A novel channel prediction method for MIMO-OFDM in high-speed environment

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Abstract

The OFDM system with multiple transceiver antennas is attractive to increase the spectrum or energy efficiency. Depending on the channel state information (CSI), precoding techniques offer array and multiplexing gain in such systems. However, if terminal equipment is fast moving, the CSI feedback would be outdated when it is applied to data transmission. This paper proposes a novel channel prediction method applicable for high-speed environment. Benefiting from the proposed structure, this method greatly improves the prediction accuracy, and meanwhile retains the low feedback overhead. Numerical results show that the performance of the proposed method outperforms conventional precoding methods in high-speed scenarios.

1 INTRODUCTION

Massive multiple-input and multiple-output (MIMO) is well known as a promising technology that could substantially improve the capacity of wireless communication networks. In the massive MIMO context, orthogonal frequency division multiplexing (OFDM) is particularly attractive because of its robustness against multi-path propagation and low complexity of data recovery at the receiver. The joint use of massive MIMO and OFDM lays a good foundation for the fifth generation (5G) of cellular networks and a topic of interest in the research community.

In most existing applications, the channel impulse response varies slowly over several OFDM symbols. However, this assumption does not hold when the terminal moves fast. High-speed railway is such a challenging scenario where the wireless channel between the base station and the terminal varies rapidly along with time. The rapidly varying channel leads to several impacts on practical communication [1–5]. The first one is inter-carrier interference (ICI), which is caused by Doppler frequency spread and severely affects detection performance. The usually used method to cope with such impact is to implement ICI cancellation algorithms at the receiver. Most of these methods suffer from the complexity that grows more than the liner with the size of OFDM block [6–8]. The linear detection scheme implemented in the time domain has the complexity of increasing with the cubic of channel length [6]. Non-linear equalizers based on iterative ICI cancellation have been studied in [7], where the better performance is achieved at the expense of increasing complexity with the number of iterations. The frequency domain equalization technique is proposed in [8] to compensate for the fading distortion with less noise enhancement in a flat Rayleigh fading channel. However, this method assumes that the pattern of ICI, corresponding to the channel variation in time domain, is invariant for all subchannels due to the frequency non-selectivity assumption, which is not the case in a multipath fading channel. To further reduce complexity, the literatures [9, 10] studied several precoding techniques at the transmitter. By inserting redundant subcarriers in the frequency domain, ICI is limited in one frequency subband, thereby...
reducing the dimension of interference suppression. The inser-
tion of redundant subcarriers could prevent interference from
contaminating adjacent subbands at the expense of extra
resource consumption.

The second challenge in high-speed environment is channel
estimation. Conventional channel estimation in OFDM system
assumes that the channel paths vary slowly within one block.
However, in high-speed railway, wireless channels may vary
dramatically even within one OFDM block, characterized by
frequency and time selectivity [11–14]. Consequently, the
orthogonality among subcarriers is destroyed and inter-carrier
interference is introduced. The use of the basis expansion
model (BEM) has received increasing attention for its ability
to approximate the time-varying channel by linearly combining
predefined basis functions [15–20]. The research results on
channel estimation and synchronization [21–23] improved the
performance of wireless communication in high-speed railway
scenarios.

The channel ageing problem is the third challenge in the
time-varying channel. For a cellular system with a massive antenna,
network coverage can be significantly increased through trans-
mit beamforming, and network capacity can be greatly boosted
through space division multiple access (SDMA). Both applica-
tions assume that channel state information is available at the
transmitter. For the frequency division duplex (FDD) system,
one of the most frequently used methods for obtaining CSI
is terminal feedback. The terminal estimates the CSI by using
downlink reference signals such as pilots and reports the estima-
tion results to the base station [24–26]. In high-speed scenarios,
the CSI may be outdated because the time interval between CSI
measurement by the terminal and CSI application by the base
station is relatively long. Hence, it is of interest to investigate
schemes to acquire high-resolution CSI in high-speed scenarios
in order to maintain the performance of massive MIMO. In [27],
the autoregressive models such as Wiener and Kalman filtering
were proposed by exploiting the correlation of channels in the
time and frequency domain. A prediction algorithm for the
time-varying channel of MIMO system was proposed in [28].
Based on the extension of the geodesic interpolation, this algo-
rithm predicted future precoders without exploiting the channel
matrix characters. A method based on Markov chain theory
was proposed in [29]. All the above methods are applicable for
moderate terminal mobility [30]. The presented results showed
that the prediction performance decreases exponentially with
the increase of feedback delay and terminal speed.

