PAPER

Multi-Level QAM Transceivers with Adaptive Power Control in Fixed Wireless Channels*

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SUMMARY This paper considers the design of quadrature amplitude modulation (QAM) transceivers for fixed wireless communications. We propose the use of power control in the QAM transmitter (Tx) to obtain BER performance robust to fading. The gain of the Tx is adaptively adjusted to keep the power of the received signal nearly constant despite of the short term fading and the second multipath. The BER performance of the proposed scheme is analytically evaluated in fixed wireless channels with flat fading and frequency selective fading. Analytic and simulation results show that the use of power control in the Tx can provide the BER performance only about 1 dB inferior to that in additive white Gaussian noise (AWGN) channel.

key words: *fading, fixed wireless channel, QAM transceivers, power control*

1. Introduction

As the demand for wireless data communications is explosively increasing, it is indispensable to improve the efficiency of spectrum usage because available radio spectrum is strictly limited. It is well known that quadrature amplitude modulation (QAM) is one of the most spectral-efficient modulation schemes. In recent, QAM schemes have been applied to fixed wireless communications [1]. Even in mobile radio systems, the use of QAM schemes is considered to increase the transmission data speed [2]. The application of QAM schemes to wireless channel requires to accurately compensate the fading effect that causes significant degradation of BER performance.

To reduce the fading effect, the use of pilot symbol assisted modulation (PSAM) schemes was proposed [3], [4]. The PSAM scheme adjusts the scale of decision grid in the receiver (Rx) by observing the level of the received pilot symbol due to channel gain variation. However, since it cannot effectively compensate the effect of channel gain variation, it may require high signal to noise power ratio (SNR) to obtain a desired BER performance. As an alternative, the use of antipodal transceivers with power control was considered in fading channel [5], [6]. In recent, a simulation result was reported that 4-QAM portable radio systems with the use of simple closed-loop power control could improve the BER performance [7]. The use of variable rate and variable power QAM scheme was proposed for wireless communications [8]. Although this scheme can increase the spectral efficiency, it may require high implementation complexity.

Since the fading characteristics in fixed wireless channels are much milder than in mobile radio channel, it may be quite feasible to employ multi-level QAM schemes. It can be possible to obtain low BER performance for data transmission at moderate SNR if the QAM parameters are optimized with the aid of adaptive power control. It is known that the rainfall can introduce excess path attenuation which depends on the rain rate, the path length and the operating frequency. For example, when the rain rate is 100 mm/h, the path length is 5 km and the operating frequency is 2 GHz, 6 GHz and 10 GHz, the excess path attenuation due to rainfall based on ITU-R Rec. PN.530-5 becomes 0.04 dB, 2.2 dB and 11 dB respectively [9]. The use of adaptive power control can also be effective for fixed wireless access (FWA) systems to compensate the rain attenuation in particular when the operating frequency is very high.

For fast data transmission over a fixed wireless channel, we consider the use of multi-level QAM transceivers whose Tx power is adaptively controlled to compensate the fading effect. Compared to the conventional fading compensation technique in the Rx, the use of power control in the Tx can not only provide the improvement of averaged BER performance but also mitigate instantaneous performance degradation due to fading. The performance of the proposed QAM transceiver is analytically evaluated and verified by computer simulation in fixed wireless channels with flat fading and frequency selective fading.

Following Introduction, the system model including the QAM transceiver with Tx power control and fixed wireless channels are described in Sect. 2. Section 3 describes analytic design of the proposed QAM transceiver. In Sect. 4, the performance of the proposed transceiver is verified by simulation, considering the effect of power control parameters. Finally, Sect. 5 summarizes the conclusion.

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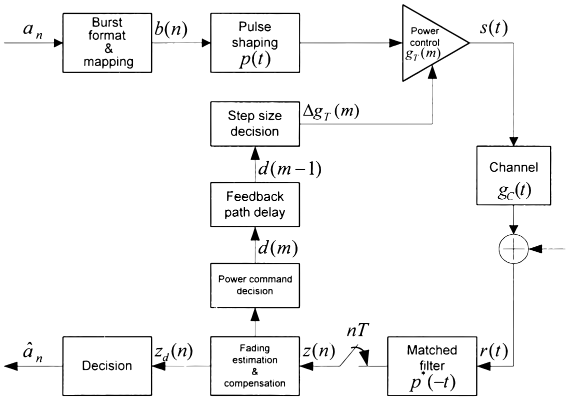

Fig. 1 Block diagram of a baseband equivalent QAM transceiver with Tx power control.

