Multiresolution MBMS Transmissions for MIMO UTRA LTE Systems

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Abstract— **Hierarchical constellations constitute a simple technique for achieving multiresolution and, therefore, are appealing for MBMS (Multimedia Broadcast and Multicast Service). In this paper we consider the use of M-QAM hierarchical constellations (Quadrature Amplitude Modulation) combined with MIMO (Multiple Input Multiple Output) for the transmission of multicast and broadcast services in UTRA (Universal Mobile Telecommunications System Terrestrial Radio Access) Long Term Evolution (LTE) systems based on Orthogonal Frequency Division Multiplexing (OFDM). Due to the demanding channel estimation requirements and the high sensitivity to interference resulting from the usage of several antennas and hierarchical constellations, an enhanced receiver based on the turbo concept is employed and its performance is evaluated.**

Keywords- MBMS, Hierarchical constellations, MIMO, LTE.

I. INTRODUCTION

It is widely recognized that OFDM modulations [1] are suitable for broadband wireless systems. For this reason they were selected for several digital broadcast systems and wireless networks [2] and are also being considered for UTRA LTE [3]. Regarding UTRA LTE, special attention is being devoted to the support of MBMS which has already been standardized in 3GPP UTRAN (UMTS Terrestrial Radio Access Network) Release-6 [4] and 7 [5]. The goal is to enable an efficient support of downlink streaming (from the base station to the mobile terminal) and download-and-play type services to large groups of users. From the radio perspective, MBMS includes point-to-point (PtP) and point-to-multipoint (PtM) modes. Regarding the PtM mode it seems attractive to employ hierarchical modulations since it is a simple and flexible enhancement technique that can increase the transmission efficiency, due to its ability to provide unequal error protection to different bits and thus provide multiresolution into a cell. By having several classes of bits with different error protection associated and to which different streams of information are mapped, a given user can attempt to demodulate only the more protected bits or also the bits that carry the additional information, depending on the propagation conditions. This type of approach is possible whenever the information can be scalable like the cases of coded voice or video signals, as studied in [6][7]. For this reason hierarchical 16-QAM and 64-QAM constellations have already been incorporated into DVB-T (Digital Video Broadcasting - Terrestrial) standards [8].

MIMO schemes have emerged as one of the most promising methods for capacity increase in a communication system [9][10] and are being considered for UTRA LTE [3]. In MIMO systems with coherent detection the channel estimation plays a crucial role since the performance of the spatial signal processing in the receiver depends on the accuracy of the channel estimates. Furthermore QAM constellations can be severely affected due to inaccurate channel estimates.

In this paper we consider the use of QAM hierarchical constellations in a UTRA LTE OFDM based system employing multiple transmitting and receiving antennas with the aim of supporting broadcast and multicast services. To deal with the high sensitivity to channel estimation errors we employ an iterative receiver capable of performing joint MIMO detection and channel estimation. This receiver is based on the approach proposed in [11] for WCDMA systems. It can apply different MIMO equalization techniques during the iterative process and obtain refined channel estimates by considering the data symbols as extra pilots, as proposed in [12].

The paper is organized as follows. First Section II introduces hierarchical constellations and defines the model of the MIMO-OFDM system considered in this study. In Section III the proposed iterative receiver structure and respective channel estimation process are described. Section IV presents some performance results obtained with the proposed scheme while the conclusions are given on Section V.

II. SYSTEM DESCRIPTION

A. M-QAM Hierarchical Signal Constellations

In hierarchical constellations there are two or more classes of bits with different error protection and to which different streams of information can be mapped. By using non uniformly spaced signal points (where the distances along the I or Q axis between adjacent symbols are different) it is possible to modify the different error protection levels. As an example, a non-uniform 16-QAM constellation can be constructed from a main

QPSK constellation where each symbol is in fact another QPSK constellation, as shown in Figure 1.

