Massive MIMO Systems in Non-Contiguous Bands with Asymmetric Traffics

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Abstract—Massive MIMO systems offer large spectrum efficiency improvements but when applied to systems with non-contiguous bands as commonly encountered in practice, their performance gains are largely offset by the overhead in acquiring channel state information (CSI). This paper presents three schemes to enable overhead-efficient massive MIMO in non-contiguous bands with frequency-selective channels. By incorporating CSI errors, CSI overhead, pilot contamination, and guard interval cost, achievable rate or throughput expressions of uplink and downlink of the proposed schemes are derived. Novel resource adaptation mechanisms between uplink and downlink are proposed and analyzed. Comparing with the conventional frequency division duplexing scheme, analytical/numerical studies elaborate substantial spectrum efficiency enhancement and flexible resource adaptation capability of the proposed schemes.

Index Terms—CSI overhead, Massive MIMO, Multi-band TDD, Non-contiguous bands, Resource adaptation, Rotating FDD

I. INTRODUCTION

Large-scale antenna technology also known as massive multi-input multi-output (MIMO) can substantially enhance spectrum and energy efficiency [1] and has emerged as an enabler for next generation wireless systems [2]. The key ingredient for massive MIMO is the base station (BS)’s knowledge of individual channel state information (CSI) of all of its user equipments (UEs) for both uplink (UL) and downlink (DL). For a system with a single contiguous band using time division duplexing (TDD), CSI of UL and DL can be obtained through the use of UL pilots and channel reciprocity. The pilot overhead is generally proportional to the number of total transmit antennas [3], and that number is smaller for UL than DL in massive MIMO systems, which justifies the approach of UL pilots. Due to the natural fit of TDD to the CSI need of massive MIMO, most existing massive MIMO works (e.g., [3], [4] and references therein) in the literature consider TDD systems.

Spectrum allocations using non-contiguous bands are quite common, e.g., see LTE frequency channels [5], due to pre-existing spectrum allocations and non-availability of a sufficient single contiguous band. For such scenarios, the UL and DL of a system operate in different bands, which is commonly known as frequency division duplexing (FDD). An advantage of FDD systems is that under a peak power constraint, they offer larger coverage than TDD systems. However, as UL and DL CSIs are different in FDD systems, acquiring individual DL CSIs for massive MIMO operation would require too excessive overhead. This overhead issue stands as a major limitation of massive MIMO in FDD mode. There are only a few works on massive MIMO with FDD in the literature. They include low complexity CSI quantization [6], precoding schemes to reduce CSI overhead by exploiting spatial correlation of the channels across BS antennas [7], [8], and open-loop and closed-loop training mechanisms exploiting temporal correlation of the channels to reduce feedback overhead in UL transmission [9]. These works also highlight importance and need of overhead-efficient massive MIMO in FDD mode. They accomplish some degree of overhead reduction, yet the overhead issue still remains.

Wireless systems generally possess asymmetric traffics where DL data rates are higher than UL data rates. In addition, the traffic loads and asymmetry vary across times. Under such scenarios, using a fixed resource allocation between UL and DL does not yield efficient resource utilization and UE support. To address this issue, recent TDD-based wireless systems introduce resource adaptation between UL and DL by means of adjusting UL duration and DL duration, i.e., with different UL and DL frame durations or with different numbers of UL and DL sub-frames of the same duration [5]. However, for FDD-based systems, such resource adaptation is infeasible. This fact is commonly viewed as one of the fundamental limitations of FDD systems. However, as mentioned above, both non-contiguous spectrum allocations and time-varying traffic asymmetry continue to exist, and hence new innovations to overcome these issues are much needed.

Motivated by the above issues, this paper considers massive MIMO in non-contiguous bands with frequency-selective channels and time-varying asymmetric traffics. The technical contributions include i) three system architectures that facilitate overhead-efficient massive MIMO in non-contiguous bands, ii) analytical performance assessment for each architecture, and iii) resource adaptation mechanisms between UL and DL of the systems with non-contiguous bands. Pilot contamination issue is incorporated. We focus on system level innovations not constrained by device implementation complexity and cost. Novelty aspects are discussed next.

First, the three system architectures have not been considered for massive MIMO systems. These architectures respectively exploit the concepts of rotating FDD (RFDD), synchronous dual band TDD (STDD), and asynchronous dual band TDD (ATDD). In RFDD, the UL and DL bands are rotated after each transmission frame by which the required CSIs for UL and DL are always obtained from the UL pilots. We independently perceived this RFDD concept and later found out that it appeared in a patent application [11] in the

1Implementation and interference management issues are open for further investigation. They are not limiting factors as even a full duplex radio, once thought to be impossible, has very recently become a reality (e.g., see [10] and references therein).
name of band switching (but no transceiver architecture was presented there). STDD applies TDD on both non-contiguous bands synchronously. Under the same peak transmit power constraint, STDD has a smaller coverage range than RFDD. ATDD also uses TDD on both bands but with a particular time offset (thus, asynchronously) between the transmission frames of the two bands. The time offset can be chosen to avoid simultaneous transmissions from the two bands which overcomes the coverage range limitation of STDD under the peak power constraint. To the best of our knowledge, STDD and ATDD in non-contiguous bands are new while the application of RFDD in massive MIMO is also new.

Next, our analytical performance assessment not only provides the performance characteristics of the three new system architectures but also incorporates system architecture dependent overhead on top of the rate analysis which is new.

Finally, to the best of our knowledge, there is no existing approach for resource adaptation between UL and DL for systems with non-contiguous bands. Such resource adaptation capability is highly desired and beneficial due to UL and DL traffic disparities and fluctuations. Thus, our resource adaptation mechanism and its resolution enhancement approach are novel and their impact on performance enhancement can be quite significant.

The rest of the paper is organized as follows. Section II describes the system and Section III presents the three system architectures for massive MIMO in non-contiguous bands. Section IV develops analytical achievable rate bounds for each architecture. Section V introduces resource adaptation mechanisms between UL and DL. Numerical performance results are presented in Section VI and conclusions are drawn in Section VII.

Notation: Vectors (matrices) are denoted by bold face small (big) letters. The superscripts $T$, $H$, and $*$ stand for the transpose, complex conjugate transpose and conjugate of a matrix or vector, respectively. $I_K$ is the $K \times K$ identity matrix and $0_I$ is the $L \times 1$ all zero vector. $\mathbb{E}\{\cdot\}$ and $||\cdot||$ denote expectation and Euclidean norm, respectively. $\mathbf{n} \sim \mathcal{CN}(0, \mathbf{C})$ means $\mathbf{n}$ has a probability density function (pdf) of the zero-mean complex Gaussian vector with covariance matrix $\mathbf{C}$. $\mathcal{CN}(\mathbf{0}, \sigma^2)$ denotes a zero-mean Gaussian pdf with variance $\sigma^2$. The $\delta_{ij}$ and $\Gamma(\cdot)$ denote the Kronecker delta and the Gamma function, respectively. $\text{diag}\{a_1, \ldots, a_K\}$ means a diagonal matrix with diagonal elements $\{a_1, \ldots, a_K\}$.

II. SYSTEM DESCRIPTION

We consider a multi-cell massive MIMO system with two non-contiguous bands which are assumed to have a sufficient frequency separation between them to avoid interference to each other (as commonly used in FDD systems). Extension to more than two non-contiguous bands is also possible but for better illustration of the proposed architectures and concepts, we will use two bands throughout the paper. The carrier frequencies of the lower band and the higher band are denoted by $f_L$ and $f_H$. We consider OFDM transmission with a sub-carrier spacing of $1/T$ Hz and a cyclic prefix interval of $T_{CP}$ seconds for frequency-selective channels. With DFT-precoding, OFDM can be converted to single carrier transmission with frequency domain equalization capability. For frequency-flat channels, OFDM with DFT-precoding without cyclic prefix yields a single carrier system. Thus, there is no loss of generality. The transmissions in the lower and higher bands can be respectively represented by individual OFDM systems with $N_1$ sub-carriers and $N_2$ sub-carriers having the same sub-carrier spacing, i.e., with bandwidths $N_1/T$ and $N_2/T$ Hz.

