Joint Channel Tracking and ICI Equalisation for VBLAST-OFDM in VANET

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Abstract

VBLAST-OFDM systems can achieve high spectral efficiency in quasi-stationary links and with channel state information (CSI) matrix knowledge. Due to the high speeds of nodes in VANET, the channel is fast fading thus raising the need for channel tracking. Furthermore Inter-Carrier-Interference (ICI) causes an error floor at high SNR even with perfect CSI knowledge. In this paper we investigate channel tracking and ICI mitigation for VBLAST-OFDM. The analysis of ICI shows that it increases with speed, number of subcarriers and/or number of transmit antennas. We then introduce a simple channel tracking algorithm for VBLAST-OFDM. Simulation results show that our algorithm reduces the BER of a 2×4 VBLAST system by 10^{-2} at 40dB SNR and 100kph speed compared to obtaining a channel estimate from a training sequence only. The change in the channel response is estimated using the channel tracking algorithm and then passed to an ICI equaliser to enhance performance and reduce the error floor caused by ICI at high SNR. Equalising five pairs of subcarriers gives 4dB improvement for 2×4 VBLAST at 180kph relative speed. The performance is enhanced as more subcarriers are included in the ICI equaliser at the expense of increased receiver complexity.

I. Introduction

Vehicle Ad-hoc Network (VANET) is an emerging technology to help reduce the number of accidents on the road and provide numerous applications such as travel information, maps, route guidance and internet provision on the roads just to name a few. Standards for Vehicle to Vehicle (V2V) and Vehicle to Roadside (V2R) are still under review in Europe and the US. The current drafts for US standards allocate seven 10MHz channels at 5.9GHz for V2V and V2R communications [1, 2]. The same band is expected to be adopted in Europe [2]. Due to the high speeds of vehicles, the communication time can be limited to a few seconds, thus high and reliable data rates are desirable to meet the demands of VANET applications.

Vertical Bell Labs lAyerd Space Time (VBLAST) exploit multiple input multiple output (MIMO) systems to provide high capacities [3, 4]. To achieve these promising high data rates, VBLAST systems require knowledge of the CSI matrix and, ideally, a flat fading channel. Since Orthogonal Frequency Division Multiplexing (OFDM) converts frequency selective channels into flat fading ones by dividing the bandwidth into a large number of narrow band channels, VBLAST-OFDM systems are gaining more interest as a promising high capacity solution for future wireless networks. However, most of the current research focuses on fixed or slowly varying links, employed in personal mobile communications, where a training sequence is sufficient to estimate the CSI. Moreover, as the speed increases, the subcarriers in OFDM spread due to the Doppler shift leading to Inter Carrier Interference (ICI) [5]. In VANET the channel varies quickly, therefore using only a training sequence with VBLAST-OFDM will lead to poor performance due to inaccurate CSI matrix and increased ICI. In this paper we propose a new channel tracking algorithm for VBLAST-OFDM in VANET. We then use this algorithm to further improve the bit error rate (BER) performance by reducing the effects of ICI. Simulation results show that ICI and inaccurate CSI cause an error floor in the BER performance of VBLAST-OFDM data reception. Using the proposed channel tracking algorithm, the CSI matrix is updated thus reducing the error floor. The error floor can be further reduced when the channel tracking algorithm is combined with ICI equalisation as we propose.

The rest of the paper is organised as follows: in the next section some of the existing channel tracking and ICI reduction schemes are reviewed. Section III is a mathematical analysis of VBLAST-OFDM system and the ICI problem. The proposed channel tracking algorithm is derived in section IV followed by our proposed ICI equalisation algorithm in section V. Section VI contains simulation results and discussion. The paper is finally concluded in section VII.

II. Related Work

Channel tracking has been the subject of many research works. In [6] the authors considered the use of Kalman filtering to track the channel for orthogonal Space Time Block Coding (STBC) MIMO systems. The authors exploited the orthogonality of the codes to reduce the complexity of the filter. In [7] a maximum likelihood channel tracking algorithm has been proposed. The authors modelled the channel as an auto regressive (AR) process using Jakes' power spectral density [8]. A combination of a Kalman filter and a minimum mean square error decision feedback equaliser (MMSE-DFE) was proposed in [9] to estimate the channel. The DFE is used to estimate the transmitted symbols and its output is fed to the Kalman filter for channel estimation. A polynomial fitting then further enhances the channel prediction. In [10] an autoregressive moving average (ARMA) filter was employed to model the channel response based on Jakes' channel power spectral density, this was then used to design a Kalman filter for channel tracking.

The ICI problem was analysed in [5, 11] for single input single output (SISO) systems and algorithms to reduce ICI were introduced in [12-15] for SISO and in [16, 17] for MIMO systems. In [12] and [15] an impulse is sent after the OFDM symbol to estimate the change in the channel during the OFDM symbol. This estimate is then used in [12] to equalise ICI and in [15] to cancel it. In [14] the authors use a different technique. They divided the subcarriers into groups. Within a group, each subcarrier carries the same data multiplied by a certain weight. The weights were designed so that the ICI from the subcarriers within the same group cancel each other. This technique was extended to STBC in [17]. A sequential decision feedback sequence estimator (SDFSE) receiver for Alamouti-OFDM systems that takes into account ICI was also introduced in [16]. The complexity of Kalman filters used in previous channel tracking algorithms increases rapidly with its order. In our method we avoid this complexity by using a bank of first order Kalman filters to track the multi-channels of the MIMO system. The impulse scheme used in [12] and [15] wastes resources and is not suitable for MIMO systems since each transmit antenna will require a separate pulse. The method in [17] shows good bit error rate (BER) performance but the data rate is reduced at least by half since two, or more, subcarriers carry the same data implying a reduction in spectral efficiency. Our algorithms are applicable to BLAST as well as STBC systems and the channel tracking uses only first order Kalman filters thus reducing the channel tracking complexity.

III. Mathematical Analysis of ICI in VBLAST-OFDM

In this section we develop a mathematical analysis for ICI in VBLAST-OFDM. The results obtained in this section will be used to derive the channel update and ICI equalisation algorithms. Starting with a SISO-OFDM system, the transmitted signal ($s(n)$) is given by [18, 19]:

$$s(n) = \frac{1}{\sqrt{N}} \sum_{p=0}^{N-1} d(p) e^{\frac{j2\pi pn}{N}} \quad \ldots\ldots\ldots\ldots (1)$$

Where *N* is the IFFT size, *d*(*p*) is the data at subcarrier *p* and *n* is the time index. Throughout this paper matrices and vectors are represented by bold upper and bold lower case characters respectively. The transmitted signal is convolved with the channel response to yield the received signal (*y*(*n*)):

$$y(n)=\sum_{l=0}^{M-1} s(n-l)\,z_l(n)+w(n) \quad \ldots\ldots\ldots (2)$$

Here $z_l(n)$ is the channel response of the path with delay index *l* at time index *n*, *M* is the number of paths and *w*(*n*) is the noise sample. The signal (*x*(*k*)) at subcarrier *k* after the FFT in the receiver is given by:

$$x(k)=\sum_{n=0}^{N-1} y(n)e^{-j\frac{2\pi}{N}kn}=\sum_{n=0}^{N-1}\sum_{l=0}^{M-1} s(n-l)\,z_l(n)e^{-j\frac{2\pi}{N}kn}+\sum_{n=0}^{N-1} w(n)e^{-j\frac{2\pi}{N}kn} \quad \ldots (3)$$

Substituting (1) in (3) we get:

$$x(k)=\frac{1}{N}\left[\sum_{n=0}^{N-1}\sum_{l=0}^{M-1}\sum_{p=0}^{N-1} d(p)e^{j\frac{2\pi}{N}p(n-l)}z_l(n)e^{-j\frac{2\pi}{N}kn}\right]+W(k) \quad \ldots\ldots\ldots (4)$$

W(*k*) is the FFT of the noise. Rearranging (4) we can write:

$$x(k)=\frac{1}{N}\left[\sum_{p=0}^{N-1} d(p)\sum_{n=0}^{N-1} e^{-j\frac{2\pi}{N}kn}e^{j\frac{2\pi}{N}pn}\cdot\sum_{l=0}^{M-1} z_l(n)e^{-j\frac{2\pi}{N}pl}\right]+W(k) \quad \ldots (5)$$