To the best of our knowledge, the existing papers did not
jointly consider channel prediction and feedback schemes for
FDD-based massive MIMO communication systems. In such
systems, the dimension of the channel matrix becomes larger,
and direct feedback of the channel matrix would consume a
large amount of uplink resources. Hence, we expect to develop
a prediction and feedback framework to improve the prediction
performance for massive MIMO systems without increasing the
feedback burden a lot. More specifically, we first represent wire-
less channels as a linear combination of transmit and receive
steering vectors. Taking into account of both the temporal and
spatial correlations of massive closely packed arrays, the com-
bining coefficients would be much less than transmit antenna.
Reporting these coefficients instead of wireless channels could
save uplink resources. Moreover, these combining coefficients
vary much slower in the time domain, and predicting these
combining coefficients would be much easier than directly pre-
dicting wireless channels. The evaluation results show good
performance of the proposal even for high terminal mobility.
The contributions of this paper are summarized as follows:

- We propose a channel prediction and feedback framework
  for massive MIMO communication systems. The channel
  matrix is represented as a linear combination of several trans-
  mit and receive steering vectors. Steering vectors and combin-
ing coefficients are reported in wideband and subband wise,
  respectively.
- We propose to predict the combining coefficients of future
  channels by polynomial approximation. In addition to
  excellent performance, polynomial approximation adapts to
  various channel modes and moving speeds by adjusting polyno-
  mial orders, which is suitable for the channel feedback
  framework.
- Expressions for the prediction error are derived, which is
  the function of polynomial orders, normalized Doppler fre-
  quency and angular spread of channel models.
- We numerically evaluate our proposal’s performance by con-
  sidering various moving speeds and channel models. The
  results demonstrate that our proposal is able to predict future
  channels accurately and outperforms conventional precoding
  methods in high-speed scenarios.

The remainder of this paper is organized as follows. Section 2
defines a system model, which lays the foundation for derivation
and analysis. The channel prediction and feedback scheme are
described in Section 3, and the performance analysis of our pro-
posal is given in Section 4. In Section 5, the simulation results
are presented.

2 | SYSTEM MODEL

2.1 | OFDM system model on transmitter

A general OFDM symbol is composed of K subcarriers. The kth
subcarrier is used to transmit symbol $d_k$, $k = 1, 2, ..., K$, from a
finite symbols constellation. Assuming the transmitter has $N_t$
antenna, a precoder $W_k = [w_k^1, w_k^2, ..., w_k^N_t]^T$ with dimension
of $N_t \times 1$ maps data symbol $d_k$ onto $N_t$ antenna, where $w_k^n$
is the precoding coefficient of the nth transmit antenna, and the
superscript $T$ stands for transpose. Therefore, the precoding
operation is represented as $s_k^n = w_k^n \cdot d_k$, $n = 1, 2, ..., N_t$, and
$k = 1, 2, ..., K$.

On the nth transmit antenna, OFDM modulator maps the
frequency-domain symbols $s_k^n$ to time-
domain signal $X^n$ through inverse discrete Fourier transform
(IDFT) of size $M$, i.e.

$$X^n = F^{-1} S^n \quad n = 1, 2, ..., N_t$$  \hspace{1cm} (1)
where $F$ is the Fourier transform matrix and the element on the $p$th row and $q$th column is denoted as

$$F_{pq} = \frac{1}{\sqrt{N}} e^{-j2\pi pq/N} (q-1)$$  \hspace{1cm} (2)

$p = 1, 2, ..., K$, $q = 1, 2, ..., M$. The superscript $H$ stands for Hermitian transpose.

### 2.2 Channel model

Channel impulse response between a pair of transmit and receive antenna includes $L$ paths. The $l$th channel path consists of $N_l$ subpaths, which are assumed to have the same arrival timing $\tau_l$, where $l = 1, 2, ..., L$. At time $t_0$, the $n$th subpath between the $m$th receive antenna and the $l$th transmit antenna is assumed to have the complex response coefficient $\eta_{ln}^{(m,n)}$, where $m, n = 1, 2, ..., N_l$. Then the channel impulse response can be expressed as

$$h_{ln}^{(m,n)}(t_0, \tau) = \sum_{l=1}^{N_l} \eta_{ln}^{(m,n)} \delta(\tau - \tau_l)$$  \hspace{1cm} (3)

Usually, $\eta_{ln}^{(m,n)}$ and $\Phi_l^{(m,n)}$ are independent of indices $s$ and $l$, but are related to $m$ and $n$ depending on the locations of the transmitter and receiver antenna.

When the terminal travels at a speed $v$, the Doppler frequency modulated on the $n$th subpath of the $l$th channel path is

$$f_{ln} = \frac{v}{c} \cos \psi_{ln}^{(m,n)}$$  \hspace{1cm} (4)

where $c$ is the wave propagation speed and $f_l$ is the carrier frequency. $\psi_{ln}^{(m,n)}$ represents the angle between the arrival subpath and the travel direction of the terminal.

