Latency-Reduced Equalizer with Model-Based Channel Estimation for Vehicle-to-Vehicle Communications

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***Abstract*—Equalization of fast time-varying channels is impacted by short coherence time and deep signal fading, especially in vehicle-to-vehicle communications due to the high mobility of user terminals. Furthermore, the stringent latency requirement of safety applications, such as collision avoidance, cannot tolerate high-complexity operations and long processing time. In order to achieve both requirements of equalization accuracy and short latency for fast varying channels, we propose a new model-based time-domain equalizer. In this equalizer, estimation and equalization are performed in two parallel parts to shorten the processing time. In main path, data symbols pass through an equalizer preset with the up-to-date channel impulse response. Since the channel variation model remains invariable for sufficiently long time, the current channel is estimated in parallel path from a number of past channel impulse responses to improve the accuracy. However, the channel variation during the long processing time of estimation leads to equalization error. Therefore, a predictor is used to update the channel response related to the processing delay, and perform the channel estimation beyond the channel coherence time. Thus, high accuracy and delay-free equalization can be achieved through this parallel structure.**

I. Introduction

Vehicular communication is receiving more and more attention for a wide range of services and applications in safety and business. It offers data exchange between vehicles and a roadside infrastructure (V2I) or between vehicles (V2V). Some of the obvious benefits for vehicular communications are the accident avoidance, the ability to notice commuters about traffic conditions, and to advance the processing at tollbooths [1]. However, the harsh signal propagation environment in V2V communications causes some challenges due to high mobility of user terminals.

In the scenario of V2V communications, channel equalization becomes more difficult than the regular outdoor environments. Since the channel coherence time depends on the relative speed of mobile stations, the channel coherence time of V2V wireless channels is much shorter than that in typical urban environments [2]. The equalization and estimation errors increase because of the significant variation of channels where the pilot period is longer than the channel coherence time. It is no surprise that the standard for V2V communications, IEEE 802.11p, which borrows most parts of its physical layer from the indoor centric IEEE 802.11a (Wi-Fi) [3], undergoes the insufficient pilot density in high mobility outdoor environments. Although the pilot density can be increased to match the channel coherence time, it will lead to lower effective data transmission rates.

Moreover, in V2V communications, the impact of vehicles as obstacles of signal propagation becomes a problem. It not only significantly decreases the received signal power, but causes high variation of the received signal power [4]. Therefore, the traditional channel estimation from a single pilot symbol increases the risk of equalization error over the fast varying and deep fading channel. The accuracy of equalization can be improved by using multiple symbols or iterative algorithms [5] [6]. Unfortunately, long processing delay and high complexity involved by these methods may exceed the latency requirement of V2V communication for the safety applications that request timely reaction, such as collision avoidance [7].

Existing channel estimaters are performed in the frequency-domain, due to the popularity of the orthogonal frequency-division multiplexing (OFDM) systems [8] [9]. Recently, time-domain techniques in channel estimation and prediction have shown to be attractive alternatives to frequency-domain techniques [10] [11], since the the frequency-domain channel, which is a combination of multiple propagation paths, is difficult to achieve accurate estimation and prediction. Furthermore, the time-domain techniques need to operate fewer parameters than is required in frequency domain. This significantly decreases the complexity of signal processing.

In this paper, we propose a novel time-domain equalization algorithm for fast varying channel in V2V communications. In order to shorten the processing delay, the equalizer is separated into two parts, i.e., main path and parallel path. Main path includes an equalizer whose coefficients are preset with the up-to-date channel impulse response from parallel path. In parallel path, the current channel impulse response are estimated from multiple past channel impulse responses, according to the characteristic of time-varying channels that the channel variation model remains invariable for sufficiently long time. The channel with the short coherence time may vary during processing time of the model-based estimation. Therefore, a predictor is used to update the channel response related to the processing delay, and to perform the channel estimation beyond the channel coherence time. The equalizer

