A general purpose low-complexity spectral equaliser for 3G and 4G systems not requiring a cyclic prefix

E. de Marinis, O. Gasparini
DUNE srl
Via Tracia, 4, 00183, Rome, Italy
demarinis@dune.191.it

M.J. Hart
Fujitsu Laboratories of Europe
Hayes Park, Hayes, London, UK, UB4 8FE
Mike.Hart@uk.fujitsu.com

Abstract— **Mid term evolved 3G and 4G systems will be required to support a variety of multiple access schemes (e.g. OFDM(A), MC-CDMA, TDMA, CDMA), as well as high mobility scenarios. A novel spectral equaliser suitable for the scope is presented, which exploits the low processing load inherent to the FFT. Converse to spectral equalisation used in OFDM systems, which requires the introduction of a cyclic prefix in the transmission format, the proposed equaliser does not require it. Hence the equaliser can work with systems using cyclic prefix as well as with systems not using it, such as UMTS. A case study is carried out for CDMA: UMTS FDD HSDPA and E-UL systems for SISO and MIMO.**

Index Terms— **3G, 4G, Cyclic Prefix, MIMO, UMTS FDD, HSDPA, E-UL, Spatial Multiplexing, Spectral Equalisation.**

I. Introduction

IN the current vision of mid term evolution of 3G [1] and 4G systems, a receiver should be capable of supporting a variety of physical access techniques (e.g. OFDM(A), MC-CDMA, TDMA, CDMA). To enable this, the receiver could embed Software Defined Radio (SDR) tools to fulfil the above task.

The systems should also perform well in the presence of high user velocities. This requirement is challenging, resulting in trade-off solutions between achievable spectral efficiency and the maximum velocity for which performance is guaranteed. Today, PHY processing is commonly implemented by segmentation of the received data into blocks of symbols, whose duration is matched to the transmitted format (e.g. sub-frame length in UMTS or in OFDM the payload duration which is the interval between successive cyclic prefix (CP) extensions). When channel coherence time becomes comparable to the window length, mobility may degrade performance. A desirable solution is to decouple the window length from the payload/sub-frame duration and let the former be much smaller than the channel coherence time.

The receiver should also cope with significant time dispersion in the channel causing Inter Symbol Interference (ISI). A common tool to recover performance impaired by ISI is equalisation at the receiver. If a CP were included in transmission, Spectral Equalisation with Cyclic Prefix (SE-CP) could be employed in the receiver, exploiting the low processing load of the FFT [2], [3]. However introduction of a CP reduces spectral efficiency and, in the case of legacy systems (as in UMTS) CP is not always available. In this paper a novel multi-access, low-complexity spectral equaliser that does not require a CP, employing short processing windows is introduced. This will be indicated as BCSE (Bias Controlled Spectral Equaliser). A case study for multiple access based on CDMA, in particular a UMTS FDD HSDPA and E-UL system (not using CP), is considered for SISO and MIMO channels.

II. Time Equalisation

Time processing for BCSE is outlined below for Spatial Multiplexing (SM) with 2-TX and 2-RX antennas (2 x 2 case). Extension to M x M case is straightforward. The received data stream is partitioned in N-sample blocks. The sampling rate is taken as the inverse of bandwidth in OFDM and TDMA, and in CDMA as the chip rate. We define:

$\mathbf{y}_j$: $[y_j(1), \ldots, y_j(N)]^T$ N- sample vector received in a block by the j^{th} RX antenna; $y_j(N)$ is the most recent sample in the block.

$\mathbf{x}_k$: N- sample vector transmitted by the k^{th} TX antenna;

ν_j: N- sample noise vector in the j^{th} RX antenna.

$\mathbf{y} = [\mathbf{y}^T_1; \mathbf{y}^T_2]^T$, $\mathbf{x} = [\mathbf{x}^T_1; \mathbf{x}^T_2]^T$, $\nu = [\nu^T_1; \nu^T_2]^T$.

