위성통신에서 도플러 및 PAR 효과에 강인한 새로운 OFDM 방식

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New OFDM Schemes Robust to Doppler and PAR Effects for Satellite Communications

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요 약

제로 포인트 콘스텔레이션(zero point constellation)을 이용하는 수정된 변조방식에 의해 평균 전력 절감 방법뿐만 아니라 콘스텔레이션 추정을 기반으로 하는 새롭고 효율적인 도플러 효과 보상 방법이 이론 해석과 적절한 시뮬레이션을 통해 제시된다. 제안된 방식은 고속 차량의 이동 단말에 대하여 전력을 제한하는 트랜스폰더(transponder)에 의해 중계되는 위성 OFDM 신호에 매우 효과적이다.

Key Words : OFDM, Doppler Effects, PAR, Nonlinear Distortion

ABSTRACT

A new and effective Doppler effects compensation method based on constellation estimation as well as an average power saving strategy by modified modulation scheme utilizing zero point constellation are presented with theoretical analysis and relevant simulations. The suggested schemes are proved to be very effective for satellite OFDM signals relayed by power limiting transponder for the mobile terminal on high speed vehicles.

I. INTRODUCTION

Orthogonal frequency division multiplexing (OFDM) is considered as one of the promising transmission techniques to accommodate ever growing mobile wireless multimedia traffics. It has already been accepted as a standard for several wireless system such as Digital Audio Broadcasting (DAB), asynchronous digital subscriber line (ADSL), wireless LAN and WiMax. However two major drawbacks, large Peak-to-Average power ratio and inter carrier interferences (ICI) due to frequency shift, may be obstacles for OFDM system transmitting the signal with power limiting amplifier to the mobile terminal on high speed vehicle such as airplane and express railroad train.

The movement of OFDM receiver on high speed vehicles causes Doppler frequency shifts which in turn destroy the orthogonal property of the sub-carriers, resulting serious ICI. Therefore Doppler shift should be mitigated or compensated for reliable communication between OFDM transmitter and the receiver on the high-speed mobile vehicles.

Unlike this paper where Line of Sight(LOS) condition between the satellite and mobile vehicle is assumed, the performance degradation due to Doppler effects of mobile OFDM system in multipath channel environment are addressed based on Jakes' model[1]. To guarantee the acceptable performance of the OFDM system with high mobility the compensation techniques are inevitable which will be able to restore the orthogonal properties between

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the sub-carriers. The Doppler effects compensation techniques are classified into two, one is the time domain techniques and the other is frequency domain techniques. The time domain compensations are performed before the FFT (Fast Fourier Transform) demodulation block by adjusting the demodulation frequency according to the frequency offset measured. On the other hand, the frequency domain compensation techniques are applied after the FFT demodulation block where the received symbols are represented as constellations in complex domain.

Joint time and frequency offset estimation based on ML (Maximum Likelihood) is addressed in [2], Based on this methods not only frequency offset but also time offset of the OFDM system can be adjusted. Since the Doppler frequency offset itself is one of the channel effects, the Doppler frequency offset compensation can be possible by estimating the channel response [3]. However channel responses contain not only frequency offset effects but also all the other channel effect such as multipath fading so the pure frequency offset value cannot be estimated separately. As a rather different approach to cancel the ICI due to frequency offset, the self cancellation scheme with the cost of bandwidth was proposed in [4], where one data symbol is modulated to subcarriers with predefined weighting that eventually brings self cancelling effects within subcarriers.

A novel Doppler effects compensation method proposed in this paper based on constellation estimation is considered not only efficient but also cost effective since no extra functional block is needed but little modification of receiver algorithm to implement such compensations.

One more thing has to be addressed in relaying the broadband signal via satellite transponder is the nonlinear problem of Traveling Wave Tube Amplifier (TWTA). Unlike the conventional M-PSK modulated signals with single carrier, the OFDM signals are very vulnerable to the nonlinear distortion due to high PAR (Peak to Average Ratio). Similarly this kind of vulnerability of non linear distortion is likely happen when multi carrier signals have to be loaded on the TWTA even though each carrier is modulated by M-PSK.

To avoid this kind of interference due to nonlinearity of TWTA, two approaches can be taken. One is loading the signal on the TWTA with sufficient back-off margin. This method is simple and very effective but the precious power of satellite transponder has to be sacrificed. The other method is to load the signal to the saturation point of TWTA with several compensation techniques.

