0}^n[:, selected\ M\ columns]$ denotes the channel from the BS 0 to its user station at the n-th tone due to($M < M_t$) transmit antenna selection and the corresponding rate-1 SFBC code designed for the M-transmit antenna case is used. Note that if the antennas at the BS form a linear array and are not sufficiently separated, choosing the $M < M_t$ antennas out of this array may reduce the possible spatial correlation. Finally, to keep the total transmit power to be constant for each tone, the normalization coefficient is chosen as $\gamma_0^n = ||(\mathbf{H}_0^n)^\dagger\ \mathbf{e}_1\ \mathbf{h}_{0,0}^n[:, selected\ M\ columns]\ \mathbf{s}_0^n||^2$, which is data and channel dependent.

Using eqn.(1) and (3), we get

$$\begin{aligned} r_0^n &= \mathbf{h}_{0,0}^n \mathbf{x}_0^n + v_0^n & (4)\\ &= \frac{1}{\sqrt{\gamma_0^n}} \mathbf{h}_{0,0}^n[:, selected\ M\ columns]\ \mathbf{s}_0^n + v_0^n & (5)\end{aligned}$$

This is very similar to previous point-to-point communication model, other than the normalization coefficient $\frac{1}{\sqrt{\gamma_0^n}}$ which is both data and channel dependent. It is hence infeasible for a user station to estimate γ_0^n; we suggest using $\sqrt{E\gamma_0^n}$ to replace $\sqrt{\gamma_0^n}$[9][11], where E denotes the expectation over the channel and data. In the following, it is assumed that $\sqrt{E\gamma_0^n}$ is applied for transmit power normalization. For any space-frequency (SF) code scheme such that all the entries of the SF code matrix are nonzero, we show that for *one* receive antenna,

$$E\gamma_0^n = E\gamma = M \times \frac{M_t - 1}{M_t(M_t - J - 1)} \tag{6}$$

Due to space limitation, the detailed derivation is omitted. Note that $E\gamma_0^n$ does not depend on the specific cell and tone index. From eqn.(6), it is clear that when $M_t = J + 1$, $E\gamma$ approaches infinity irrespective of the value $M(M_t)$. In addition, eqn.(5) shows that increasing $E\gamma$ at the transmitter is equivalent to enhancing the noise power. Therefore, to reduce $E\gamma$, we generally let $M_t > J+1$ (e.g. when $J = 6$, let $M_t = 8$ or even larger).

For the case where rate-1 SFBC is combined with nulling-beamformer (we call it ZFSFBC in the sequel), the code matrix of rate-1 SFBC only has nonzero values on diagonal positions (see Fig.4) and only one symbol is transmitted from a transmit antennas at each tone; therefore, $E\gamma = E\gamma_0^n = E||(\mathbf{H}_0^n)^\dagger\ \mathbf{e}_1\ h_{0,0,i}^n\ s_{0,i}^n||^2$ where $s_{0,i}^n$ is the nonzero symbol in the column vector $\mathbf{s}_0^n$ corresponding to i-th antenna at the n-th tone and $h_{0,0,i}^n$ is the corresponding channel coefficient in the channel row vector $\mathbf{h}_{0,0}^n[:, selected\ M\ columns]$. As a result, $E\gamma$ for ZFSFBC is independent of M and equals

$$E\gamma_{ZFSFBC} = \frac{M_t - 1}{M_t(M_t - J - 1)}. \tag{7}$$

For $M_t > J + 1$, $E\gamma < 1$, which is equivalent to increasing the signal power with respect to the noise. Thus, as verified by simulations, ZF-SFBC has a somewhat improved performance than original rate-1 SFBC for the same M and K.

It is worth noting that for the case $M_r > 1$, existing high rate schemes such as VBLAST can also be used assuming CSI of the reference channel between BS and desired user. However, for the pseudo-inverse to exist, $M_t \geq (J + 1)M_r$ would now be necessary, where $J = 6$ is number of nearest cochannel interference. In addition, $M_r \geq M$ is generally needed to separate M spatially independent substreams for each user with BLAST type of receiver. Moreover, the $\mathbf{e}_1$ in Eqn. (3) should be replaced by $[\mathbf{e}_1\ \cdots\ \mathbf{e}_{M_r}]$ and $E\gamma$ needs to be recalculated.

A. Key System Implementation Issues

ZFSFBC for future multi-cell MIMO architecture needs the support of simple but effective channel estimation method. *Although in this paper, we only study downlink schemes, the actual channel estimation is conducted on the uplink; the BS transmitter uses these channel estimates based on reciprocity of TDD for the downlink beamforming. Separately, downlink training sequences are sent for the desired user stations to obtain CSI of the reference channel at the user that is needed for sphere decoding.*

For low-mobility or fixed wireless communications, a quasi-static channel can be assumed for each OFDM frame. As is standard in packetized communications, we use a preamble of one or more known OFDM symbols (i.e. training symbols) at the beginning of the frame and thus only preamble based channel estimation methods are considered.