Taking time $t_0$ as a reference, the channel response of the $l$th channel path between a pair of receive and transmit antenna $(m, n)$, i.e. the $n$th receive antenna and the $l$th transmit antenna, can be represented as

$$h_{ln}^{(m,n)}(t, \tau) = \sum_{l=1}^{N_l} \eta_{ln}^{(m,n)} e^{j2\pi f_{ln}^{(m,n)}(t-t_0)} \delta(\tau - \tau_l)$$  \hspace{1cm} (5)

Considering an OFDM communication system with $N_t$ transmit antenna and $N_r$ receive antenna, the channel response matrix on a subcarrier $k$ is

$$H_k(t) = \begin{bmatrix} H_k^{1,1} & H_k^{1,2} & \cdots & H_k^{1,N_r} \\ H_k^{2,1} & H_k^{2,2} & \cdots & H_k^{2,N_r} \\ \vdots & \vdots & \ddots & \vdots \\ H_k^{N_r,1} & H_k^{N_r,2} & \cdots & H_k^{N_r,N_r} \end{bmatrix}$$  \hspace{1cm} (6)

where $H_k^{m,n}$ represents the frequency effect of multipath propagation between a pair of receive and transmit antenna $(m, n)$ on a subcarrier $k$ at a moment $t$, which is expressed as

$$H_k^{m,n} = \sum_{l=1}^{L} h_l^{(m,n)}(t, \tau) e^{-j2\pi / \tau_t}$$  \hspace{1cm} (7)

### 2.3 OFDM system model on receiver

At the receiver, when the length of cyclic prefix attached to each OFDM symbol is longer than the maximum delay of the channel, the inter-symbol interference can be avoided. After CP removal and DFT, the received signal on the $k$th subcarrier can be denoted as

$$Y_k(t) = H_k(t) \times W_k \times d_k + U_k$$  \hspace{1cm} (8)

where $H_k(t) \in \mathbb{C}^{N_r \times N_t}$ is the channel matrix in frequency domain, and $U_k \in \mathbb{C}^{N_t \times 1}$ is the additive Gaussian noise vector with zero mean and covariance $\sigma_u^2 I_{N_t}$, i.e. $U_k \sim N(0, \sigma_u^2 I_{N_t})$.

The data symbols $d_k$ can be estimated by using conventional detection algorithms such as zero-forcing (ZF) and minimum mean square error (MMSE).

### 3 PROPOSED CHANNEL PREDICTION AND FEEDBACK MECHANISM

We assume $W_k$ as the energy-constrained beamforming vector for the $k$th subcarrier, i.e. $\|W_k\|_2 \leq 1$. If the base station acquired channel frequency response $H_k$, the optimal beamforming vector $W_k$ should be designed to maximize the signal-to-noise ratio (SNR) on the terminal side. It is well known that when $W_k$ equals the dominant unit-norm right singular vector of $H_k$, the SNR on the receiver can be maximized. The singular value decomposition of $H_k$ is given by $H_k = U \Lambda W$, where $U$ and $W$ are $N_r \times N_r$ and $N_t \times N_t$ unitary matrices of left and right singular vectors, respectively. The precoder vector $W_k$ is one column of $W$ corresponding to the maximal singular value.

In time division duplex (TDD) communication systems, downlink and uplink work on the same frequency. The reciprocity between downlink and uplink channels usually holds. The downlink channel response $H_k$ can be obtained by estimating the uplink channel. While in FDD systems, downlink and uplink operate on different frequencies. It is usually impossible to acquire CSI by exploiting channel reciprocity. The terminal needs to feedback channel response or enable CSI acquisition. When the base station is equipped with massive antenna and applied to scenarios with fast-moving terminals, two key factors have to be taken into account, i.e. the payloads carrying the CSI and the CSI ageing problem when the base station decides to apply it for data transmission.

To solve the above problems, we propose a method that compensates the CSI variation by channel prediction. By applying this method, the future CSI could be accurately acquired without increasing feedback burden a lot, thereby resulting in a better transmit rate.
3.1 Derivation of the proposed method

The channels between antenna are usually correlated because the antenna spacing at the base station and terminal is relatively close, e.g. 0.5λ. For a subpath, taking the response on antenna pair (1, 1) as reference, i.e. \( \eta_{l,s} = \eta_{l,s}^{(1,1)} \), the channel responses on other antenna pairs can be represented by the multiplication of reference and coefficients determined by antenna locations, e.g. \( \eta_{l,s}^{(m,a)} = \eta_{l,s} \exp(i \varphi_{l,s}^{(m,a)} + \varphi_{l,s}^{(a)}) \), where

\[
\varphi_{l,s}^{(m)}(m) = 2\pi \frac{d_{l,s}}{\lambda} (m - 1) \sin \theta_{l,s}^{(\infty)}
\]

\[
\varphi_{l,s}^{(a)}(n) = 2\pi \frac{d_{l,s}}{\lambda} (n - 1) \sin \theta_{l,s}^{(\infty)}
\]

denote the relative phases on the \( m \)th receive antenna and the \( n \)th transmit antenna. \( \theta_{l,s}^{(\infty)} \) and \( \theta_{l,s}^{(\infty)} \) stand for the angle of departure from the transmitter and arrival to the receiver, respectively. The antenna spacing of the transmitter and receiver are represented by \( d_{l,s}^{\infty} \) and \( d_{l,s}^{\infty} \).