In typical OFDM systems, *M* should not exceed the cyclic prefix (*CP*) [18, 19]. Since the cyclic prefix is usually a fraction of the FFT size, we assume *N* > *M* and hence the *N* point FFT of the channel response $z_l(n)$ is:

$$H(n,p)=\sum_{l=0}^{N-1} z_l(n)e^{-j\frac{2\pi}{N}pl}=\sum_{l=0}^{M-1} z_l(n)e^{-j\frac{2\pi}{N}pl}+\sum_{l=M}^{N-1} 0\,e^{-j\frac{2\pi}{N}pl} \quad \ldots\ldots (6)$$

Equation (5) then reduces to:

$$x(k)=\frac{1}{N}\sum_{n=0}^{N-1}\sum_{p=0}^{N-1} d(p)H(n,p)e^{j\frac{2\pi}{N}np}e^{-j\frac{2\pi}{N}nk}+W(k) \quad \ldots\ldots\ldots (7)$$

If we assume the channel does not vary within the OFDM symbol we have:

$$H(n,p)=H(p) \quad \ldots\ldots\ldots\ldots\ldots\ldots (8)$$

Equation (7) becomes:

$$x(k)=\frac{1}{N}\sum_{p=0}^{N-1} d(p)H(p)\sum_{n=0}^{N-1} e^{j\frac{2\pi}{N}n(p-k)}+W(k) \quad \ldots (9)$$

$$x(k)=\sum_{p=0}^{N-1} d(p)H(p)\,\delta(p-k)+W(k) \quad \ldots\ldots\ldots (10)$$

$\delta(p\text{-}k)$ is the shifted Dirac-Delta function defined as:

$$\delta(p-k)=\begin{cases}1 & p=k\\ 0 & otherwise\end{cases} \quad \ldots(11)$$

Equation (10) then clearly reduces to:

$$X(k)=d(k)H(k)+W(k)\ \ldots\ldots\ldots\ldots(12)$$

However if the channel experiences fast fading, equation (8) is not satisfied, we rewrite equation (7) as:

$$X(k)=\frac{1}{N}\sum_{n=0}^{N-1}d(k)h(n,k)+\frac{1}{N}\sum_{\substack{p=0\\p\neq k}}^{N-1}\sum_{n=0}^{N-1}d(p)h(n,p)e^{j\frac{2\pi}{N}n(p-k)}+W(k)\ \ldots\ldots(13)$$

$$X(k)=\frac{1}{N}d(k)\sum_{n=0}^{N-1}h(n,k)+\frac{1}{N}\sum_{\substack{p=0\\p\neq k}}^{N-1}d(p)\sum_{n=0}^{N-1}h(n,p)e^{j\frac{2\pi}{N}n(p-k)}+W(k)\ \ldots\ldots(14)$$

Note that the first term in (14) is the desired data multiplied by the channel response at frequency k averaged over the OFDM symbol duration while the second term is the ICI. Since the OFDM duration is usually short compared to the channel coherence time we can approximate the change in the channel with a linear equation as [5, 12]:

$$h(n,k)=h(0,k)+a(k)n\ \ldots\ldots\ldots\ldots(15)$$

Where $h(0,k)$ is the channel response at the beginning of the OFDM symbol and $a(k)$ is the change (slope) of the channel response at frequency k over one symbol period (T_s). Here we assumed the slope to be constant for the duration of an OFDM symbol. This approximation is valid for $f_d\times T_{OFDM} < 0.1$, where T_{OFDM} is the OFDM symbol duration and f_d the Doppler shift [5, 12]. Using (15) in (14) we get:

$$X(k)=\frac{1}{N}d(k)\sum_{n=0}^{N-1}\left[h(0,k)+a(k)n\right]+\frac{1}{N}\sum_{\substack{p=0\\p\neq k}}^{N-1}d(p)\sum_{n=0}^{N-1}\left[h(0,p)+a(p)n\right]e^{j\frac{2\pi}{N}n(p-k)}+W(k)\ \ldots(16)$$

$$X(k)=d(k)\left[h(0,k)+\frac{N-1}{2}a(k)\right]+\frac{1}{N}\sum_{\substack{p=0\\p\neq k}}^{N-1}d(p)h(0,p)\delta(p-k)+\frac{1}{N}\sum_{\substack{p=0\\p\neq k}}^{N-1}d(p)a(p)\sum_{n=0}^{N-1}n\,e^{j\frac{2\pi}{N}n(p-k)}+W(k)\ \ldots(17)$$

$$X(k)=d(k)\left[h(0,k)+\frac{N-1}{2}a(k)\right]+\frac{1}{N}\sum_{\substack{p=0\\p\neq k}}^{N-1}d(p)a(p)\sum_{n=0}^{N-1}n\,e^{j\frac{2\pi}{N}n(p-k)}+W(k)\ \ldots\ldots\ldots\ldots(18)$$

Equation (18) can be generalised for an $A\times B$ MIMO system as:

$$x_r(k)=\sum_{t=1}^{A}\left[d_t(k)\left[h_{tr}(0,k)+\frac{N-1}{2}a_{tr}(k)\right]+\frac{1}{N}\sum_{\substack{p=0\\p\neq k}}^{N-1}d_t(p)a_{tr}(p)\sum_{n=0}^{N-1}n\,e^{j\frac{2\pi}{N}n(p-k)}\right]+W_r(k)\ \ldots(19)$$

The subscript i represents the transmit antenna and r the receive antenna ($r \in \{1, \ldots, B\}$). Let:

$$C_{pk} = \frac{1}{N}\sum_{n=0}^{N-1} n\, e^{j\frac{2\pi n(p-k)}{N}} \quad \ldots\ldots 2)$$

The value of C_{pk} affects the contribution of the individual subcarrier in the ICI. Fig (1) and (2) show the imaginary and absolute square value of C_{pk} for k = 30 and N = 64 subcarriers. The value of k shifts the figures to the left or right without changing the shape. The real part of C_{pk} was found to be –½ regardless of the subcarrier index except when p=k, i.e. the desired subcarrier, where its value is (N-1)/2 as can be seen from equation (18). The contribution of ICI mainly comes from the few adjacent subcarriers as can be seen from Fig (1) and (2). By cancelling or equalising the ICI contribution of these subcarriers the interference can be reduced considerably thus improving the BER performance.