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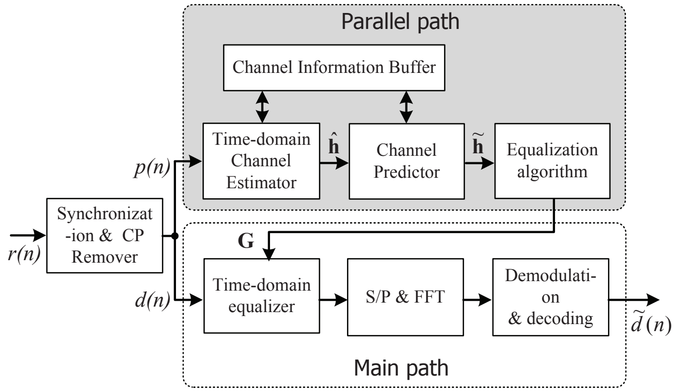

Fig. 1: Structure of proposed equalizer for fast time-varying channel.

coefficients are trained by the predicted channel, and set the equalizer in main path. With this structure, the equalizer reduces equalization error by predicting channel variation from multiple past channel estimation, and realize delay-free equalization.

The remainder of this paper is organized as follows. The system and channel models are described in Section II. The structure of the purposed time-domain equalizer is presented in Section III. Section IV presents the performance evaluation of the proposed equalization algorithm by a series of simulation, while Section V concludes the paper.

II. System Model

The OFDM transmission model with N subcarriers is used in this paper. The input bits are initially mapped to the frequency domain transmitted signal $\{S(k)\}$, $k = 0, 1, \ldots, N-1$. The combination of data and pilot form the N-subcarrier OFDM symbol $S(k)$, which is subsequently transformed into a time domain sequence using the N-point IFFT. i.e.,

$$s(n) = \sum_{l=0}^{N-1} S(k) e^{\frac{j2\pi lnk}{N}}, n = 0, 1, \ldots, N-1. \quad (1)$$

In order to avoid intersymbol interference (ISI), and to allow effectively circular convolution between the output sequence and the channel, the last ν samples of the time domain symbol $\{s(n)\}$, $n = 0, 1, \ldots, N-1$ are used as a guard interval (GI) and prepended to the block to form the transmit sequence $\{s(N-1-\nu), \ldots, s(N-1), s(0), \ldots s(N-1)\}$. This is called cyclic prefix (CP). The wireless channel between the transmitter and the receiver is denoted by $\mathbf{h}$. We consider a baseband-equivalent, discrete-time model of the channel $\mathbf{h} = \{h(0), h(1), \ldots, h(L-1)\}$. $\{s(n)\}$ is filtered by the channel filter with impulse response $\mathbf{h}$, and superimpose an additive white Gaussian noise term $\omega(n)$ with variance σ_ω^2. Then the received signal is obtained by,

$$r(n) = \sum_{l=0}^{L-1} h(l) s(n-l) + \omega(n). \quad (2)$$

It is assumed that CP is longer than the delay spread of the channel, and the effects of inter-carrier interference can be ignored.

Involving CP can force the convolution in Eqn. (2) to be circular on bolck of size N. Then the received signals after the removal of CP can be written in the following matrix form

$$\mathbf{r} = \mathbf{H}\mathbf{s} + \mathbf{w} \quad (3)$$

where $\mathbf{s} = [s(0), s(1), \ldots s(N-1)]^T$, $\mathbf{H}$ is an N-by-N matrix which is given by

$$\begin{bmatrix} h(0) & 0 & \cdots & 0 & h(L-1) & \cdots & h(1) \\ h(1) & h(0) & \cdots & 0 & h(L-1) & \cdots & h(2) \\ \vdots & & \ddots & & & \ddots & \vdots \\ 0 & \cdots & 0 & h(L-1) & \cdots & h(1) & h(0) \end{bmatrix}, \quad (4)$$

and $\mathbf{w}$ denots the Gaussian noise vector.