As a result, Equation (5) can be rewritten as

\[
h_{l,s}^{(m,a)}(t, \tau) = \sum_{i=1}^{N_{l}} \eta_{l,s} \exp(i 2\pi f_{i}(t-\tau)) \exp(i \varphi_{l,s}^{(m,a)} + \varphi_{l,s}^{(a)}) \delta(\tau - \tau_{i})
\]

\[
= \sum_{i=1}^{N_{l}} a_{l,s}(t)(m) \mathbf{u}_{l,s}(m) \mathbf{v}_{l,s}^{T}(n) \delta(\tau - \tau_{i})
\]

where \( a_{l,s}(t) = \eta_{l,s} \exp(i 2\pi f_{i}(t-\tau_{i})) \) denotes the time-varying complex gain of the \( m \)th subpath of the \( h \)th channel path. \( \mathbf{u}_{l,s}(m) = \exp(i \varphi_{l,s}^{(m,a)}) \) and \( \mathbf{v}_{l,s}(n) = \exp(i \varphi_{l,s}^{(a)}) \) correspond to the \( m \)th and \( n \)th element of receive and transmit steering vectors, respectively. This equation is the extended Saleh-Valenzuela geometric model [31].

Correspondingly, the channel of the \( k \)th subcarrier at time \( t \) is

\[
H_{k}^{m,a}(t) = \sum_{i=1}^{L} h_{l,s}^{(m,a)}(t, \tau) e^{-j2\pi f_{i} \tau_{t}}
\]

\[
\approx \sum_{i=1}^{L} \sum_{l,s=1}^{N_{l}} a_{l,s}(t)(m) \mathbf{u}_{l,s}(m) \mathbf{v}_{l,s}^{T}(n) e^{-j2\pi f_{i} \tau_{t}}
\]

Then channel responses of all antenna pairs construct the channel matrix corresponding to the \( k \)th subcarrier, which is expressed as

\[
H_{k}(t) = \begin{bmatrix}
H_{k}^{1,1}(t) & H_{k}^{1,2}(t) & \cdots & H_{k}^{1,N_{l}}(t) \\
H_{k}^{2,1}(t) & H_{k}^{2,2}(t) & \cdots & H_{k}^{2,N_{l}}(t) \\
\vdots & \vdots & \ddots & \vdots \\
H_{k}^{N_{l},1}(t) & H_{k}^{N_{l},2}(t) & \cdots & H_{k}^{N_{l},N_{l}}(t)
\end{bmatrix}
\]

\[
= \sum_{l,s=1}^{N_{l}} \sum_{i=1}^{L} a_{l,s}(t) e^{-j2\pi f_{i} \tau_{t}} \mathbf{u}_{l,s} \mathbf{v}_{l,s}^{T}
\]

Equation (13) lays the foundation for channel prediction. Obviously, within a relatively short time, \( \mathbf{u}_{l,s}, \mathbf{v}_{l,s}, \) and \( e^{-j2\pi f_{i} \tau_{t}} \) can be treated as time-invariant, and variable \( a_{l,s}(t) \) determines the time-varying characteristics of the channel. According to the expression \( a_{l,s}(t) = \eta_{l,s} \exp(i 2\pi f_{i}(t-\tau_{i})) \), the phase of \( a_{l,s}(t) \) varies linearly along time \( t \), while the amplitude of \( a_{l,s}(t) \) remains constant. In other words, if each \( \mathbf{u}_{l,s}, \mathbf{v}_{l,s}, \) and \( e^{-j2\pi f_{i} \tau_{t}} \) are perfectly known, channel prediction would be very simple, i.e. \( a_{l,s}(t) \) at each time instant is estimated by linear interpolation based on most recently acquired channels. Then future channel \( H_{k}(t) \) is obtained via Equation (13).

However, in practice, due to the limited aperture of antenna array, the angular resolution is insufficient to distinguish these channel subpaths. One alternative is to divide these subpaths into \( N_{g} \) clusters according to their departure or arrival directions. The \( g \)th cluster contains the subpaths with departure angles ranging from \( \theta_{l,s}^{(\infty)} - \frac{\Delta \theta}{2} \) to \( \theta_{l,s}^{(\infty)} + \frac{\Delta \theta}{2} \), and arrival angles ranging from \( \theta_{l,s}^{(\infty)} - \frac{\Delta \theta}{2} \) to \( \theta_{l,s}^{(\infty)} + \frac{\Delta \theta}{2} \). \( \Delta \theta_{k} \) and \( \Delta \theta_{g} \) stand for the central angles of departure and arrival subpaths of the \( g \)th cluster. \( \Delta \theta_{k} \) and \( \Delta \theta_{g} \) are the spatial angular spreads determined by beamwidth that transmitter and receiver can form. Based on the above assumptions, Equation (13) is rewritten as

\[
H_{k}(t) = \sum_{g=1}^{N_{g}} \sum_{l,s=1}^{N_{l}} a_{l,s}(t) e^{-j2\pi f_{i} \tau_{t}} \mathbf{u}_{l,s} \mathbf{v}_{l,s}^{T} = \sum_{g=1}^{N_{g}} \mathbf{b}_{k}^{g}(t) \mathbf{u}_{l,s} \mathbf{v}_{l,s}^{T}
\]
where

\[ \hat{c}_{g,p}^k = \sum_{l \in \text{(cluster g)}} \frac{n_{Lc}}{p} e^{-j2\pi f_{lt} f_{0}^p} \]  