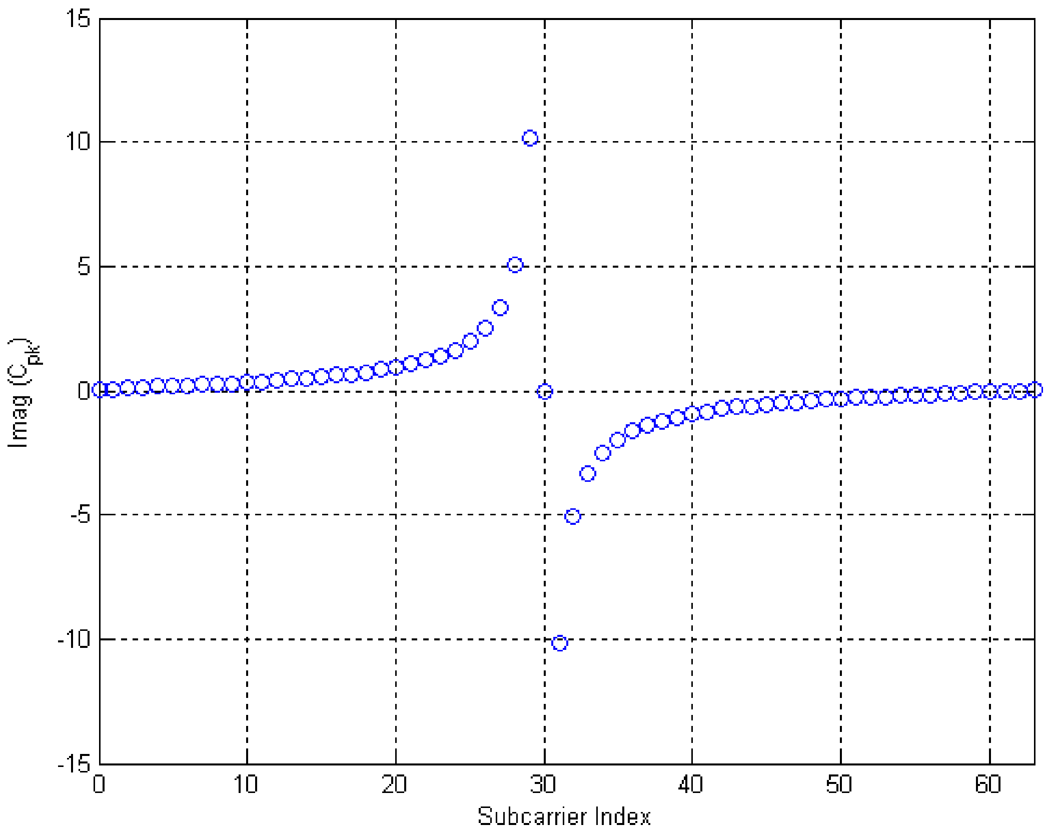

Fig (1): Imaginary part of Subcarrier Contribution (C_{pk}) to ICI

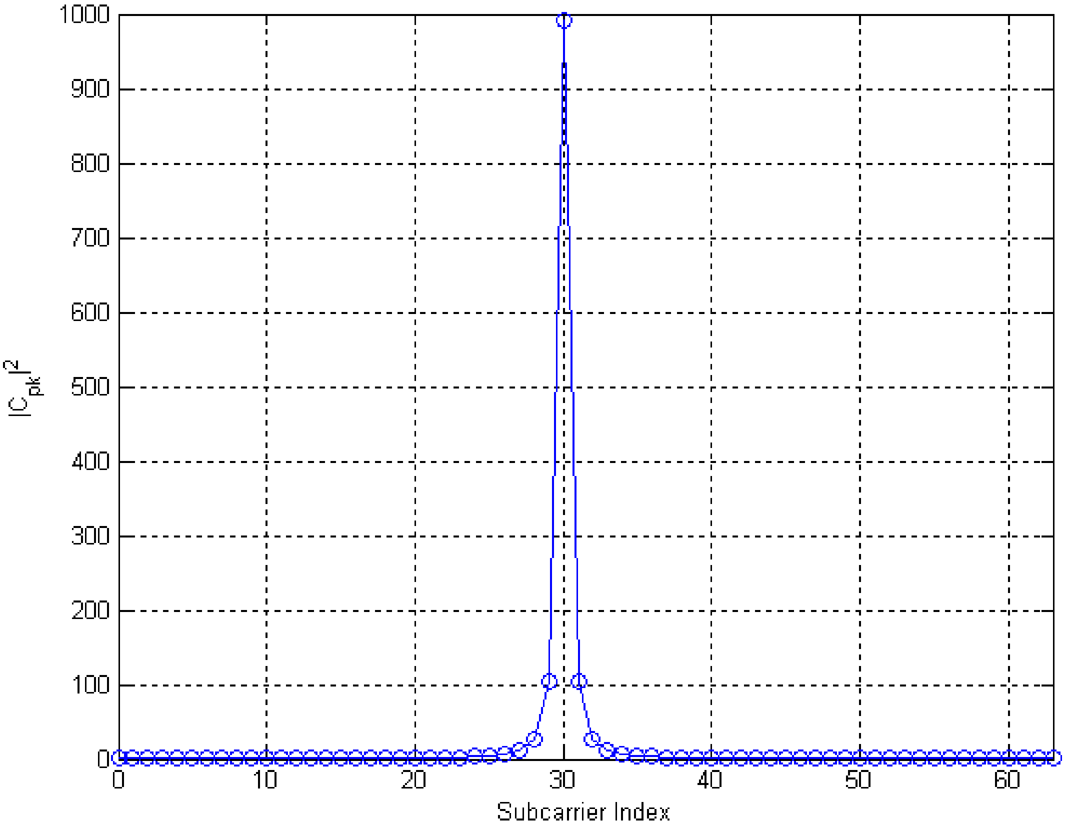

Fig (2): $|C_{pk}|^2$ vs Subcarrier Index for 64 subcarriers and k = 30

The ICI power is plotted in Fig (3) for a single transmit antenna, fixed bandwidth and various FFT sizes. As can be seen the size of the FFT has a major impact on performance. At high speeds using a small number of subcarriers gives better performance because the number of subcarriers, and therefore interferers, is small and the OFDM symbol duration is short leading to smaller channel variation within the OFDM symbol. Both factors will reduce the ICI power. The use of a small FFT size, however, leads to wider bandwidth per subcarrier and, if the coherence bandwidth of the channel is small, the subcarriers may experience frequency selective fading and inter-symbol-inference (ISI) instead of flat fading. For low speeds and fixed links using more subcarriers is feasible to ensure a flat fading channel and possibly to provide OFDM multiple access (OFDMA) to serve a larger number of users. The number of transmit antennas (A) will also cause an increase in the ICI power by $10\times\log_{10}(A)$ as shown in Fig (4) since the power in the interfering subcarriers increases.