A. Wireless Channel Model

The wireless channel is represented by a time-varying linear filter, with a time-varying impulse response $h(t, \tau)$ in time and delay domains. The complex baseband representation of the time-varying multipath channel is given as

$$h(t, \tau) = \sum_{l=0}^{L-1} a_l(t) \delta(\tau - \tau_l(t)), \quad (5)$$

where $\tau_l(t)$ is the delay, $a_l(t)$ is the complex amplitude of the lth multipath tap, and L is the number of propagation paths. We consider a Wide Sense Stationary - Uncorrelated Scattering (WSSUS) channel. The WSSUS characteristic implies that the channel correlation is dependent only on the time difference Δt in time domain and the multipath components are uncorrelated in the delay domain. Therefore, the Doppler shift and the channel variation of different multipath taps are uncorrelated. Since Doppler frequency and time are dual variables, channel variation in time domain is directly related to its spectrum in Doppler frequency domain. The variation of a given multipath component, in time domain, that experiences a Doppler shift of f_d Hz is:

$$h(n, l) = h(l) e^{j2\pi f_d(l) n}, \quad (6)$$

where n is the discrete-time index, and the channel envelope $h(l)$ remains invariant over a number of symbols.

III. Proposed Channel Equalization Technique

In order to achieve both requirements of equalization accuracy and short latency for fast varying channels, we introduce a time-domain equalizer shown in Figure 1 based on the channel variation model. In order to limit the processing time, this equalizer is separated into two parts. Main path includes a delay-free equalizer and reception blocks, while the model-based channel estimation and prediction are performed in parallel path. After equalization, the signals are transformed to frequency domain by Fast Fourier transform (FFT) to be detected and demodulated.

A. Time-domain channel estimation

In order to reduce the complexity and processing delay in parallel path, the channel estimation method considered in this paper operates in time domain, which is based on the time-domain correlation between the received signal and pilot sequence embedded in the OFDM signal [10]. The major advantage of this technique is its robustness to strong interference and frequency offset. The time domain pilot sequence is derived from the frequency domain pilots and known to the receiver. In this method, the transmitted signals can be divided into two different sets, i.e., the data symbol and the pilot symbol. The combined signal is $s(n) = d(n) + p(n)$.

After passing through a multipath channel, the received signal can be written as

$$r(n) = \sum_{l=0}^{L-1} h(l)d(n-l) + \sum_{l=0}^{L-1} h(l)p(n-l) + \omega(n). \quad (7)$$

The correlation between the the received signal $r(n)$ and the time domain pilot sequence $p(n)$ is $R_{rp} = r(n) \bigotimes p(n)$, where $\bigotimes$ denotes the convolution operation, then the channel response can be estimated by

$$\hat{h}(n) = \frac{R'_{rp}}{R'_{rp\_max}}, \quad (8)$$

where R'_{rp} is the truncated R_{rp} which includes the significant part of the energy, and the subscript max denotes the index of the path with highest amplitude.

The channel impulse response estimated from pilot symbols are send to channel information buffer for channel variation modeling.

B. Model-based channel estimation and prediction

The performance of equalizer is sensitive to the channel estimation error caused by channel variation and low signal-to-noise ratio (SNR). However, the channel estimation from a single training symbol cannot achieve high accuracy for fast varying channels. Although the channel response is varying, the channel variation model built from multiple past channel impulse responses remains invariable for sufficiently long time. Therefore, under the assumption that the channel correlation is dependent only on the time difference Δt in the time domain, the current channel can be generated by a linear filter, which linearly combines the past channel impulse responses, as follows [11],

$$\hat{h}(n,l) = \sum_{m=1}^{M} \psi(m,l)\hat{h}(n-m,l), \quad (9)$$

where $\Psi = \{\psi(m,l)\}$ is the set of M coefficients, M is the modeling order, and n is the time indices. The training algorithm adopted in this paper operates in time domain on each multipath cluster, rather than the overall channel impulse responses. And we assume that the variation of components within a cluster are highly correlated, while the variation of different clusters are uncorrelated [12]. The Eqn. (9) can be considered as a FIR filter with M coefficients. The optimal filter coefficients can be determined to minimize the mean square error (MSE), i.e., MSE $= E[|h(n) - \hat{h}(n)|^2]$, with [13]:

$$\Psi_{op} = R_{hh}^{-1} r_{hh}, \quad (10)$$

where

$$R_{hh}(i,j) = E[\hat{h}(n-1-i,l)\hat{h}^*(n-1-j,l)]] \quad (11)$$

is the correlation matrix, calculated for the lth multipath component, and

$$r_{hh}(i) = E[\hat{h}(n-i,l)\hat{h}^*(n,l)]] \quad (12)$$

is the cross-correlation vector between past channel impulse responses and currant channel impulse responses.