$\mathbf{h}_{j,k}$ = $[h_{j,k}(0), \ldots, h_{j,k}(L), 0, \ldots, 0]$: N-sample, base-band channel impulse response of spread L ($L < N$), for the link between the k^{th} TX antenna and the j^{th} RX antenna. We assume the channel is perfectly known at the receiver.

$\mathbf{H}_{\mathrm{T},j,k}$: N x N Toeplitz matrix with first column $[h_{j,k}(0), \ldots, h_{j,k}(L), 0, \ldots, 0]$ and first row $[h_{j,k}(0), 0, \ldots, 0]$;

The work has been developed in the frame of the IST-ROMANTIK project (contract IST-2001-32549).

$\mathbf{H}_{U,j,k}$: N x N Toeplitz, upper triangular, matrix with first column [0, …, 0] and first row [0, …, 0, $h_{j,k}(L)$, …, $h_{j,k}(1)$];

$\mathbf{H}_{C,j,k}$: N x N circulant (then Toeplitz) matrix, with first column [$h_{j,k}(0)$, …, $h_{j,k}(L)$, $\mathbf{0}$];

$\mathbf{H}_P$ = [$\mathbf{H}_{P,1,1}$, $\mathbf{H}_{P,1,2}$; $\mathbf{H}_{P,2,1}$, $\mathbf{H}_{P,2,2}$], block matrix; where P stands for {T, U, C}.

Since $\mathbf{H}_T = \mathbf{H}_C - \mathbf{H}_U$, the n^{th} received block (n is the block index, increasing with increasing time) can be written as [2]:

$$\mathbf{y}(n) = \mathbf{H}_C\mathbf{x}(n) + \mathbf{H}_U\left(\mathbf{x}(n-1) - \mathbf{x}(n)\right) + \mathbf{v}(n)\,, \qquad (1)$$

Eq (1) indicates that the received vector in block n has a contribution from transmitted vectors in blocks n and $n-1$.

Received data blocks are sequentially processed in an N-sample sliding window to derive soft estimates of the transmitted vectors, so that the n-1 block estimate is available before processing the n^{th} block. As preliminary processing for block n, the transmitted vector soft estimate from block n–1, indicated by $\hat{\hat{\mathbf{x}}}_{ZF}(n-1)$, is multiplied by $\mathbf{H}_U$, then subtracted from $\mathbf{y}(n)$. This is a kind of Soft Feedback Equalization (SFE). Errorless estimation of $\mathbf{x}(n\text{-}1)$ is assumed for the moment for sake of explanation. This assumption will be removed in Sec. IV. We have:

$$\mathbf{y}_S(n) \equiv \mathbf{y}(n) - \mathbf{H}_U\mathbf{x}(n-1) = \mathbf{H}_C\left(\mathbf{I} - \mathbf{H}_C^{-1}\mathbf{H}_U\right)\mathbf{x}(n) + \mathbf{v}(n)\,. \qquad (2)$$

BCSE employs two estimation steps. Discarding index n and defining $\mathbf{B} = \mathbf{H}_C^{-1}\mathbf{H}_U$ (bias matrix), the first order (biased) estimate of the transmitted vector forms the first step. This is defined as the output of the zero forcing equaliser, given by:

$$\hat{\mathbf{x}}_{ZF} = \mathbf{H}_C^{-1}\mathbf{y}_S = \left(\mathbf{I} - \mathbf{B}\right)\mathbf{x} + \mathbf{H}_C^{-1}\mathbf{v} \qquad (3)$$

$\hat{\mathbf{x}}_{ZF}$ is the unbiased estimate of $\mathbf{x}_b = (\mathbf{I}-\mathbf{B})\mathbf{x}$.

A heuristic step to reduce bias is to add a correction term to the first order estimate to obtain the second order estimate:

$$\hat{\hat{\mathbf{x}}}_{ZF} = \hat{\mathbf{x}}_{ZF} + \mathbf{B}\hat{\mathbf{x}}_{ZF} = \left(\mathbf{I} - \mathbf{B}^2\right)\mathbf{x} + \left(\mathbf{I} + \mathbf{B}\right)\mathbf{H}_C^{-1}\mathbf{v}\,, \qquad (4)$$

Then $\hat{\hat{\mathbf{x}}}_{ZF}$ is obtained by an approximate inversion of (3), that is by expanding $(\mathbf{I}\text{-}\mathbf{B})^{-1}$ and truncating it at the first term.