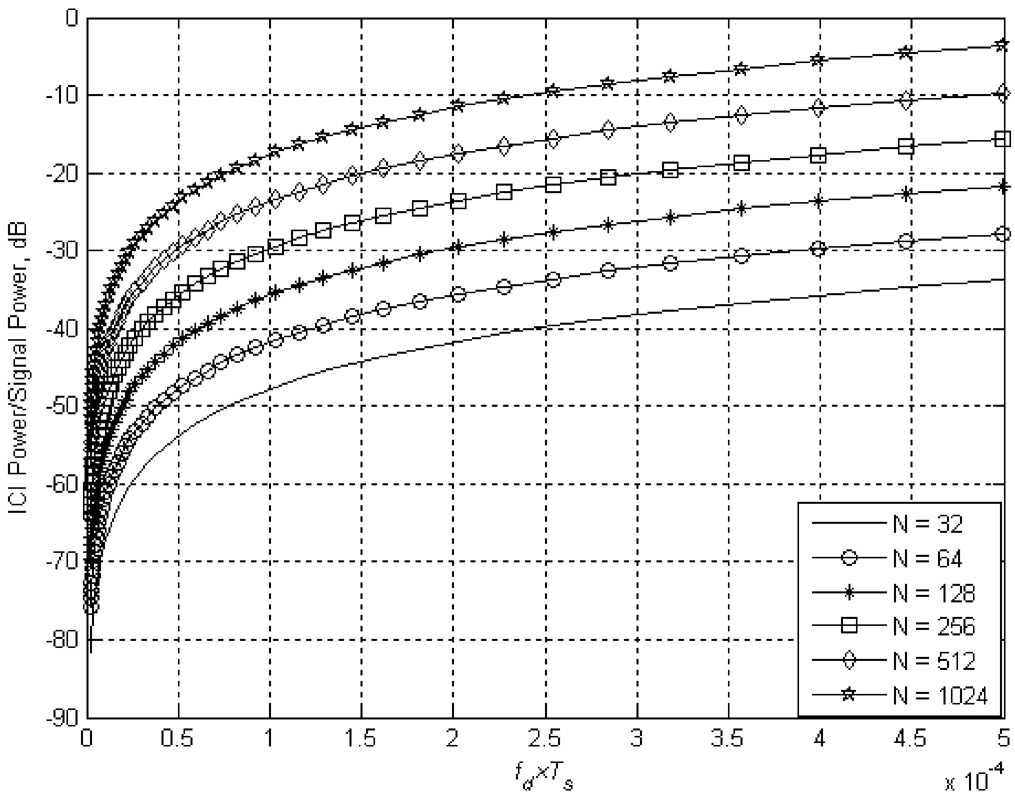

Fig (3): SISO ICI to Signal ratio vs. normalised Doppler Shift

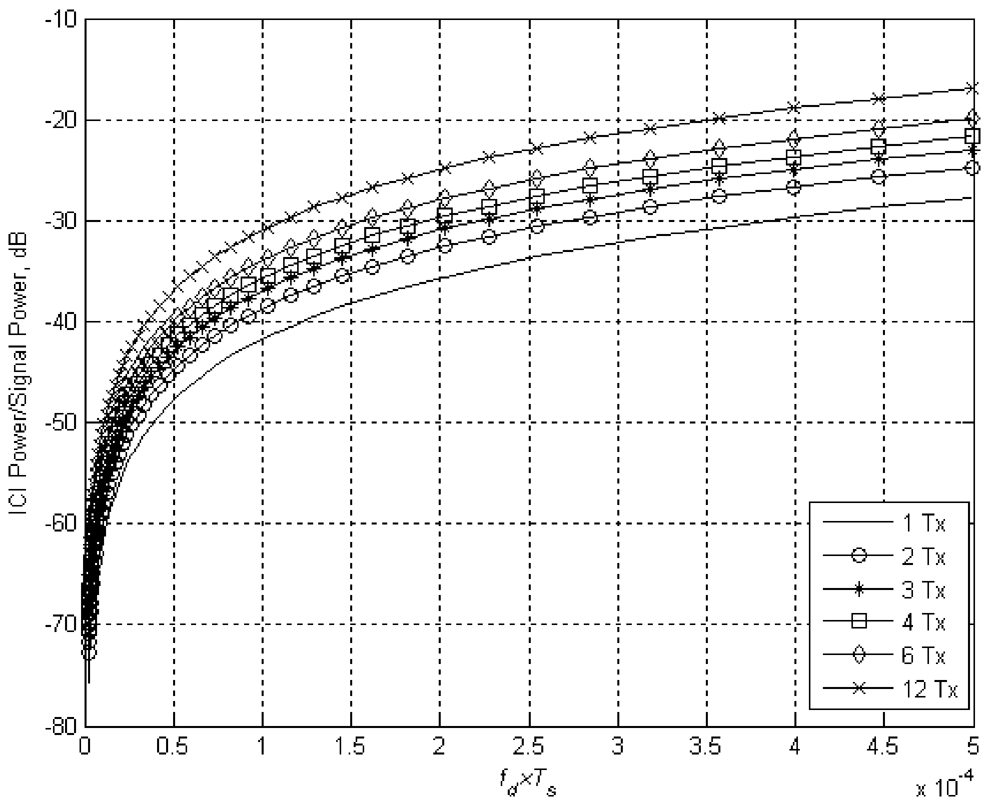

Fig (4): MIMO ICI to Signal ratio vs. normalised Doppler Shift, *N*=64

The ICI becomes dominant at high SNR leading to an error floor for large Doppler shifts even for perfect CSI knowledge. Fig (5) shows the BER performance of a 2×4 VBLAST-OFDM using QPSK, 10MHz bandwidth, 512 subcarriers and perfect channel knowledge. As can be seen the error floor, due to IC, increases with speed and consequently limits the performance of the system. Fig (6) compares the performance of the same system with 64, 128, 256 and 512 subcarriers and a speed of 180km/h. We observe from the figure that using a smaller number of subcarriers improves the BER performance of the system. This analysis of ICI in VBLAST-OFDM proves that it becomes necessary to use ICI equalisation/cancellation at high speeds, with large number of subcarriers and/or large number of transmit antennas.

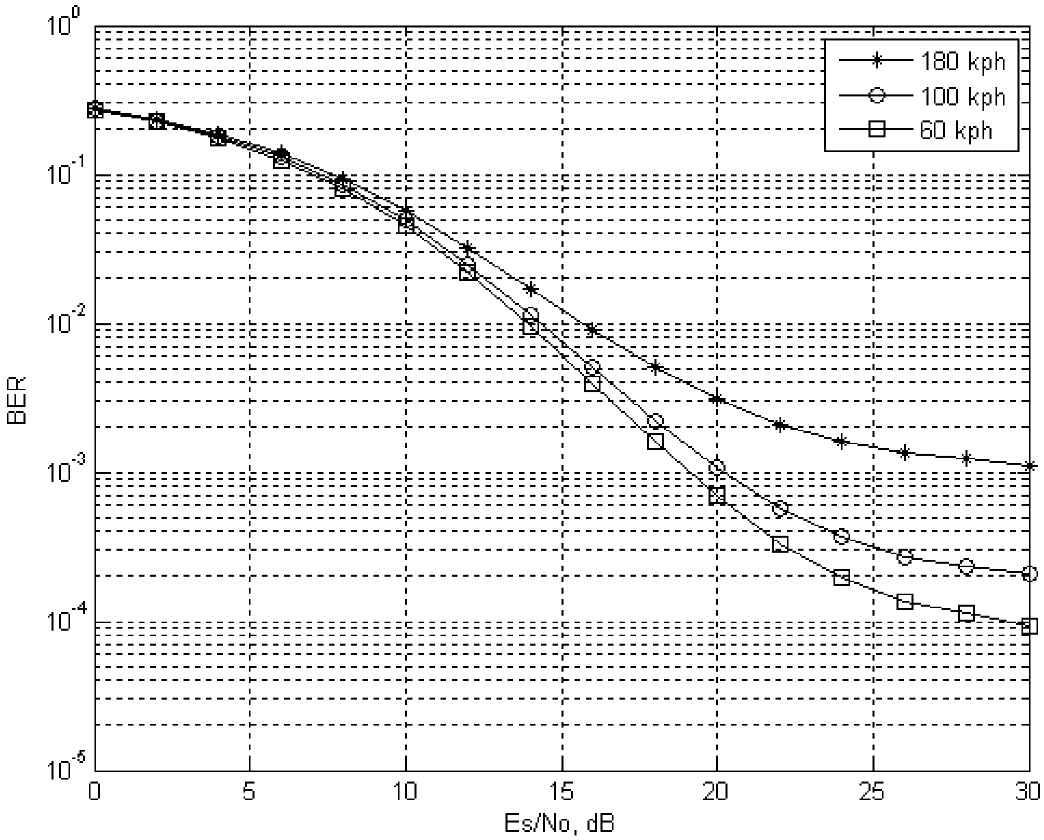

Fig (5): BER Performance of 2×4 VBLAST-OFDM for 512 subcarriers (Perfect CSI)

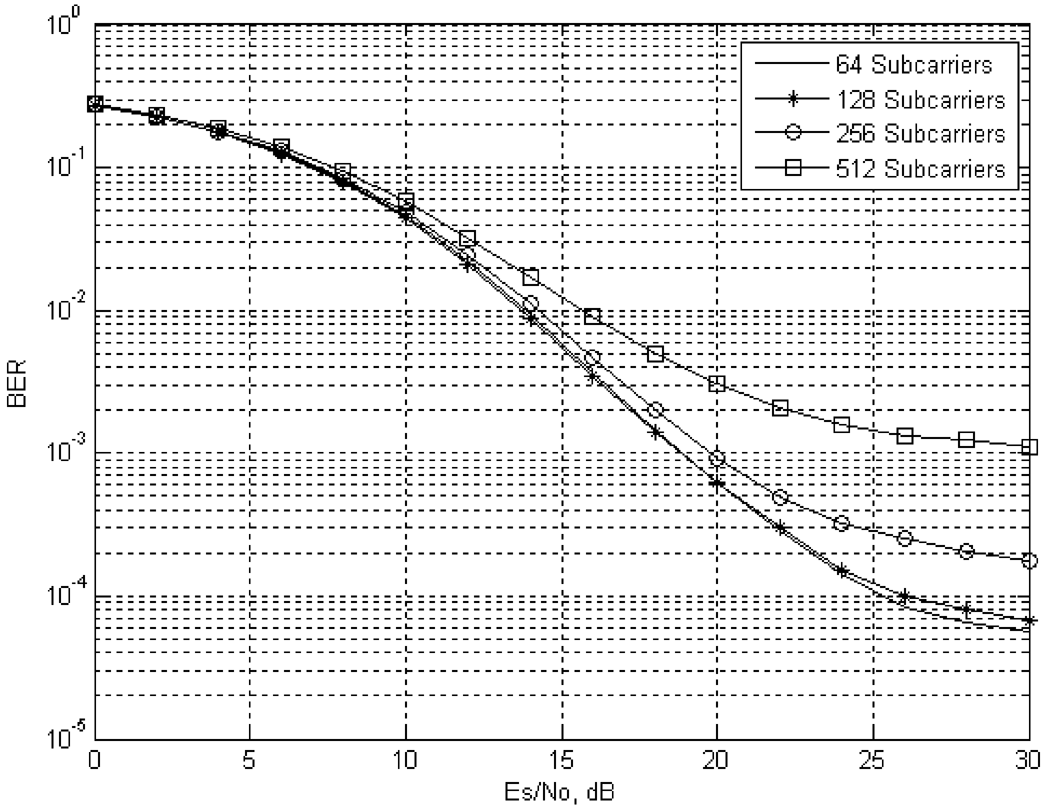

Fig (6): BER Comparison of 2×4 VBLAST-OFDM BER for 180kph (Perfect CSI)

