High Performance Receiver of Maximum LLR's for DFT-S-OFDMA and DFT-S-OFDM-SDMA

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Abstract- *For the 3GPP LTE uplink transmissions, the DFT spread OFDM technique has been adopted as the air interface in order to reduce the peak-to-average-power ratio (PAPR). In this scheme, each data symbol is spread over many tones by a Discrete Fourier Transform (DFT) operation at the transmitter before being sent to the Orthogonal Frequency Division Multiplexing (OFDM) modulator. Moreover, more than one user can be scheduled over the same frequency and time resource block (RB) via space-division multiple-access (SDMA). The conventional receiver technique for such DFT-spread OFDM systems involves tone-by-tone single tap equalization followed by an inverse DFT operation. In this paper, we propose a more powerful receiver technique for DFT-spread OFDM systems that consists of an efficient linear pre-filter and a two-symbol soft output demodulator. The proposed method can be applied to both single-user per RB (DFT-S-OFDMA) and multiple users per RB (DFT-S-OFDMSDMA) systems and it offers significant performance gains over the conventional method, especially in the high-rate regime, with little attendant increase in computational complexity.*

I. INTRODUCTION

Discrete Fourier Transform-Spread-Orthogonal Frequency Division Multiple Access (DFT-SOFDMA) has emerged as the preferred uplink air interface for the next generation cellular systems such as those envisaged by the 3GPP LTE . DFT-S-OFDMA is essentially a modified form of OFDMA where scheduled users transmit their data simultaneously on non-overlapping (orthogonal) sets of sub carriers (frequencies). The key difference from OFDMA is that each user spreads its coded and modulated information bits using a DFT matrix and the spread (preceoded) symbols are then mapped to its allocated sub carriers. The main advantage of this multiple access technique is that it results in considerably lower PAPR at each transmitter (user) compared to the classical OFDMA technique . Upcoming cellular systems will employ antenna arrays at the base station (a.k.a Node-B) and possibly at the user equipment (UE) as well. A promising scheme, also adopted in the 3GPP LTE that is enabled due to the use of antenna arrays at the base station is the SDMA scheme which is sometimes referred to as the virtual Multi Input Multi-Output (MIMO) scheme. In SDMA multiple single-antenna users are scheduled over the same frequency and time resource blocks in order to boost the system throughput. Since different users are geographically separated, their channel responses seen at the base-station antenna array will be independent and hence capable of supporting high rate communications. Henceforth, the DFT-S-OFDM based uplink employing SDMA will be referred to as the DFT-S-OFDM SDMA uplink.

In DFT-S-OFDM systems, which encompass both DFT-SOFDMA and DFT-S-OFDM-SDMA, as a consequence of the DFT spreading system at the transmitter, the signal arrives at the base-station with substantial inter symbol interference and the received sufficient statistics can be modeled as a consequence of the DFT spreading operation as the channel output of a large MIMO system. The conventional receiver technique involves tone-by-tone single-tap equalization followed by an inverse DFT operation . Such a simple receiver suffices for the DFT-S-OFDMA case in the low-rate regime when there is enough receive diversity and where the available frequency diversity can be garnered by the underlying outer code. However, it results in a degraded performance at higher rates as well as with SDMA. Unfortunately, the large dimension of the equivalent MIMO model in DFT-S OFDM systems precludes us from leveraging the sphere decoder which has an exponential complexity in the problem dimension . Furthermore, the stringent complexity constraints in practical systems also rule out the near-optimal MIMO receivers developed for the narrowband channels, see for instance . Other promising equalizers for the DFT-S-OFDM systems are the decision feedback equalizers (DFEs). The two most promising DFEs are the hybrid DFE proposed in , where the feed forward filter is realized in the frequency domain and the feedback filter is realized in the time domain, and the iterative block DFE with soft decision feedback proposed in , where even the cancelation is performed in the frequency domain. However, even the DFE whose iterative process does not include decoding the outer code, is substantially more complex and has higher latency especially in the SDMA case, than the conventional receiver and hence is not considered in this paper. We remark however, that the iterative soft cancelation can be readily added to the receivers considered here in order to obtain further performance improvement salbeit with higher complexity and latency.