Bias of $\hat{\hat{\mathbf{x}}}_{ZF}$ is related to $\mathbf{B}^2$: by comparing $\hat{\mathbf{x}}_{ZF}$ and $\hat{\hat{\mathbf{x}}}_{ZF}$, when $|\mathbf{B}| \ll 1$ bias decreases, whereas noise increase is negligible. However, when $|\mathbf{B}| > 1$, both bias and noise in $\hat{\hat{\mathbf{x}}}_{ZF}$ are magnified. However, since magnification does not have the same result on the various components of $\hat{\hat{\mathbf{x}}}_{ZF}$, performance can be improved by assuming that only a part of the components in the output of a processing window are valid, as indicated in the following. To justify BCSE in the simplest multipath case, consider the following SISO channel $\mathbf{h}$ = [1, -a]. The first column of $\mathbf{H}_C^{-1}$ is $\mathbf{a}_N = (1\text{-}\,a^N)^{-1}[1, a, \ldots, a^{N-1}]^T$, the bias matrix is $\mathbf{B}=-a(\mathbf{0},\ldots,\mathbf{0}, \mathbf{a}_N)$ and $\mathbf{B}^2=-(a^N/1\text{-}a^N)\mathbf{B}$. Maximum magnitudes M_1 and M_2 among the elements of $\mathbf{B}$ and $\mathbf{B}^2$ are, for $|a|<1$: $M_1=|a/(1\text{-}a^N)| \sim |a|$; and $M_2= |a^{N+1}/(1\text{-}a^N)| \sim |a^{N+1}| << 1$.

Maximum bias, which occurs for the 1st elements of $\hat{\mathbf{x}}_{ZF}$ and $\hat{\hat{\mathbf{x}}}_{ZF}$, decreases in $\hat{\hat{\mathbf{x}}}_{ZF}$ by increasing N. For $|a|>1$ we have: $M_1= |a^N/(1\text{-}a^N)| \sim 1$; and $M_2 = |a^{2N}/(1\text{-}a^N)^2| \sim 1$.

Maximum bias which occurs in last elements of $\hat{\mathbf{x}}_{ZF}$ and $\hat{\hat{\mathbf{x}}}_{ZF}$ does not decrease when passing from the 1st to the 2nd order estimate. The percentage bias in the k^{th} component $\hat{\hat{x}}_{ZF,k}$ is defined as $\left|\hat{\hat{x}}_{ZF,k} - x_k\right| \big/ \left|x_N\right| = \left|a^{N+k} \big/ (1-a^N)^2\right|$ and increases with k.

To control the bias effect, only an initial part of $\hat{\hat{\mathbf{x}}}_{ZF}$, up to $k_{\max}$, which has to be defined, is assumed valid.

A closed-form analysis for a generic channel has not been carried out, but simulation shows that the qualitative behaviour of bias on the second order estimate holds also for more complex, generic channels. Now a percentage error vector estimate is defined from (1) as $\boldsymbol{\varepsilon} = (\mathbf{H}_C^{-1}\mathbf{y}(n)\text{-}\mathbf{x}(n))/\bar{\sigma}_x = \mathbf{B}\left(\hat{\mathbf{x}}(n-1) - \hat{\mathbf{x}}(n)\right)(|\mathbf{h}|/\bar{\sigma}_y)$ where $\bar{\sigma}_x$ is the estimate of the standard deviation of the elements of $\mathbf{x}$, $|\mathbf{h}|=(\mathbf{h}^H\mathbf{h})^{1/2}$, $\bar{\sigma}_y = (1/N)(\mathbf{y}^H\mathbf{y})^{1/2}$ and hence ($\bar{\sigma}_y/|\mathbf{h}|$) is an estimate of $\bar{\sigma}_x$. The test for acceptance of the initial part of the window is: first index k_0 is defined as maximum index for which $|\boldsymbol{\varepsilon}(k)| \le T$ for all $k \le k_0$, T being a fixed adimensional threshold to be defined; then the index $k_{\max}$ is defined as min(k_0, N-L): only $\hat{\hat{\mathbf{x}}}_{ZF}$ components with index not smaller than $k_{\max}$ are assumed as a valid while other components are discarded. Next the N sample processing window slides along the received data samples, so that the start of the new window position is the first discarded element in the previous window position.