(18)

In practical systems, it is unnecessary to represent \( \hat{b}_{g,p}^k(t) \) with infinite polynomial orders. Given the polynomial order \( P \), the channel at the future time \( t \) can be predicted as

\[ H_g(k) \approx \sum_{g=0}^{N_g-1} \sum_{p=0}^{P-1} \hat{c}_{g,p}^k(t-t_0)^p u_g v_p^T \]  

(19)

Precise estimate of \( u_g \) and \( v_p \) can be achieved via conventional super resolution DoA estimation methods, such as multiple signal classification (MUSIC) or spatial filtering. Moreover, estimates of \( u_g \) and \( v_p \) can be wideband wise, while the calculation of \( c_{g,p}^k \) should be as per subcarrier theoretically. Considering the flatness of frequency channel response, one \( c_{g,p}^k \) actually applies to several consecutive subcarriers. Hence, one way to reduce the overhead of feedback \( \hat{c}_{g,p}^k \) is to group several consecutive subcarriers into one subband, and report one set of \{ \( c_{g,p}^k \) \} for each subband.

In summary, the proposed channel prediction and reporting process include the following steps:

S0: Receive downlink pilot at time \( t_0, t_1, \ldots, t_{P-1} \), and estimate the channel matrix \( H_k(t_0), H_k(t_1), \ldots, H_k(t_{P-1}) \) for \( k = 1, 2, \ldots, K \).
S1: Determine the \( N_g \) transmit and receive steering vectors. In this step, the correlation matrices are first calculated as \( R_1 = \sum_{k=1}^{K} \sum_{l=0}^{P-1} (H_k(t))^H H_k(t) \) and \( R_2 = \sum_{k=1}^{K} \sum_{l=0}^{P-1} H_k(t)(H_k(t))^H \). Then based on \( R_1 \) and \( R_2 \), the \( N_g \) central departure angles \( \Theta_k^d \) and arrival angles \( \Theta_k^a \) are calculated by MUSIC or spatial filtering. Finally, the steering vectors \( u_g \) and \( v_p \) are constructed, where \( g = 1, 2, \ldots, N_g \).
S2: Divide the frequency responds \( H_k(t) \) into \( S \) subbands, where each subband includes \( G \) subcarriers, and \( G \) is determined based on the flatness of the frequency channel. The channel of the middle subcarrier of each subband represents the associated subband.
S3: For each subband, calculate \( \hat{b}_{g,p}^k(t) \) according to Equation (14), where \( g = 1, 2, \ldots, N_g, l = l_0, l_1, \ldots, l_{P-1} \). Then train the coefficients \( \hat{c}_{g,p}^k \) for each subband, where \( P \) is the order of polynomial approximation, \( k = \frac{G - 3G}{2}, \ldots, (S - 1)G + \frac{G}{2} \).
S4: Report the \( N_g \) transmit and receive steering vectors \( u_g \) and \( v_p \) and coefficients \( \hat{c}^k_{g,p} \) for each subband, \( g = 0, 1, \ldots, N_g - 1, k = \frac{G - 3G}{2}, \ldots, (S - 1)G + \frac{G}{2} \).
S5: The base station predicts the future channel using Equation (19) and calculates the precoding vectors for data transmission.

### 3.2 Overhead of channel feedback

In our proposal, the feedback of steering vectors \( u_g \) and \( v_p \) can be achieved by reporting each angle of departure and arrival, i.e. \( \Theta_k^d \) and \( \Theta_k^a \). Assuming each angle is quantized at a resolution of 1 degree, reporting one angle requires 8 bits. Hence, the payload required to give the feedback \( \Theta_k^d \) and \( \Theta_k^a \) is 16 \( N_g \). In addition, we assume that each \( \hat{c}_{g,p}^k \) is quantized with 8 bits. Then the payload is \( 8PN_gS \) given \( P \) polynomial order and \( S \) subbands. Thus, the total overhead of the proposed channel feedback scheme is \( 16N_g + 8PS \).

As a comparison, we show the payload of direct feedback of channel matrices as well. There are a total of \( S \) channel matrices to be reported. One channel matrix has the dimension \( N_h \times N_h \). Feedback of each matrix element requires 8 bits, and \( 8N_h^2 \) bits are required for each matrix. There are total \( S \) subbands to be reported. Thus, the total payload would be \( 8SN_h^2 \). Considering the case of \( N_h = 16, N_h = 8, N_g = 4, P = 3 \), and \( S = 10 \), the overhead of directly reporting channel matrices is about 10 times over the proposal.