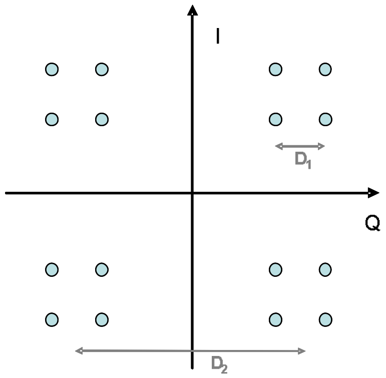

Figure 1. Non-uniform 16-QAM constellation.

The basic idea is that the constellation can be viewed as a 16-QAM constellation if the channel conditions are good enough or as a QPSK constellation otherwise. In the latter situation, the received bit rate is reduced to half. These constellations can be characterized by the parameter $k_1=D_1/D_2$ ($0<k_1\leq 0.5$), as shown in Figure 1. If $k_1=0.5$, the resulting constellation corresponds to a uniform 16-QAM. This approach can be naturally extended to any QAM constellation size M where the number of possible classes of bits with different error protection is $1/2\cdot\log_2 M$.

B. *Transmitted Signals*

In Figure 2 we show a transmitter chain that incorporates QAM hierarchical constellations into a UTRA LTE based MIMO-OFDM transmission. In the proposed scheme, there are $1/2\cdot\log_2 M$ parallel chains for the different input bit streams that will have unequal error protection. For 16-QAM we can use two parallel chains while for 64-QAM we can use three chains. Each stream is encoded, interleaved and mapped into the constellation symbols in the modulation mappers according to the importance attributed to the chain. Pilot symbols are inserted into the modulated data sequence which is then converted to the time domain using an IDFT (Inverse Discrete Fourier Transform). The resulting stream is then split into several smaller streams which are transmitted simultaneously by M_{tx} transmitting antennas. Note that in the proposed scheme the coding is not performed independently for each different antenna. Instead each data sequence is encoded and divided equally among the transmitting antennas by the Serial to Parallel block. The objective is to try to obtain some diversity for the same encoded block.

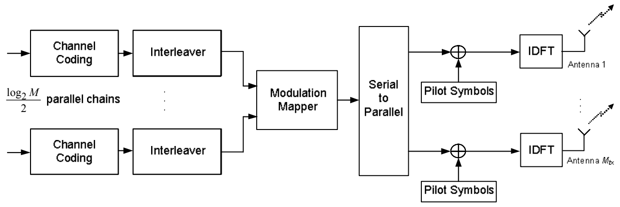

Figure 2. Transmitter chain.

In this paper we consider the frame structure of Figure 3 for a MIMO-OFDM system with N carriers. According to this structure the pilot symbols are multiplexed with the data symbols using a spacing of ΔN_T symbols in the time domain and ΔN_F OFDM blocks in the frequency domain. To avoid interference between pilots of different transmitting antennas, FDM (Frequency Division Multiplexing) is employed for the pilots, which means that pilot symbols cannot be transmitted over the same subcarrier in different antennas. Data symbols are not transmitted on subcarriers reserved for pilots in any antenna, therefore, the minimum allowed spacing in the frequency domain is $(\Delta N_F)_{\min} = M_{tx}+1$.

Before being transmitted, the sequences of symbols are converted to the time domain through $\left\{x_{i,l}^m, i=0,1,...,N-1\right\} = \mathrm{IDFT}\left\{S_{k,l}^m, k=0,1,...,N-1\right\}$, where $S_{k,l}^m$ is the symbol transmitted by the k^{th} subcarrier of the l^{th} OFDM block using antenna m. The transmitted OFDM signals are then expressed as

$$x^m(t) = \sum_l \sum_{i=-N_G}^{N-1} x_{i,l}^m \cdot h_T(t - i\cdot T_s), \qquad (1)$$

with T_s denoting the symbol duration, N_G the number of samples at the cyclic prefix ($x_{-i,l}^m = x_{N-i,l}^m$, i= 1,...,N_G) and $h_T(t)$ the adopted pulse shaping filter.