Each BS has $M$ antennas and each cell has $K$ single-antenna UEs. $^2$ User scheduling based on channel knowledge can enhance system sum rate, e.g. see [12], but we do not consider such issue. The frequency reuse factor is 1, and there are $N_{cell} - 1$ interfering neighbor cells. Each BS uses maximum ratio combining (MRC) receiver and maximum ratio transmission (MRT). We note that our beamforming uses estimates of channel gains as in MRC/MRT but there also exist arrival/departure angle based beamforming and related estimation schemes in the literature, e.g., [13]. We consider a practical constraint that the transmitters are peak power constrained. As we are interested in achievable rate (or its bound), we assume that each UE transmits on all available subcarriers. As massive MIMO averages out the small scale fading effect, processing across subcarriers to counter frequency-selectivity is not needed.

Each frame consists of an UL sub-frame with duration $T_{UL}$, a receive-to-transmit guard interval $T_{RTG}$ (for circuit switching time to change from receive mode to transmit mode), a DL sub-frame with duration $T_{DL}$, and a transmit-to-receive guard interval $T_{TRG}$ (for two-way propagation and circuit switching time), or it can start with a DL sub-frame followed by $T_{TRG}$, an UL sub-frame, and $T_{RTG}$. The frame length is generally determined by the latency requirement, channel time variation level, and frame overhead, and typically fixed in cellular standards. Thus, we consider a fixed frame duration, but $T_{UL}$ and $T_{DL}$ can be changed for resource adaptation between UL and DL. Suppose the first sub-frame has $N_{sym1}$ OFDM symbols and the second sub-frame has $N_{sym2}$ OFDM symbols. Then, depending on the resource adaptation, $N_{sym1}$ and $N_{sym2}$ can be changed but $N_{sym1} + N_{sym2} \equiv N_{sym}$ is fixed.

We consider quasi-static channels where channel gains remain the same within each frame. Small-scale fading channels are assumed to be independent and identically distributed across BS antennas and across UEs with $L_1$ and $L_2$ channel taps having Rayleigh envelopes in the lower and higher bands, respectively, (frequency-flat if $L_i = 1$ and frequency-selective if $L_i > 1$). For conciseness, in the following, we will use $L$ channel taps and $N$ sub-carriers without distinguishing between the lower and upper bands. For example, if the considered band is the lower band, it implies that $L = L_1$ and $N = N_1$.

For CSI acquisition, each of the $K$ UEs in a cell is assigned with one of the $K$ orthogonal pilot signals in each band to

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$^2$For a UE with multiple antennas, independent channels can be viewed as single-antenna virtual UEs. For correlated channels, an appropriate precoding can transform those correlated channels into (possibly a smaller number of) orthogonal channels which can be treated as single-antenna virtual UEs.
transmit at the beginning of its UL transmission. The same pilot set is reused in other cells. The minimum pilot overhead is \( K_L^1 \) tones for the lower band and \( K_L^2 \) tones for the higher band. We adopt optimal orthogonal pilot designs from [14] which yield minimum overhead if \( \log_2(L) \) is an integer. If minimum pilot overhead is desired for the case where \( \log_2(L) \) is not an integer, the pilot design with approximately equal pilot tone spacings from [15] can be used at a cost of marginal performance degradation. If \( KL > N \), more than one OFDM symbol is needed for pilot transmission, and time division multiplexing (TDM) pilot design or code division multiplexing across time (CDM-T) pilot design from [14] can be applied. Within each OFDM symbol carrying pilots, frequency division multiplexing (FDM) or code division multiplexing in frequency domain (CDM-F) pilot designs from [14] can be applied. Thus, without loss of generality, we will present pilot signal model of one OFDM symbol. Suppose the \( K \) pilot signals in time-domain are denoted by \( \{ x_k : k = 1, 2, \cdots, K \} \). Define \( X_k \) to be an \( N \times L \) circulant matrix with the first column given by \( x_k \), for \( k = 1, \cdots, K \), and the \( m \)-th column given by the \( m \) times downward cyclic shifted version of \( x_k \). With the orthogonal pilot design, we have \( X_j X_k = E_0 a_{mk} \delta_{jk} \mathbf{I}_L \), \( N \geq L \). We define the pilot power in time domain as \( q \triangleq E_x^2 / N \).

The large-scale fading (path loss and shadowing) model is \( \beta_{lk} = \alpha \psi / (d_{0}^{a} d)^{\nu} \) with \( \alpha = (\log_{10}(\psi))^{2} \) [16], where \( \lambda \) is carrier wavelength, \( d \) and \( d_0 \) are the distance and the reference minimum distance of user to BS, \( \nu \) is the path loss exponent and \( \psi \) is log-normal shadow fading, i.e. \( 10 \log_{10}(\psi) \sim N \left( 0, \sigma_{\text{shadow}}^2 \right) \). Note that we can normalize \( \beta \) and noise by \( a \) without affecting Signal to Noise Ratio (SNR), hence in the following signal model we use \( \beta = \beta / a \) and the normalized noise.

The channel impulse response between UE \( k \) in cell \( l \) and antenna \( m \) of the BS in cell \( i \) (will be simply called BS \( i \)) is given by an \( L \times 1 \) vector \( g_{ilk} = \sqrt{\beta_{ilk}} h_{ilk} \), where \( \beta_{ilk} \) is a real variable responsible for large-scale fading and \( h_{ilk} \) is a complex \( L \times 1 \) vector responsible for small scale fading. \( \beta_{ilk} \) is assumed to be the same for all channels between UE \( k \) in cell \( l \) and all the antennas at BS \( i \) due to the assumption of co-located BS antennas, thus the index \( m \) is omitted. We assume \( h_{ilk} \) to be Rayleigh fading with uncorrelated taps, i.e., \( h_{ilk} \sim \mathcal{CN}(0, \Sigma) \) with \( \Sigma = \text{diag}([\sigma_1^2, \cdots, \sigma_L^2]) \) and \( \text{tr}(\Sigma) = 1 \). In the same way as [17]–[19], we assume large scale fading coefficients \( \{ \beta_{ilk}, l = 1, \cdots, N_{\text{cell}}, k = 1, \cdots, K \} \) are known at all BSs, and this assumption is justified since large scale fading coefficients change very slowly compared with small-scale fading coefficients and they can be reliably estimated.

The \( N \times 1 \) received pilot vector in time domain at antenna \( m \) of BS \( i \) is given by

\[
y_{im} = \sum_{l=1}^{N_{\text{cell}}} \sum_{j=1}^{K} X_{lj} g_{ilkj} + n_{im} \tag{1}
\]

where \( n_{im} \) is AWGN with \( n_{im} \sim \mathcal{CN}(0, \sigma_{\text{noise}}^2 \mathbf{I}_N) \) with \( \sigma_{\text{noise}}^2 = 1 \). BS \( i \) applies minimum mean squared error (MMSE) channel estimation as [21]

\[
\hat{g}_{ilk} = \Phi_{\text{ilk}} X_{lj}^H y_{im} \tag{2}
\]

where \( \Phi_{\text{ilk}} \triangleq \frac{1}{E_a} \left( \frac{1}{\beta_{ilk}} \Sigma^{-1} + \Psi_{ilk}^{-1} \right)^{-1} \Psi_{ilk}^{-1}, \) with \( \Psi_{ilk} \triangleq \left( \sum_{l'=1}^{L} \beta_{ilk'} \right)^{-1} \). After more simplification, we have

\[
\Phi_{\text{ilk}} = \beta_{ilk} \text{diag}(\eta_{lk}, 1, \cdots, \eta_{lk}) \tag{3}
\]

where \( \eta_{lk,m} \triangleq \frac{\sigma_2^2}{1 + \sigma_2^2 \text{tr}(\Sigma) / \sigma_2^2 \beta_{ilk} + 1 / (E_a \sigma_2^2)} \). Hence, from (2) and (3), we conclude

\[
\hat{g}_{ilk} = \frac{\beta_{ilk}}{\beta_{ilk}} g_{ilk} \tag{4}
\]

The role of (4) is to facilitate the rate analysis by decomposing the inter-cell interference into a correlated term and an uncorrelated term with respect to the desired user’s channel estimate (more details in the Appendix).