IV. Proposed channel update algorithm

In this section we extend the channel update algorithm introduced in [20] to VBLAST-OFDM. Let:

$$h_{ri}(k)=h_{ri}(0,k)+\frac{N-1}{2}a_{ri}(k)\;\ldots\ldots(22)$$

Substituting equations (20) and (21) in (19) we may express the signal after FFT received by antenna r at subcarrier k in VBLAST-OFDM by:

$$x_r(k)=\sum_{i=1}^{A}\left[d_i(k)h_{ri}(k)+\frac{1}{N}\sum_{\substack{p=0\\p\neq k}}^{N-1}a_{ri}(p)d_i(p)C_{pk}\right]+W_r(k)\;\ldots(23)$$

The receiver then with knowledge of an estimate ($\hat{h}_{ri}(k)$) of $h_{ri}(k)$ decodes the OFDM symbol using the VBLAST algorithm and, assuming correct decoding, formulates the term:

$$x_r(F,k)-\sum_{i=1}^{A}\hat{h}_{ri}(F,k)d_i(F,k)=\sum_{i=1}^{A}\left[d_i(F,k)\varepsilon_{ri}(F,k)+\frac{1}{N}\sum_{\substack{p=0\\p\neq k}}^{N-1}a_{ri}(F,p)d_i(F,p)C_{pk}\right]+W_r(F,k)\;\ldots(24)$$

Where $\varepsilon_{ri}(F, k)$ is the error between the estimated and exact channel response for transmit antenna *i* and receive antenna *r* at subcarrier *k*. The variable *F* (OFDM symbol index) was added to indicate that these variables change between different OFDM symbols. Define **ΔH** as:

$$\Delta\mathbf{H}(F,k)=\left(\mathbf{x}(F,k)-\hat{\mathbf{H}}(F,k)\mathbf{s}(F,k)\right)\mathbf{s}^{+}(F,k)\ldots\ldots$$

Where **x**(F, k) is the length B column vector of received signal, $\hat{\mathbf{H}}(F, k)$ is the estimated channel matrix, **s**(F, k) is the length A column vector of transmitted symbols and $\mathbf{s}^{+}(F, k)$ is the length A row vector of the reciprocal of the decoded symbols, all at subcarrier k of OFDM symbol F. The element $\Delta h_{ri}(F, k)$ of the matrix $\Delta\mathbf{H}(F, k)$ at any column (transmit antenna) i and row (receive antenna) r is then:

$$\Delta h_{ri}(F,k)=\varepsilon_{ri}(F,k)+\sum_{\substack{b=1\\ b\neq i}}^{A}\frac{a_b(F,k)}{a_i(F,k)}\varepsilon_{rb}(k)+\sum_{b=1}^{A}\frac{1}{N}\sum_{\substack{p=0\\ p\neq k}}^{N-1}\frac{a_b(F,p)G_{rb}(F,p)}{a_i(F,k)}C_{pk}+\frac{W_r(F,k)}{a_i(F,k)}\ldots$$

The *i*th column vector ($\hat{\mathbf{H}}_i(F, k)$) of the estimated channel matrix ($\hat{\mathbf{H}}(F, k)$) is updated to produce the new estimate ($\hat{\mathbf{H}}_i(F+1, k)$) using the equation:

$$\hat{\mathbf{H}}_i(F+1,k)=\hat{\mathbf{H}}_i(F,k)+\gamma_i\times\Delta\mathbf{H}_i(F,k)\ldots\ldots$$

The parameter γ_i is chosen to minimise the mean square error (MSE) between the exact and estimated channel response $E|\mathbf{H}_i(F+1, k)-\hat{\mathbf{H}}_i(F+1, k)|^2$. The optimum ($\gamma_i^{opt}$) value of γ_i is calculated from (see Appendix I):

$$N_T=\sum_{\substack{b=1\\ b\neq i}}^{A}\sigma_b^2(k)+2A(\pi f_d T_s)^2\sum_{\substack{p=0\\ p\neq k}}^{N-1}\left|C_{pk}\right|^2+\frac{N_{oise}}{P}\ldots$$

$$\gamma_i^{opt}=\sqrt[3]{\frac{\left[\pi f_d T_s(N+CP)\right]^2}{N_T}}\ldots\ldots\ldots$$

$$\sigma_i^2(k)=2\left[\frac{\pi f_d T_s(N+CP)}{\gamma_i^{opt}}\right]^2+\frac{\gamma_i^{opt}}{2}N_T\ldots\ldots$$

Where P is the symbol power per antenna, N_{oise} is the noise power, T_s is the symbol rate, f_d is the maximum Doppler shift, and CP is the length of the cyclic prefix. γ_i^{opt} is calculated recursively by first setting $\sigma_i^2(k)$ (MSE) to zero and calculating an initial γ_i^{opt}. This new γ_i^{opt} is then used to update $\sigma_i(k)$. The process is then repeated again. The algorithm converges very quickly. At high SNR and high Doppler shift, the system becomes dominated by ICI. Looking at equation (22), the interference can be reduced by cancelling the contribution of detected streams and/or equalisation.

V. ICI equalisation

In [12] an equalisation technique to reduce the effects of ICI in single antenna systems was introduced. An estimate of the change (slope) in the channel is obtained by an impulse and then used to formulate a channel matrix. The signal is then decoded using the inverse of this channel matrix. In [15] a two steps decoding receiver was proposed. First an initial estimate of the transmitted symbols is obtained by an ordinary one tap equaliser. A decision feedback filter is then used to cancel the ICI from the subcarriers before decoding the signal again. In both methods the values of $a(F, k)$ are estimated by sending an impulse at the end of the OFDM symbol and calculating the difference in channel response before and after the OFDM symbol. Sending an impulse after the OFDM symbol wastes resources and is not suitable for MIMO systems since each transmit antenna will need a separate impulse. We adapt the equalisation scheme derived in [12] and extend it to VBLAST. However, instead of sending an impulse, we use our proposed channel update algorithm and the linear assumption of the change in channel (equation (15)) to calculate an estimate of $a_{ri}(F, k)$. This estimate is then used to formulate a channel matrix analogous to that of [12]. By decoding the signal using this matrix, VBLAST reduces ICI because the pseudo inverse process eliminates the contribution of the other columns of the matrix [4]. Equation (22) can be written in matrix form as:

$$\mathbf{x}=\mathbf{H}'\mathbf{d}+\mathbf{w} \quad \ldots\ldots (23)$$

x is the vector of received signals, **d** is the vector of transmitted symbols, **w** is the noise vector and **H′** is the matrix with the elements:

$$H'(r+k\times B,\, i+p\times A)=\begin{cases} h_{ri}(F,k) & \text{if } p=k \\ a_{ri}(F,p)\,C_{pk} & \text{if } p\neq k \end{cases} \quad \ldots (24)$$

for receive antenna r ($=\{1,\ldots,L\}$), transmit antenna i ($=\{1,\ldots,A\}$), and transmit subcarrier p and receive subcarrier k (p and k $=\{0,\ldots,N-1\}$). An estimate of the slope $a_{ri}(F, k)$ is given by:

$$\hat{a}_{ri}(F,k)=\frac{h_{ri}(F,k)-h_{ri}(F-1,k)}{N+CP}=\frac{[illegible]}{N+CP} \quad \ldots (25)$$