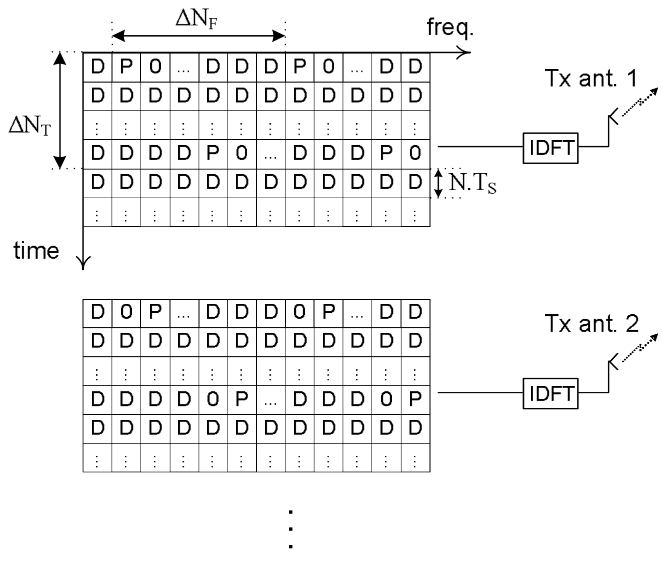

Figure 3. Frame structure for a OFDM transmission (P – pilot symbol, D – data symbol, T_S – symbol duration).

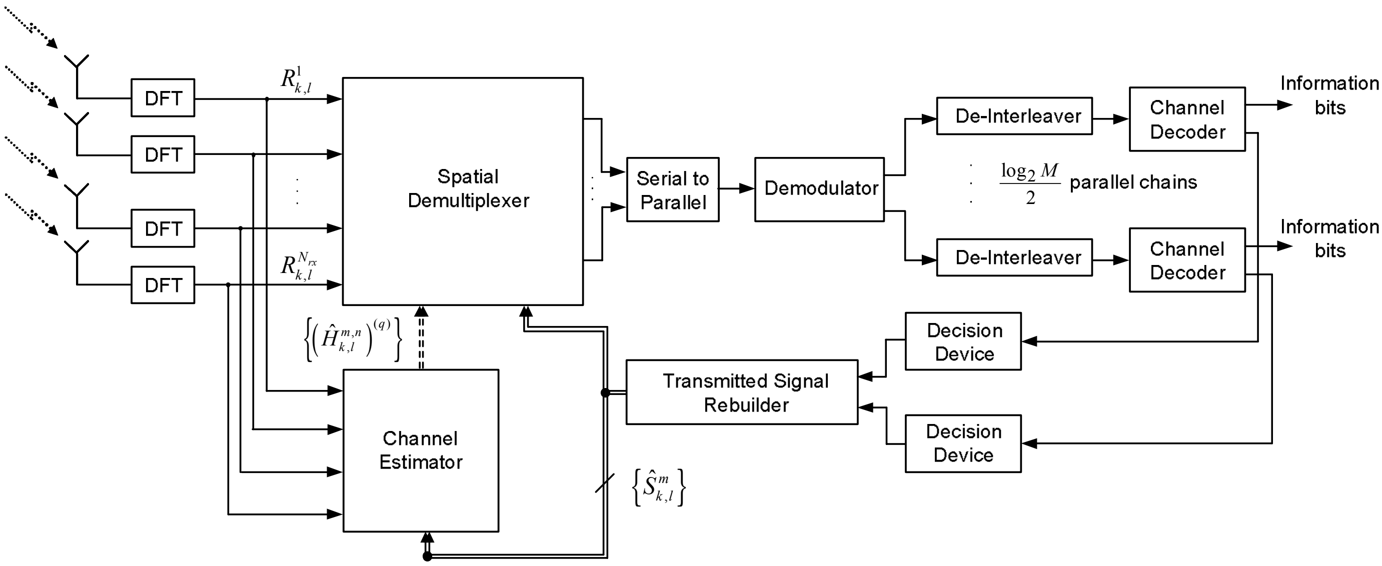

Figure 4. Iterative receiver structure.