In this paper we consider receiver design for such DFTS- OFDM systems. To realize our goal of obtaining efficient receivers, we first propose a very efficient implementation of a soft-output demodulator for the narrowband MIMO model with two input symbols. We then design a novel receiver, henceforth referred to as the group soft demodulator, which can be used in DFT-S-OFDM systems with and without SDMA. This receiver groups the input symbols into multiple pairs, and demodulates each pair using the efficient soft output demodulator, after suppressing the interference from other pairs via linear Minimum Mean Squared Error (MMSE) filtering. Particular care is taken in designing the interference suppression filters in order to minimize the computational complexity.

II. A TWO-SYMBOL SOFT-OUTPUT DEMODULATOR

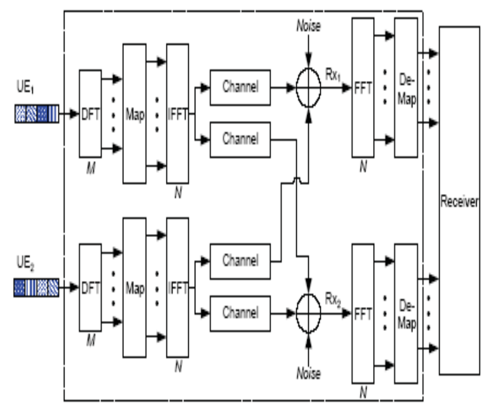
In this section, we derive an efficient two-symbol soft output demodulator, which is one of the key ingredients of the receiver techniques discussed in the next section. We consider the following 2-input nR-output narrowband signal model where $\boldsymbol{y}$ CnR is the received signal vector, $\boldsymbol{H} = [\boldsymbol{h}1, \boldsymbol{h}2]$ $CnR\times 2$ is the $nR\times 2$ channel matrix ($nR \geq 2$); $\boldsymbol{n}$ $Nc(\mathbf{0}, \boldsymbol{I})$ is the noise vector with independent identically distributed (i.i.d) zero mean, unit variance complex proper Gaussian elements; $\boldsymbol{s} = [s1, s2]T$ is the symbol vector with symbols from two quadrature amplitude modulation (QAM) constellations$S1$, $S2$ of sizes $|S1|$ and $|S2|$, respectively.

where $A2$ denotes the set of all the $\_|S2|$ symbols in the corresponding pulse amplitude modulation (PAM) constellation.2 Note that the two minimizations in can be done in parallel using simple slicing (rounding) operations with O complexity each. Moreover using the fact that $l11$ is positive along with the symmetry of the QAM constellation$S1$, we have that the sets $\{l11sR1,j, j = 1, . . . ,\_|S1|\}$ and $\{l11sI 1,j, j = 1, . . . ,\_|S1|\}$ are identical, a fact that can be exploited to efficiently compute the first term in (6) after expanding it as $|z1-l11s1,j |2 = |zR1 -l11sR1,j |2+|zI1 -l11sI1,j |2$.Further, using $\{Q(s1,j)\}$ the maximum-likelihood (ML) decision, denoted by $[s1,\hat{}j , s2,\hat{}k]$, can be readily determined as Similarly for demodulating symbol $s2$, we obtain another modified QR decomposition $\boldsymbol{H} = \_\boldsymbol{h}1\_\boldsymbol{VR}$, with $\boldsymbol{V} = [\boldsymbol{v}1, \boldsymbol{v}2]$ being a semi-unitary matrix and $\boldsymbol{R}$ being upper triangular with positive diagonal elements.

