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A novel OFDM transmission scheme with length-adaptive Cyclic Prefix*

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Abstract: Conventional OFDM transmission system uses a fixed-length Cyclic Prefix to counteract Inter-Symbol Interferences (ISI) caused by channel delay spreading under wireless mobile environment. This may cause considerable performance deterioration when the CP length is less than the channel RMS delay spread, or may decrease the system power and spectrum efficiency when it is much larger. A novel Orthogonal Frequency Division Multiplexing (OFDM) transmission scheme is proposed in this paper to adapt the CP length to the variation of channel delay spread. AOFDM-VCPL utilizes the preamble or pilot sub-carriers of each OFDM packet to estimate the channel RMS delay spread; and then uses a criterion to calculate the CP length , which finally affects the OFDM transmitter. As illustrated in the simulation section, by deploying this scheme in a typical wireless environment, the system can transmit at data rate 11.5 Mb/s higher than conventional non-adaptive system while gaining a 0.65 dB power saving at the same BER performance.

Key words: Adaptive, OFDM, Cyclic Prefix, RMS delay spread
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INTRODUCTION

In broadband wireless communication, the delay spread caused by channel multi-path fading will usually introduce serious Inter-Symbol Interference (ISI) to the receiver. To counteract the ISI, the high-efficiency Orthogonal Frequency Division Multiplexing (OFDM) modulation first splits the high-rate data stream into a number of parallel sub-streams and modulates them onto different orthogonal sub-carriers and thus lower the symbol rate, and then add a Cyclic Prefix (CP) to the head of each symbol to reduce the influence of adjacent symbol interference.

Although the CPs are crucial to OFDM system, they introduce significant overhead. For example, in 802.11a wireless LAN, a fixed proportion of 1/5 of the energy and time is spent on CPs. As a system design rule, the CP length should be about two times the RMS (Root-Mean-Squared) delay spread (Van Nee and Prasad, 2000). Obviously, the RMS delay spread is not constant in a wireless mobile communication environment. Conventional OFDM system usually chooses a fixed CP length based on the average or even maximum delay spread the mobile terminal may experience. According to a measurement conducted in (Van Nee and Prasad, 2000), when the mobile terminal is in an office building room, the RMS delay spread is about 35 ns, but when the mobile terminal moves into a factory, the RMS delay spread will change to 300 ns. If the receiver is designed based on the measurement in

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the office, it will undergo severe ISI when the user moves to a factory. On the other hand, if the receiver is designed according to the measurement in the factory, some of the guard interval is unnecessary which will consume the scarce spectral and power resources but achieve no extra gain.

So it is natural to think that if we can estimate the RMS delay spread and change the length of CP accordingly, the overhead of CPs will be reduced when delay spread is small and the ISI will still be eliminated when delay spread becomes large.

Based on this consideration, we designed a new OFDM system with adaptive CP length. We call it AOFDM-VCPL (Adaptive OFDM with Variable CP Length). The goal of this system is to maximize the spectral efficiency without wasting power consumption or sacrificing BER performance, as all previously proposed adaptive OFDM systems (Souryal *et al*., 2002; Keller and Hanzo, 2000; Moon *et al*., 2002) do. The AOFDM-VCPL does the following: 1) estimates the channel parameters; 2) chooses the modulation parameters for the next transmission based on the estimated channel parameters; 3) signals the employed modulation parameters. In this paper, an RMS delay spread estimation structure is derived in detail. It is based on the ML (Maximum Likelihood) criteria using pilots in the OFDM symbols. The adapting strategy and signaling method are also addressed.

The rest of this paper is organized as follows. In section II, we describe the various aspects of the AOFDM-VCPL system. The system configuration is described first, and then the RMS delay spread estimation method is discussed in detail. The method to decide the length of the CP and signal the decided length is also discussed in this section. Performance analysis and simulation result is given in Section III.

SYSTEM MODEL

System Description

Fig.1 shows the AOFDM-VCPL system. This system works in TDD burst mode. The transceiver B receives the data packet transmitted by A, estimates the RMS delay spread using this packet, and decides the CP length of the next data packet that will be transmitted to A soon.