For any specific time slot and frequency tone, Eqn.(1) serves as the receive signal model for uplink channel estimation ($\mathbf{H}_{0,j}$ is replaced by $\mathbf{h}_{0,j}$ and $\mathbf{x}_j^n$ replaced by x_j^n respectively), where $\mathbf{h}_{0,j}$ is the uplink channel from user in the j-th cochannel cell to reference BS and x_j^n is the training symbol on the n-th tone from user in the j-th cell. Eqn.(5) serves as the reception model for downlink channel estimation, where $\mathbf{s}_0^n$ is regarded as the training symbol vector sent from the reference BS to the desired user at the n-th tone.

Among preamble based channel estimation methods for MIMO-OFDM systems, [12] used LS method to obtain the channel estimates *given training sequences* and then selected the training sequences based on minimizing the MSE. It was shown that specially designed *local* orthogonal sequences are optimal for this time domain estimator architecture; further the estimates are attained with low complexity within *one* OFDM training symbol time if the constraint of $B\,L < N_c$ is satisfied, where B denotes total number of transmit antennas (*from single user in point-to-point MIMO or from all cochannel users involved in the local joint training in multiuser MIMO*), N_c is number of tones of the OFDM systems, and L is number of sampled nonzero taps of the channel impulse response. The training sequences we adopted have the following form: let $t_1[n,k]$'s for $k = 0, \cdots, N_c - 1$ be any training sequence at training time n that is good for timing and frequency synchronization and possibly other properties in OFDM systems. Then $t_i[n,k] = t_1[n,k]W_{N_c}^{-\bar{L}(i-1)k}$ for $i = 2, \cdots, B$, where $W_{N_c} = exp(-j(2\pi/N_c))$ and $\bar{L} = \lfloor N_c/B \rfloor \geq L$[12]. The above condition implies a system with sufficiently large N_c; in our simulations, we use FFT size of 256 consistent with the 802.16 standard. However, for *quasi-static* fading channels, it is possible in principle to *increase the FFT size even further* as long orthogonality between sub-carriers can be maintained (typically by adjusting the OFDM frame duration). Moreover, we demonstrate that the channel estimator is robust (suffers negligible degradation) to mild underestimation of the number of channel taps. These two facts imply that the channel estimator in [12] can be extended to the scenarios with a) increasing number of cochannel users or b) cochannel users each with more transmit antennas. Finally, when $B < N_c/L$ still can't be satisfied, we can use the following simple scheduling method to ensure that the necessary condition is satisfied: partition the 1st tier cochannel users into several groups with N_c/L user stations per group; each group sends their training sequences during their respective allocated one OFDM symbol duration. In this way, channel estimates of all these user stations can be obtained in a few OFDM symbol durations. We use [12] for both local uplink intercell joint channel estimation (assume perfect pathloss information is known at the BSs) and downlink intracell channel estimation.

In a multi-cellular system, the optimal training sequences so derived can be intelligently re-used. It is intuitive that the co-channel cells should be assigned different orthogonal training sequences [12] (see Fig. 5), but the cells in the same cluster with different frequency subchannels could reuse the same set of training sequences. In addition, all the user stations within each cell can share the same training sequence since they are orthogonal in time. For the uplink training sequence reuse, we have some additional assumptions: (1) each $J + 1$ (with $J = 6$) local co-channel cells are synchronized and (2) each BS has knowledge of all the training sequences of its J 1st tier co-channel cells and can thus estimate the channels between itself and nearest $J + 1$ co-channel users.

During the training stage, all the cells perform *joint* uplink training based on the training sequence assignment pattern. Each BS estimates the channels between itself and the 1st-tier co-channel cells assuming that signals from secondary co-channel interferers can be neglected. For the downlink training, the same training sequence can be used for all co-channel cells assuming that null beamformer is (near) perfect and cancels any co-channel interference.

B. Simulation results of null-beamfoming SFBC (ZFSFBC)

In the simulation, we use $M_t \geq 8$ transmit antennas at each BS and one receive antenna at each user station. Rayleigh fading multipath channels with i.i.d taps and an exponential power delay profile was assumed with r.m.s. delay spread of 50ns. With the sampling rate of 20MHz, this is equivalent to $L = 1 + 10\frac{\tau_{rms}}{T_s} = 11$ taps for each channel. To investigate the validity of the assumption that only one tier interferers need to be considered, we leave the 2nd tier interferers uncancelled by null-beamformer. We show in Fig. 6 that performance of ZFSFBC with uncancelled 2nd-tier interferers as a function of reuse factor N; as expected, performance improves with increasing N. Thus for $M_t = 8$, the second-tier of interferers can be neglected for larger reuse factor $N > 9$, i.e., the diversity of our rate-1 SFBC is preserved with the help of null-beamformer. Moreover, in this case the ZF-SFBC has slightly better performance than the original rate-1 SFBC with the same code matrix due to the fact that $E\gamma < 1$ when $M_t > J + 1$. Further, it is seen in Fig. 7 that increasing the number of transmit antennas at each BS helps to reduce the reuse factor needed to retain the diversity of original rate-1 SFBC.