2. System Model

2.1 QAM Transceiver with Tx Power Control

Consider the use of a QAM transmission system over a fixed wireless channel. Figure 1 depicts the block diagram of a baseband equivalent QAM transceiver using adaptive power control. We propose the use of power control in the Tx to compensate gain variation due to fading.

Assuming that the gain of the transmitter is controlled at a frame rate equal to $1/T_F$, a baseband equivalent transmitted signal $s(t)$ can be represented as

$$\begin{aligned} s(t) &= \sum_{n=-\infty}^{\infty} b(n) g_T(m) p(t-nT_S) \\ &= \sum_{n=-\infty}^{\infty} b(n) \prod_{i=-\infty}^{m} \Delta g_T(i) p(t-nT_S), \end{aligned} \tag{1}$$

where T_S is the symbol duration, $b(n)$ denotes the nth symbol data, $g_T(m)$ is the gain of the Tx, $\Delta g_T(m)$ is the incremental gain of the Tx gain controller at time $t = mT_F$, $p(t)$ is the impulse response of the pulse shaping filter and $m = \lfloor n/M \rfloor$. Here M is the number of symbols for each frame and $\lfloor x \rfloor$ denotes the largest integer less than or equal to x.

For simplicity of description, assume a fixed wireless channel that has a two-ray multipath propagation with a delay of lT_S. Then, the impulse response of the channel can be represented by

$$g_C(t) = g_{C1}(t)\delta(t) + g_{C2}(t)\delta(t-lT_S), \tag{2}$$

where $g_{C1}(t)$ is the channel gain of the direct path, $g_{C2}(t)$ is the channel gain of the second multipath and $\delta(t)$ is Dirac delta function. The received baseband signal can be represented by

$$\begin{aligned} r(t) &= s(t) * g_C(t) + \nu(t) \\ &= g_{C1}(t)s(t) + g_{C2}(t)s(t-lT_S) + \nu(t), \end{aligned} \tag{3}$$

where $*$ denotes the convolution operation and $\nu(t)$ is additive white Gaussian noise (AWGN) term with a two-sided power spectral density of $N_0/2$. The output $z(n)$ of the matched filter sampled at time $t = nT_S$ can be represented by

$$\begin{aligned} z(n) = b(n) &\prod_{i=-\infty}^{m} \Delta g_T(i) g_{C1}(n) \\ &+ b(n-lT_S) \prod_{i=-\infty}^{m} \Delta g_T(i) g_{C2}(n) + \nu(n). \end{aligned} \tag{4}$$

We further assume that there is no intersymbol interference (ISI) for ease of description. At each frame time $t = mT_F$, the amount of power control is determined by comparing the power of the received pilot signal with that of the reference pilot signal in order to keep the power of the received signal constant. We assume that the Rx compensates the residual gain variation $g_o(n) = \prod_{i=-\infty}^{m} \Delta g_T(i) g_{C1}(n)$ due to imperfect power control in the Tx by using a conventional fading compensation technique [4]. Given a fading estimate $\hat{g}_o(n)$, the input to the decision device after residual fading compensation is given by

$$z_d(n) = b(n)\frac{g_o(n)}{\hat{g}_o(n)} + \frac{\nu(n)}{\hat{g}_o(n)}. \tag{5}$$

In the case of frequency selective fading (i.e., $g_{C2}(n) \neq 0$), an adaptive equalizer can be used for combating the ISI in time dispersive channels. When a decision feedback equalizer (DFE) is employed, its output can be represented as

$$\begin{aligned} z_d(n) = &\sum_{i=-N_F+1}^{0} C_i(n) z\left(n - \frac{iT_S}{p}\right) \\ &+ \sum_{k=1}^{N_B} B_k(n)\hat{b}(n-kT_S), \end{aligned} \tag{6}$$

where $C_i(n)$ and $B_k(n)$ are respectively the coefficients of the feed-forward filter (FFF) with N_F taps and the feedback filter (FBF) with N_B taps, p is the positive integer which means T_S-spaced equalizer or fractionally-spaced equalizer and $\hat{b}(n)$ is the decoded data. Since $C_0(n) = 1/g_o(n)$, the gain of the Tx can be determined from the main tap coefficient of the equalizer in order to keep the received signal gain of the direct path constant.