Next, due to the MMSE property under the Gaussian signal model, the channel estimate \( \hat{g}_{ilk} \) and its error \( e_{ilk} \triangleq g_{ilk} - \hat{g}_{ilk} \) are independent and they are distributed as \( g_{ilk} \sim \mathcal{CN}(0, \Sigma_{g_{ilk}}) \) and \( e_{ilk} \sim \mathcal{CN}(0, \Sigma_{e_{ilk}}) \). \( \Sigma_{g_{ilk}} \) and \( \Sigma_{e_{ilk}} \) are \( L \times L \) diagonal matrices with \( j \)th diagonal elements given by \( (\Sigma_{g_{ilk}})_{jj} = \beta_{ilk} \sigma_2^2 \) and \( (\Sigma_{e_{ilk}})_{jj} = \beta_{ilk} \sigma_2^2 \Sigma^{-1} / \sigma_2^2 \beta_{ilk} + 1 / E_a \). The \( N \times 1 \) frequency domain channel vector for the link from UE \( k \) in cell \( l \) to antenna \( m \) of BS \( i \) is given by \( g_{ilkm} = \sqrt{N F} F_L g_{ilk} \), where \( F_L \) is first \( L \) columns of the \( N \)-point unitary DFT matrix. The corresponding frequency domain channel vector estimate and its error vector are \( \hat{g}_{ilkm} = \sqrt{N F} \hat{F}_L g_{ilk} \) and \( e_{ilkm} = \sqrt{N F} \hat{F}_L e_{ilk} \). Again, due to the MMSE property, \( g_{ilkm} \) and \( e_{ilkm} \) are independent and distributed as \( g_{ilkm} \sim \mathcal{CN}(0, \hat{\Sigma}_{g_{ilkm}}) \) and \( e_{ilkm} \sim \mathcal{CN}(0, \hat{\Sigma}_{e_{ilkm}}) \), where their covariance matrices are given by \( \hat{\Sigma}_{g_{ilkm}} = N F L \Sigma_{g_{ilk}} g_{ilk}^H \) and \( \hat{\Sigma}_{e_{ilkm}} = N F L \Sigma_{e_{ilk}} g_{ilk}^H \). Note that \( \hat{\Sigma}_{g_{ilkm}} \) and \( \hat{\Sigma}_{e_{ilkm}} \) individually have the same diagonal elements which are given by

\[
[
(\hat{\Sigma}_{g_{ilkm}})_{nn} = \text{tr}(\hat{\Sigma}_{g_{ilkm}}) \triangleq \zeta_{ilk} \tag{5}
\]

\[
(\hat{\Sigma}_{e_{ilkm}})_{nn} = \text{tr}(\hat{\Sigma}_{e_{ilkm}}) \triangleq \eta_{ilk} \tag{6}
\]

The channel gain, its estimate, and estimation error on subcarrier \( n \), given by the \( n \)th elements of the corresponding vectors, are respectively denoted by \( g_{ilkmn}, \hat{g}_{ilkmn}, \) and \( e_{ilkmn} \), with the following relationship

\[
g_{ilkmn} = \hat{g}_{ilkmn} + e_{ilkmn} \tag{7}
\]

and are distributed as \( g_{ilkmn} \sim \mathcal{CN}(0, \zeta_{ilk}) \), \( \hat{g}_{ilkmn} \sim \mathcal{CN}(0, \zeta_{ilk}) \) and \( e_{ilkmn} \sim \mathcal{CN}(0, \eta_{ilk}) \). The statistics of the channel, channel estimate and channel estimation error are the same across subcarriers. Due to the large number

\[\text{The assumption of each base station knowing individual } \{ \beta_{ilk} \} \text{ of other cells is in fact not necessary and the required information for the MMSE channel estimator can be easily obtained as in [20] which shows almost the same estimation performance as the MMSE with perfect knowledge of } \{ \beta_{ilk} \}.\]
of BS antennas, after precoding/beam-forming, the relative fluctuations of the effective channel gains across the subcarriers are substantially suppressed. Thus, frequency diversity exploitation is not much relevant and massive MIMO systems just apply per-subcarrier processing only. Under such system setup, the signal model can be decoupled for each subcarrier. The channel for each subcarrier is frequency-flat, and per-subcarrier channel statistics are the same across subcarriers. Thus, the signal model is the same on each subcarrier. Consequently, the achievable rate expression is the same across subcarriers, and we can omit the subcarrier index. $g_{ilk,m}$, $\hat{g}_{ilk,m}$, and $e_{ilk,m}$ will be denoted by $g_{ilk}$, $\hat{g}_{ilk}$, and $e_{ilk}$. In the rate analysis, we can simply consider a signal model based on a subcarrier similar to frequency-flat channel but the corresponding channel estimation MSE which is different from the frequency-flat channel needs to be used. Define $g_{ilk} \triangleq [g_{ilk}, g_{il,k+1}, \ldots, g_{ilk,M}]^T$ which represents the gains of the channels from UE $k$ in cell $l$ to all the antennas of BS $i$ and has $CN(0_M, \beta_{ilk}I_M)$ distribution. $\hat{g}_{ilk}$ and $e_{ilk}$ are defined in the same way with respective distributions $CN(0_M, \zeta_{ilk}I_M)$ and $CN(0_M, \eta_{ilk}I_M)$. The overall $M \times K$ channel matrix from the $K$ UEs of cell $l$ to BS $i$ is denoted by $G_{il}$ whose $k$th column is $g_{ilk}$. Similarly, the $M \times K$ matrix of channel estimates is denoted by $\hat{G}_{il}$ whose $k$th column is $\hat{g}_{ilk}$. Let $s_i^l$ represent a $K \times 1$ UL data vector on a subcarrier in cell $l$ where its $k$th element is UL data of UE $k$, $E[s_i^l]=0_K$ and $\mathbb{E}[s_i^l(s_i^l)^H]=I_K$. Then, the corresponding UL received signal vector on a subcarrier collected from the $M$ antennas of BS $i$ is

$$y_i^l = \sum_{k=1}^{N_{cell}} \sqrt{P_k}G_{il}s_i^l + n_i^l \quad (8)$$

where $n_i^l \sim CN(0_M, \sigma^2_{\text{noise}}I_M)$ is AWGN term with $\sigma^2_{\text{noise}} = 1$ and $P_k$ is the UL transmit data power. BS $i$ applies MRC processing on the received signal on each subcarrier as

$$r_i^l = \frac{\hat{G}_{il}^H y_i^l}{\sqrt{\lambda_i}} \quad (9)$$

where the $k$th element of $r_i^l$ denoted by $r_{ik}^l$ represents the decision variable for UE $k$ of cell $i$.

Define the DL complex data vector on a subcarrier for the $K$ UEs of cell $l$ as $s_i^d = [s_i^{d1}, \ldots, s_i^{dk}, \ldots, s_i^{dK}]^T$ with $E[s_i^d]=0_K$ and $\mathbb{E}[s_i^d(s_i^d)^H]=I_K$. For the DL transmission, on each subcarrier, BS $l$ linearly precodes the data vector $s_i^d$ through an $M \times K$ precoding matrix $W_i$ and transmits them. The $k$th column of $W_i$ is denoted by $w_{ik}$ with $\|w_{ik}\|=1$. We define $w_{ik} \triangleq \frac{\hat{g}_{ilk}}{\|\hat{g}_{ilk}\|}$ for MRT precoding. The received signal at UE $k$ in cell $i$ is given by

$$r_{ik}^d = \sqrt{P_d} \sum_{l=1}^{N_{cell}} g_{ilk}^H W_i s_i^d + n_{ik}^d \quad (10)$$

where $P_d$ is the DL data transmit power $\sigma^2_{\text{noise}} \sim CN(0, \sigma^2_{\text{noise}})$ is AWGN with $\sigma^2_{\text{noise}} = 1$. The total transmit power at each BS is $K P_d$. Similar to [19], the UE’s receiver uses the mean effective channel gain as its channel knowledge for data detection, thus, no DL training is used.

### III. Transceiver System Architectures

We present three transceiver system architectures for massive MIMO in non-contiguous bands.