In the simulations a value of T=0.5 was selected by trial and error. Index $k_{\max}$ is a random variable, which depends on channel realisation. For SISO, T=0.5 and N=128, the average value of ($k_{\max}$) is found to be slightly smaller than N-L and its standard deviation is only a few chips, allowing the sub-optimal selection $k_{\max}$= N-L, which simplifies processing without performance loss. In this case the percentage of rejected samples in a processing block is about L/N.

III. Spectral Processing

Estimates $\hat{\mathbf{x}}_{ZF}$ and $\hat{\hat{\mathbf{x}}}_{ZF}$ require inversion of $\mathbf{H}_C$, which can be efficiently carried out in the spectral domain [2]: the spectral factorisation of the block circulant matrix $\mathbf{H}_C$ is $\mathbf{H}_C = \mathbf{W}_2\boldsymbol{\Lambda}\mathbf{W}_2^H$, where:

$\mathbf{W}_2$ = [$\mathbf{W}$, $\mathbf{0}$; $\mathbf{0}$, $\mathbf{W}$]; $\boldsymbol{\Lambda}$ = [$\boldsymbol{\Lambda}_{11}$, $\boldsymbol{\Lambda}_{12}$; $\boldsymbol{\Lambda}_{21}$, $\boldsymbol{\Lambda}_{22}$];

$\mathbf{W}$: NxN IFFT (Inverse Fourier Transform) matrix, with elements $w_{m,n} = (N)^{-1/2}\exp[j2\pi(m\text{-}1)(n\text{-}1)/N]$ (m, n=1,…, N);

$\mathbf{\Lambda}_{j,k}$: is a diagonal matrix, whose diagonal $\lambda_{j,k} = [\lambda_{j,k}(1), \ldots, \lambda_{j,k}(N)]^T$ is given by the relation $\lambda_{j,k} = N^{1/2}\,\mathbf{W}^H\,\mathbf{h}_{j,k}$,. For SISO spectral factorisation is identical to eigendecomposition.

Performing a FFT on the last part of (2) we have $\mathbf{y}_{S,f} = \mathbf{\Lambda}\mathbf{x}_{b,f} + \mathbf{v}_f$, where $\mathbf{y}_{S,f} = \mathbf{W}_2^H \mathbf{y}_S$, $\nu_f = \mathbf{W}_2^H\,\nu$ and $\mathbf{x}_{b,f} = \mathbf{W}_2^H\,\mathbf{x}_b$. The estimate of $\mathbf{x}_{b,f}$ (indicated as $\hat{\mathbf{x}}_f$) is obtained by spectral inversion:

$$\hat{\mathbf{x}}_f = \mathbf{D}\,\mathbf{y}_{S,f} = \begin{cases} \mathbf{\Lambda}^{-1}\mathbf{y}_{S,f} & \text{SM-ZF} \\ \left(\mathbf{\Lambda}^H\mathbf{\Lambda} + \delta^2\mathbf{I}\right)^{-1}\mathbf{\Lambda}^H\mathbf{y}_{S,f} & \text{SM-MSE} \end{cases}, \qquad (5)$$

Where $\delta^2 = \sigma^2/\sigma_x^2$, with $\sigma^2\mathbf{I}$ and $\sigma_x^2\,\mathbf{I}$ are the covariance matrices of noise ν and of $\mathbf{x}_k$, and SM-ZF and SM-MSE stand for Spectral MIMO Zero Forcing and Minimum mean Squared Error equaliser, respectively.

Due to the block-diagonal structure of $\mathbf{D}$, the estimation problem in (5) reduces to N frequency-wise separate estimations, each involving a 2x2 (or M x M) matrix inversion.