The command bit for power control is delivered to the Tx via a feedback path. Then, the transmitter calculates the incremental gain $\Delta g_T(m)$. This power control can effectively keep the power of the received signal nearly constant despite of the short term fading and the second multipath. Figure 2 illustrates the gains of the Tx signal, the received QAM signal and the channel, when the maximum power control range is set to ± 6 dB with respect to the average Tx power level.

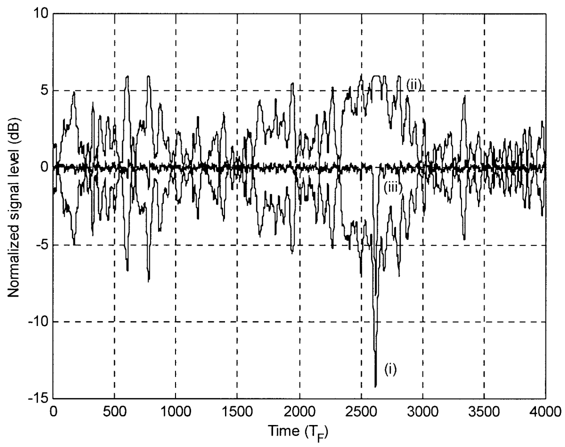

Fig. 2 Simulated waveform of power controlled signals of QAM transceivers in a typical fixed wireless channel: (i) Channel gain, (ii) Power controlled Tx signal, (iii) Rx gain after power control.

2.2 Fixed Wireless Channel

Although there have been very active studies on the characteristics of wireless channel, few results were reported in public domain. A fixed wireless channel model was presented for hilly environment by modifying the GSM HTX model [10]. Recently, the gain of a fixed wireless channel was modeled as a product of two Ricean random variables using experimental data measured in various channel environments [11].

The behavior of the channel gain was characterized by two gain terms; the fast gain variation and the slow gain variation terms. The fast gain term has a Ricean factor K_f whose gain variance σ_f is from 1.2 dB up to 15 dB with a Doppler frequency f_d of 0.3–0.4 Hz, and the slow gain term has a Ricean factor K_s whose gain variance σ_s is from 1 dB up to 12 dB with f_d of 0.03–0.04 Hz. The probability density function (pdf) of the channel gain is jointly determined by the above two gain variation terms. The Ricean factors of the two gain terms can be empirically determined. A typical fixed wireless channel can be modeled as one that has the fast gain term with σ_f of 10 dB with of 0.3 Hz and the slow gain term with σ_s of 8 dB with of 0.03 Hz, which means that K_f and K_s have an equal value of about 13.

When the channel has Ricean fading, its envelope density function $f_R(r)$ is given by [12]

$$f_R(r) = \frac{r}{\sigma^2}\exp\left\{-\frac{r^2+a^2}{2\sigma^2}\right\} I_0\left(\frac{ar}{\sigma^2}\right) = \frac{r}{\sigma^2}\exp\left\{-K-\frac{r^2}{2\sigma^2}\right\} I_0\left(r\sqrt{\frac{2K}{\sigma^2}}\right), \quad (7)$$

where a is the amplitude of the direct wave, σ^2 is the total power of indirect waves, K is the Ricean factor equal to $K = a^2/2\sigma^2$ and $I_0(\cdot)$ denotes the zeroth order modified Bessel function of the first kind. The normalized (power) gain density function $f_{G_R}(g)$ of a Ricean fading channel is expressed by

$$f_{G_R}(g) = (1+K)\exp\{-K-g(1+K)\} \cdot I_0\left(2\sqrt{g(K^2+K)}\right). \quad (8)$$

Since two Ricean random variables corresponding to the fast gain variation and the slow gain variation are statistically independent, the normalized (power) gain density function of a fixed wireless channel can be calculated by

$$f_G(g) = \int_0^\infty \frac{1}{x}(1+K_f)\exp\{-K_f-x(1+K_f)\} \cdot (1+K_s)\exp\left\{-K_s-\frac{g}{x}(1+K_s)\right\} \cdot I_0\left(2\sqrt{x(K_f^2+K_f)}\right) \cdot I_0\left(2\sqrt{\frac{g}{x}(K_s^2+K_s)}\right) dx. \quad (9)$$

In the presence of AWGN, the instantaneous SNR γ can be expressed in terms of the signal-to-noise power ratio.