#### A. Rotating FDD (RFDD)

In this architecture, the UL band and the DL band are alternated in each transmission sub-frame, i.e., the carrier frequencies of the UL and DL bands are alternated or rotated between $f_L$ and $f_H$ before the beginning of each transmission sub-frame. The UL band in the current sub-frame becomes the DL band in the next sub-frame, so the CSI of the DL for the next sub-frame can be obtained from the pilots in the current UL sub-frame. This effectively solves the CSI overhead issue of massive MIMO in the conventional FDD scheme. Fig. 1 presents the system architecture for a transceiver with RFDD where the difference from the conventional FDD architecture is the adaptation/switching in the duplex filter and the frequency synthesizer. Fig. 2 shows an example architecture for the adaptive duplex filter which swaps the bandpass filters (BPFs) for the transmit and receive signals through a switching circuit. Basically it has two modes where mode 1 receives in the higher band and transmits in the lower band while mode 2 receives in the lower band and transmits in the higher band. The switching circuit facilitates changing between the two modes at the beginning of every sub-frame as instructed by the control signal. An adaptive frequency synthesizer for RFDD is illustrated in Fig. 3 where the transmit carrier frequency $f_1(t)$ and the receive carrier frequency $f_2(t)$ are generated by a simple switching circuit and a dual output frequency synthesizer.

The transmit and receive timelines of RFDD in the two bands are shown in Fig. 4 for BS and UEs. The BS timeline serves as the reference and is designed based on the target coverage range or the UE at the cell edge (UEedge), while UE $k$ with propagation delay $\tau_k$ adjusts its transmit and receive timelines through synchronization (ranging [22]) to conform to the BS reference timeline. $T_{UL,H}$ and $T_{DL,H}$ represent the frame lengths of UL and DL in the higher band, and $T_{UL,L}$ and $T_{DL,L}$ are those in the lower band. For RFDD, we set $T_{DL,H} = T_{UL,L}$ and $T_{UL,H} = T_{DL,L}$ so that the transmissions from the two bands do not overlap and there is always some guard interval for switching the bands.

#### B. Synchronous Dual Band TDD (STDD)

STDD applies TDD for both bands synchronously. The corresponding architecture is shown in Fig. 5 where the two carrier frequencies $f_L$ and $f_H$ are generated by a dual output frequency synthesizer. The BS reference timelines are shown in Fig. 6 where we set $T_{DL,H} = T_{DL,L}$ and $T_{UL,H} = T_{UL,L}$ for synchronous operation between the two bands. The receive branch may require a dual-band band-pass filter (not shown in the figure) to filter out other signals. Due to TDD, required
CSIs for both UL and DL can be obtained from UL pilots. The peak transmit power is shared between the two bands, which somewhat limits the coverage range or data rate under the peak transmit power constraint.

C. Asynchronous Dual Band TDD (ATDD)

ATDD allows asynchronous TDD transmissions on the two bands, and an architecture for ATDD is shown in Fig. 7. The required CSIs for massive MIMO are obtained in the same way as in STDD. As the TDD transmissions of the two bands need not be synchronous, depending on the setting of the transmission timeline of the two bands, there may be overlap of one band’s transmit interval with the other band’s receive interval for a transceiver. For such a situation, a duplex filter is needed for each receiving band to suppress the transmit

Fig. 1. Transceiver system architecture for RFDD

Fig. 2. An adaptive duplex filter (DF) architecture for RFDD

Fig. 3. An adaptive frequency synthesizer for RFDD

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Fig. 4. Frame timeline for RFDD

Fig. 5. Transceiver system architecture for STDD

Fig. 6. Frame timeline for STDD
signal of the other band. There can also be time-overlap of transmissions from the two bands during which coverage range or data rate would be reduced under the peak transmit power constraint. ATDD’s transmission timeline can be set as in STDD (no duplex filter), as in RFDD (no rate sacrifice under peak power constraint), or between them depending on system requirements (e.g., UL and DL traffic loads as will be discussed in Section V). An example of the BS reference timeline without overlapped transmissions (i.e., overcoming the peak power constraint issue of synchronous architecture) is shown in Fig. 8, for which duplex filters are required.

The architecture in Fig. 7 is developed with adaptability for various scenarios in mind, thus it includes switches to bypass duplex filters when they are not needed (to avoid unnecessary insertion loss). If desired, the architecture can be simplified for a particular fixed scenario.

D. Characteristics Comparison

Here, we describe some key characteristics of the three architectures. RFDD needs an adaptive duplex filter. STDD experiences more limited coverage range or data rate than RFDD under the peak transmit power constraint. ATDD has most general transmission patterns (including those of RFDD and STDD), and may have similar coverage and data rate limitation of STDD when its transmission patterns follow STDD. ATDD offers more flexible resource adaptation than the other two, as will be discussed in a later section. The delays of feedback signaling (e.g., ARQ) in RFDD and ATDD can be smaller than those in STDD since the transmission frames of the two bands in the former architectures are consecutive across time except the guard intervals (see the figures of their frame timeline), so feedback can be sent on the earlier transmission band of the two bands.

The proposed schemes are in general not limited to non-contiguous bands and could be applied to a single contiguous band by dividing it into two adjacent bands. In this case, STDD will be the same as TDD with a single band, while RFDD and ATDD would need to handle i) self-interference at each transceiver as there will be significant leakage from the transmit chain to the receive chain and ii) intra-system multi-user interference arisen from duplexing (unable to sufficiently suppress transmit signal of other user at the receiver of a user). There exists a similar concept for a single contiguous band [23] which considers multi-point to multi-point transmission, but due to the above mentioned issues, the communication range is rather limited. Thus, although such extension of the proposed work is theoretically feasible, it will require substantial research which is beyond the scope of this paper.

IV. PERFORMANCE ANALYSIS

We first present rate expressions, and then considering system overheads including pilots, cyclic prefix and guard intervals, we present throughput for different schemes.

A. Rate Expression

There exist several rate analyses of massive MIMO systems [17]–[19], [24]–[27] in the literature but for our considered systems with multi-cell and pilot contamination, relevant works are [26] which derives both UL and DL rates and [27] which concerns with DL rate only. However, both [26] and [27] consider frequency-flat channels and single band TDD systems while our system has frequency-selective channels and two non-contiguous bands. We note that the DL rates in [27] and [26] are different due to differences in signal models (different constraints in precoding and different desired received signal term). Our DL rate expression is adopted from our work in [27] by additionally incorporating frequency-selective channels, their channel estimation MSEs which could be different for the two non-contiguous bands, and related CSI acquisition overhead. We also derive our UL rate expression but later find that it can also be obtained from [26] although the approaches in derivation are different.

As discussed in Section II, for the considered system the signal model is the same on each subcarrier. Thus, the sum rate over all the \( N \) subcarriers is \( N \) times the rate on a subcarrier. We will present the rate expressions per subcarrier in the following. The lower bound of the achievable rate expression on a subcarrier for UL transmission is given in (11), where \( p_u \) is the UL transmit data power per subcarrier, and \( \zeta_{ilk} \) is the variance of the channel estimate on the considered subcarrier which is given in (5). Although (11) can be obtained from [26], we obtain it by a different approach which could be beneficial for other research developments and hence our proof for (11) is briefly given in Appendix A.

The lower bound of the achievable rate on a subcarrier of DL is given in (12), where \( c_1(M) = \left( \frac{\Gamma(M+\frac{1}{2})}{\Gamma(M)} \right)^2 \), and \( c_2(M) = M - c_1(M) \). Note that for a large \( M \), \( c_2(M) \approx 0.25 \).
\[ R_{ik,n}^u = \log_2 \left( 1 + \frac{(M-1)\zeta_{ik}}{(M-2)\sum_{l=1, l \neq i}^{N_{\text{cell}}} \zeta_l + \sum_{l=1}^{N_{\text{cell}}} N_{l,l} \beta_{l,l} - \zeta_{ik} + \frac{1}{p_l}} \right) \] (11)

\[ R_{ik,n}^d = \log_2 \left( 1 + \frac{c_1(M)\zeta_{ik}}{(c_2(M) - 1)\zeta_{ik} + K\sum_{l=1}^{N_{\text{cell}}} \beta_{l,l} + (M-1)\sum_{l=1, l \neq i}^{N_{\text{cell}}} \zeta_l + \frac{1}{p_l}} \right) \] (12)

and \( c_1(M) \approx M \) [19]. We observe from the plot of \( c_2(M) \) versus \( M \) that \( c_2(M) \) monotonically increases and saturates to 0.25 and it is well approximated by 0.25 even at a medium value of \( M \), e.g., \( c_2(60) = 0.2495 \). Thus, the approximate expressions of \( c_1(M) \) and \( c_2(M) \) are accurate for a moderate or large \( M \). The power \( p_1 \) is the DL transmit signal power per subcarrier per user. The proof for (12) can be obtained from [27] by replacing the statistics of the channel estimate of frequency-flat channels in [27] with those of frequency-selective channels presented in (5), thus it is omitted.