Let k be the subcarrier to be decoded and q the number of subcarrier pairs of equal interference power, one to left and another to the right of k, to be equalised. Our equalisation scheme assumes the ICI from subcarriers beyond q is zero. Define the matrix $\hat{\mathbf{H}}'(k)$ with elements:

$$G(r+v\times B,\, l+p\times A) = G_{rl}(k+v-q,\, k+p-q) = \begin{cases} h_{rl}(k+v-q), & \text{if } v=p \\ \alpha_{rl}(v,p)\, C_{k+p-q,\,k+v-q}, & \text{if } v\neq p \end{cases} \quad \ldots\ldots (33)$$

$v, p \in \{0, \ldots, 2q\}$ and $0 \leq k-q,\ k+q \leq N-1$. $\hat{\mathbf{H}}'(k)$, which is an estimated subset of $\mathbf{H}'(k)$, then becomes:

$$\hat{\mathbf{H}}'(k) = \begin{bmatrix} G_{11}(k-q,k-q) & \cdots & G_{1A}(k-q,k-q) & G_{11}(k-q,k-q+1) & \cdots & G_{1A}(k-q,k+q) \\ \vdots & & \vdots & \vdots & & \vdots \\ G_{B1}(k-q,k-q) & \cdots & G_{BA}(k-q,k-q) & G_{B1}(k-q,k-q+1) & \cdots & G_{BA}(k-q,k+q) \\ \vdots & & \vdots & \vdots & & \vdots \\ G_{11}(k,k-q) & \cdots & G_{1A}(k,k-q) & G_{11}(k,k-q+1) & \cdots & G_{1A}(k,k+q) \\ \vdots & & \vdots & \vdots & & \vdots \\ G_{B1}(k,k-q) & \cdots & G_{BA}(k,k-q) & G_{B1}(k,k-q+1) & \cdots & G_{BA}(k,k+q) \\ \vdots & & \vdots & \vdots & & \vdots \\ G_{B1}(k+q,k-q) & \cdots & G_{BA}(k+q,k-q) & G_{B1}(k+q,k-q+1) & \cdots & G_{BA}(k+q,k+q) \end{bmatrix} \quad (34)$$

and the vector of received signal:

$$\mathbf{X}(k) = [X_1(k-q) \cdots X_B(k-q)\ X_1(k-q+1) \cdots X_1(k) \cdots X_B(k+q)]^T \ldots\ldots\ldots (35)$$

Let:

$$\mathbf{D}(k) = [d_1(k-q) \cdots d_A(k-q)\ d_1(k-q+1) \cdots d_1(k) \cdots d_A(k+q)]^T \ldots\ldots\ldots (36)$$

which is the vector of symbols transmitted from the antennas at subcarriers *k*-*q* to *k*+*q*. Using equations (34) to (36) we approximate equation (30) by:

$$\mathbf{X}(k) = \hat{\mathbf{H}}'(k)\mathbf{D}(k) \ldots\ldots\ldots\ldots (37)$$

Clearly left multiplying (37) by the inverse (or pseudo inverse) of $\hat{\mathbf{H}}'(k)$ provides an estimate of $d_i(k)$ while minimising the ICI from the other subcarriers. A better approach is to use VBLAST. VBLAST decodes the symbols recursively [4], therefore we decode only the symbols at subcarrier *k*. Separate $\hat{\mathbf{H}}'(k)$ matrices should be formed for the other subcarriers. Using higher values of *q* reduces ICI by taking more subcarriers into account but also increases the size of the $\hat{\mathbf{H}}'(k)$ matrix and, therefore, the pseudo inverse and hardware complexity.

VI. Simulation Results

We simulated 2×4 and 3×4 VBLAST-OFDM systems with the elliptical channel model proposed in [20, 21] and 10MHz bandwidth with 64 subcarriers at frequency of 5.9GHz as specified in the ASTM standard for VANET communications [1, 2]. In the simulations perfect CSI is provided to the receiver every 10 OFDM symbols. This CSI is then either held constant (no update case) or updated every symbol using the derived algorithm (with update case). For the perfect CSI case, the perfect channel information at the beginning of the OFDM symbol is provided to the receiver every symbol.

Fig (7) and (8) show the BER performance with 64 subcarriers. As can be seen at low SNR the performance without update is better than when using the update algorithm. This is due to the high noise power which affects the algorithm in two ways. First the high noise is directly affecting the channel estimate of the update algorithm, second at low SNR the probability of error is higher, and since the algorithm assumes correct decoding, the estimate of the channel will not be accurate. At high SNR the BER performance of the channel update algorithm is superior to the no update case, approximately 10^{-1} and 10^{-2} at 40dB for 3×4 and 2×4 respectively. Without update, the BER drops as the speed increases due to the change in the channel and the ICI. The proposed channel update tracks the changes in the channel and takes into account the ICI, thus reducing the error and improving the BER. Both methods (i.e. with and without update) however experience an error floor due to the ICI.