Therefore, the SNR density function $f_\Gamma(\gamma)$ of the received signal over a fixed wireless channel can be calculated by

$$f_\Gamma(\gamma) = \int_0^\infty \frac{1}{x}\frac{1+K_f}{\gamma_0}\exp\{-K_f-x(1+K_f)\} \cdot (1+K_s)\exp\left\{-K_s-\frac{\gamma}{\gamma_0 x}(1+K_s)\right\} \cdot I_0\left(2\sqrt{x(K_f^2+K_f)}\right) \cdot I_0\left(2\sqrt{\frac{\gamma}{\gamma_0 x}(K_s^2+K_s)}\right) dx, \quad (10)$$

where γ_0 is the average energy per bit to the noise spectral density ratio defined by E_b/N_0.

3. Performance Analysis

The BER performance of Gray-coded QAM signals with a lattice type signal constellation in an AWGN channel can be approximated by [12]

$$P_G(\gamma) = a\,\mathrm{erfc}(\sqrt{b\gamma}) - c\,\mathrm{erfc}^2(\sqrt{b\gamma}) \approx a\,\mathrm{erfc}(\sqrt{b\gamma}), \quad (11)$$

where erfc(·) denotes the complementary error function, $a = 3/8$, $b = 2/5$, $c = 9/64$ for 16-QAM, $a = 7/24$, $b = 1/7$, $c = 49/384$ for 64-QAM, and $a = 15/64$, $b = 4/85$, $c = 225/2048$ for 256-QAM. If the gain variation is perfectly compensated only in the receiver, the BER of the QAM receiver in a fixed wireless channel can be calculated by

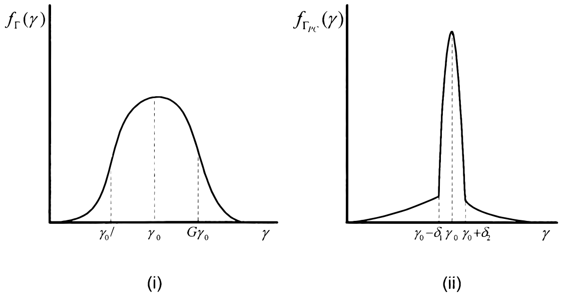

Fig. 3 SNR distribution of the received signal in a fixed wireless channel: (i) Without Tx power control, (ii) With Tx power control.

$$P_F(\gamma_0) = \int_0^{\infty} P_G(\gamma) f_\Gamma(\gamma) d\gamma = \int_0^{\infty} a \ \mathrm{erfc}(\sqrt{b\gamma}) f_\Gamma(\gamma) d\gamma. \quad (12)$$

In the proposed scheme, the transmitter power is controlled so as to maintain a constant power in the receiver. Assuming that the maximum power control range is $\pm 10 \log G$ (dB) with respect to the reference power P_T, the transmitted signal power P is controlled by

$$P = \begin{cases} GP_T, & g < 1/G \\ P_T/g, & 1/G \le g \le G \\ P_T/G, & g > G, \end{cases} \quad (13)$$

where g is the power gain of a fixed wireless channel.

To analyze the BER performance of the proposed scheme, it is required to calculate the pdf of the received SNR after power control. Figure 3 illustrates the distribution of the received SNR in a fixed wireless channel. When the variation of the channel gain of the direct path is within the power control range, neglecting the noise effect, the new incremental gain $\Delta g_T(m+1)$ can be approximately determined by

$$\Delta g_T(m+1) \approx \frac{1}{g_T(m) \mid g_{C1}(m) \mid}. \quad (14)$$

Since the gain of the Tx at $t = (m+1)T_F$ is

$$g_T(m+1) = g_T(m)\Delta g_T(m+1) = 1/ \mid g_{C1}(m) \mid, \quad (15)$$

the gain of the received signal is

$$g_o(m+1) = g_T(m+1) g_{C1}(m+1) = g_{C1}(m+1)/ \mid g_{C1}(m) \mid. \quad (16)$$

If Eq. (16) is expressed in dB scale, the gain of the received signal is equal to the difference between the consecutive channel gain terms. The distribution of $g_o(m)$ in a typical fixed wireless channel is illustrated in Fig. 4 when $f_d T_F = 0.0108$ and 0.0216. It can be seen that the pdf of $g_o(m)$ can be empirically approximated by a log-normal random variable. When the channel