The QAM symbol $s1$ ($s2$), corresponds to log2$|S1|$ (log2$|S2|$) bits, respectively, i.e., it is represented by a log2$|S1|$ (log2$|S2|$) length bit-vector. The $|S1| + |S2| -$ 1metrics$\{Q(s1,j), Q(s2,k)\}$ defined above are sufficient to determine the max-log log likelihood ratio (LLR) for each coded bit. To see this, consider the log2$|S1|$ bits associated with $s1$.Then letting $\lambda 1,\_$ denote the max-log LLR for the _-th bit $b1,\_$ associated with $s1$ and assuming equal *a priori* bit probabilities we have which can be simplified to (15).

It can be verified that one of the two terms in the RHS of for each _ is equal to $\hat{}\lambda$, which is the metric associated with the ML decision $[s1,\hat{}j , s2,\hat{}k]$and was computed in .Indeed, if the lth bit associated with $s1,\hat{}j$ is 1 then is equal to if the lth bit associated with $s2,\hat{}k$ is 0.

It is easily seen that when the two QAM constellations are identical, i.e., $S1 = S2 = S$, the complexity of the proposed method for computing the max-log LLR for each coded bit is $O(|S|/$ log2$|S|)$ instead of $O(|S|2/$ log2$|S|)$ assign the conventional method.



The LLR computations remain unchanged. Moreover, the complexity involved in computing the metrics can also be reduced by exploiting the structure of the underlying constellations. For instance, when $S2$ is a unit average energy Phase Shift Keying (PSK) constellation $\{\exp(i\theta j)\}, \theta j$ $\{0, 2\pi|S2|, \cdot \cdot \cdot , 2\pi(|S2|-1)|S2| \}$, we can determine $\{Q(s1,j)\}$ efficiently by first expanding $Q(s1,j)$as in (5). Then, we can obtain the polar representation of $q1j$as $q1j = a1j$ exp$(ib1j)$, with $a1j > 0$ and $b1j$ $[0, 2\pi)$. The minimizing $s2$ in (5) can be determined with $O(1)$ complexity as exp$(i2\pi(_\beta 1j_ + 1)/|S2|)$, where _._ denotes the floor operation and $\beta 1j = |S2|b1j2\pi - 12$. Finally, in some cases only the LLRs for the coded bits associated with one of the two symbols (say symbol 1) need to be generated. In such a case, we can implement a *partial* soft-output demodulator by first computing $\{Q(s1,j)\}$ and $\hat{}\lambda$ respectively and then computing $\{\lambda 1,\_\}$

III. RECEIVER DESIGN FOR DFT-S-OFDM SYSTEMS

In this section, we consider receiver designs for the DFT spread OFDM uplink systems, where the base station is equipped with nR receive antennas and each user has a single transmit antenna and is assigned M sub carriers out of the N total available sub carriers. A schematic of a DFT-S-OFDM uplink with two users is shown in Fig. 1. In the DFT-SOFDMA case the two users are assigned non-overlapping sets of sub carriers, whereas in the DFT-S-OFDM-SDMA case they are assigned overlapping sets of subcarriers.3

A. DFT-S-OFDMA Receivers

Without loss of generality, we assume that the user of interest is assigned tones 1 through M. Upon multiplying the sampled received observations at each receive antenna by an N-point DFT matrix, the (frequency domain) received signal vector on the mth tone can be modeled.

Effective(frequency domain) channel response vector for the m th tone, the DFT-spread input symbol is xm and $\boldsymbol{n}m$ $Nc(\mathbf{0}, \boldsymbol{I})$ is the noise vector. Moreover, we denote the M transmitted (unit average energy) QAM symbols by $\boldsymbol{s} = [s1, s2, \cdots, sM]T$ .Let $\boldsymbol{F}$ be the $M \times M$ DFT matrix with its $(k, n)th$ element given by $Fk,n = (1/\sqrt{M})e{-}j2\pi(k{-}1)(n{-}1)/M$. Collecting the received signal vectors for all M tones, we have $\mathbb{C}nRM\times M$ is a block-diagonal matrix with $\boldsymbol{h}m$ as its m-th diagonal block. Note that $\boldsymbol{n}$ $Nc(\mathbf{0}, \boldsymbol{I})$ and since each element of $\boldsymbol{s}$ has zero-mean and unit variance, all the elements of $\boldsymbol{x}$ have zero mean. Further, assuming that the elements of $\boldsymbol{s}$ are mutually uncorrelated4 and using the fact that $\boldsymbol{F}$ is unitary, we have that all the elements of $\boldsymbol{x}$ are mutually uncorrelated and also have unit variance. In what follows we discuss two receivers for obtaining the soft estimates, i.e., LLRs of the coded bits associated with each symbol sm