Once $\hat{\mathbf{x}}_f$ is available for each frequency bin, the 1st order BCSE estimate is, after IFFT: $\hat{\mathbf{x}} = \mathbf{W}_2\hat{\mathbf{x}}_f = \mathbf{W}_2\,\mathbf{D}\,\mathbf{W}_2^H\mathbf{y}_S$.

Introducing spectral factorisation of $\mathbf{H}_C$, the bias correction term $\mathbf{B}\,\hat{\mathbf{x}}$ in the second part of (4) becomes $\mathbf{B}\,\hat{\mathbf{x}} = \mathbf{W}_2\,\mathbf{D}\,\mathbf{W}_2^H\,\mathbf{H}_U\;\hat{\mathbf{x}}$. The term is efficiently computed by pre-multiplying first $\hat{\mathbf{x}}$ by $\mathbf{H}_U$ (entailing $M^2L^2/2$ complex multiplications and additions), followed by the same spectral inversion (5) employed to obtain $\hat{\mathbf{x}}$. In conclusion, we have $\hat{\hat{\mathbf{x}}} = \hat{\mathbf{x}} + \mathbf{W}_2\,\mathbf{D}\,\mathbf{W}_2^H\;\mathbf{H}_U\;\hat{\mathbf{x}}$, with $\mathbf{D}$ given by (5).

Figure 1 shows the BCSE processing. For SISO, BCSE comprises the following sequence of operations:

1. The second-step estimate contribution from block n-1 is subtracted from the received N-vector in block n. The resulting N-vector is spectrally equalised, resulting in the 1st step estimate of the transmitted length N-vector.
2. The vector above is pre-multiplied by the upper triangular matrix $\mathbf{H}_U$. This amounts to: (i) setting to zero the initial N-L components of the 1st step estimate (ii) filtering the resulting N-vector with $\mathbf{H}_U$ matrix, to obtain a vector with only the initial L components different from zero.
3. The vector above is spectrally equalised, resulting in a N-vector with all components generally different from zero.
4. Sum the vector above with the first step N-vector estimate.
5. Discard the last N-L components of the sum vector above (k_{max}= N-L is assumed).

These receiver operations are reminiscent of operations carried out at TX and RX when CP is present (as in SE-CP).

Remark: if a CP were employed (as in SE-CP), Bias Matrix would vanish and $\hat{\mathbf{x}}$, now unbiased, would suffice.

IV. Case study: UMTS HSDPA and E-UL

A. Downlink: UMTS-FDD-HSDPA

In this standard a block (also indicated as subframe), not to be confused with N-length block considered so far, comprises 7680 chips (2 ms). In UMTS-FDD HSDPA [4], RAKE performance, satisfactory for low/medium bit rates, fails to attain operative Block Error Rate values (BLER) for long time dispersion channels (e.g. ITU Vehicular Type A (VA), time spread ~2.6 μs, corresponding to 11 chips) and high bit rates. We define the operative region as BLER < 10%; good performance can be achieved by BCSE, which has been evaluated for SISO and MIMO Spatial Multiplexing with a standard compliant link-level UMTS-FDD-HSDPA simulator. In parallel the HSDPA chain has been modified to include a CP to implement a SE-CP system, which is not standard compliant, for benchmarking BCSE which does not use CP against a system using it. Performance was measured by averaging over 2000 channel realizations.

Converse to the assumption of errorless estimation of the previous (n-1)th block, employed to derive eq. (2), simulations include estimation errors.

For MIMO evaluation, 2x2 Spatial Multiplexing was adopted to determine the performance of SM-MSE (defined in (5)), V-BLAST and S-BLAST, which is an extension of the narrowband V-BLAST to wideband channels [5], [6]. MIMO channels are assumed spatially uncorrelated, and the same codes are re-used among the transmit antennas. The assumed processing window comprises 128 chips. Comparative performance, measured as BLER against Eb/No, was determined for two different modulation and coding schemes (MCS1 is QPSK with ½ rate and MCS2 is 16 QAM with ¾ rate) for a Vehicular Type A (VA) channel.