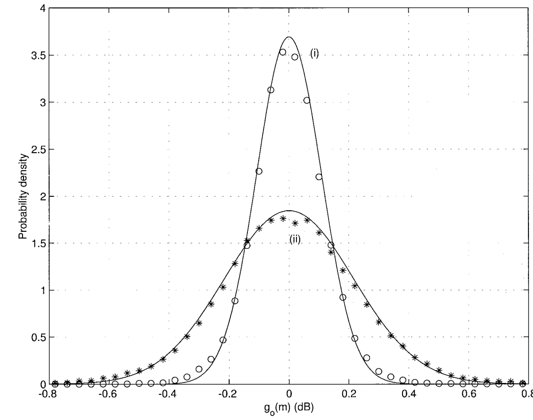

Fig. 4 Distribution of $g_o(m)$ in a typical fixed wireless channel: (i) solid line: log-normal distribution ($\sigma = 0.108$ dB), symbol ($\circ$): distribution of $g_o(m)$ ($f_d T_F = 0.0108$), (ii) solid line: log-normal distribution ($\sigma = 0.216$ dB), symbol ($*$): distribution of $g_o(m)$ ($f_d T_F = 0.0216$).

gain variation is smaller than the maximum power control range, the pdf of SNR of the received signal can be approximated by

$$f_{\Gamma_{PC}}(\gamma) = \frac{10}{\ln 10}\frac{1}{\gamma}\frac{C}{\sqrt{2\pi}\sigma}\exp\left\{-\frac{(10\log\gamma - 10\log\gamma_o)^2}{2\sigma^2}\right\}, \quad \gamma_0 - \delta_1 \le \gamma \le \gamma_0 + \delta_2, \quad (17)$$

where σ can be empirically determined by a value equal to $\sigma_f f_d T_F$, $C = \int_{\gamma_0/G}^{G\gamma_0} f_\Gamma(\gamma) d\gamma$, and δ_1 and δ_2 are the desired SNR range after power control.

When the gain variation of the channel is larger than the maximum power control range, the gain of the received signal is only level-shifted by an amount of the maximum gain control. The pdf of the received SNR beyond the maximum power control range is given by

$$f_{\Gamma_{PC}}(\gamma) = \begin{cases} f_\Gamma(\gamma/G)/G, & \gamma < \gamma_0 - \delta_1 \\ G f_\Gamma(G\gamma), & \gamma > \gamma_0 + \delta_2. \end{cases} \quad (18)$$

Therefore, the BER of QAM signals with the use of Tx power control in a fixed wireless channel can be approximately calculated by

$$P(\gamma_0) = \frac{a}{G}\int_0^{\gamma_0-\delta_1} \mathrm{erfc}(\sqrt{b\gamma}) f_\Gamma(\frac{\gamma}{G}) d\gamma + Ga\int_{\gamma_0+\delta_2}^{\infty} \mathrm{erfc}(\sqrt{b\gamma}) f_\Gamma(G\gamma) d\gamma + a\frac{10}{\ln 10}\frac{C}{\sqrt{2\pi}\sigma}\int_{\gamma_0-\delta_1}^{\gamma_0+\delta_2} \mathrm{erfc}(\sqrt{b\gamma})\frac{1}{\gamma} \cdot \exp\left\{-\frac{(10\log\gamma - 10\log\gamma_o)^2}{2\sigma^2}\right\} d\gamma. \quad (19)$$

4. Performance Evaluation

To evaluate the performance, the proposed scheme is applied to a TDMA based multi-level QAM transceiver operating in 2 GHz band with a channel spacing of 200 kHz and a symbol rate of 160 kbaud, where each frame has a length of 36 ms and comprises of sixteen time slots. Each time slot contains a preamble for frame synchronization and equalizer training, pilot signal for power control and user data. In the case of frequency selective fading, we assume a two-ray fixed wireless channel whose delayed path to the direct path average power ratio (DDR) is −15 dB and −20 dB, and the maximum delay time $\tau_{\max}$ is equal to T_S.

When a channel has frequency selective fading, we consider the use of a DFE that comprises of a $T_S/2$-spaced feed-forward filter (FFF) and a T_S-spaced feedback filter (FBF). Since it is assumed $\tau_{\max} = T_S$, the use of a DFE with a 4-tap FFF and 1-tap FBF can be sufficient to handle the channel dispersion. When the size of the training sequence is not long enough, it may not be appropriate to use a gradient-type training algorithm, such as the least mean square (LMS) method for equalizer adaptation. We consider the use of the recursive least squares (RLS) algorithm suitable for the use of a small number of preamble symbols for synchronization and equalizer training. Although the RLS algorithm has a computational complexity proportional to the number of the equalizer tap size, it can be applied to the proposed scheme without large implementation complexity because the equalizer needs the use of a small tap size.