### 3.3 Calculation complexity

We detail the calculation complexity for the proposed channel prediction method. In step S1, the calculation of correlation matrices has the complexity of \( P \times K \times O((N_g)^2 + (N_h)^2) \). In step S2, the complexity is about \( O((N_g)^2 + (N_h)^2) \) assuming MUSIC algorithm is adopted to estimate angles of departure and arrival. The computational complexity to estimate the combing coefficients in step S3 is \( P \times G \times O((N_g)^2) \). The step of calculating polynomial coefficients has the complexity of \( N_g \times G \times O((P)^2) \). Taking into account \( N_g = 16, N_h = 8, N_g = 4, P = 3 \), and \( S = 10 \), the above computational complexity is affordable for 5G terminals.

### 4 PERFORMANCE ANALYSIS

The time-varying channel is approximated by \( P \) order polynomial

\[ H_k(t) \approx \sum_{g=0}^{N_g-1} \sum_{p=0}^{P-1} \hat{c}_{g,p}^k(t-t_0)^p u_g v_p^T \]

where \( \hat{c}_{g,p}^k = \sum_{l \in \text{(cluster g)}} \frac{n_{Lc}}{p} e^{-j2\pi f_{lt} f_{0}^p} \), and \( f_{lt} = \frac{\nu}{c} \cos \psi_{lt} \)

represents the Doppler frequency generated by the \( l \)th subpath of the \( k \)th channel path. Let \( \hat{c}_{g,p}^k = \eta_{Lc} e^{-j2\pi f_{lt} f_{0}^p} \), we have

\[ \hat{c}_{g,p}^k = \sum_{l \in \text{(cluster g)}} \frac{n_{Lc}}{p} e^{-j2\pi f_{lt} f_{0}^p} \cos \psi_{lt}^p \]  

(20)
The covariance of $\phi_{g,p1}(\phi_{g,p2})$ is

$$E\left\{\phi_{g,p1}(\phi_{g,p2})^*\right\} = \left(\frac{j2\pi f_d}{c}\right)^{p_1+p_2} \times \sum_{l_1 \in \text{(cluster \_g \_p1)}} \sum_{s_1 \in \text{(cluster \_g \_p1)}} E\left\{|\phi_{l_1,s_1}|^2\right\}, \text{ if } l_1 = l_2 \text{ and } s_1 = s_2$$

$$E\left\{\phi_{g,p1}(\phi_{g,p2})^*\right\} = \left\{E\left\{|\phi_{l_1,s_1}|^2\right\}, \text{ if } l_1 = l_2 \text{ and } s_1 = s_2 \right. \nonumber \left. \quad 0, \quad \text{ others} \right\}$$

(21)

Since $\eta_{l_1,s_1}^k$ and $\eta_{l_2,s_2}^k$ are independent, the covariance of $\eta_{l_1,s_1}^k$ and $\eta_{l_2,s_2}^k$ is

$$E\left\{\eta_{l_1,s_1}^k(\eta_{l_2,s_2}^k)^*\right\} = \left\{E\left\{|\eta_{l_1,s_1}|^2\right\}, \text{ if } l_1 = l_2 \text{ and } s_1 = s_2 \right. \nonumber \left. \quad 0, \quad \text{ others} \right\}$$

Generally, $\eta_{l_1,s_1}^k$ is assumed to be uniformly distributed within $\{\psi_g - \frac{\Delta \theta}{2} \leq \psi_k \leq \frac{\Delta \theta}{2}\}$. When $\Delta \theta$ is small, $\cos \psi_{l_1,s_1}$ is approximated by $\cos \psi_g + (\psi_{l_1} - \psi_g) \sin \psi_g$. Therefore, the covariance of $\cos \psi_{l_1,s_1}$ is

$$E\left\{\cos \psi_{l_1,s_1}(\cos \psi_{l_2,s_2})^*\right\} \approx E\left\{|\cos \psi_{l_1,s_1}|^2\right\}$$

$$= \frac{2}{\Delta \theta (p_1 + p_2 + 1) \sin \psi_g} \times \left(\left(\cos \left(\psi_g + \frac{\Delta \theta}{2}\right)\right)^{p_1+p_2+1} - \left(\cos \left(\psi_g - \frac{\Delta \theta}{2}\right)\right)^{p_1+p_2+1}\right)$$

As a result

$$E\left\{\phi_{g,p1}(\phi_{g,p2})^*\right\} = \left(\frac{j2\pi f_d}{c}\right)^{p_1+p_2} \frac{\sigma_s^2}{\Delta \theta (p_1 + p_2 + 1) \sin \psi_g} \times \left(\left(\cos \left(\psi_g + \frac{\Delta \theta}{2}\right)\right)^{p_1+p_2+1} - \left(\cos \left(\psi_g - \frac{\Delta \theta}{2}\right)\right)^{p_1+p_2+1}\right)$$

From the derivation, the covariance of $\phi_{g,p}$ decreases significantly with the increase in polynomial order $p$. This proves that discarding higher-order polynomial coefficients will only lead to trivial performance loss. Moreover, given a polynomial order $p$, the faster the terminal moves, the greater the covariance of $\phi_{g,p}$. It is anticipated that a smaller polynomial order will be sufficient when the moving speed is not so large.