We make a minor modification to the well-known IEEE 802.11a wireless LAN standard to get the system configuration of our own. In our systems, the sample time T_s is set to be 50 ns and the data duration is 3.2 μs. It is the same with 802.11a. But the duration of CP is variable based on the channel quality as compared with the 802.11a where the CP length is fixed to be 800 ns. There are 52 subcarrieres with subcarrier spacing 312.5 kHz, in which 48 subcarriers are used to transmit information and 4 subcarriers are used as pilots. Adding 12 virtual subcarriers, we get 64 samples in the data field of an OFDM symbol. In the information bearing subcarriers, uncoded 64QAM is employed. In the pilots, BPSK is used. Fig.2 shows the data packet structure. The preamble part is used to assist the packet capture and synchronization. The signal field is used to signal the used number of CP samples and the number of symbols of this burst. In the preamble and signal part, the length of CP is fixed to be 800 ns. We assume that the channel varies very slowly. This assumption is the base of

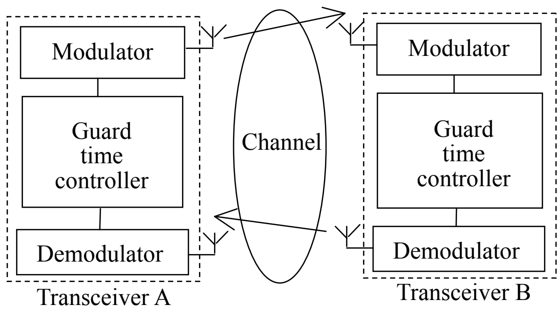

Fig.1 The AOFDM-VCPL system

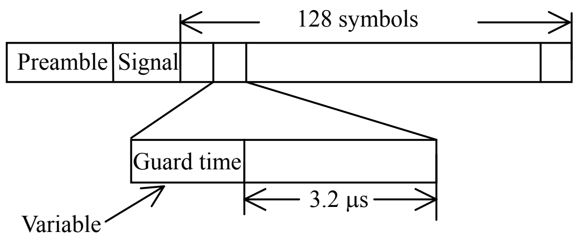

Fig.2 AOFDM-VCPL data packet structure

any adaptive system (Van Nee and Prasad, 2000; Souryal *et al*., 2002; Keller and Hanzo, 2000).

Fig.3 is the block diagram of the transceiver in Fig.1. Input data $\boldsymbol{S}(n)$ is first serial/parallel converted, and then the converted N-dimensional vector of frequency domain data $\boldsymbol{S}=[S(0), \ldots, S(N-1)]$ is fed to the IFFT part. We assume that there are some known symbols (pilots) multiplexed in the vector $\boldsymbol{S}$. These pilots are used to estimate the channel parameters. After IFFT, the time-domain N-dimensional vector $\boldsymbol{s}=[s(0), \ldots, s(N-1)]$ is parallel/serial converted. Next, an N_G point cyclic prefix is added to the time-domain data $\boldsymbol{s}$. The value of N_G is set as two times the estimated RMS delay spread. The resulting extended vector drives a linear modulator with impulse response $g(t)$. The modulated signal $s(t)$ is sent out in the end.

In the receiver part, the received signal $r(t)$ is first passed through a matched filter, and then it is sampled at rate $1/T_s$ and serial to parallel converted. Next, the N_G point cyclic prefix is removed, and the received samples are passed to an N-point FFT unit. Then the multiplexed pilots are extracted out and fed to the channel estimator part which estimates the instant channel response and RMS delay spread using the "a priori" knowledge of the pilot symbols. The channel response estimation result is used in the coherent demodulation part, and the RMS delay spread estimation is used to decide the CP length of the next transmission.

Estimating the RMS delay spread

Consider a frequency-selective mobile radio channel whose baseband CIR (Channel Impulse Response) can be described by (Steele, 1992):