III. ITERATIVE RECEIVER

A. Receiver Structure

To achieve reliable channel estimation and data detection we employ a receiver capable of jointly performing these tasks through iterative processing. The structure of the iterative receiver is shown in Figure 4 where N_{rx} receiving antennas are employed. According to the figure, the signal, which is considered to be sampled and with the cyclic prefix removed, is converted to the frequency domain after an appropriate size-N DFT operation. If the cyclic prefix is longer than the overall channel impulse response the resulting sequence received in antenna n can be expressed as

$$R^{n}_{k,l} = \sum_{m=1}^{M_{tx}} S^{m}_{k,l} H^{m,n}_{k,l} + N^{n}_{k,l}, \tag{2}$$

with $H^{m,n}_{k,l}$ denoting the overall channel frequency response between transmit antenna m and receiving antenna n for the kth frequency of the lth time block and $N^{m}_{k,l}$ denoting the corresponding channel noise.

The sequences of samples (2) enter the MIMO equalizer (Spatial Demultiplexer block) which separates the simultaneous transmitted streams. This can be accomplished with an MMSE (Minimum Mean Squared Error) equalizer [13], a ZF (Zero Forcing) equalizer [13], a Maximum Likelihood Soft Output criterion (MLSO) or an interference canceller (IC) [11]. It is possible to perform some of the receiver iterations using one spatial demultiplexing technique, like the MMSE, and the others using a different one, like the IC, as was studied in [11]. In any case, after MIMO equalization the demultiplexed symbol sequences are serialized and pass through the demodulator, de-interleaver and channel decoder blocks. This channel decoder has two outputs. One is the estimated information sequence and the other is the sequence of log-likelihood ratio (LLR) estimates of the code symbols. These LLRs go through the Decision Device, which outputs either soft-decision or hard decision estimates of the code symbols, and enter the Transmitted Signal Rebuilder which performs the same operations of the transmitter (interleaving, modulation, conversion of serial to parallel streams). The reconstructed symbol sequences are then used for a refinement of the channel estimates and also for possible improvement of the spatial demultiplexing task (in case of employing an IC) for the subsequent iteration.

The possible MIMO equalization techniques are now going to be briefly described. Using matrix notation the MMSE estimates of the transmitted symbols in subcarrier k and OFDM block l is given by

$$\hat{\mathbf{S}}_{k,l} = \hat{\mathbf{H}}_{k,l}^{H} \cdot \left(\hat{\mathbf{H}}_{k,l}\hat{\mathbf{H}}_{k,l}^{H} + \sigma^2\mathbf{I}\right)^{-1} \mathbf{R}_{k,l} \tag{3}$$

where $\hat{\mathbf{S}}_{k,l}$ is the M_{tx}×1 estimated transmitted signal vector with one different transmit antenna in each position, $\hat{\mathbf{H}}_{k,l}$ is the N_{rx}×M_{tx} channel matrix estimate with each column representing a different transmit antenna and each line representing a different receive antenna, $\mathbf{R}_{k,l}$ is the N_{rx}×1 received signal vector with one different receive antenna in each position and σ^2 is the noise variance. The ZF estimate can be simply obtained by setting σ to 0 in (3).

In the MLSO criterion we use the following estimate for each symbol

$$\hat{S}^{m}_{k,l} = E\left[S^{m}_{k,l} \middle| \mathbf{R}_{k,l}\right]$$

$$= \sum_{s_i \in \Lambda} s_i \cdot \frac{P\left(S^{m}_{k,l} = s_i\right)}{p\left(\mathbf{R}_{k,l}\right)} p\left(\mathbf{R}_{k,l} \middle| S^{m}_{k,l} = s_i\right), \tag{4}$$

where s_i corresponds to a constellation symbol from the modulation alphabet Λ, $E[\cdot]$ is the expected value, $P(\cdot)$