4.1 Effect of the Maximum Power Control Range

The BER performance can be improved by increasing the dynamic range of the power control. The performance of the proposed QAM transceivers in a typical fixed wireless channel with flat fading is compared with that in AWGN channel in terms of the maximum power control range. As can be seen in Fig. 5, the use of the proposed power control scheme can improve the receiver performance about 3 dB and 6 dB at BERs of 10^{-3} and 10^{-5}, respectively, resulting in performance inferior to the AWGN channel case by less than 1 dB.

The larger the maximum power control range, the better the receiver performance. As the maximum gain of the power control increases, however, the average power of the transmitter also needs to be increased. Assuming perfect power control, the pdf of the transmitted signal power due to power control can be represented by

$$f_P(p) = \begin{cases} A\delta(p - \frac{P_T}{G}) + C\frac{P_T}{p^2} f_G(\frac{P_T}{p}) \\ \quad +B\delta(p - GP_T), \; \frac{P_T}{G} \le p \le GP_T \\ 0, \quad \text{otherwise,} \end{cases} \quad (20)$$

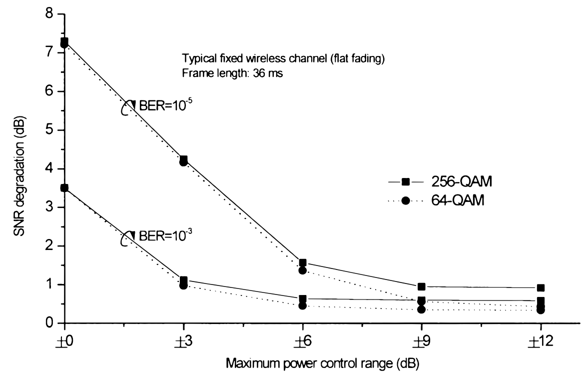

Fig. 5 Performance degradation of proposed QAM transceiver versus power control range.

Table 1 Increase of the average transmitter power due to power control in a typical fixed wireless channel.

Power control range	± 0 dB	± 3 dB	± 6 dB	± 9 dB	± 12 dB
$\overline{P}/P_T$	0 dB	0.83 dB	1.27 dB	1.38 dB	1.40 dB

where $A = \int_{G\gamma_0}^{\infty} f_\Gamma(\gamma)d\gamma$, $B = \int_0^{\gamma_0/G} f_\Gamma(\gamma)d\gamma$ and $C = \int_{\gamma_0/G}^{G\gamma_0} f_\Gamma(\gamma)d\gamma$. The average Tx signal power $\overline{P}$ can be calculated as

$$\begin{aligned} \overline{P} &= \int_{P_T/G}^{GP_T} p f_P(p) dp \\ &= A\frac{P_T}{G} + \int_{P_T/G}^{GP_T} C\frac{P_T}{p} f_G\left(\frac{P_T}{p}\right) dp + BGP_T. \end{aligned} \quad (21)$$

Table 1 summarizes the increase of the average Tx signal power due to the use of power control in a typical fixed wireless channel. Although the use of Tx power control significantly improves the BER performance, it may cause to increase the peak to average ratio (PAR) as well as the average Tx signal power. Also, it should be considered that the increase of power control range may require the use of a Tx amplifier with large gain control range.

It may be desirable to determine the power control range by considering the tradeoff between the effect of power control and the cost of Tx amplifier. For example, when the desired BER is 10^{-5}, the power control with ±6 dB and ±9 dB can improve the receiver performance by about 5.8 dB and 6.2 dB, respectively. Therefore, the use of ±6 dB can use a Tx amplifier similar to one used with no power control, but the use of ±9 dB requires a Tx amplifier with an increased gain control range of about 3 dB.

4.2 Effect of the Step Size for Power Control

We consider a power control scheme that incrementally updates the Tx power at a frame rate of $1/T_F$ using a single bit from the feedback path. The power control bit is generated by the receiver and is sent to transmitter via a feedback path. The power control bit $d(m)$ at frame time $t = mT_F$ is given by

$$d(m) = \begin{cases} 1, & P_{\rm ps} \le P_{\rm ref} \text{ or } \mid C_0(m) \mid \ge 1 \\ -1, & P_{\rm ps} > P_{\rm ref} \text{ or } \mid C_0(m) \mid < 1, \end{cases} \tag{22}$$

where $P_{\rm ps}$ is the received pilot symbol power, $P_{\rm ref}$ is the reference power of the pilot symbol and $C_0(m)$ is the coefficient of the equalizer main tap.