B. GROUP SOFT DEMODULATOR*:*

We next propose a new receiver technique that makes use of the efficient two-symbol soft output demodulator developed in Section II, henceforth referred to as the two-symbol max-log LLR demodulator. The basic idea is to divide symbol vector $\boldsymbol{s}$ into M 2 groups, each with two symbols. When demodulating a particular two-symbol group, as interference and apply a properly designed linear filter to suppress them. In particular, assume that the M symbols repaired as $\boldsymbol{s}k\_ = [s2k{-}1, s2k]T$, $k = 1, \ldots, M$ 2 . A key benefit of this pairing will be highlighted in the sequel. We can then write (22) as where $\boldsymbol{G}k$ $\mathbb{C}nRM\times 2$ contains the $(2k - 1)$-th and $2k$-th columns of the matrix $\boldsymbol{G}$, given by Moreover $\boldsymbol{G}k$ $\mathbb{C}nRM\times(M{-}2)$ consists of the remaining columns of $\boldsymbol{G}$ after removing $\boldsymbol{G}k$, and $\boldsymbol{s}k$ contains the remaining symbols in $\boldsymbol{s}$ after removing $\boldsymbol{s}k$.

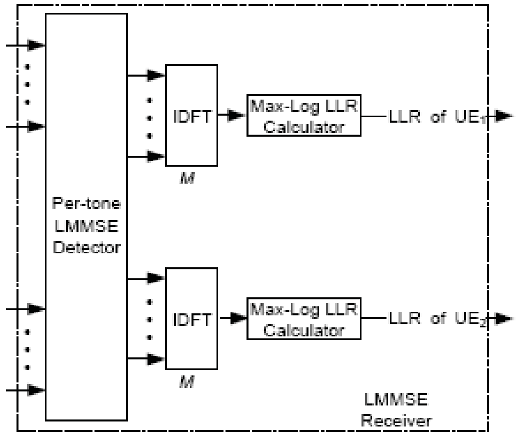
We first apply a linear filter to the vector of received observations $\boldsymbol{y}$ to suppress the interference term $\boldsymbol{G}k\boldsymbol{s}k$. In the Appendix, we show that the group linear MMSE filter (referred to simply as the prefilter hereafter) for the k-th pair, which is designed to whiten the noise plus interference followed by maximum ratio combining, can also be written as