As mentioned before PAR problem in OFDM system is caused by multicarrier transmission. Because of the amplifier nonlinearity, the peaks are distorted nonlinearly which generates inter modulation product causing both in-band distortion; responsible for Inter Symbolic Interference (ISI), and Out of Band (OOB) radiation; responsible for Adjacent Channel Interference (ACI) [5]. The nonlinearity of the TWTA causes spectral spreading of the OFDM signal besides it also causes wrapping i.e. amplitude and phase distortion and clustering of the signal constellation in each sub-channel at the amplifier output [6].

There are many techniques present for the remedial of the PAR problem [7] among which the major categories are signal distortion, signal scrambling and coding to reduce peak power[7-9].

Although several techniques have already been presented, most of them have some form of expenses, either as of bandwidth or system complexity. Clipping is the simplest, but it is a nonlinear process so causes in-band distortion, which degrades the SER performance, and OOB noise [7]. Signal scrambling [8] and coding [9] are used to generate the low PAR OFDM symbols.

However performance enhancements of the most works are brought by the expenses of bandwidth or system complexity. So method of MMZPSK(Modified M point Zero Padded PSK) is proposed to reduce the average power of the signal compared to normal PSK therefore the probability that the peaks are within the non-linear region of TWTA is also reduced. Without any expenses in terms of bandwidth or complexity, an enhancement in BER performance can be achieved based on MMZPSK modulation particularly there is no enough back-off margin of TWTA.

Throughout this paper, simple but effective schemes to reduce interferences because of the nonlinearity of power amplifier and Doppler frequency offset are presented with detail analytical derivations and relevant results of computer simulation. The suggested schemes are very efficient and cost-effective particularly for the broadband mobile internet networks linked by satellite transponder.

Ⅱ. OFDM SIGNAL ON SATELLITE TRANSPONDER

Satellite downlink OFDM signals are relayed to the mobile users via satellite transponder which has nonlinear characteristics in the vicinity of saturation. The discrete time domain OFDM signals are generated by taking Inverse Fast Fourier Transform (IFFT) to the frequency domain symbol loaded on the lth sub-carrier.

$$x_{s(g)}(n) = \frac{1}{\sqrt{N}} \sum_{l=0}^{N-1} X_{s(g)}(l) e^{j2\pi ln/N}, \tag{1}$$

Where subscript and designate satellite and ground respectively specifying the location of signal of interest and is the total number of subcarrier. To avoid the harmful inter-modulations between the subcarriers due to the nonlinearity of TWTA, TWTA is normally operated in back-off condition particularly when TWTA is loaded with multiple carriers. However for the OFDM signal with large peak to average power ratio (PAR), the normal 3 dB back-off of TWTA may not be enough to prevent the inter-modulation interference due to non-linearity of the TWTA.

To get the probability that the signal will be located in the vicinity of the saturation point of the of TWTA and so generate non-linear interferences, equation (4) is derived by substituting equation (3) into (2), that is introduced as the complementary cumulative distribution function (CCDF) of the PAR [10].

$$\Pr(PAR > \gamma) = 1 - \exp[-\sqrt{\frac{\pi\gamma}{3}} N e^{-\gamma}], \tag{2}$$

$$PAR = \frac{\max|x_s(n)|^2}{E\left[|x_s(n)|^2\right]}, \tag{3}$$

$$P_{r(}\max|x_s(n)|^2 > \gamma E\left[|x_s(n)|^2\right] = 1 - \exp[-\sqrt{\frac{\pi\gamma}{3}} N e^{-\gamma}], \tag{4}$$

Based on equation (4), probability that the maximum power of the signal are in the nonlinear region of amplifier therefore generate inter-modulation interferences can be estimated. Where the back-off power margin is set as γ. So beyond the threshold value set by, $\gamma_{th} = \gamma \cdot E\left[|x_s(t)|^2\right]$, power amplifier is operating nonlinearly to generate the inter-modulation interferences between the subcarriers. The derivation of the m th order inter-modulation products when the number of carriers N is large, which is reasonable assumption for normal mobile OFDM system such as WiMax, is given as [11],

$$M_{m,N} = M(\underbrace{1,1,1,...,1,}_{\frac{m+1}{2}} \underbrace{-1,-1,...,-1,}_{\frac{m-1}{2}} \underbrace{0,0,...,0}_{N-m}), \tag{5}$$