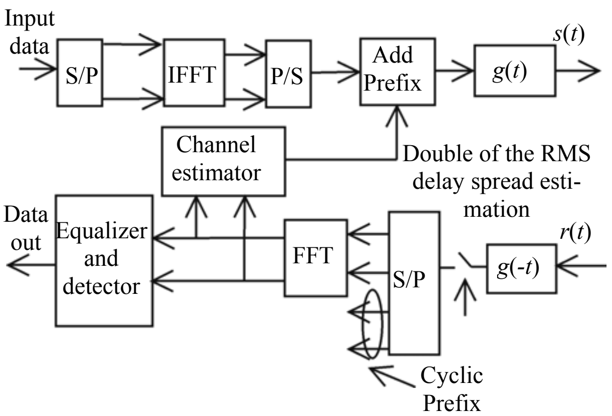

Fig.3 Block diagram of the AOFDM transceiver

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

where τ_l is the lth path delay and $\gamma_l(t)$ is the corresponding complex amplitude. $\gamma_l(t)$ is assumed to be wide-sense-stationary (WSS) complex Gaussian processes and independent between different paths. L is the number of independent paths unknown to us. We assume that the channel variations are negligible over an OFDM block, so we can drop the identifier t for simplicity. Let $\boldsymbol{h}=[h(0),\ldots,h(N-1)]^{\mathrm{T}}$ represent the T_s-spaced samples of the overall CIR. If the delays τ_l are integer times of T_s, which means $\tau_l=nT_s$ for all l, then only first L elements have none-zero value. Otherwise, all the N entries have none-zero value because of the energy leak effect (Van de Beek *et al*., 1995), but the values of the last N-L elements are very small. Denoting

$$H(n)=\sum_{l=0}^{N-1}h(l)\exp(-j2\pi nl/N) \tag{2}$$

in fact $\boldsymbol{H}=[H(0),\ldots,H(N-1)]^{\mathrm{T}}$ is the FFT of $\boldsymbol{h}$.

Then the output of the FFT part could be written as:

$$R(n)=S(n)H(n)+w(n), \tag{3}$$

where $S(n)$ is the transmitted symbol on the nth sub-carrier and $w(n)$ is the white Gaussian noise with zero mean and variance σ^2.

In order to change the length of the cyclic prefix adaptively, we must estimate the T_s normalized RMS delay spread:

$$\tau_{RMS}=\sqrt{\frac{\sum_l |h(l)|^2 l^2}{\sum_l |h(l)|^2}-\left(\frac{\sum_l |h(l)|^2 l}{\sum_l |h(l)|^2}\right)^2} \tag{4}$$

From Eq.(4), we can see that if we could get an estimation of the CIR $\boldsymbol{h}$ then the estimation of the RMS delay spread $\hat{\tau}_{\text{RMS}}$ can be computed.

As mentioned above, some pilots are multiplexed into the data stream to assist the channel estimation. Suppose that there are N_p pilots $\{p_n; 0\le n\le N_\text{p}-1\}$ at known positions $\{i_n; 0\le n\le N_\text{p}-1\}$. Let $\boldsymbol{R}_\text{p}=[R(i_0), \ldots, R(i_{N_\text{p}-1})]^\text{T}$ represent the $N_\text{p}\times 1$ vector whose elements are the outputs of the pilot positions. Using Eqs.(2) and (3), we can get:

$$\boldsymbol{R}_\text{p} = \boldsymbol{S}_\text{p}\boldsymbol{F}\boldsymbol{h} + \boldsymbol{w}, \tag{5}$$

in which $\boldsymbol{S}_\text{p}$ is a diagonal matrix with transmitted pilot symbols on its diagonal:

$$\boldsymbol{S}_\text{p} = \text{diag}\{S(i_0),\ldots,S(i_{N_\text{p}-1})\} \tag{6}$$

$\boldsymbol{F}$ is an $N_\text{p}\times N$ matrix with entries

$$[\boldsymbol{F}]_{n,k}=\exp(-j2\pi k i_n/N) \tag{7}$$

$\boldsymbol{w}$ is vector composed of Gaussian noise.

The receiver knows the pilot matrix $\boldsymbol{S}_\text{p}$. Because BPSK signal is transmitted on the pilot symbols, so pre-multiplying both sides of Eq.(5) by $\boldsymbol{S}_\text{P}^\text{H}$ produces:

$$\boldsymbol{X}=\boldsymbol{F}\boldsymbol{h}+\boldsymbol{n}, \tag{8}$$

where $\boldsymbol{X}=\boldsymbol{S}_\text{P}^\text{H}\boldsymbol{R}$, and $\boldsymbol{n}=\boldsymbol{S}_\text{P}^\text{H}\boldsymbol{w}$ whose elements are still independent Gaussian noise with zero means and variance σ^2.