represents a probability and $p(\cdot)$ a probability density function (PDF). Considering equiprobable symbols $P\left(S_{k,l}^{m}=s_i\right)=1/M$, where M is the constellation size. The PDF values required in (4) can be computed as

$$p\left(\mathbf{R}_{k,l}\left|S_{k,l}^{m}=s_i\right.\right)=\frac{1}{M^{M_{tx}-1}}\sum_{\mathbf{S}_{k,l}^{\text{interf}}\in\Lambda^{M_{tx}-1}}p\left(\mathbf{R}_{k,l}\left|S_{k,l}^{m}=s_i,\mathbf{S}_{k,l}^{\text{interf}}\right.\right) \quad (5)$$

with

$$p\left(\mathbf{R}_{k,l}\left|S_{k,l}^{m}=s_i,\mathbf{S}_{k,l}^{\text{interf}}\right.\right)=\frac{1}{\left(2\pi\sigma^2\right)^{N_{rx}}}\exp\left[\sum_{n=1}^{N_{rx}}-\frac{\left|R_{k,l}^{n}-\hat{\mathbf{H}}_{k,l}(n,:)\cdot\mathbf{s}\right|^2}{2\sigma^2}\right],\quad (6)$$

where $\mathbf{S}_{k,l}^{\text{interf}}$ is a (M_{tx}-1)×1 vector representing a possible combination of symbols transmitted simultaneously by all antennas except antenna m, $\mathbf{s}$ is a M_{tx}×1 vector composed by $\mathbf{S}_{k,l}^{\text{interf}}$ and s_i and $\hat{\mathbf{H}}_{k,l}(n,:)$ is the n^{th} line of channel matrix $\hat{\mathbf{H}}_{k,l}$.

An IC can also be used inside the Spatial Demultiplexer block, but usually is only recommendable after the first receiver iteration [11]. In this case, in iteration q, for each transmit antenna m and receive antenna n, the IC subtracts the interference caused by all the other antennas. This can be represented as

$$\left(R_{k,l}^{n,m}\right)^{(q)}=R_{k,l}^{n}-\sum_{\substack{m'=1\\m'\neq m}}^{M_{tx}}\left(\hat{S}_{k,l}^{m'}\right)^{(q-1)}\hat{H}_{k,l}^{m',n}, \quad (7)$$

where $\left(\hat{S}_{k,l}^{m}\right)^{(q-1)}$ represents the data symbols estimates of the previous iteration for transmit antenna m, subcarrier k and OFDM block l.

B. Channel Estimation

To obtain the frequency channel response estimates for each transmitting/receiving antenna pair the receiver applies the following steps in each iteration:

(1) The channel estimate between transmit antenna m and receive antenna n for each pilot symbol position, is simply computed as

$$\tilde{H}_{k,l}^{m,n}=\frac{\left(S_{k,l}^{m,Pilot}\right)^{*}}{\left|S_{k,l}^{m,Pilot}\right|^{2}}R_{k,l}^{n}. \quad (8)$$

where $S_{k,l}^{m,Pilot}$ corresponds to a pilot symbol transmitted in the k^{th} subcarrier of the l^{th} OFDM block using antenna m. Obviously not all indexes k an l will correspond to a pilot symbol since $\Delta N_T>1$ and $\Delta N_F>1$.

(2) Channel estimates for the same subcarrier k, transmit antenna m and receive antenna n but in time domain positions (index l) that do not carry a pilot symbol can be obtained through interpolation using a finite impulse response (FIR) filter with length W as follows:

$$\hat{H}_{k,l+t}^{m,n}=\sum_{j=-\lfloor(W-1)/2\rfloor}^{\lfloor W/2\rfloor}h_t^j\tilde{H}_{k,l+j\cdot\Delta N_T}^{m,n} \quad (9)$$

where t is the OFDM block index relative to the last one carrying a pilot (which is block with index l) and h_t^j are the interpolation coefficients of the estimation filter which depend on the channel estimation algorithm employed. There are several proposed algorithms in the literature like the optimal Wiener filter interpolator [14] or the low pass sinc interpolator [15].