Figure 2 shows results for a SISO VA channel, MCS1, 15 transmitted codes, 8.81 mbps throughput. For operative BLER, negligible performance difference between BCSE (affected by bias) and SE-CP (not affected) results for N=W= 32 chips. The N=W=64 chips window (i.e. ~6 times the channel spread) is an acceptable tradeoff between performance and window length minimisation. Ineffectiveness of 1st step estimate for such short windows indicates that 2nd step estimate is necessary.

Figure 3 shows results for a MIMO VA channel, MCS1, 8 and 15 codes. Because of the channel time spread, V-BLAST never achieves an operative BLER, whereas both S-BLAST and SM-MSE do, with the former outperforming the latter.

In Figure 4, where a higher order modulation and coding rate (MCS 2) is adopted with respect to Figure 3, the V-BLAST performance is even worse, but again S-BLAST and SM-MSE both achieve operable BLER, with closer performance to each other.

B. Uplink: UMTS-FDD-E-UL

Figure 5 shows E-UL [7] performance for one and eight antennas at the Base Station (BS), respectively. Among the spreading factors (SF) available for the E-UL (64, 32, 16, 8, 4, 2), a SF=32 was selected. A set of Mobile Equipment (ME) with 8 or 24 total number of antennas is assumed, with 1, 3, 24, 31 employed codes, which can be re-used or not by distinct MEs. Although the maximum number of multi-codes any user can employ in a TTI is 4 (e.g. category 6 ME, Table 5.1g in [8]), here this number is raised, to demonstrate that the proposed processing performs also with multi-code configurations above the standard specifications. The variable in the abscissa accounts for the BS antenna gain.

When a single antenna is employed at the BS, classical Multi User Detection (MUD) [6] is employed; whereas when the number of antennas at BS equals to the overall number of MEs antennas, SM-MSE processing is employed, which requires significantly lower processing load than MUD. The plot labels in Figure 5 indicate:

1 – 8 MEs, 3 codes per ME, 8 BS antennas, no code reuse;
2 – 8 MEs, 3 codes per ME, 8 BS antennas, code reuse;
3 – 8 MEs, 24 codes per ME, 8 BS antennas, code reuse;
4 – 8 MEs, 31 codes per ME, 8 BS antennas, code reuse;
5 – 24 MEs, 1 code per ME, 1 BS antenna, no code reuse;
6 – 8 MEs, 3 codes per ME, 1 BS antenna, no code reuse.

In plot 4 all the 31 available codes are exploited and re-used among all MEs: MIMO processing separates the distinct ME antenna data streams, so each ME antenna can reach full-rate.

When passing from 3 to 31 (i.e. full-rate) employed (and reused) codes per ME (plots 2 and 4, respectively) a ~3 dB Eb/No loss is paid, but a ~10 times throughput increase is gained. Considering instead performance for fixed throughput, this can be attained by the full-rate case with an Eb/No gain with respect to the case of a lower number of codes per ME: so there is no Eb/No loss. For a fixed number of codes assigned to each user (plots 1 and 2, both with 3 codes per ME), if codes are not re-used there is an additional benefit in terms of Eb/No.

The performance difference between 1 and 8 BS antennas is only due to spatial diversity which arises from inadvertent (partial) exploitation of all ME-BS channels involved in BCSE processing (basically matrix inversions), which however is not designed to exploit maximal diversity gain.

C. Processing load

BCSE processing load per block is a random variable, correlated to $k_{\max}$. For comparison, the (deterministic) operation count, evaluated for OFDM with processing window of 512 samples, is found to be about 56 flops/chip whereas the average BCSE load is about 247 flops/chip (i.e. about 4 times that of OFDM) and is mainly due to the extra FFT and IFFT's, performed once (FFT) in OFDM receiver and four times in BCSE.

D. Final remarks

In the absence of CP, as a specific time structure is not present in the data, for equalisation purposes there is no special requirement on synchronism among the ME receptions.

Spectral equalisation not requiring CP has been introduced before in [2] and [3]; in both symbol hard-decision and feedback is employed.