In the scenario that the channel varies slowly, the model can be simplified to average multiple channel response, i.e.,

$$\hat{h}(n,l) = \frac{1}{M} \sum_{m=0}^{M} \hat{h}(n-m,l). \quad (13)$$

Following the channel estimator, a predictor is used to update the channel response according to the processing delay, and to perform the channel estimation beyond the channel coherence time. In practice, the operations in parallel path will take longer time than the symbol duration. Therefore, the channel impulse response may vary during the time difference Δt over the channels with short coherence time. On the other hand, the insufficient pilot density fails to track the channel variation. In order to accommodate the equalizer with the current channel conditions, the predictor is used to generate the impulse channel response $\hat{\mathbf{h}}(t + \Delta t)$ from the past channel response $\hat{\mathbf{h}}(t)$ in the absence of pilot symbols. The processing time of parallel path t_p can be self tested when the reception equipment is turned on. Then the total time gap is $\Delta t = t_p + P \times$ symbol duration, where P is the prediction range. The updated channel impulse response after prediction becomes

$$\hat{h}(t + \Delta t, l) = \sum_{m=1}^{M+1} \psi(m,l)\hat{h}(n + \frac{\Delta t}{T_s} - m, l), \quad (14)$$

where T_s is the sampling time.

C. The equalization algorithm with predicted channel

The equalizer coefficients are generated based on the predicted channel $\hat{\mathbf{h}}(t + \Delta t)$. As we mentioned, the processing time of the parallel path t_p need to be tested by reception equipments itself. Therefore, we adopt the equalization algorithm without recurrence, where the processing time is a constant and can be easily determined according to different chips.

Let us denote that the N-by-N tap weight coefficients matrix of the equalizer is $\mathbf{G}$, equalized signals vector is $\mathbf{y} = [y(0), y(1), \ldots, y(N-1)]^T$, and

$$\mathbf{y} = \mathbf{GHs} + \mathbf{Gw}. \quad (15)$$

The equalization coefficients are adjusted to minimize the error $\mathbf{e} = \mathbf{y} - \mathbf{s}$. The minimum mean squared error MMSE

TABLE I: The channel model used in simulations

Tap	Delay (ns)	Average power (dB)
1	1	0
2	100	-10
3	200	-17.8
4	300	-21.1
5	400	-26.3

algorithm minimize the mean-square error (MSE) which is formulated as

$$\mathbf{G} = \arg\min E[|\mathbf{y} - \mathbf{s}|^2], \tag{16}$$

where expectation $E[\cdot]$ is taken over the equalized data sequence and additive noise. Since the MSE has the concave surface, the condition $\frac{\partial E[|\mathbf{y}-\mathbf{s}|^2]}{\partial \mathbf{G}} = 0$ is satisfied. Solving this equation, we obtain the optimum weight vector [14]

$$\mathbf{G} = (\hat{\mathbf{H}}^H \hat{\mathbf{H}} + \sigma_\omega^2 \mathbf{I}_N)^{-1} \hat{\mathbf{H}}^H, \tag{17}$$

where $(\cdot)^H$ denotes the conjugate transpose operation, and $\mathbf{I}_N$ is the N-by-N identity matrix. In this algorithm, there is an insertion of the noise power in the pseudo inverse matrix in order to reduces the noise enhancement.

The overall equalization algorithm can be described as follow,

Step 1 The channel impulse response $\hat{\mathbf{h}}(n)$ is estimated through the cross correlation of the received signal $\mathbf{r}$ and the pilot $\mathbf{p}$, and send to channel information buffer.

Step 2 The time-domain equalizer is set with the coefficients trained by the MMSE algorithm.