**B. System Throughput**

In this section, considering DL and UL rate expressions as well as overhead due to pilots, CP and guard intervals, we will present system throughputs. The UL pilot overhead is \( KL_1 \) tones for the lower band and \( KL_2 \) tones for the upper band. The total guard interval is \( T_G = T_{\text{TRG}} + T_{\text{RTG}} \). The total time-frequency resource amount of the system for a combined UL sub-frame and DL sub-frame is \( W T_{\text{tot}} \) where \( W = \frac{1}{T}(N_1 + N_2) \) and \( T_{\text{tot}} = T_G + N_{\text{sym1}}(T+T_{\text{CP}}) + N_{\text{sym2}}(T+T_{\text{CP}}) \). We use a normalized throughput which is the achievable data rate of the considered link (UL, DL, or UL+DL) normalized by \( W T_{\text{tot}} \). Such normalization is adopted due to two reasons.

First, the sum of the normalized throughputs of UL and DL gives the overall normalized throughput of the system (UL+DL). Second, they also serve as convenient metric values in resource adaptation between UL and DL.

Let \( R_{ik,L}^u, R_{ik,L}^d, R_{ik,H}^u, \) and \( R_{ik,H}^d \) denote the UL and DL achievable rate on a subcarrier in the lower band and the higher band, respectively. Note that for frequency-flat channels the rates in the two bands are equal, but for frequency-selective channels they can be different because of different channel estimation MSEs due to different numbers of channel taps.

First, we present the normalized throughputs of RFDD and STDD. Their UL and DL normalized throughputs for UE \( k \) in cell \( i \) are given by

\[ \rho_{ik}^{u,\text{RFDD}} = \frac{(N_{\text{sym1}}N_1 - KL_1)R_{ik,L}^u + (N_{\text{sym2}}N_2 - KL_2)R_{ik,H}^u}{(N_1 + N_2)(N_{\text{sym}}) + (N_1 + N_2)(\frac{\zeta_{ik}}{p_l} + \frac{T_{\text{CP}}}{T}N_{\text{sym}})} \] (13)

\[ \rho_{ik}^{d,\text{RFDD}} = \frac{(N_{\text{sym2}}N_2)R_{ik,L}^d + (N_{\text{sym1}}N_1)R_{ik,H}^d}{(N_1 + N_2)(N_{\text{sym}}) + (N_1 + N_2)(\frac{\zeta_{ik}}{p_l} + \frac{T_{\text{CP}}}{T}N_{\text{sym}})} \] (14)

Suppose the system has a constraint of maximum transmit power per UE, i.e., \( E \). Then, the transmit power per subcarrier is given by \( \frac{E}{N_{\text{sym}}} \), where \( N_{\text{sym}} \) is the number of subcarriers simultaneously used. For example for RFDD, data and pilot transmit powers per subcarrier for UL and DL are \( q = p^u = p^d = \frac{E}{N_{\text{sym}}} \) in the lower band and \( q = p^u = p^d = \frac{E}{N_{\text{sym}}} \) in the upper band. But for STDD, due to simultaneous transmissions in the two bands, pilot and data transmit powers per subcarrier for UL and DL are given by \( q = p^u = p^d = \frac{E}{N_{\text{sym}}} = \frac{E}{N_{\text{sym}}} \).

Next, we present normalized throughputs for ATDD. We assume the total numbers of OFDM symbols per frame in the lower and higher bands are the same and equal to \( N_{\text{sym}} \). But the number of OFDM symbols for an UL sub-frame is \( N_{\text{sym},L1} \) in the lower band and \( N_{\text{sym},H1} \) in the higher band.

The corresponding numbers for a DL sub-frame are \( N_{\text{sym},L2} = N_{\text{sym}} - N_{\text{sym},L1} \) and \( N_{\text{sym},H2} = N_{\text{sym}} - N_{\text{sym},H1} \). The two quantities \( N_{\text{sym},L1} \) and \( N_{\text{sym},H1} \) can be set independently. An additional parameter to adjust for ATDD is the delay between the frame starts of the two bands. This delay influences the transmission overlap between the two bands which can affect the transmit powers of the two bands under a peak transmit power constraint.

As we mentioned before, under per-UE peak transmit power constraint, if there is time-overlap in transmissions in the two bands, the transmit power per subcarrier is reduced. Thus, depending on the parameter setting, ATDD may have two intervals with different subcarrier powers within a frame. To account for it, we introduce modified notations for the rate expressions as \( R_{ik,L}^{u,\text{ATDD}}(A,B) \), \( R_{ik,L}^{d,\text{ATDD}}(A,B) \), \( R_{ik,H}^{u,\text{ATDD}}(A,B) \) and \( R_{ik,H}^{d,\text{ATDD}}(A,B) \) where \( A \) and \( B \) represent pilot power per subcarrier and data power per subcarrier, respectively.

We consider the scenario without UL transmission overlap between the two bands. The reason is that the UL transmission overlap would reduce the UL pilot power, which will degrade channel estimation accuracy and both UL and DL rates. From Fig. 8, we obtain a general condition to avoid UL transmission overlap as \( T_{UL,H} + T_{\text{RTG}} \leq T_{DL,L} + T_{\text{TRG}} \) or \( T_{UL,L} + T_{\text{RTG}} \leq T_{DL,H} + T_{\text{TRG}} \). As can be observed in Fig. 8, the above condition guarantees that each RX frame of BS (i.e., TX frame of UE) in the higher band can be sandwiched between two RX frames of BS (i.e., TX frames...
OFDM systems typically have $2\tau_{\text{edge}} < T + T_{\text{CP}}$, thus the above condition reduces to $N_{\text{sym},H1} \leq N_{\text{sym},L2}$, i.e., the UL sub-frame length is not larger than the DL sub-frame length. The sub-frame length condition is fortunately in line with typical DL-dominated traffic loads. Next, the time offset between the two bands should be set to avoid UL transmission overlap.

Under UL transmission non-overlap case with a maximum transmit power constraint, the normalized throughputs of UL and DL of ATDD read as (18) and (19) shown at the top of next page.

Similar to the throughput computation of RFDD, STDD and ATDD, by incorporating pilot and guard interval overhead, we can obtain the normalized throughputs of UL and DL of a conventional FDD scheme with massive MIMO as

$$\rho_{ik}^{\text{u,FDD}} = \frac{(N_{\text{sym}}N_1 - ML_2 - KL_1)R_{ik,L}}{(N_1 + N_2)(N_{\text{sym}})(1 + \frac{T_{\text{CP}}}{T})}$$

$$\rho_{ik}^{\text{d,FDD}} = \frac{(N_{\text{sym}}N_2 - ML_2)R_{ik,H}}{(N_1 + N_2)(N_{\text{sym}})(1 + \frac{T_{\text{CP}}}{T})}. \quad (22)$$

The above FDD rates are substantially affected by the large pilot overhead due to the large number of BS antennas. $R_{ik,H}$ in the above is the same as that given in (12) as we assume perfect feedback of the UEs’ DL MMSE channel estimates through the UL. However, for practical FDD systems, BS’s knowledge of its DL channel would be degraded by additional errors in extracting DL channel estimates, thus its actual DL rate would be worse than (12) and the above DL throughput represents an optimistic result.

The total normalized throughput of the system (UL+DL) is given by

$$\rho_{ik} = \rho_{ik}^{\text{u}} + \rho_{ik}^{\text{d}} \quad (23)$$

where we use corresponding UL and DL throughput expressions of RFDD, STDD, ATDD or FDD to obtain the corresponding total normalized throughput.