From Eq.(8) we can get our estimation of $\boldsymbol{h}$ with ML criteria. Denoting $\hat{\boldsymbol{h}}$ the estimation of $\boldsymbol{h}$, we can get $\hat{\boldsymbol{h}}$ by maximizing the conditional probability distribution function $f(\boldsymbol{X}\,|\,\hat{\boldsymbol{h}})$ under ML criteria. Because the elements of $\boldsymbol{n}$ are independent Gaussian noise with zero means and variance σ^2, $f(\boldsymbol{X}\,|\,\hat{\boldsymbol{h}})$ can be written as:

$$f(\boldsymbol{X}\,|\,\hat{\boldsymbol{h}}) = \prod_{n=0}^{N_\text{p}-1}\frac{1}{\sqrt{2\pi}\sigma}\exp\left(\frac{(X(n)-[\boldsymbol{F}\hat{\boldsymbol{h}}]_n}{2\sigma^2}\right)^2 \tag{9}$$

Solving the function $\dfrac{\partial f(\boldsymbol{X}\,|\,\hat{\boldsymbol{h}})}{\partial\hat{\boldsymbol{h}}}=0$, yields:

$$\hat{\boldsymbol{h}}_\text{ML} = \boldsymbol{E}^{-1}\boldsymbol{F}^\text{H}\boldsymbol{X} \tag{10}$$

where $\boldsymbol{E}$ is a square matrix:

$$\boldsymbol{E}=\boldsymbol{F}^\text{H}\boldsymbol{F} \tag{11}$$

Rewriting Eq.(10) in scalar form:

$$\hat{h}_\text{ML}(n) = \sum_{m=0}^{N_\text{p}-1}X(m)\sum_{n=0}^{N-1}[\boldsymbol{E}^{-1}]_{k,n}\exp(j2\pi n i_m/N) \tag{12}$$

Substituting Eq.(12) into Eq.(4), yields the estimation of τ_RMS:

$$\hat{\tau}_\text{RMS} = \sqrt{\frac{\sum_n\left|\hat{h}(n)\right|^2 n^2}{\sum_n\left|\hat{h}(n)\right|^2} - \left(\frac{\sum_n\left|\hat{h}(n)\right|^2 n}{\sum_n\left|\hat{h}(n)\right|^2}\right)^2} \tag{13}$$

Substituting Eq.(12) into Eq.(2), we can get the frequency domain channel gain $\hat{\boldsymbol{H}}$, which also is the ML estimation of $\boldsymbol{H}$. In any OFDM using coherent demodulation, we must get the estimation of the frequency domain gain. So we can get the RMS delay spread estimation without adding much computation burden in an OFDM receiver employing coherent demodulation scheme. The only computation burden added is the computation of Eq.(13). Fig.4 is the block diagram of the proposed

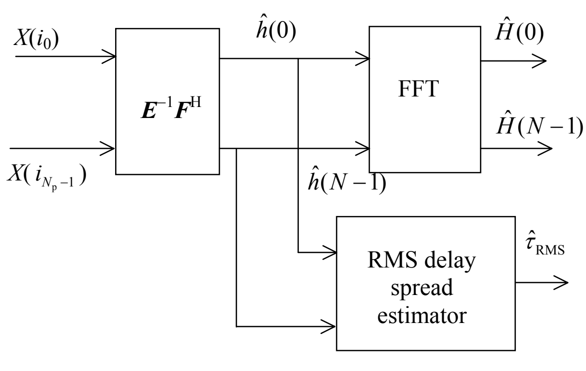

Fig.4 Block diagram of the channel parameter estimation structure

RMS delay spread and frequency domain channel gain estimation structure.

Choosing modulation parameters

In the AOFDM-VCPL, we will change the length of the CP based on the estimation result. In the subsection above we have derived the estimation structure of $\hat{\tau}_{\mathrm{RMS}}$. After we got the value $\hat{\tau}_{\mathrm{RMS}}$, we can set the duration of the CP of the next transmission as $2\lceil \hat{\tau}_{\mathrm{RMS}} \rceil$ according to the design rule. In which $\lceil x \rceil$ means the smallest integer greater than x.

Signaling

In the AOFDM-VCPL, the transceiver must signal the chosen CP length which can be done by using the signal field in the data packet as described in Section 2.1. This is similar to the IEEE 802.11 Standard. The number of the CP sample can be presented in a few bits; this will not add much burden to the system.

SIMULATION RESULTS

In order to analysis the performance of the proposed estimation method and the gain that can be achieved by using AOFDM-VCPL over conventional OFDM system, computer simulations were performed.

Simulation setup

The system of AOFDM-VCPL has been described above. Many researchers have conducted the RMS delay spread measurements on various positions (Rapport and McGillm, 1989; Nobles and Halsall, 1997). Here we use the data of Rapport. The measurement was conducted in a factory. Fig.5 shows the CDF (Cumulative Distribution Function) of the measured RMS delay spread. From the figure, we can point out that the RMS delay spread varied significantly in the various positions of the same factory. This further validates the proposed scheme in this paper. In the paper (Rapport and McGillm, 1989), the authors also pointed out that the measured RMS delay spread could be approximately modeled by a uniform distribution as shown by the dashed line in Fig.5.