(3) After the first iteration the data estimates can also be used as pilots for channel estimation refinement.

IV. Numerical Results

To study the behaviour of the proposed MIMO-OFDM scheme and respective iterative receiver, several simulations were performed for a 16-QAM (k_1=0.4) hierarchical constellation. Two classes of bits with different error protection were used. Each individual information stream was encoded with a block size chosen so that the final encoded and modulated stream fitted a sub-frame composed of 7 OFDM blocks (corresponding to a 0.5ms duration). All the parameters used for these simulations were based on UTRA LTE 3GPP documents [16] and [17], for a 10MHz bandwidth. Table 1 shows the respective parameters.

Transmission BW	10 MHz
CP length	72
FFT size	1024
Number of occupied sub-carriers	600
Sub-frame duration (ms)	0.5
Sub-carrier spacing (kHz)	15
symbol duration (ns)	130
pilot power (dB)	0
OFDM symbols per sub-frame	7

Table 1. Simulation parameters for 10MHz bandwidth.

The channel impulse response is based on Vehicular A environment [19] with Rayleigh fading assumed for the different paths. A velocity of 30 km/h was employed unless otherwise stated. The channel encoders were rate-1/2 turbo codes based on two identical recursive convolutional codes characterized by G(D) = [1 $(1+D^2+D^3)/(1+D+D^3)$] [18]. A random interleaver was used within the turbo encoders. Table 2 shows the different combinations of receiver and turbo decoder iterations applied as well as the respective MIMO decoding methods used.

Receiver method	Turbo decoder iterations per receiver loop	Iterative Receiver Iterations
1	12	1 with MMSE
2	3	1 with MMSE+3 with IC
3	12	1 with MLSO

Table 2. Different receiver method applied for the simulations.

Most of the BER (Bit Error Rate) results presented next will be shown as a function of E_S/N_0, where E_S is the average symbol energy and N_0 is the single sided noise power spectral density. For channel estimation purposes, pilot symbols were distributed using a spacing of $\Delta N_F = 6 + M_{tx}$ and $\Delta N_T = 4$ or 7 (the two possible configurations proposed in [16]) and a sinc filter interpolation with length W=2 was used at the receiver. In the graphs legends, MPB designates most protected bits, IPB means intermediate protected bits and LPB corresponds to least protected bits.

Figure 5 compares the performance of the different receiver methods of Table 2 for a MIMO 2x2 transmission employing a 16-QAM (with k_I=0.4) hierarchical constellation. It is visible that, although the receiver with the MMSE equalizer alone performs worse than when using the MLSO equalizer, the performance can be substantially improved and achieve lower BLERs than with the MLSO when applying also an IC in the last receiver iterations. For the remainder of the paper the receiver configuration considered will be method 2 (MMSE+IC).

Figure 6 shows the behaviour of the receiver for different velocities. According to the results, the performance is almost insensitive to velocity until 120 km/h, being visible only a small degradation in the performance of the LPB. For higher velocities the performance quickly degrades for the LPB but does not change significantly for the MPB even at 300 km/h.

Figure 7 compares the performance of a MIMO 2x2 transmission employing a 16-QAM (with k_I=0.4) hierarchical constellation with the two possible pilot spacings in the time domain. It is visible that both cases have similar performances (close to the perfect estimation curves), which means that it is possible to adopt the larger pilot spacing and therefore increase the transmission efficiency if one or two antennas (we verified that the conclusion is also true even for four transmit antennas). It is important to remember however that reducing the number of pilot symbols will sacrifice the system robustness for higher velocities as will be shown further ahead.

V. CONCLUSIONS

In this paper we have studied the use of QAM hierarchical constellations with the aim of supporting multicast and broadcast transmissions in a MIMO-OFDM system similar to the one being considered for UTRA LTE.