The spectrum efficiency expression of each link (UL or DL) is given by the achievable rate normalized by the resource amount of the considered individual link. If we split the guard time overhead cost equally between UL and DL, the spectrum efficiency expressions for the UL and DL of RFDD, STDD and ATDD are given as follows:

$$\theta_{ik}^{\text{u,RFDD}} = \frac{(N_{\text{sym}}N_1 - KL_1)R_{ik,L}}{N_1(N_{\text{sym}})(1 + \frac{T_{\text{CP}}}{T} + \frac{T_{\text{G}}}{2})} + \frac{(N_{\text{sym}}N_2 - KL_2)R_{ik,H}}{N_2(N_{\text{sym}})(1 + \frac{T_{\text{CP}}}{T} + \frac{T_{\text{G}}}{2})} \quad (24)$$

$$\theta_{ik}^{\text{d,RFDD}} = \frac{(N_{\text{sym}}N_2 - KL_2)R_{ik,L}}{N_2(N_{\text{sym}})(1 + \frac{T_{\text{CP}}}{T} + \frac{T_{\text{G}}}{2})} + \frac{(N_{\text{sym}}N_1 - KL_1)R_{ik,H}}{N_1(N_{\text{sym}})(1 + \frac{T_{\text{CP}}}{T} + \frac{T_{\text{G}}}{2})} \quad (25)$$

Note that the spectrum efficiency of the overall system (UL+DL) is not necessarily the average of the UL and DL spectrum efficiencies, and a proper weighted average should be applied especially when the bandwidths of the two bands are different. On the other hand, the normalized throughput of the overall system given in (23) represents the spectrum efficiency of the overall system.

V. RESOURCE ADAPTATION BETWEEN UPLINK AND DOWNLINK

An important characteristic of a system is its capability of resource adaptation between UL and DL since UL and DL traffic loads in practice are asymmetric and time-varying as well. For systems with non-contiguous bands, FDD is conventionally applied but no resource adaptation mechanism is available in the literature. Here, we present resource adaptation mechanisms for such systems by means of RFDD, STDD, and ATDD schemes. As presented earlier, the UL and DL throughput depend on the numbers of OFDM symbols in the UL and DL subframes, i.e., $N_{\text{sym1}}$ and $N_{\text{sym2}} = N_{\text{sym}} - N_{\text{sym1}}$. Thus, the adaptation is accomplished by changing them.

For RFDD, the time-frequency resource amounts (numbers of tones across time and frequency) allocated to UL and DL, respectively denoted by $\kappa_{\text{u,RFDD}}$ and $\kappa_{\text{d,RFDD}}$, are given by

$$\kappa_{\text{u,RFDD}} = N_{\text{sym}}N_1 + N_{\text{sym2}}N_2$$

$$\kappa_{\text{d,RFDD}} = N_{\text{sym}}N_2 + N_{\text{sym1}}(N_1 - N_2) \quad (30)$$

From (30) and (31), we can see that RFDD can perform resource adaptation between UL and DL if $N_1$ and $N_2$...
are different or equivalently if the two bands have different bandwidths.

The corresponding resource amounts for STDD are
\[
\kappa^{u,\text{STDD}} = N_{\text{sym}}(N_1 + N_2)
\]
\[
\kappa^{d,\text{STDD}} = (N_{\text{sym}} - N_{\text{sym}})(N_1 + N_2)
\]
and those for ATDD are
\[
\kappa^{u,\text{ATDD}} = N_{\text{sym}, L_1}N_1 + N_{\text{sym}, H_1}N_2
\]
\[
\kappa^{d,\text{ATDD}} = (N_{\text{sym}} - N_{\text{sym}, L_1})N_1 + (N_{\text{sym}} - N_{\text{sym}, H_1})N_2.
\]

We can see from (30)-(35) that RFDD and STDD have one parameter \(N_{\text{sym}}\) to adjust for resource adaptation, but ATDD has two parameters \(N_{\text{sym}, L_1}\) and \(N_{\text{sym}, H_1}\) for adjustment. Thus, ATDD exhibits more flexible resource adaptation than the other two. However, if the system has a peak transmit power constraint, the ATDD’s adaptation should avoid some of the resource allocation settings with transmission overlap in the two bands which suffer substantial rate loss due to the transmit power sharing between simultaneous transmissions in the two bands. In the numerical result section, we will show results for the case without UL transmission overlap.

As the adjustment parameters are discrete integers within certain ranges, there are finite numbers of possible operation points in resource adaptation. For RFDD and STDD, there are \(N_{\text{sym}} - 1\) operation points. For ATDD, by checking the number of points of integer parameter pairs \((N_{\text{sym}, H_1}, N_{\text{sym}, L_2})\) satisfying the condition \(1 \leq N_{\text{sym}, H_1} \leq N_{\text{sym}, L_2} \leq N - 1\) (from the sentence following (17) and the range of the parameters), we obtain \(0.5N_{\text{sym}}(N_{\text{sym}} - 1)\) operation points. But as mentioned before, the ATDD adaptation should exclude some points with substantial rate loss incurred by simultaneous transmissions from the bands. The selection of the operation points of ATDD for resource adaptation can easily be done by numerical evaluation of the throughput expressions.

As can be seen from (30)-(35), the bandwidths of the two bands (represented by \(N_1\) and \(N_2\)) could also influence the operation points or range of the resource adaptation. Thus, for future systems, the allocation/setting of the bandwidths of the two bands should also consider the targeted range of resource adaptation so that the systems are more capable of handling traffic fluctuation and disparity between UL and DL.

The above numbers of operation points for RFDD, STDD and ATDD are reasonably sufficient for practical applications. If a finer granularity in resource adaptation is desired, we propose to change the adjustment parameter(s) over a few successive frames. For convenience of presentation, first we define operation point as ratio of the amount of DL resources and the amount of UL+DL resources, i.e. \(\frac{N_2}{N_1+N_2}\), over the considered number of successive frames. For example, for RFDD, setting two different values of \(N_{\text{sym}}\), say \((A_1, A_2)\), for every two successive frames, the operation points for adaptation are given by

\[
\kappa^{u,\text{RFDD}} = N_{\text{sym}}N_2 + \frac{A_1 + A_2}{2}(N_1 - N_2)
\]
\[
\kappa^{d,\text{RFDD}} = N_{\text{sym}}N_1 + \frac{A_1 + A_2}{2}(N_1 - N_2)
\]

instead of \(N_{\text{sym}}N_2 + A_1(N_1 - N_2)\). Then the adaptation resolution of the operation point improves from \((N_1 - N_2)/(N_{\text{sym}}(N_1 + N_2))\) to \(0.5(N_1 - N_2)/(N_{\text{sym}}(N_1 + N_2))\). Similarly, for every \(V\) successive frames, setting values of \(N_{\text{sym}}\) for the \(V\) frames to be \((A_1, A_2, \ldots, A_V)\) yields

\[
\kappa^{u,\text{RFDD}} = N_{\text{sym}}N_2 + \sum_{i=1}^{V} A_i(N_1 - N_2)
\]
\[
\kappa^{d,\text{RFDD}} = N_{\text{sym}}N_1 + \sum_{i=1}^{V} A_i(N_1 - N_2)
\]

where \(\kappa^{u,\text{RFDD}}\) and \(\kappa^{d,\text{RFDD}}\) represent the total resource amounts over \(V\) successive frames for DL and UL, respectively. Again, the operation point resolution for adaptation is improved by a factor of \(V\).