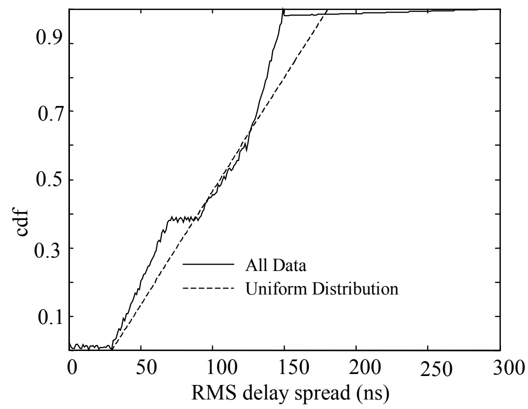

Fig.5 The CDF of the measured RMS delay spread

Performance of the proposed RMS estimation method

Substituting Eq.(8) into Eq.(10), $\hat{\boldsymbol{h}}_{\mathrm{ML}}$ yields:

$$\hat{\boldsymbol{h}}_{\mathrm{ML}} = \boldsymbol{E}^{-1}\boldsymbol{F}^{\mathrm{H}}\boldsymbol{F}\boldsymbol{h} + \boldsymbol{E}^{-1}\boldsymbol{F}^{\mathrm{H}}\boldsymbol{n} \tag{14}$$

Bearing in mind that $\boldsymbol{n}$ is zero mean Gaussian noise, and averaging $\hat{\boldsymbol{h}}_{\mathrm{ML}}$ over noise, we can prove that the estimation $\hat{\boldsymbol{h}}_{\mathrm{ML}}$ is unbiased. And because the entries of $\boldsymbol{h}$ are independent (Steele, 1992), it can be proved that $\hat{\tau}_{\mathrm{RMS}}$ is an unbiased estimation of τ_{RMS}.

The MSE (Mean Square Error) is another criteria for evaluating the effectiveness of estimation. The mean square error is defined as:

$$MSE = E\{|\hat{\tau}_{\mathrm{RMS}} - \tau_{\mathrm{RMS}}|^2\} \tag{15}$$

Fig.6 shows the MSE at various SNR. We get it by transmitting 1000 symbols and calculate the MSE using Eq.(15). It shows that when SNR is 10 dB, MSE is about 10^{-2}. Under this order of the estimation error, we can choose the duration of the CP almost without error. It can be explained by an example. Suppose that the T_s normalized RMS delay spread τ_{RMS} is 4.6; we should use 10 samples in the cyclic prefix according to the design rule.

Then with MSE equaling to 10^{-2}, $\hat{\tau}_{\mathrm{RMS}}$ will fall to the range [4.0, 5.0] almost with probability 1. So based on the adaptation method, we will still choose 10 samples in the CP.

Performance of the AOFDM-VCPL system

Fig.7 shows the relative amount of power and time spent on CP at various positions. The RMS delay spread varies depending on where the receiver is. The curve with * present the conventional OFDM with fixed CP length. So no matter where the receiver is, the overhead is the same. The curve marked with Δ presents the AOFDM-VCPL system. It presents the power loss when a transceiver is at different positions with different RMS delay spreads. The power spent on CP varies according to the estimated RS delay spread. Compared with the fixed OFDM, it can achieve 1 dB gain. This 1 dB power saving is very valuable to mobile terminal that use battery as its power source.

Fig.8 is the maximum achievable uncoded data rates the system can get using 64QAM as its modulation method. The system with fixed CP length can achieve 72 Mb/s. It is fixed no matter where the transceiver is. But for the AOFDM-VCPL, its maximum uncoded data rates can be as high as 87.27 Mb/s. It achieves 15.27 Mb/s gain over fixed scheme. It is very desirable considering that the spectrum is very scarce.