where $\boldsymbol{R}{-}1 = [\mathrm{E}\{\boldsymbol{y}\boldsymbol{y}\dagger\}]{-}1$ $\mathbb{C}nRM\times nRM$ is a *block diagonal matrix* with $\boldsymbol{I} - \boldsymbol{h}m\boldsymbol{h}\dagger m/(1 + \_\boldsymbol{h}m\_2)$ as its m-th diagonal block. Consequently, $\boldsymbol{R}{-}1$ can be easily computed and can be used to compute all the $M/2$ prefilters. $\boldsymbol{Q}$ $\mathbb{C}2\times 2$ in (31) is given by the following Cholesky decomposition .with $\boldsymbol{z}k$ $\mathbb{C}2\times 1$, $\boldsymbol{T}$ $\mathbb{C}2\times 2$, and $\mathrm{E}\{\tilde{}\boldsymbol{n}\tilde{}\boldsymbol{n}\dagger\} = \boldsymbol{I}$. Note that as a consequence of the chosen pairing, $\boldsymbol{G}\dagger k\boldsymbol{R}{-}1\boldsymbol{G}k$ is independent of k and therefore $\boldsymbol{Q}$ and $\boldsymbol{T}$ in are also independent of k. Hence we need to calculate them only once for all the M 2 pairs. Note that in order to obtain we have assumed that $\boldsymbol{Q}{-}1$ exists and then to be able to use the efficient two-symbol max-log LLR demodulator, we need that $\boldsymbol{T}$ be non-singular, so that its QR decomposition (in which the diagonal elements of the triangular matrix are strictly positive) exists. In this regard, we note that the spectral norm (maximum eigenvalue) of $\boldsymbol{G}\dagger k\boldsymbol{R}{-}1\boldsymbol{G}k$ is strictly less than 1. Thus, $\boldsymbol{I} - \boldsymbol{G}\dagger k\boldsymbol{R}{-}1\boldsymbol{G}k$ is positive definite, using which we can conclude that $\boldsymbol{Q}{-}1$ exists and $\boldsymbol{T}$ is non-singular if and only if the 2×2 matrix $\boldsymbol{G}\dagger k\boldsymbol{R}{-}1\boldsymbol{G}k$ has full rank. Next, using and the fact that any two vectors in the set $\{\boldsymbol{e}m\}Mm = 1$ are *linearly independent*, we have that the rank of $\boldsymbol{G}\dagger k\boldsymbol{R}{-}1\boldsymbol{G}k$ is 2 if *at least two vectors out of* $\{\boldsymbol{h}m\}Mm = 1$ *are non-zero*. Such a mild condition will clearly be satisfied in practice. The complexity of computing the LLRs can be substantially reduced by observing the following. we can expand $\boldsymbol{G}\dagger k\boldsymbol{R}{-}1\boldsymbol{y}$ as

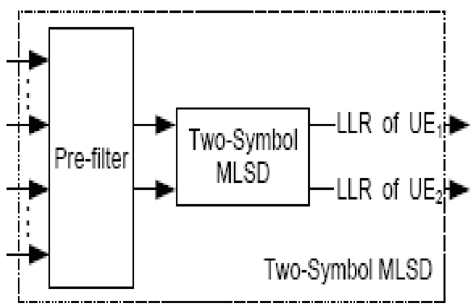
C. DFT-S-OFDM-SDMA RECEIVERS

We now assume that there are two users and for the m-th tone, the effective (frequency domain) channel response vector corresponding the k-th user is $\boldsymbol{h}(k)$ $\mathbb{C}nR$ and the DFT-spread symbol is $x(k)$ m , $k = 1, 2$. The received signal vector on the m-th tone is now given by

1) Conventional Receiver*:* The linear MMSE estimate of$x(k)$ m , $k = 1, 2$, based on ym



2) Group Soft Demodulator*:* In a similar way as in Section III-A2, we now group the $2M$ transmitted symbols into M pairs, where the m-th pair is given by $\boldsymbol{s}m$ _= [$s(1)$ m , $s(2)$ m]T .