In multi-cell environments, not to cause additional intercell interferences, adjacent cells should adopt the same resource adaptation between UL and DL which can be determined by a mobile switching center which those cells are connected to. This practical constraint is the same as in TDD systems where BSs synchronize to each other and apply the same resource adaptation in terms of sub-frame configurations for TDD. Note that such resource adaptation is done based on medium-term traffic statistics (i.e., at a much larger time scale than channel adaptation), and due to similarity of traffic load patterns across cells, the UL and DL traffic load disparity is relatively similar across adjacent cells. Thus, even though individual cells would have different instantaneous UL and DL traffic load disparities, the adaptation would still offer overall improvement of system capacity. This aspect will be illustrated in the next section.
VI. NUMERICAL RESULTS AND DISCUSSION

In this section, we present the massive MIMO performance of RFDD, STDD, ATDD, and FDD in frequency-selective Rayleigh fading channels with exponential power delay profile and 3 dB decay per tap. Most of the results are obtained under the peak transmit power constraint, but we also include a set of results without such constraint for completeness. We assume pilots and data tones have the same power for each scheme. Under the constraint of per-UE peak transmit power $E$, RFDD and FDD have the transmit power per subcarrier of $E/N_1$ and $E/N_2$ in the two bands. But STDD has the transmit power per subcarrier of $E/(N_1 + N_2)$ in both of the bands. For ATDD, the transmit power per subcarrier is the same as STDD (or RFDD) during the time with (or without) transmission overlap between the two bands. When no peak transmit power constraint is imposed, all schemes use the same transmit power per subcarrier. We use a typical multicell structure with $L = 7$ cells and frequency reuse factor of 1. We assume the users in each cell are uniformly located in the disk with radius 1000 m around BS not closer than $d_0 = 100$ m to BS, and the distance of user $k$ in cell $l$ to BS $i$ is denoted by $d_{ik}$. So, $\{\beta_{ik}\}$ are independently generated by $\beta_{ik} = \psi/(d_{ik}^\nu)$ where $\nu = 3.8$, $10 \log_{10}(\psi) \sim \mathcal{N}(\sigma_{\text{shadow, dB}}^2, \sigma_{\text{noise, dB}}^2)$ with $\sigma_{\text{shadow, dB}} = 8$. The SNR is defined as $\text{SNR} = p E_{\beta_{ik}} / \sigma_{\text{noise}}^2$, where $p$ is transmit power and the expectation is performed over shadow fading and user locations. Under our simulation setting, $E_{\beta_{ik}} = -12.2 \text{dB}$ (which corresponds to $E_{\beta_{ik}} = -84.6 \text{dB}$ at 1 GHz). There are $K = 10$ UEs in each cell. The OFDM parameters are chosen to be consistent with LTE parameters, i.e., $N_{\text{sym1}} = 14$, $T = 1/15 \text{ (ms)}$, $T_{\text{CP}} = 4.7 \mu s$, and $T_G = 6.7 \mu s$ (based on the round-trip delay of a cell with radius 1000 m). Results are averaged over 10,000 realizations of random large scale fading coefficients.

A. Performance Analysis

Here, we present the overall normalized system throughput (spectrum efficiency) given in (23) under the system setting of $N_1 = N_2 = 128$ subcarriers, $L_1 = L_2 = 8$ channel taps, the same resource amount between UL and DL, $N_{\text{sym1}} = N_{\text{sym2}} = 7$ symbols, and $M = 60$ BS antennas. Under the above setting, ATDD can have two settings ATDD1 and ATDD2 where their transmission timelines are the same as that of RFDD and STDD, respectively.\(^7\)

Fig. 9 shows the spectrum efficiency versus the per-UE peak transmit power $E$ (under the setting of $E_{\beta_{ik}} = -12.2 \text{dB}, \sigma_{\text{noise}}^2 = 1$) for RFDD, STDD, ATDD, and FDD. RFDD and ATDD1 clearly outperform all the other schemes. STDD and ATDD2 have 3 dB SNR loss due to the peak power constraint of $E$ if compared to RFDD, ATDD1 and FDD. Thus, the performance of STDD and ATDD2 is worse than FDD at low $E$ but substantially better than FDD at high $E$ where FDD’s higher CSI overhead cost outweighs SNR loss of STDD and ATDD2. The performance of STDD and ATDD2 approach that of RFDD and ATDD1 at high $E$ where the performance gets saturated by pilot contamination and CSI errors.

Next, we show the spectrum efficiency versus per-UE SNR (i.e., without the peak transmit power constraint) in Fig. 10. In this case, RFDD, STDD, ATDD1, and ATDD2 have the same performance and they substantially outperform FDD due to FDD’s much higher CSI overhead.

The effects of the number of BS antennas ($M$) on RFDD, STDD, ATDD, and FDD are illustrated in Fig. 11 under a fixed $E = 30 \text{ dB (}E_{\beta_{ik}} = -12.2 \text{dB}, \sigma_{\text{noise}}^2 = 1\)$. A larger $M$ yields better spectrum efficiency for RFDD, STDD, and ATDD. The same is true for FDD up to $M = 70$ (insignificantly increase from $M = 50$ to $M = 70$) but beyond which FDD’s performance degrades with the increase of $M$ due to

\(^7\)With the same total bandwidth, the performance of TDD in a system with a single band is the same as that of STDD in two non-contiguous bands.
the increased CSI overhead toll. Also recall that the presented performance of FDD is optimistic due to the assumption of perfect CSI feedback through UL.

B. Resource Adaptation between UL and DL

In this subsection, we present the resource adaptation capability of RFDD, STDD, and ATDD. Note that FDD does not have such capability. Furthermore, the effects of the frame length ($N_{sym}$) and the bandwidths of the two bands ($N_1$ and $N_2$) on the resource adaptation capability are also illustrated. We assume $E = 30$ dB ($\mathbb{E}\{\beta_{i,k}\} = -12.2$ dB, $\sigma_{noise}^2 = 1$), and $M = 60$ BS antennas. The adaptation is done through $N_{sym,1}$ for RFDD and STDD, and through $N_{sym,L1}$ and $N_{sym,H1}$ for ATDD. As mentioned before, the frame timeline of ATDD is set to avoid UL transmission overlap.

The (un-normalized) throughputs per sub-frame (i.e., the numerators of (13)-(16), (18), and (19)) versus the allowable normalized DL resource amount (i.e., $\frac{B_{UL}}{N_{sym}}$) are shown in Fig. 12, Fig. 13, and Fig. 14 for RFDD, STDD, and ATDD, respectively. The left sub-plots show effects of changing the number of subcarriers with $N_{sym}=14$. The right sub-plots presents effects of changing $N_{sym}$ with $N_1 = 128$, $N_2 = 512$ subcarriers and $L_1 = 8$, $L_2 = 32$ channel taps. We observe the following:

- As expected, the operation points and ranges for resource adaptation differ across the three schemes, and ATDD offers a lot more operation points for resource adaptation than RFDD and STDD as it has more degrees of freedom in adaptation.
- Different bandwidth ratios ($N_1/N_2$) do not affect the operation points and range of adaptation of STDD but they do influence those of RFDD and ATDD. The smaller ratio of the smaller to the larger bandwidth offers a wider resource adaptation range for RFDD and ATDD which could be exploited in spectrum allocation of future systems.
- A larger number of OFDM symbols per frame does not affect the adaptation range but it can improve the resolution of resource adaptation for all the three schemes.
- ATDD’s performance curves show slight dispersion (more obvious for UL). This is caused by two factors: 1) the normalized rates (especially for UL) can be quite different for the two bands due to their different channel estimation MSEs and 2) the adaptation can be done by changing the UL subframe length of the lower band ($N_{sym,L1}$), of the higher band ($N_{sym,H1}$) or both. For the two bands with different bandwidths, the gradual increases of the resource amount ratio between DL and DL+UL (x-axis in the figure) correspond to different combinations of $N_{sym,L1}$ and $N_{sym,H1}$ which together with the different normalized rates of the two bands result in slight dispersion of the rate curves.

Fig. 15 compares the performance and adaptability of RFDD, STDD and ATDD. It can be seen that ATDD provides more adaption points but STDD has the widest adaptation range. The average throughputs of ATDD and RFDD are higher than that of STDD.

Fig. 16 illustrates further enhancement of resource adaptation granularity by means of changing the value of $N_{sym}$ within every 10 consecutive frames under the same frame length $N_{sym}$. The approach can be applied to all RFDD, STDD, and ATDD schemes, but we use RFDD with $N_1 = 128$ and $N_2 = 512$ as an example. By comparing Fig. 16 and the corresponding case in Fig. 12, we can observe that this approach yields a very fine granularity of resource adaptation.

Next we present the performance of resource adaptation in multicell environments. We assume the same total resource amount for RFDD and FDD, with $N_{sym} = 14$ and $N_1 + N_2 =$...
1000. We consider three traffic scenarios where the average DL resource amount is 2, 5, and 10 times the average UL resource amount. The actual requested resource amounts for UL and DL for different cells are modeled to be independent and identically distributed as $R^* = (1 + 0.1L) R^*$ where * represents either UL or DL, and $R_{DL}/R_{UL} = 2, 5,$ or 10 depending on the scenario. We assume users in each cell have the same ratio of UL and DL requested resource amounts (as we are only concerned with system performance, not individual user performance), and the transmit power on each subcarrier is set to 30dB with $E[\beta_{nk}^2] = \sigma_n^2$. As FDD cannot adapt resources between UL and DL, to show performances of different FDD spectrum allocation settings, we consider 3 FDD settings where $N_1$ and $N_2$ for FDD are designed to match to $R_{DL}/R_{UL}$ of the three scenarios. For RFDD, we use only one setting of $N_1 = 23$ and $N_2 = 977$, (obtained from (30) and (31)) with $\kappa_{UL}^{RFDD}/\kappa_{DL}^{RFDD} = 10$ and $N_{sym1} = 13$) and apply three settings of sub-frame length adaptation based on $R_{DL}/R_{UL}$. The RFDD adaptation is the same across all cells and is managed by the mobile switching center.