BCSE is akin to system in [3], where the major difference is that BCSE uses two estimation steps, whereas the latter uses a single step; this is conjectured to be the reason why [3] requires a significantly longer processing window than BCSE. BCSE was derived by modifying the recursive approach in [2]; the significant difference between BCSE and [2] is that the former discards the last L samples of the 2nd step N-vector estimate, in which bias effect may be the largest.

The short BCSE processing window is beneficial in scenarios with very high mobility; estimated performance remains good for speeds up to 5000km/h, provided that channel variation can be properly tracked [6].

As a final remark, analytical investigation of why BCSE performs well for a generic channel, although not carried out so far, is deemed as important and requires future work.

V. CONCLUSIONS

A novel multi-access equaliser is presented, having the following features.

- Provides low complexity channel equalization for SISO and MIMO, achieved by spectral processing.
- It does not require a CP in the transmission format, avoiding the associated spectral efficiency reduction.
- It is suitable for a number of access systems, either using CP or not, enabling inter-working at PHY level between distinct access systems.
- Employs short processing windows (6-10 times the channel time spread), so does not restrict the maximum velocity at which it can operate.

REFERENCES

[1] 3GPP TSG RAN Future evolution workshop, Totonto, 2nd-3rd; Nov., 2004, www.3gpp.org/ftp/workshop/2004_11_RAN_Future_Evo/.
[2] Barbarossa, G. Scutari, D. Scamolla, "Effcient Space-Time Coding for Wideband CDMA Systems Using Variable-Length Cyclic Prefix", *European Wireless 2002*, Florence (Italy), Febr. 2002.
[3] D.D. Falconer, S. Lek Ariyavisitakul, "Broadband Wireless Using Single Carrier and Frequency Domain Equalization", WPMC '02, Honolulu, Oct. 2002
[4] 3GPP TR 25.858, "High Speed Dowlink Packet Access: Physical Layer Aspects (Release 5)," V5.0.0, March 2003.
[5] de Marinis E., Hart M., "Performance aspects of advanced spectral processing MIMO receivers for multi-code reception", 3rd Generation Partnership Project, 3GPP RAN1, TDOC R1-040930, meeting 38, Prague (CZ), August 16th, 2004.
[6] de Marinis E., Gasparini O. and Iarossi M., "Multi User Detection downlink/uplink algorithms for CDMA-CPE in SISO/MISO/MIMO architectures", IST-2001-32549 ROMANTIK D411-D412-D432 (parts 1 and 2), Version 1, Sept. 2003.
[7] 3GPP RTS/TSG R-0125.211 v 6.40, 25.212 v 6.40, 25.213 v 6.20, 25.214 v 6.50, 2004.
[8] 3GPP TS 25.306 v6.41, "Radio access capabilities", 2004