$\boldsymbol{G}m$ CnRM×2(M−1) consists of the remaining columns of $\boldsymbol{G}$ after removing $\boldsymbol{G}m$, and $\boldsymbol{s}m$ comprises of the remaining elements of $\boldsymbol{s}$ after removing $\boldsymbol{s}m$. written as its m-th diagonal block and $\boldsymbol{Q}$ C2×2 with with $\boldsymbol{z}m$ C2×1, $\boldsymbol{T}$ ∈C2×2, and E{˜$\boldsymbol{n}$˜$\boldsymbol{n}$†} = $\boldsymbol{I}$. Note that as a result of the chosen pairing, $\boldsymbol{G}$†$m\boldsymbol{R}$−1$\boldsymbol{G}m$ is independent of m and hence $\boldsymbol{Q}$ and $\boldsymbol{T}$ are independent of m too,. Therefore we need to calculate them only once for all the Mpairs. Furthermore, it can be verified that $\boldsymbol{Q}$−1 exists and $\boldsymbol{T}$is non-singular if the 2×2 matrix $\boldsymbol{G}$†$m\boldsymbol{R}$−1$\boldsymbol{G}m$ has full rank. Next, it can be shown that the rank of $\boldsymbol{G}$†$m\boldsymbol{R}$−1$\boldsymbol{G}m$ is equal to that of the n RM × 2 matrix [$\boldsymbol{H}T$1, · · · ,$\boldsymbol{H}TM$]T. Thus, we need a mild assumption that the two columns of the matrix [$\boldsymbol{H}T$1, · · · ,$\boldsymbol{H}TM$]T be linearly independent. We can efficiently compute the LLRs by expanding$\boldsymbol{G}$†$m\boldsymbol{R}$−1$\boldsymbol{y}$ where _$\boldsymbol{s}m$ = [_s(1)m , _s(2)m]T and_$\boldsymbol{s}$(k) = [_s(k)1 , · · · , _s(k)M T, k = 1, 2 are the linear MMSE estimates Then, using the matrix $\boldsymbol{T}$ in we can compute ˜$\boldsymbol{U}$ and ˜ $\boldsymbol{V}$ and multiply each _$\boldsymbol{s}m$,where m = 1, . . .,M, with the matrices ˜$\boldsymbol{U}$ †$\boldsymbol{Q}$−1 and ˜ $\boldsymbol{V}$ †$\boldsymbol{Q}$−1to obtain soft statistics similar to the ones in and. We can then compute the LLRs for the bits associated with the symbols [s(1) m , s(2)m]T , A block diagram for the proposed receiver is shown in Fig. 3.A few comments are now in order about another candidate receiver for coded multi-user systems, namely the successive interference cancelation (SIC) receiver . We recall that the SIC receiver can improve performance whenever *post-decoding feedback* is applicable, i.e., when a decoded codeword is re-constructed and then subtracted, prior to the decoding of remaining code words. However, if we employ predecoding feedback, where detected symbols are subtracted, error propagation can degrade performance. In the scenario of interest in this paper, which is each frame each scheduled user transmits its information using a single codeword. Thus, in the DFT-S-OFDMA case post-decoding feedback is not possible. On the other hand, in the DFT-S-OFDM-SDMA case post-decoding feedback is indeed possible. To realize an *improved SIC* receiver in a DFT-SOFDM-SDMA uplink with two users, we can first obtain the M "whitened" soft statistics. Then using the partial max-log LLR demodulator as described towards the end of Section II, we can generate the LLRs only for coded bits of user 1. After decoding the codeword of user 1, we can subtract the re-constructed codeword. Assuming perfect feedback, we then have a single user DFT-S-OFDMA system in which we have to decode the codeword of user 2 and we can employ the conventional DFT-S-OFDMA receiver (or the group soft demodulator for the DFT-S-OFDMA uplink) to decode the codeword of user 2.

3) Receiver Algorithms for More Than Two UEs: The receivers described in the previous sections can be extended in a simple manner to a DFT-S-OFDM-SDMA system with more than two users. Suppose $K = 2a + 1$ is the number of users. Then we can use the two-user group soft demodulator to decode each one of the a user pairs and either the conventional receiver or the single-user group soft demodulator to decode the remaining user. In particular, we can assume that the users have been divided into a + 1 groups with the first a groups containing two users each. Consider the decoding of the first group and without loss of generality, let it comprise of users 1 and 2. Next, for multiple users the channel model can be modified .

The $\boldsymbol{G}$int, $\boldsymbol{s}$int denote the effective channel matrix and the symbol vector corresponding to users 3 to K which are regarded as Gaussian interferers. Then, it is readily verified that the only change we need to make in the two-user group soft demodulator is in the covariance matrix $\boldsymbol{R}$, which should be updated to modification, the two-user group soft demodulator can be used to demodulate the a pairs (in parallel if the hardware complexity is feasible).