To further illustrate the power attenuation of polynomial coefficients with the order $p$, we perform simulations considering different Doppler frequencies. In the simulation, normalized Doppler frequency is defined as $f_d^\alpha = \frac{\Delta f}{\Delta f}$. As the metric of terminal velocity, where $\Delta f$ stands for subcarrier spacing and $f_d = \frac{\Delta f}{\Delta f}$. When a terminal is configured with 16 or 8 antenna, one receive beam may cover the subpaths with 8-degree or 16-degree angular spread (AS). The normalized power of an order-$p$ polynomial coefficient is illustrated in Figure 1.

From Figure 1, we observe that if the normalized frequency $f_d^\alpha \leq 0.1$, polynomial with order $p = 3$ can collect more than 99% of energy. The wireless channel can be well represented by polynomial approximation. Furthermore, by comparing the results of 8-degree and 16-degree angular spread, we observe
that the less the subpaths spread in spatial domain within one beam, the more steeply the polynomial coefficients descend with the increasing order $p$. Hence, taking advantage of massive antenna at the base station and terminal, it is effective to decompose the channel subpaths into several closely spaced clusters, and predict the channel variation per cluster.

5 | NUMERICAL RESULT

In this section, numerical results are presented to verify the performance of our proposed channel prediction method.

5.1 | System setup

The OFDM system is set up based on the following parameters.

The carrier frequency of 2 GHz is considered. Within 20 MHz bandwidth, $K = 1200$ subcarriers are available for data transmission, and the subcarrier spacing is $\Delta_f = 15$ KHz.

The duration of an OFDM symbol is about 71.4 s. One subframe consists of 14 OFDM symbols. Throughout the evaluations, we assume that the channels on subframe $n$ and before are perfectly known. Based on these known channels, the terminal predicts the channel on subframe $n + \Delta n$, where $\Delta n = 4$.

The terminal moving at the speed of 200 km/h, 350 km/h and 500 km/h are evaluated.

The channel has $L = 4$ non-line-of-sight (NLOS) taps [32]. Each channel tap is composed of 20 subpaths. The arrival directions of these subpaths are equally distributed within $0 \sim 2\pi$. Moreover, for each subpath, it is assumed that $E\{b_{n,t}(t)b_{n,t}(t + \Delta t)^\dagger\} = \sigma^2 f_0(2\pi f_{l,s}(\Delta t))$, where $f_0(\bullet)$ denotes the zero-order Bessel function of the first kind.

The base station has 16 antennas with 0.5$\lambda$ antenna spacing. The corresponding beamwidth is about $\Delta \Theta_{bc} = 8^\circ$. The terminals with 8-antenna and 16-antenna are taken into account, which corresponds to beamwidths of about $\Delta \Theta = 16^\circ$ and $\Delta \Theta = 8^\circ$.

The additive noise is zero-mean complex white Gaussian, and the SNR refers to the average energy per symbol received on all antennas over one-side noise spectral density.

The performance is evaluated in two ways.

1. Normalized correlation ratio defined as

$$\eta = \frac{\|W_{\text{ideal}}^H W\|}{\|W_{\text{ideal}}\|_2}$$

where $W_{\text{ideal}}$ is the eigenvector of the ideal channel on subframe $n + \Delta n$. $W$ is the estimated eigenvector, which is based on the predicted channel on subframe $n + \Delta n$.

2. Block error rate (BLER). Based on the estimated eigenvector of $W$, the performance in BLER is evaluated. In the simulator, the frequency band of 20MHz is allocated to the terminal. The coding rate of the data channel is set to be 0.6, and 16QAM modulation is assumed.

5.2 | Simulation results

To illustrate the performance of the proposed method, the normalized correlation ratio is first evaluated. The polynomial order for prediction is set to 1, 2, and 3, respectively. Order-1 method is actually precoder without any prediction, i.e. the eigenvector of the channel on subframe $n$. The proposed methods with Order-2 and Order-3 predict channel on subframe $n + \Delta n$ using the channel estimates on subframe $n$ and before.

In Figure 2, the normalized correlation ratio is illustrated by varying the polynomial orders and the terminal velocity. The terminal has 16 antennas, corresponding to a beamwidth of 8 degrees. We observe that for the terminal velocity of 200 km/h, channel prediction with Order-2 is very close to the ideal precoder, i.e. over 95% close to the ideal precoder. When the terminal velocity is greater than 350 km/h, the channel prediction with Order-2 could not predict future channel with sufficient accuracy. The channel prediction with Order-3 can provide more than 95% similarity to the ideal precoder.