It was verified through simulations that the iterative receivers schemes studied are able to achieve good performances for all the bit streams, including those with lower error protection levels even for very high velocities. Therefore the proposed transmitter/receiver scheme can provide unequal error protection which is adequate for supporting MBMS transmissions in UTRA LTE. It was also observed that if a maximum of four transmit antennas are being used with 16-QAM hierarchical modulations, the option with only half of the pilots symbols proposed for UTRA LTE ($\Delta N_T = 7$) is adequate, unless a higher robustness is desired for high velocities for all information streams.

VI. ACNKOWLEDGMENTS

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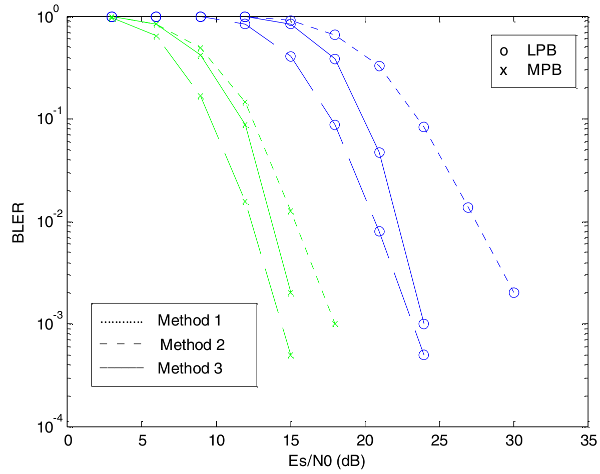

Figure 5. 16-QAM (k_I=0.4) hierarchical constellation, v=30 Km/h, 2x2 MIMO transmission using several receiver methods. $\Delta N_T = 4$.

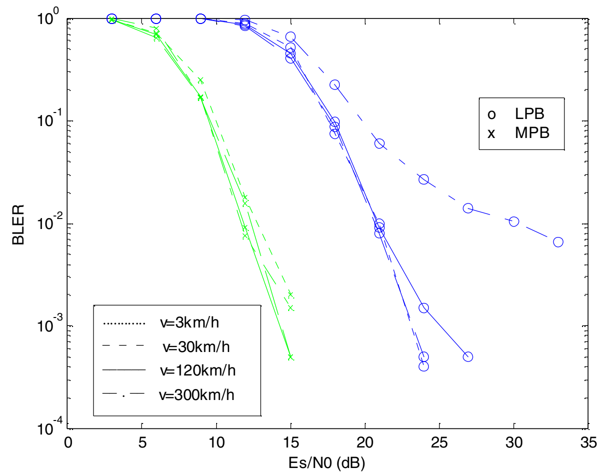

Figure 6. 16-QAM (k_I=0.4) hierarchical constellation. 2x2 MIMO transmission using receiver method 2 for several velocities. $\Delta N_T = 4$.

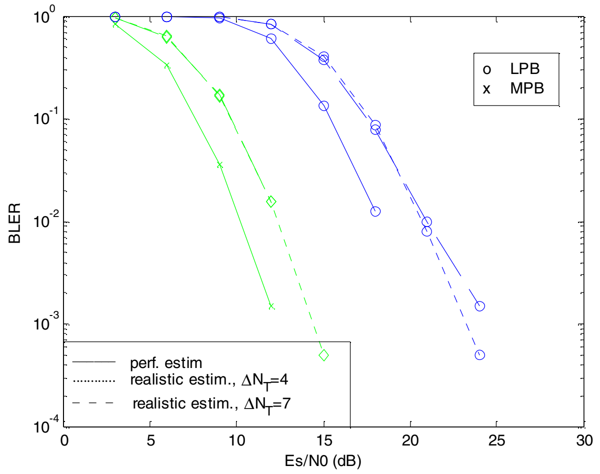

Figure 7. 16-QAM (k_1=0.4) hierarchical constellation, v=30 Km/h, 2x2 MIMO transmission using receiver method 2 for several pilot spacings.