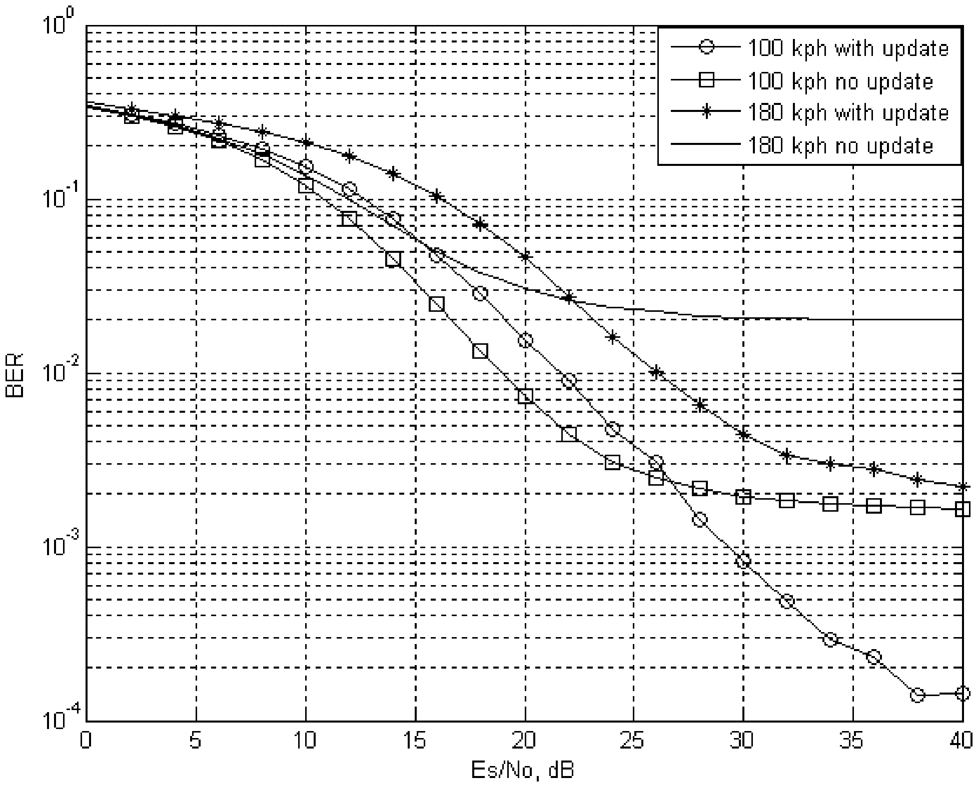

Fig (7): BER performance of 3×4 with and without channel update

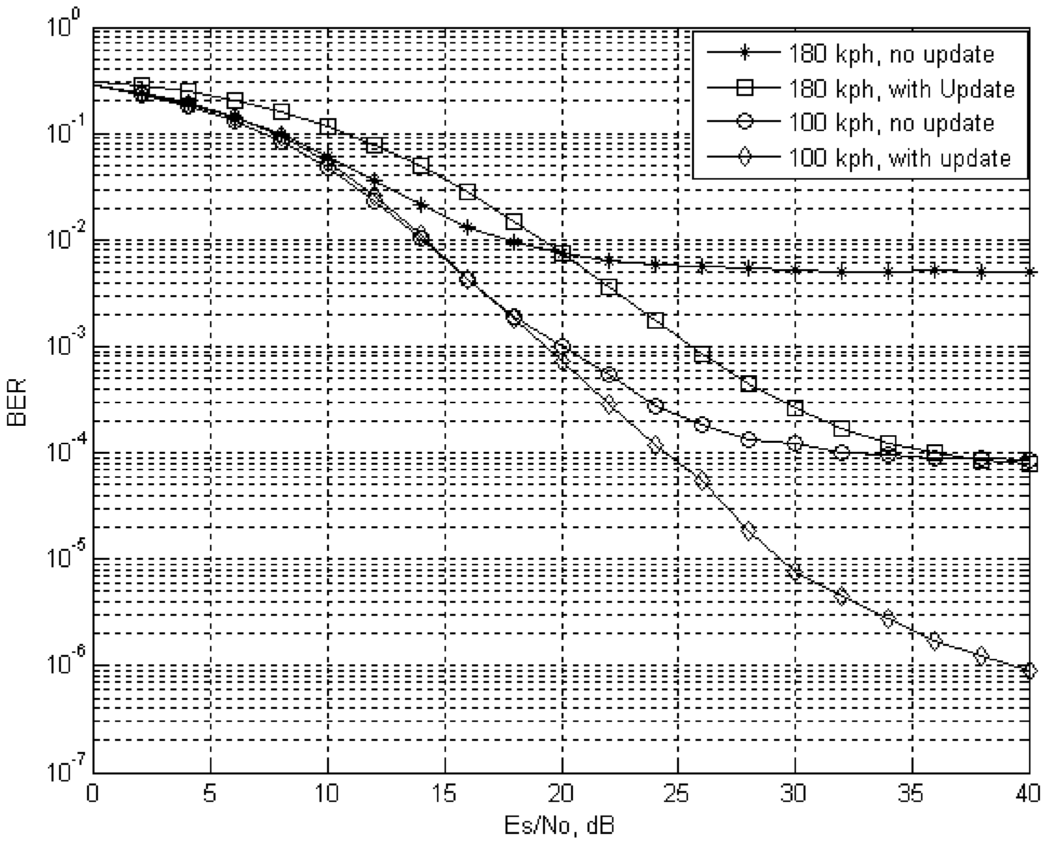

Fig (8): BER performance of 2×4 with and without channel update

Fig (9) and (10) compare the BER performance with and without the proposed channel update and ICI equalisation. As can be seen the performance improves as the number of subcarriers considered in the equaliser increases. As more subcarriers are considered, more ICI power is cancelled in the decoding process thus providing more reliable decisions. This however, comes at the price of increased receiver complexity since the size of the channel matrix becomes larger and, therefore, the pseudo inverse and decoding processes become more complicated [4]. By equalising five pairs we gain approximately 4dB for 2×4 and 3dB for 3×4 systems.

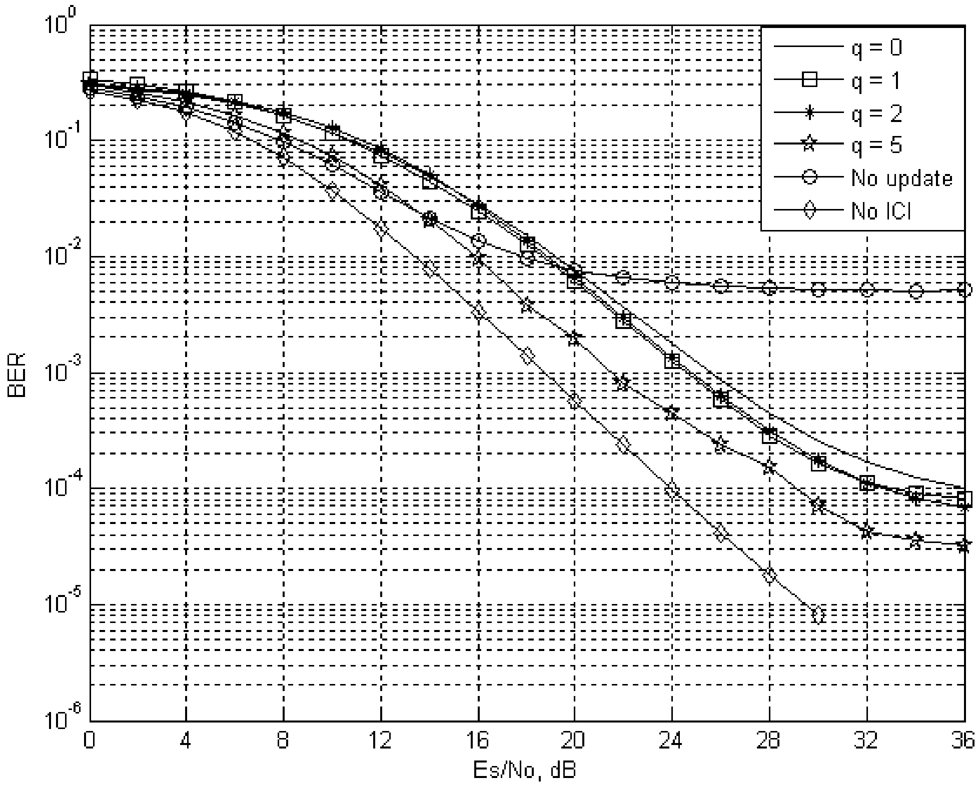

Fig (9): BER performance of 2×4 VBLAST-OFDM, 180 km/h

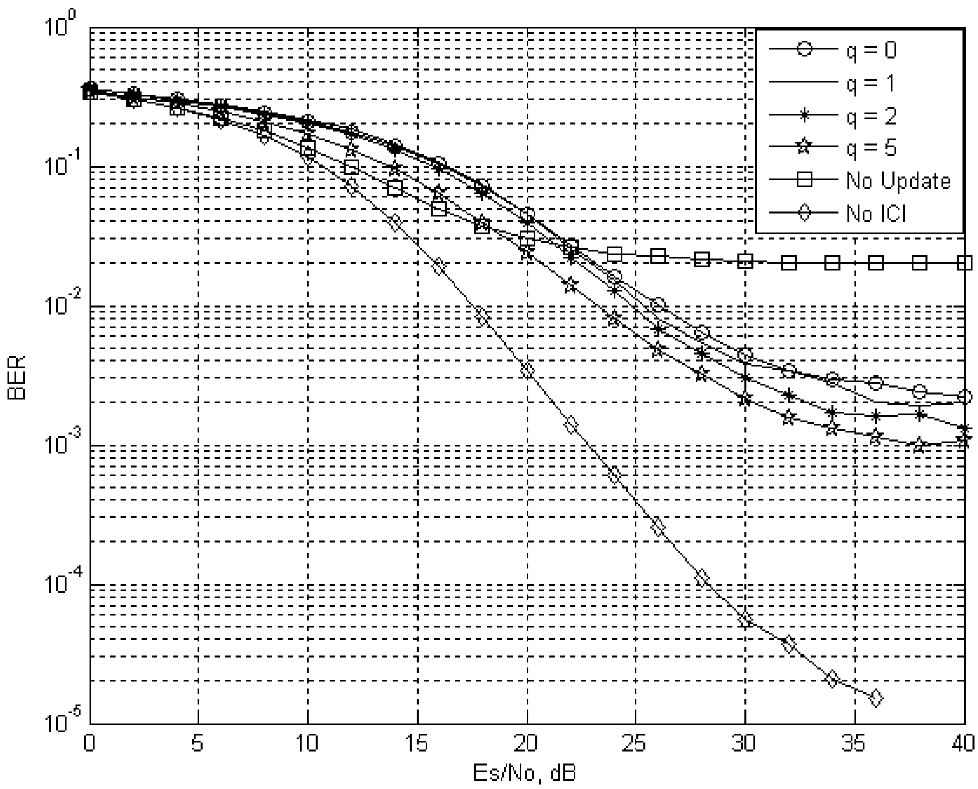

Fig (10): BER Performance of 3×4 VBLAST-OFDM, 180 km/h

VII. Conclusion

In this paper we have analysed the ICI problem and introduced algorithms for channel tracking and ICI equalisation in VBLAST-OFDM systems. ICI was found to increase with the number of subcarriers used, antennas transmitting and speed. The analysis showed that ICI causes an error floor at high SNR. A channel update algorithm was derived and it showed improvement in BER performance when compared to the ordinary training based channel estimation. The algorithm uses a bank of first order Kalman filters and thus has low complexity. An equalisation technique to reduce ICI for SISO was extended to VBLAST and evaluated. Our channel tracking algorithm updates the channel and estimates the slope of change in the channel. This estimate is then fed to the equaliser to reduce ICI. The equalisation scheme reduces the error floor as the number of subcarriers included in the equaliser increases. This, however, comes at the expense of more receiver complexity.