Step 3 The current channel is estimated according to the past channel impulse response in the buffer, i.e., $\hat{\mathbf{h}}(n), \hat{\mathbf{h}}(n-1), \ldots, \hat{\mathbf{h}}(n-M)$.

Step 4 The accommodated channel impulse response $\tilde{\mathbf{h}}(n)$ is calculated by the time-domain predictor with special time delay parameter Δt.

Step 5 The channel variation is $\Delta\mathbf{h} = \tilde{\mathbf{h}}(n) - \hat{\mathbf{h}}(n)$. If $|\Delta\mathbf{h}| \geq \eta$, where η is the threshold value for judging channel variation, equalization coefficients are regenerated with the update channel impulse response $\tilde{\mathbf{h}}(n)$. Otherwise, the coefficients are regenerated with the average of past channel impulse response. Then return to *Step* 2 to reset the equalizer.

IV. Simulation

A. Simulation parameters

The performance of the proposed equalization algorithm is shown by a series of simulation results. We consider an OFDM system with symbols modulated by QPSK. The total number of subcarrier is $N = 64$, and the CP length is 8. The channel models are time varying multipath channels with Rayleigh fading in each path. The channel parameters are listed in Table I [15]. The channel length L considered in the simulation is 6. The normalized Doppler frequency is obtained as $f_d T_s$, which is different for different multipaths. The length of channel information buffer is 20, 50 and 100 respectively,

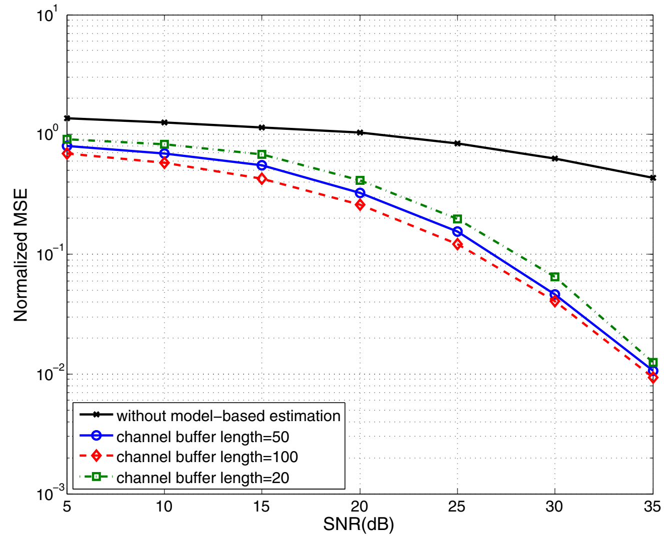

Fig. 2: Normalized mean squared error (MSE) of channel estimation with and without model-based algorithm over different lengths of the channel information buffer.

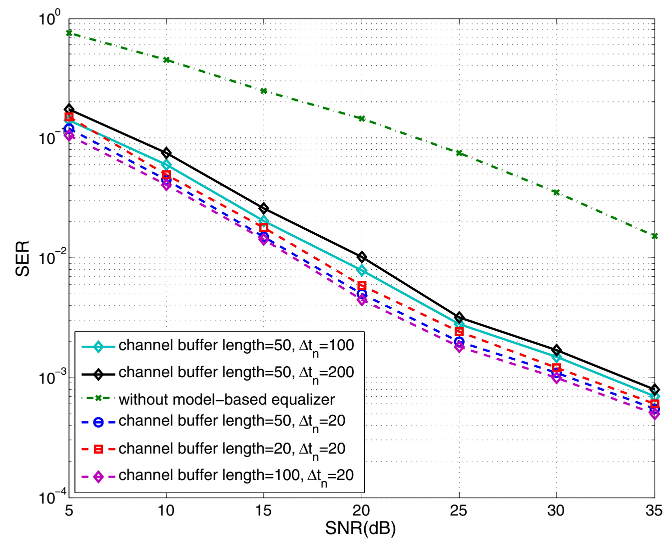

Fig. 3: Symbol error rate of received signals with model-based equalization over different lengths of the channel information buffer and prediction ranges.

which means the current channel impulse response is estimated from 20, 50 and 100 past training symbols. Besides, the normalized prediction range $\Delta t_n = \frac{\Delta t}{T_s}$ equals to 50, 100 and 200 OFDM symbols respectively. $h(\tilde{l})$ in Eqn. (5) remains invariable during Δt_n.