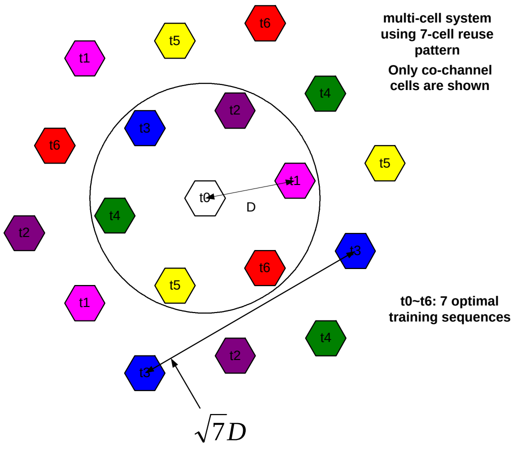

Fig. 5. Training sequences reuse for uplink joint channel estimation

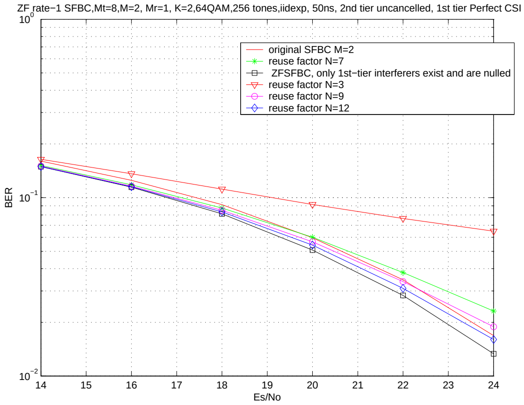

Fig. 6. The effect of different reuse factors on ZFSFBC with uncancelled 2nd-tier interferers

In Fig. 8, we investigate the effect of channel estimation error on performance of ZFSFBC. There is 1dB loss due to channel under-estimation error when $L = 7$ is assumed by the channel estimator although the *true* channel has 11 taps. This indicates that for the exponential power delay profile, modest underestimation of the number of taps doesn't affect the diversity order. However, when L is underestimated beyond a threshold (e.g. $L = 5$ in the plot), the diversity of ZFSFBC is impaired. In addition, with larger reuse factor, the channel estimation accuracy is also improved.

In Fig. 9, we investigate the impact of FFT size on the channel estimation accuracy. When $N_c > BL$, larger N_c

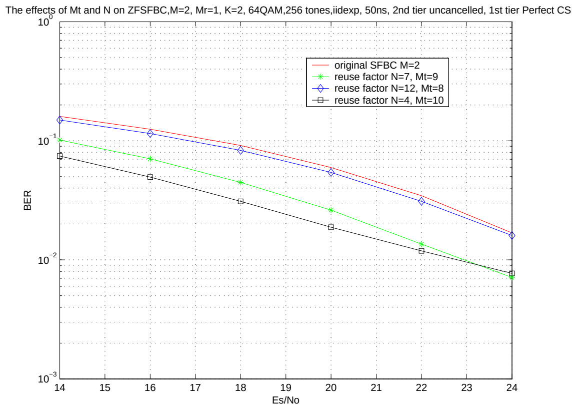

Fig. 7. Performance of ZFSFBC w.r.t. N and Mt with uncancelled 2nd-tier interferers

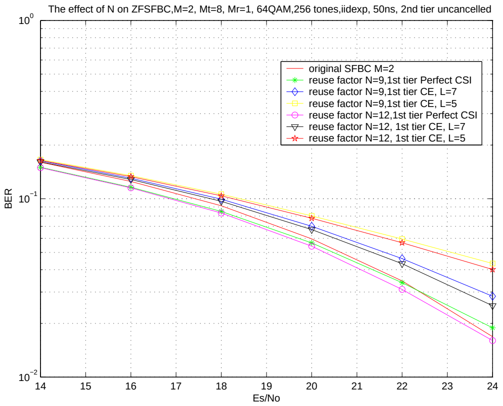

Fig. 8. Performance of ZFSFBC under CE error with uncancelled 2nd-tier interferers

improves the channel estimation accuracy as expected since more observation variables are available (with large N_c) to estimate the$(B\ L)$ unknown variables.

V. CONCLUSION

In this paper, a novel downlink scheme is proposed for multicell MIMO-OFDM system which can effectively avoid causing interference to co-channel cells with only local BS coordination and thus can retain the desirable performance of point-to-point communication in the multicell scenario.

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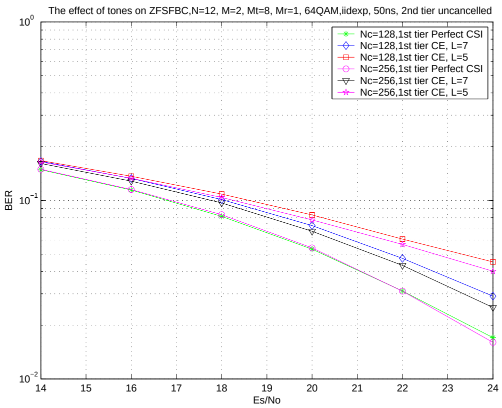

Fig. 9. The effect of different FFT size on ZFSFBC under CE error with uncancelled 2nd-tier interferers

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