D. COMPLEXITY ANALYSIS

In this section a complexity analysis for the group soft demodulator and the conventional receiver is performed. We assume that there are nR receive antennas at the base station, which communicates simultaneously with two UEs (each with a single transmit antenna) via SDMA and that the DFT block size for each user is M. Further, we suppose that each codeword spans B DFT blocks over which the channel coefficients remain constant. Consequently, the filters have to be computed only once in every B blocks. For simplicity, we assume that each user employs an identical QAM constellation *1) Conventional Receiver:* First, we examine the complexity of the conventional receiver and consider the common terms, which are calculated only once for all the B Fetlocks and count the number of real multiplications. The filters $\{\_\boldsymbol{H}\dagger m\boldsymbol{H}m + \boldsymbol{I}\_{-1}\ \boldsymbol{H}\dagger m\}$, for $m = 1, ...,M$, as wellas the terms $\boldsymbol{D}(k)$, $\alpha(k)$, $E[|n(k)i|2]$ for $k = 1, 2$ are calculated once per B blocks and together involve about $M(28nR+40)$ real multiplications. Next, consider the operations that are performed for each DFT block. For each DFT block, we need to determine $\{\_x(k)\ m\ \}Mm{=}1$, $1 \le k \le 2$, as well as perform two M point DFTs. Together, these involve about $4BM(2nR + \log 2M)$ real multiplications. Thus, the filtering operations over B sub blocks require about $4BM(\log 2M + 2nR) + M(28nR + 40)$ real multiplications. Next, for the conventional receiver, to compute the LLRs, we need 1, 5, and 9 real multiplications per-bit for QPSK, 16-QAM and 64-QAM, respectively complexity to be $4BM(\log 2M + 2nR + 8) + M(28nR + 40)$ real multiplications. From the above analysis, we can conclude that both the group soft demodulator and the conventional receiver incur almost the same complexity in the filtering operations. The remaining operations are those that are involved in computing the LLRs. For the two-symbol max log LLR demodulator, according to Section II, we see that obtaining the LLR of each bit needs about $1+(2|S|+5\_|S|)/\log 2|S|$ real multiplications.

IV. SIMULATION RESULTS

In this section, we present simulation results for the various receivers considered in this paper. In all simulations, the high-rate codes are obtained by puncturing a rate- 13 mother code and the block size is chosen to be 768 information bits. Random inter leavers are used between the two component convolution encoders within the turbo encoder, and between the modulator and the turbo encoder, respectively. Furthermore, we assume that the channel is block fading ,i.e., the channel impulse responses stay constant during one coded block, and vary independently from block to block. The parameters of the simulated DFT-spread OFDM system are summarized in Table II. The WINNER C2 channel is used to generate the channel impulse responses. Its power delay profile is given in Table III. *DFT-S-OFDMA:* We first consider the DFT-S-OFDMA systems. In this scenario, simulations were performed for systems with coding rates 1⁄2,

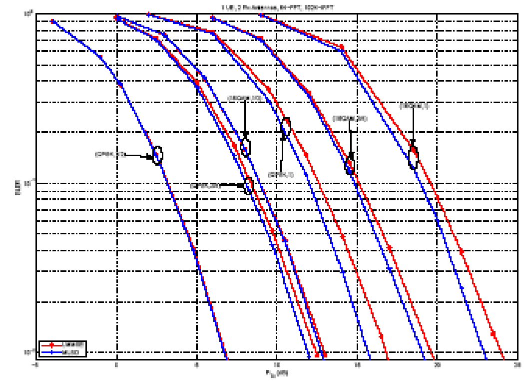


Fig. 4. Performance comparison between the conventional receiver and the group soft demodulator in a DFT-S-OFDMA system.