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Appendix I

Here we derive the optimum value (γ_i^{opt}) of the update parameter (γ_i). Equation (26) can be expanded for the elements within the matrices as:

$$h_{ri}(l+1,k) = h_{ri}(l,k) + \gamma_i \Delta h_{ri}(l,k) \quad \ldots\ldots\ldots (I.1)$$

The mean square error in the estimation is found by:

$$\begin{aligned} E\left[\left|\varepsilon_{ri}(l+1,k)\right|^2\right] &= E\left[\left|h_{ri}(l+1,k) - \hat{h}_{ri}(l+1,k)\right|^2\right] \\ &= E\left[\left|h_{ri}(F,k) + (N+CP)a_{ri}(F,k) - \hat{h}_{ri}(F,k) - \gamma_i \Delta h_{ri}(F,k)\right|^2\right] \quad \ldots (I.2) \end{aligned}$$

CP is the cyclic prefix. To calculate the mean square error we need to find the autocorrelation function of $a_{ri}(F, k)$. This is given by [5]:

$$E\left[a_{ri}(t,k)\, a_{ri}(t+\tau,k)\right] = \frac{\partial^2 AC(\tau)}{\partial \tau^2} \quad \ldots\ldots\ldots (I.3)$$

Where $AC(\tau)$ is the autocorrelation function of the channel for a delay τ, The autocorrelation function for Rayleigh fading channels is [8]:

$$AC(\tau) = J_0(2\pi f_d \tau) \quad \ldots\ldots\ldots\ldots\ldots\ldots (I.4)$$

Where J_0 is the zero order Bessel function and f_d is the maximum Doppler shift. The zero order Bessel function $J_0(x)$ can be approximated for small values of x by [5, 8, 22]:

$$J_0(x) \approx 1 - \left(\frac{x}{2}\right)^2 \quad \text{(I.4)}$$

Using this approximation, equation (I.3) for discrete time can be written as:

$$\frac{E[\alpha_i(l,k)\alpha_i^*(l+m,k)]}{\sigma_H^2} = J_0(2\pi f_d m T_s) \approx 1 - (\pi f_d m T_s)^2 \quad \text{(I.5)}$$

T_s is the symbol duration. mT_s is used instead of the delay (τ) for discrete time.

Now substituting (I.1) and (25) in (I.2) we get:

$$E\left[|\varepsilon_i(l+1,k)|^2\right] = E\left[\left|(1-\gamma)\varepsilon_i(l,k) + (1-C)\alpha_i(l,k) - \gamma\left(\sum_{u=1,u\neq i}^{A}\frac{\alpha_u(l,k)}{\alpha_i(l,k)}\varepsilon_u(l,k) + \frac{1}{N}\sum_{u=1}^{A}\sum_{p=0,p\neq k}^{N-1}\frac{\alpha_u(l,p)}{\alpha_i(l,k)}a_u(l,p)C_{pk} + \frac{W(l,k)}{\alpha_i(l,k)}\right)\right|^2\right] \quad \text{(I.7)}$$

Let:

$$g_i(l,k) = \sum_{u=1,u\neq i}^{A}\frac{\alpha_u(l,k)}{\alpha_i(l,k)}\varepsilon_u(l,k) + \frac{1}{N}\sum_{u=1}^{A}\sum_{p=0,p\neq k}^{N-1}\frac{\alpha_u(l,p)}{\alpha_i(l,k)}a_u(l,p)C_{pk} + \frac{W(l,k)}{\alpha_i(l,k)} \quad \text{(I.8)}$$

Assuming the data on each antenna and subcarrier are independent and identically distributed (i.i.d) white data with equal average power (P), we can treat the terms in (I.8) as noise with average power [22]:

$$N_T = E\left[|g_i(l,k)|^2\right] = \sum_{u=1,u\neq i}^{A}\sigma_u^2(k) + 2A(\pi f_d T_s)^2\sum_{p=0,p\neq k}^{N-1}|C_{pk}|^2 + \frac{N_{oise}}{P} \quad \text{(I.9)}$$

Where N_{oise} is the noise power and:

$$\sigma_i^2(l,k) = E\left[|\varepsilon_i(l,k)|^2\right] \quad \text{(I.10)}$$

We then rewrite (I.7) as:

$$E\left[|\varepsilon_i(l+1,k)|^2\right] = E\left[|(1-\gamma)\varepsilon_i(l,k) + (1-C)\alpha_i(l,k)|^2\right] + \gamma^2 N_T \quad \text{(I.11)}$$

This can be split into two uncorrelated errors as [22]:

$$E\left[|\varepsilon_i^C(l+1,k)|^2\right] = E\left[|(1-\gamma)\varepsilon_i^C(l,k) + (1-C)\alpha_{ik}|^2\right] \quad \text{(I.12)}$$

$$E\left[|\varepsilon_i^N(l+1,k)|^2\right] = E\left[|(1-\gamma)\varepsilon_i^N(l,k)|^2\right] + \gamma^2 N_T \quad \text{(I.13)}$$

Where ε_i^C is the error due to the change in the channel and ε_i^N is the error due to the noise and ICI. Equations (I.12) and (I.13) can be written as:

$$E\left[\left|\varepsilon_i^C(F+1,k)\right|^2\right] = E\left[(N_T-C)^2\sum_{u=0}\sum_{v=0}(1-\gamma_i)^u(1-\gamma_i)^v a_{ik}(u)\bar{a}_{ik}(v)\right] \quad (I.14)$$

$$E\left[\left|\varepsilon_i^N(F+1,k)\right|^2\right] = E\left[\gamma_i^2\sum_{u=0}\sum_{v=0}(1-\gamma_i)^u(1-\gamma_i)^v g(u,k)\bar{g}(v,k)\right] \quad (I.15)$$

($\bar{\ }$) represents complex conjugate. Since the noise and data are assumed white, for large F equation (I.15) becomes:

$$E\left[\left|\varepsilon_i^N(F+1,k)\right|^2\right] = \frac{\gamma_i}{2-\gamma_i}N_T \approx \frac{\gamma_i}{2}N_T \quad (I.16)$$

Using equation (I.6), for large F equation (I.14) becomes:

$$E\left[\left|\varepsilon_i^C(F+1,k)\right|^2\right] = 2\left(\frac{\pi f_d T_s (N_T-C)}{\gamma_i}\right)^2 \quad (I.17)$$

Combining (I.16) and (I.17) we have:

$$E\left[\left|\varepsilon_i(F+1,k)\right|^2\right] \approx 2\left(\frac{\pi f_d T_s (N_T-C)}{\gamma_i}\right)^2 + \frac{\gamma_i}{2}N_T \quad (I.18)$$

The optimum value of γ_i, denoted (γ_i^{opt}), for transmit antenna i is obtained by differentiating the mean square error (equation (I.18)) and setting the derivative equal to zero to minimise the MSE. We then find:

$$\gamma_i^{opt} = \sqrt[3]{\frac{8\left[\pi f_d T_s (N_T-C)\right]^2}{N_T}} \quad (I.19)$$

Substituting this value in equation (I.18) we get:

$$\sigma_i^2(F,k) = E\left[\left|\varepsilon_i(F,k)\right|^2\right] = 2\left[\frac{\pi f_d T_s (N_T-C)}{\gamma_i^{opt}}\right]^2 + \frac{\gamma_i^{opt}}{2}N_T \quad (I.20)$$

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