B. Numerical results

Figure 2 presents the normalized mean squared error (MSE) of channel estimation with and without model-based algorithm over different length of the channel information buffer from 20 to 100 OFDM symbols. One observes that the MSEs are sig-

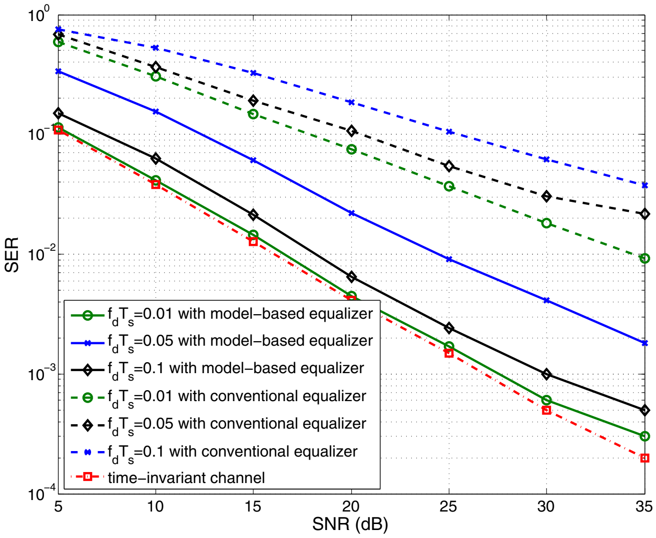

Fig. 4: Symbol error rates of received signals with and without model-based equalization over different channel variation rates.

nificantly reduced when the model-based channel estimation is preformed. Moreover, the MSEs with long buffer length are lower than the ones with short buffer, since the longer length of channel buffer implies the more training symbols which helps to improve the accuracy of the estimation.

Figure 3 shows the average symbol error rates (SER) of received signals equalized with model-based channel estimation over different lengths of the channel information buffer and prediction ranges. In order to compare the performance of systems with and without the proposed equalizer, we simulated the SERs of received signals using the same equalization algorithm but without the model-based estimation and prediction. The length of the channel buffer is from 20 to 100 symbols, the prediction ranges is from 20 to 200 symbols, and the maximum $f_dT_s = 0.03$. It is obvious that the SERs is improved by the proposed equalizer which is expected, as seen form the MSE characteristics. Besides, the equalization performance degrades as the prediction range increases. This result shows that the equalizer works better over the short term prediction, because long prediction range leads to the accumulation of estimation error.

Figure 4 shows the average SERs of received signals equalized with and without proposed algorithm over different channel variation rates. We considered time-invariant, slowly time-varying and rapidly time-varying channels to evaluate the performance of the propose algorithm, where “time-invariant” means that the channel coherence time is much longer than OFDM symbol duration. The f_dT_s are 0.01, 0.05 and 0.1, respectively. The prediction range is 20 symbols, and the buffer length is 50 symbols. Comparing to the regular equalizer, the proposed equalizer can reduce the equalization error by tracking the channel variation under different channel variation rates. Furthermore, when the channel variation is not too fast, such as $f_dT_s = 0.01$, the performance of the proposed equalization approaches to that over time-invariant channel.

V. Conclusion

A novel time-domain equalizer has been proposed in this paper for fast time-varying V2V channels. The performance of the equalizer is improved by estimating the channel from the channel variation model. The equalizer is separated into two parts to reduce the processing delay. Parallel path can track the channel variation without increasing the density of pilot symbols and improve the estimation accuracy. In main path, data symbols are equalized by a delay-free equalizer which preset by parallel path. The performance of the proposed equalizer has been evaluated by a series of simulations for different lengths of the channel information buffer, prediction ranges and channel variation rates. The average SERs of received signals equalized with multiple-symbol channel estimation are remarkably enhanced over fast varying channels.

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