3⁄4 and 1. The size of the DFT spreading matrix (M) is taken to be64 whereas the OFDMA system has 1024 tones (i.e., the IDFT used at the transmitter is of size 1024×1024). The results are presented in Fig. 4, where we can see that the proposed group soft demodulator offers no gain over the conventional receiver in the low-rate regimes

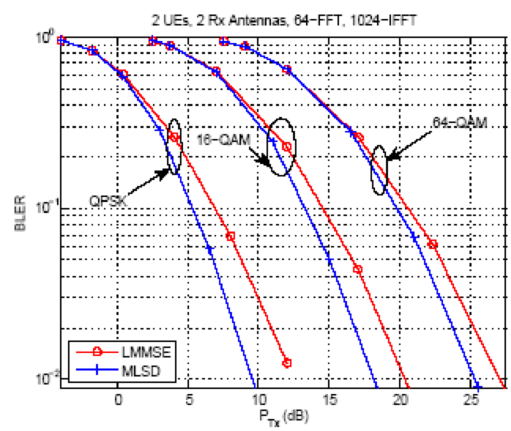


Fig. 5. Performance comparison between the conventional receiver and the group soft demodulator in a DFT-S-OFDM-SDMA system with coding rate 1/2.

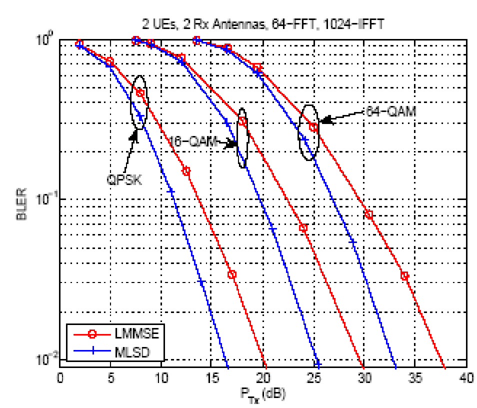


Fig. 6. Performance comparison between the conventional receiver and the group soft demodulator in a DFT-S-OFDM-SDMA system with coding rate 3/4.

The coding rate for each user is 1/2. It is seen that the group soft demodulator offers a significant gain over the conventional receiver. At 10^{-2} BLER, the gains are 3dB, 2.2dB, and 1.8dB for QPSK, 16-QAM and 64-QAM, respectively. Note that the loss of receive diversity in the SDMA case exposes the limitation of the conventional receiver. Moreover, an even larger gain is achieved at higher coding rates, where the outer code cannot capture all the available frequency diversity. One example is shown in Fig. 6, where the coding rate is 3/4, and the gains are 3.73dB, 4.3dB and 4.76dB for QPSK, 16-QAMand 64-QAM, respectively.

V. CONCLUSIONS

We have proposed low-complexity high-performance receiver algorithms for computing the LLRs in DFT-spread OFDM systems. Two ingredients of the proposed method.

TABLE II
SIMULATION PARAMETERS

Parameter	Assumption
Bandwidth	10.0 MHz
Total number of subcarriers	1024
Sub-carrier spacing	15.0 kHz
Number of occupied sub-carriers	64
Number of information bits per block	768
Tone mapping method	Localized
Number of antennas at Node-B	2
Number of antennas at user	1
Number of users	1 or 2
Channel model	WINNER C2 (r.f. Table III)
Channel estimation	Ideal

TABLE III
POWER DELAY PROFILE OF WINNER C2 CHANNEL

Power (dB)	$-0.5, 0, -3.4, -2.8, -4.6, -0.9, -6.7, -4.5, -9.0, -7.8, -7.4, -8.4, -11.0, -9.0, -5.1, -6.7, -12.1, -13.2, -13.7, -19.8$
Delay (ns)	$0, 5, 135, 160, 215, 260, 385, 400, 530, 540, 650, 670, 720, 750, 800, 945, 1035, 1185, 1390, 1470$

are an efficient linear prefilter for interference suppression, and a two-symbol max-log LLR demodulator. The proposed technique has been applied to both the DFT-S-OFDMA and the DFT-S-OFDM-SDMA uplink and it offers significant gain over the conventional receiver that is based on tone-by-tone equalization and inverse DFT, with little attendant increase in complexity.

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