Fig. 17 compares the throughput performance of three FDD schemes (3 settings) and RFDD in multicell environments with the three traffic scenarios mentioned above. The results are obtained from 10,000 realizations of random large scale fading.
coefficients and 1000 realizations of DL and UL resource requests. If we neglect CSI overhead cost, the FDD schemes perform well only when their spectrum allocation matches to the traffic scenario, but RFDD performs approximately the same as the best of the 3 FDD schemes at each traffic scenario. This illustrates the advantage of the proposed resource adaptation in multicell environments. When CSI overhead cost is included, RFDD’s performance is slightly reduced but FDD schemes suffer substantial performance loss.

VII. CONCLUSIONS

We have presented three schemes (RFDD, STDD, and ATDD) for overhead-efficient massive MIMO systems in non-contiguous bands with frequency-selective channels. Incorporating CSI estimation errors, pilot contamination, CSI overhead and guard interval cost, we have derived their achievable rate or throughput expressions for both UL and DL. Compared with optimistic performance of the conventional FDD (obtained by assuming perfect feedback of CSI estimate through UL), the proposed schemes show several advantages. Under the peak transmit power constraint, STDD could have smaller throughput than FDD for low to moderate values of the transmit peak power and the number of BS antennas but it has substantially higher throughput than FDD as the peak transmit power or the number of BS antennas increase. Without the peak transmit power constraint, STDD always outperforms FDD. Both RFDD and ATDD with a proper setting substantially outperform FDD for all scenarios. Next, we have also developed resource adaptation schemes between UL and DL and presented their operating points and ranges of resource adaptation with reference to adaptation parameters, bandwidths and frame length. We have also proposed a scheme for granularity enhancement of resource adaptation by changing the numbers of OFDM symbols per frame devoted to UL and DL over several successive frames. The conventional FDD is incapable of resource adaptation but practical systems show highly asymmetric and dynamic traffics. The proposed flexible resource adaptation for the proposed schemes provides a practically appealing additional advantage over the FDD. Overall, the proposed strategies offer promising solutions for overhead-efficient massive MIMO systems in non-contiguous bands.

APPENDIX A

This Appendix presents the proof for (11). The following lemma shows how some interference terms are correlated. This characteristic is critical in computing interference power in the proof.

Lemma 1. The estimated channel between UE $k$ in cell $l$ and BS $i$ on a subcarrier is related to the channel estimate between UE $k$ in cell $i$ and BS $i$ as

$$\hat{g}_{ilk} = \frac{\beta_{ilk}}{\beta_{iik}} g_{ilk}. \quad (38)$$

Proof. From (4) and definitions of $\hat{g}_{ilk}$ and $g_{ilk}$ in Section II, we arrive at (38). $\square$

From (9), the UL received signal of UE $k$ in cell $i$ can be given as

$$r_{ik}^u = \sqrt{p_i} |g_{ilk}|^2 u_{ik} + I_{ik}, \quad (39)$$

where $I_{ik} \triangleq \sum_{j=1}^{5} I_{ik,j}$ with $I_{ik,1} \triangleq \sqrt{p_i} g_{iik}^H g_{iik} s_{ij}^u$, $I_{ik,2} \triangleq \sum_{j=1, j \neq k}^{5} \sqrt{p_i} g_{ijkl}^H e_{ij}^u$, $I_{ik,3} \triangleq \sum_{l=1, l \neq i}^{K} \sum_{j=1, j \neq k}^{K} \sqrt{p_i} g_{ilk}^H g_{ikl} s_{ij}^u$, $I_{ik,4} \triangleq \sum_{j=1, j \neq k}^{5} \sqrt{p_i} g_{ijkl}^H e_{ij}^u$, and $I_{ik,5} \triangleq \sqrt{p_i} g_{iik} n_i$. Based on [28, Appendix I], the lower bound on achievable rate is given by

$$\hat{R}_{ik}^u = \mathbb{E} \{ \log_2 (1 + \frac{p_i |g_{iik}|^2}{\mathbb{E} |I_{ik}^u| g_{iik}}) \} \quad (40)$$

where the outer expectation is over $g_{iik}$, the expectation in the denominator is over all other random variables, and $\mathbb{E} |I_{ik}^u| g_{iik} = \sum_{i=1}^{5} \mathbb{E} |I_{ik,j}| g_{iik}$. From Section II, we have $g_{iik} \sim \mathcal{CN}(0, \beta_{iik} I_{M})$, $g_{ijkl} \sim \mathcal{CN}(0, \beta_{ijkl} I_{M})$, and $e_{ij} \sim \mathcal{CN}(0, \beta_{ijkl} I_{M})$. Conditioned on $g_{iik}$ and due to the independence of $g_{iik}$ from $e_{ijkl}$, $g_{iik}$ and $g_{ijkl}$, (9) can be written as

$$\hat{R}_{ik}^u = \mathbb{E} \{ \log_2 (1 + \frac{p_i |g_{iik}|^2}{\mathbb{E} |I_{ik}^u| g_{iik}}) \}. \quad (41)$$

Applying the above results, the power of five interference terms defined below (39) can be obtained as $\mathbb{E} |I_{ik,1}^u| g_{iik} = p_i |g_{iik}|^2 \eta_{iik}$, $\mathbb{E} |I_{ik,2}^u| g_{iik} = p_i \sum_{j=1, j \neq k}^{5} |g_{ijk}|^2 \beta_{iij}$, $\mathbb{E} |I_{ik,3}^u| g_{iik} = p_i \sum_{l=1, l \neq i}^{K} \sum_{j=1, j \neq k}^{K} |g_{ilk}|^2 \beta_{ijk}$, $\mathbb{E} |I_{ik,4}^u| g_{iik} = p_i \sum_{j=1, j \neq k}^{5} |g_{ijkl}|^2 \beta_{ijkl}$, $\mathbb{E} |I_{ik,5}^u| g_{iik} = p_i \sum_{j=1, j \neq k}^{5} |g_{ijkl}|^2 \beta_{ijkl}$, and finally, $\hat{R}_{ik}^u = \mathbb{E} \{ \log_2 (1 + \frac{p_i |g_{iik}|^2}{\mathbb{E} |I_{ik}^u| g_{iik}}) \}$. Then, we can write

$$\hat{R}_{ik}^u = \mathbb{E} \{ \log_2 (1 + \frac{1}{X}) \}$$

where $X$ is given by

$$X = \frac{1}{\mathbb{E} |I_{ik}^u||g_{iik}|^2} (\frac{\eta_{iik}}{\beta_{iik}} + \sum_{j=1, j \neq k}^{K} \beta_{iij} + \sum_{l=1, l \neq i}^{K} \sum_{j=1, j \neq k}^{K} \beta_{ijkl} + \sum_{l=1, l \neq i}^{K} \eta_{iik} + \frac{1}{\beta_{iij}}) + \sum_{l=1, l \neq i}^{K} \sum_{j=1, j \neq k}^{K} \frac{1}{\beta_{ijkl}}. \quad (42)$$

Next, with the use of Jensen’s inequality for a convex function, i.e., $\mathbb{E} \{ \log_2 (1 + \frac{1}{X}) \} \geq \log_2 (1 + \frac{1}{\mathbb{E} X})$, and the fact
that
\[
\mathbb{E}\left\{ \frac{1}{\| \mathbf{R} \|^2} \right\} = \frac{1}{(M-1) \lambda_{\text{min}}},
\]
and after some simplification using
\[
\zeta_{\text{ilk}} + \nu_{\text{ilk}} = \beta_{\text{ilk}},
\]
we arrive at the lower bound on achievable rate
\[
R_{\text{ilk}}
\]
as given in (11).

REFERENCES