If the step size is fixed for power control, the Tx power is changed by $\pm\delta_T$ dB according to the power control bit. The use of a large step size may cause large gain fluctuation when the channel gain varies very slowly and the use of a small step size may not provide sufficient power control when the channel gain varies too fast. We consider the use of a multiplicative adaptation algorithm, called by constant factor delta modulation (CFDM) with a 2-bit memory [13], to adaptively change the step size. The incremental gain is determined by

$$\delta_T(m) = \begin{cases} 0.4\delta_T(m-1)(\text{dB}), \\ d(m)=\mp 1, d(m-1)=\pm 1, d(m-2)=\pm 1 \\ 0.9\delta_T(m-1)(\text{dB}), \\ d(m)=\mp 1, d(m-1)=\pm 1, d(m-2)=\mp 1 \\ 1.5\delta_T(m-1)(\text{dB}), \\ d(m)=\mp 1, d(m-1)=\mp 1, d(m-2)=\pm 1 \\ 2.0\delta_T(m-1)(\text{dB}), \\ d(m)=\mp 1, d(m-1)=\mp 1, d(m-2)=\mp 1. \end{cases} \tag{23}$$

The gain variation of the received signal with maximum power control of ± 6 dB and ± 9 dB is depicted in Fig. 6 when 256-QAM signal with $T_F = 36$ ms is sent over a typical fixed wireless channel. Here the gain variation means the standard deviation of the received signal gain. The use of a too small power control step size, say 0.1 dB, may not properly follow up gain change due to fast channel variation, resulting in large gain variation of the received signal. It can be seen that the use of adaptive step size can provide better performance than the use of a fixed step size.

4.3 Effect of the Frame Length

Since the Tx power is controlled at a frame rate, the receiver performance will be affected by the frame length. Figure 7 depicts the SNR degradation of the proposed

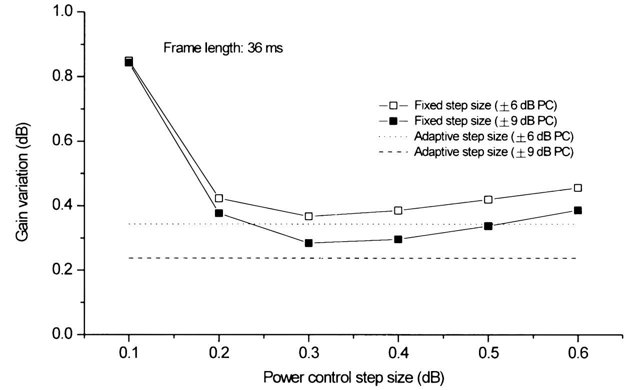

Fig. 6 Gain variation of 256-QAM signal power after the Tx power control.

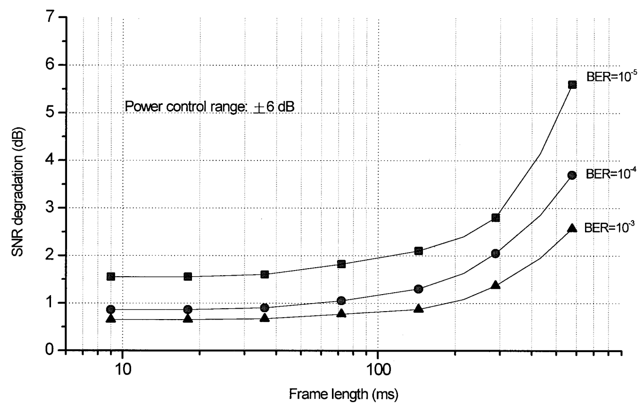

Fig. 7 Performance degradation of proposed 256-QAM transceiver versus frame length.

QAM transceiver in a typical fixed wireless channel when the maximum power control range is ± 6 dB and the adaptive step size is used. It can be seen that the performance is rapidly deteriorated if the length of the frame is larger than 70 ms. The use of a small frame length can improve SNR performance, but it may reduce the overall throughput performance due to increased overhead.

4.4 BER Performance of Multi-Level QAM

Figure 8 depicts the BER performance of a conventional QAM receiver without power control in a typical fixed wireless channel. In the case of flat fading, it can be seen that BER performance is approximately 7 dB inferior to that in an AWGN channel at a BER of 10^{-5}. In the case of frequency selective fading with DDR $= -15$ dB, however, high level QAM shows large performance degradation compared to low level QAM as the received E_b/N_0 increases.