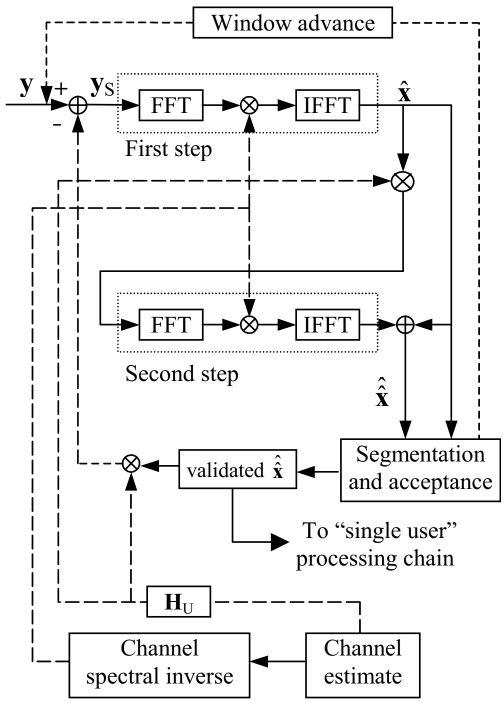

Figure 1. BCSE processing

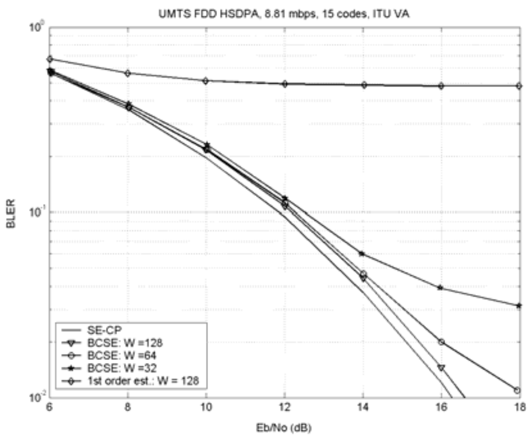

Figure 2. SISO BLER vs. Eb/N0, 16QAM, Spreading Factor (SF) 16, ITU Vehicular A channel. SE-CP and BCSE with different lengths of the sliding processing window. First step estimation is also reported.

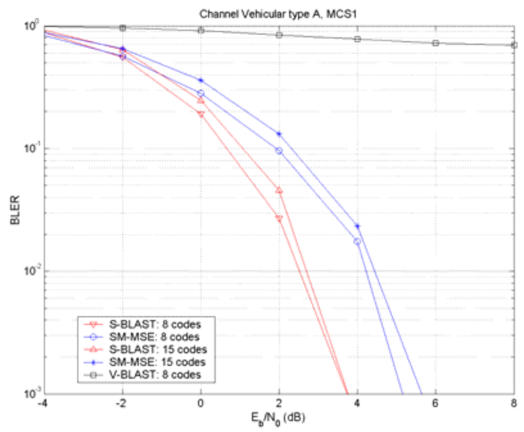

Figure 3. UMTSS FDD HSDPA, MIMO 2x2 BLER vs. Eb/N0 for MCS1 (QPSK ½ rate), Channel VA, S-BLAST, SM-MSE and V-BLAST, 3.624 and 6.795 mbps services, 8 and 15 codes respectively.

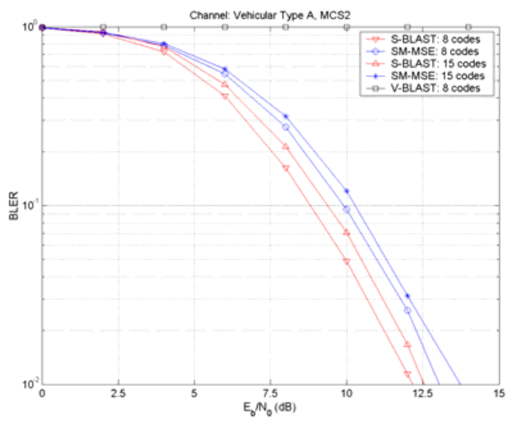

Figure 4. UMTS FDD HSDPA, BLER vs. E_b/N_0 for MCS2 (16QAM ¾ rate), Channel VA, S-BLAST, SM-MSE and V-BLAST 11.28 and 21.15 mbps services, 8 and 15 codes respectively.

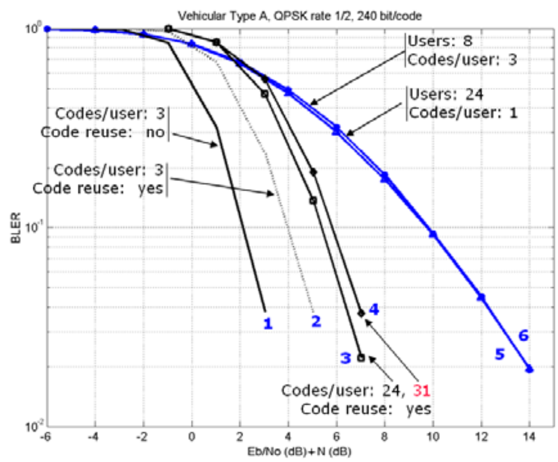

Figure 5. Uplink BLER plots. UMTS FDD E-UL, Spreading Factor SF=32, QPSK rate ½, 240 bit/code. Base Station with a single or eight antennas, with and without multi-code reuse. *N* indicates the number of antennas at the Base Station.