Fig.9 shows the BER performance of the AOFDM-VCPL system and the conventional non-adaptive OFDM system. The BER is obtained

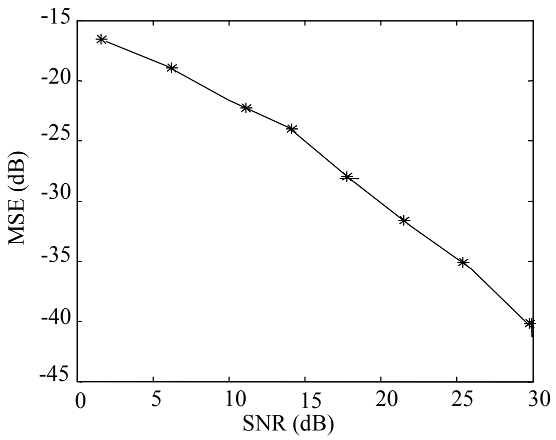

Fig.6 MSE versus SNR

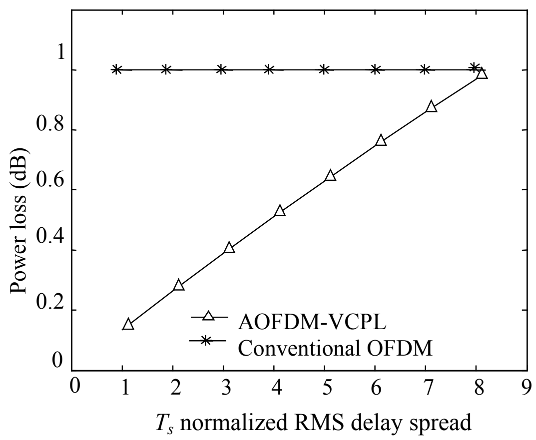

Fig.7 Power Loss versus delay spread

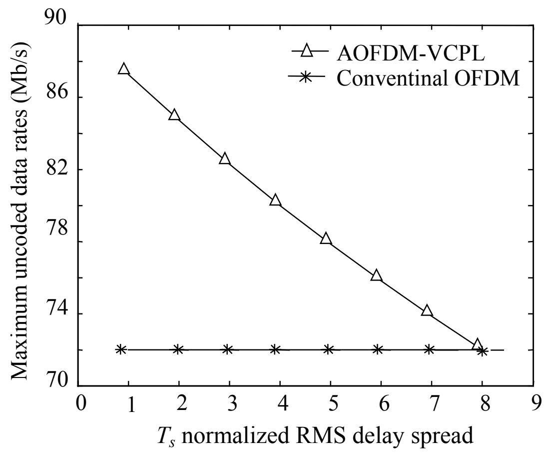

Fig.8 Maximum achievable data rates versus delay spread

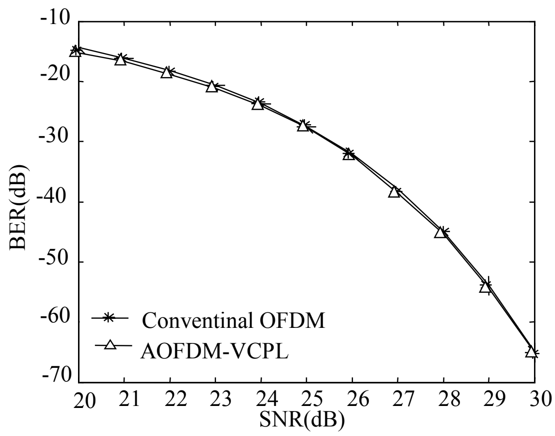

Fig.9 BER versus SNR

under the assumption that synchronization and carrier recovery are perfect and only noise and channel estimation error is considered. It can be pointed out from Fig.9 that the gain in power loss and throughput can be achieved without sacrificing the BER performance.

Table 1 shows the average gain that the AOFDM-VCPL system could get by assuming that the receiver is uniformly distributed in the factory environment described above in Section 3.1. Fig.5 is the RMS delay spread that the receiver will encounter when it is uniformly distributed in the factory. The gains the system could get at various positions of the factory have been described in the sections above. Averaging gains at a factory is given in Table 1. It can be pointed out the proposed system could transmit at a data rate 11.5 Mb/s higher than the non-adaptive system while saving 0.65 dB power at the same BER rate.

Table 1 Average gain achieved in a factory

Power loss (dB)	Data rate (Mb/s)
0.6454	11.536

CONCLUSION

An adaptive OFDM system with variable CP length is described in detail in this paper. The RMS delay spread delay estimator is designed. The CP length is changed based on the estimation result. Simulation result showed that the RMS delay spread estimation is unbiased and that the variance of the estimation is small. Simulation result also showed that this system can achieve up to 0.65 dB gain in power and can transmit at a bit rate 11.5 Mb/s higher than that of the non-adaptive OFDM system without sacrificing the BER performance.

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