Figure 9 shows the BER performance of the proposed QAM transceiver in a typical fixed wireless channel with flat fading. When the maximum power control of ± 6 dB, the performance is about 1.5 dB inferior

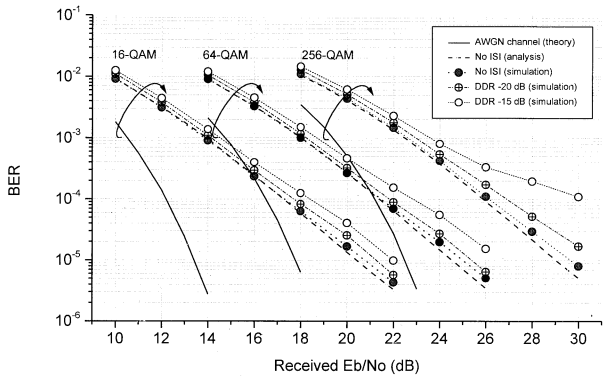

Fig. 8 BER performance without power control in a typical fixed wireless channel.

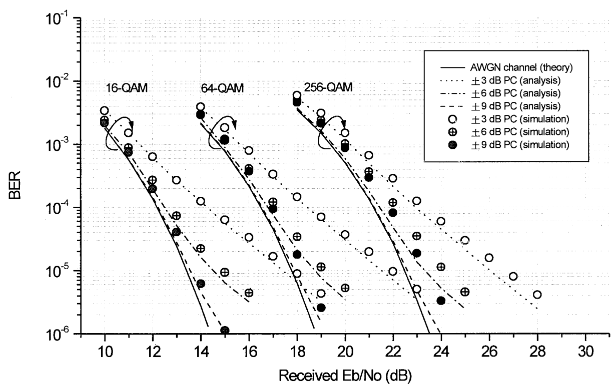

Fig. 9 BER performance of the proposed QAM transceiver in a typical fixed wireless channel with flat fading.

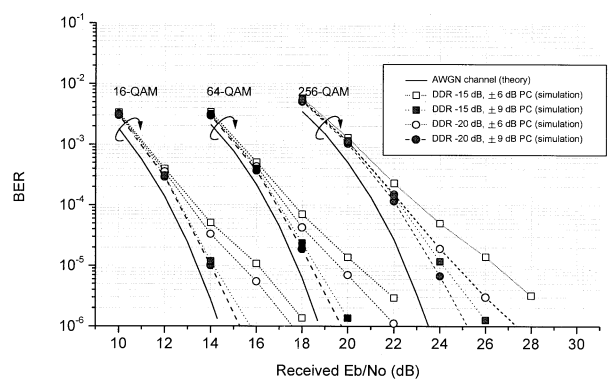

Fig. 10 BER performance of the proposed QAM transceiver in a typical fixed wireless channel with frequency selective fading.

to that in the AWGN channel. If the maximum power control range is increased to ± 9 dB, it can be seen that the BER performance is worse than that in the AWGN channel by a fractional dB. It can also be seen that the analytical results agree well with the simulation results although small BER performance deviation is observed due to imperfect gain control of residual gain variation when high level QAM is used.

Figure 10 shows the BER performance of the proposed QAM transceiver in a typical fixed wireless channel with frequency selective fading. Simulation results show that the performance of the proposed QAM transceiver is quite sensitive to the maximum power control range. When the DDR $= -15$ dB, the use of the maximum power control of ± 6 dB and ± 9 dB provides the BER performance inferior to that in the AWGN channel at a BER of 10^{-5} by 3–4 dB and 1–1.5 dB, respectively.

5. Conclusion

We have proposed the use of power control in the Tx for employment of multi-level QAM schemes in fixed wireless channels. The use of power control in the Tx, which adapts itself to the short-term fading, can provide improved BER performance. With the characterization of fixed wireless channel, the BER performance of the proposed multi-level QAM transceivers is analytically derived. Analytic and simulation results show that the BER performance of proposed QAM transceiver in a typical fixed wireless channel is inferior to that in AWGN channel by less than 1 dB.

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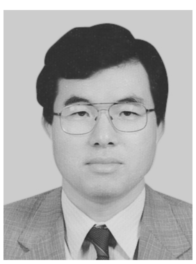

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