VII. References

[1] L.Cimini, "Analysis and Simulation of a Digital Mobile Channel using Orthogonal Frequency Division Multiplexing", IEEE Trans. on Comm., Vol. 33, No. 7, July 1985.

[2] R.van Nee and R.Prasad, "OFDM for Wireless Multimedia Communications", Artech House Publ., 2000.

[3] 3GPPP TR 25.814-v7.1.0, "3rd Generation Partenership Project: Technical Specification Group Radio Access Network; Physical Layers Aspects for Evolved UTRA", September 2006.

[4] 3GPP, 25.346 - v.6.9.0, 'Introduction of Multimedia Broadcast/Multicast Service (MBMS) in RAN', 3GPP, June 2006.

[5] 3GPP, 25.346 - v.7.2.0, 'Introduction of Multimedia Broadcast/Multicast Service (MBMS) in RAN', 3GPP, September 2006.

[6] T. Cover, "Broadcast channels", IEEE Trans. On Inform. Theory, vol.IT-18, pp.2-14, January 1972.

[7] K. Ramchandran, A. Ortega, K. M. Uz, and M. Vetterli, "Multiresolution broadcast for digital HDTV using joint source/channel coding", IEEE J. Select. Areas Commun., vol. 11, January 1993.

[8] 'Digital video broadcasting (DVB) framing structure, channel coding and modulation for digital terrestrial television (DVB-T)', Mar. 1997. ETSI, European Telecommunication Standard ETS 300 744.

[9] E. Telatar, "Capacity of multiantenna Gaussian channels," *AT&T Bell Laboratories*, *Technical Memorandum*, June 1995.

[10] G.J. Foschini and M.J. Gans, "On limits of wireless communications in a fading environment when using multiple antennas," *Wireless Pers.Commun.*, vol. 6, pp. 311–335, Mar. 1998.

[11] N. Souto, R. Dinis, F. Cercas J. Silva, A. Correia "An Iterative Receiver for WCDMA Systems with MIMO Transmissions and Hierarchical Constellations", Proc. IEEE International Symposium on Spread Spectrum Techniques and Applications - ISSSTA 2006, August 2006, Manaus, Brasil.

[12] M. C. Valenti, "Iterative Channel Estimation and Decoding of Pilot Symbol Assisted Turbo Codes Over Flat-Fading Channels", IEEE Journal on Selected Areas in Comm., vol. 19, no. 9, pp. 1697-1705, September, 2001.

[13] S.M. Kay, Fundamentals of Statistical Signal Processing: Estimation Theory. Englewood Cliffs, NJ: Prentice-Hall, 1993.

[14] J. K. Cavers, "An analysis of Pilot Symbol Assisted Modulation for Rayleigh Fading Channels", IEEE Trans. On Veh. Tech., vol. 40, no. 4, pp.686-693, Nov. 1991.

[15] Y.-S. Kim, C.-J. Kim, G.-Y. Jeong, Y.-J. Bang, H.-K. Park and S. S. Choi, "New Rayleigh fading channel estimator based on PSAM channel sounding technique", in Proc. IEEE International Conf. on Comm., pp.1518-1520, Montreal, Canada, June 1997.

[16] 3GPP "Feasibility study for evolved Universal Terrestrial Radio Access (UTRA) and Universal Terrestrial Radio Access Network (UTRAN)" Tech rep. 25.912 v7.1.0, http://www.3gpp.org

[17] 3GPP "Feasibility Study for Orthogonal Frequency Division Multiplexing (OFDM) for UTRAN enhancement" Tech rep. 25.892 v6.0.0 http://www.3gpp.org

[18] 3GPP, 25.212-v6.2.0, 'Multiplexing and Channel Coding (FDD)'.

[19] 'Selection procedures for the choice of radio transmission technologies of UMTS', ETSI, TR 101 112 v3.2.0, Sophia Antipolis, France, 1998.