In Figure 3, the terminal has 8 antennas, corresponding to a beamwidth of 16 degrees. This configuration is more challenging to our proposal. We observe that for terminal velocity below 350 km/h, acceptable performance is still achieved with channel prediction of Order-3, i.e. more than 95% close to ideal precoder. For terminal velocity of 500 km/h, performance loss is observed. This is because one beam basis in our proposal covers too many spatial channel subpaths. Large order of polynomial approximation should be exploited to solve this issue.

![Figure 2](image-url) Normalized correlation ratio for applying different polynomial orders for prediction, and different terminal velocities. The channel has $L = 4$ significant non-line-of-sight (NLOS) taps, and one receiver beam covers subpaths with 8 degree angular spread.
At the terminal side, which corresponds to a beamwidth of 8 degrees. We observe that the conventional precoding method suffers from severe performance loss due to CSI ageing problems. Our proposal essentially overcomes this issue and shows better performance over the conventional method. We also observe that the performance with Order = 2 and Order = 3 are very close to the ideal case, i.e. only 0.5–1 dB worse than the ideal channel. It proves the validity of our proposal.

Moreover, we evaluate the case that the terminal is equipped with 8 antennas and the results are shown in Figure 5. Moreover, the case that channel on subframe $n+4$ is ideally known is evaluated for fairness. It is observed that although our proposal with Order = 2 still outperforms the conventional method, there are about 2–3 dB performance loss compared with Order = 3, and 3.5 dB worse than the ideal case. The reason is that one beam in our proposal covers much wider subpaths than that in Figure 3. Polynomial approximation with large order should be applied in such cases. Additionally, terminal velocity is set to 350 km/h. Figures 6 and 7 show the simulation results of the terminal with 16 and 8 antennas, respectively.

From Figure 6, we observe that the channel prediction with Order = 2 is about 3 dB better than the conventional method at the BLER of 0.1. Our proposal with Order = 3 is 5 dB and 2 dB better over convention and the proposed prediction with Order = 2, respectively. Compared with an ideal channel, the proposal with Order = 3 is only 0.7 dB worse at the BLER of 0.1. In Figure 7, although the prediction with Order = 2 is still 2 dB better than the conventional method, it suffers from the prediction inaccuracy and has the error floor phenomena. For the channel prediction with Order = 3, the curve drops sharply with the increase of SNR. It illustrates that in the case of high terminal mobility and large angle spread, polynomial approximation with large order should be applied in order to accurately predict the channel variation.

In order to verify our proposal under extreme channel conditions, the terminal velocity is set to 500 km/h. Moreover, the
FIGURE 6  The terminal velocity is 350 km/h. The channel has $L = 4$ significant NLOS taps, and one beam covers subpaths with 8 degree angular spread

FIGURE 7  The terminal velocity is 350 km/h. The channel has $L = 4$ significant NLOS taps, and one beam covers subpaths with 16 degree angular spread

FIGURE 8  The terminal velocity is 500 km/h. The channel has $L = 4$ significant NLOS taps, and one beam covers subpaths with 8 degree angular spread

FIGURE 9  The terminal velocity is 500 km/h. The channel has $L = 4$ significant NLOS taps, and one beam covers subpaths with 16 degree angular spread

differential. The simulation results are shown in Figures 8 and 9, respectively.

We observe from Figure 8 that the conventional method suffers from severe error floor and cannot reach 0.1 BLER. Our proposal with Order $= 2$ is about 3.5 dB better than the baseline, and the prediction with Order $= 3$ is 5.5 dB better than the conventional method. In Figure 9, both the conventional method and our proposal with Order $= 2$ suffer from error floor, which is similar to that in Figure 7. Our proposal with Order $= 3$ greatly alleviates this issue and shows much better performance. In both the figures, our proposal with Order $= 3$ is about 1.4 dB worse than the ideal channel at the BLER of 0.1.

These evaluation results show that our proposal adapts to various terminal speeds by adopting different polynomial orders. It demonstrates the robustness and adaptability of our proposals.

By evaluating the performance of various terminal velocities, we observe that BLER of 500 km/h and 350 km/h is much worse than that of 200 km/h, even adopting our proposal with the higher polynomial order. The reason behind is that after solving the channel ageing problem, the inter-carrier interference becomes the bottleneck for performance improvement. Therefore, advanced receivers are required to suppress such interference.

6 | CONCLUSIONS

In this paper, the channel prediction and feedback mechanism for fast time-varying channels are proposed. With the proposed framework, the future channel can be accurately
predicted meanwhile the feedback overhead is retained at a low level. The performance is verified under the conditions of various terminal velocities. The channel ageing problem is greatly alleviated by our proposal.

We also observe that in high-speed scenarios, only solving the channel ageing problem cannot achieve the expected performance. Other technologies such as low-complexity ICI algorithm should be considered in our next research step.

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