Where M is expressed as,

$$M(k_1,k_2,...,k_N) = \int_0^\infty r \prod_{i=1}^{N} J_{k_l}(|X(l)|r)dr \int_0^\infty \rho G[\rho] \exp\{jF[\rho]\} J_1(r\rho) d\rho, \tag{6}$$

where $G[\cdot]$and $F[\cdot]$ represent the amplitude and phase transfer characteristics of the TWTA, respectively, J_k is Bessel function of the first kind of order k. In order to evaluate the inter-modulation power, the distribution of the inter-modulation products is derived as in [12]:

$$D_{m,N}(\nu) = \frac{N^{m-1}}{\left(\frac{m+1}{2}\right)!\left(\frac{m-1}{2}\right)!(m-1)!} \sum_{k=0}^{\lfloor m/2-\nu \rfloor} (-1)^k \binom{m}{k} \left(\frac{m}{2} - k - \nu\right)^{m-1}, \tag{7}$$

Where normalized frequency ν is associated specific subcarrier frequency ω_l by:

$$\nu = \frac{\omega_l - (\omega_0 - \omega_{N-1})/2}{\omega_0 - \omega_{N-1}}, \tag{8}$$

and $\lfloor \cdot \rfloor$ above the summation sign means the integer part of the inside value. Finally the most dominant the 3th and 5th interference power at ν due to inter-modulation can be expressed as,

$$I_\nu = \sum_{m=3,5} D_{m,N}(\nu)|M_{m,N}|^2, \tag{9}$$

Since $M_{1,N}$ is representing the desired signal, the output Signal to Noise plus Interference Ratio (SNIR) of the overdriven TWTA can be expressed as,

$$\frac{S}{N} = \frac{|M_{1,N}|^2}{N + I_\nu}, \ N: AWGN, \tag{10}$$

In order to avoid such inter-modulation interferences, the TWTA is operated below the saturation point specified by back-off of the TWTA. In equation (4), γ acts as back-off value of the TWTA so the probability of the saturation of maximum value can be reduced by increasing the back-off value γ. However since the normal PARs of the commercial OFDM systems are greater than 10 dB, it is not feasible to design the TWTA operation without any saturation. The back-off value γ should be decided by compromising between the power efficiency of TWTA, which is measured by the ratio of real operating power to the maximum power available, and interference level of inter-modulation. The TWTA power of satellite is limited and expensive resources so the back-off value of the TWTA should be reduced as possible.

The regions of operation of TWTA can be categorized into linear operation region and nonlinear delimited by the value of γ_{th}. Therefore the total output SNIR of satellite TWTA backoffed by γ can be expressed as follow:

$$\left(\frac{S}{\tilde{N}}\right)_s = [\left\{P(\breve{\gamma}_{th})\left(\frac{|G(|X_s(l)|)|^2}{N}\right)\right\}^{-1} + \left\{P(\hat{\gamma}_{th})\frac{S}{\tilde{N}}\right\}^{-1}]\}, \tag{11}$$

where

$$P(\check{\gamma}_{th}) = \Pr(\max|x_s(n)|^2 \le \gamma_{th})$$
$$P(\breve{\gamma}_{th}) = \Pr(\max|x_s(n)|^2 \le \gamma_{th})$$

Ⅲ. DOPPLER EFFECTS ON THE SATELLITE DOWNLINK OFDM SIGNAL

The downlink OFDM signal with SNIR as in equation (11) is transmitted from the satellite to the mobile terminals which may be on the high speed vehicles such as airplane or express railroad train. The received subcarriers can be represented as,

$$X_g(l) = X'(l) + i_\nu'(l) + N_g, \tag{12}$$

Where $X_g'(l)$ and $i_\nu'(l)$ are attenuated terms of $P(\check{\gamma}_{th})G(|X_s(l)|) + P(\hat{\gamma}_{th})|M_{1,N}|$, $P(\hat{\gamma}_{th}) \cdot \sqrt{I_\nu(l)}$ respectively because of free space loss, N_g designates AWGN at receiver terminal and l of $i_\nu'(l)$ is the denormalized subcarriers index of ν. The ICI due to Doppler frequency shift and the desired signal on the l th carrier can be expressed in equation (13) and (14)

$$i_g(l) = \sum_{j=0, j\neq l}^{N-1} X_g(j) H(j-l)$$
$$\cong \sum_{j=0, j\neq l}^{N-1} X_g'(j) H(j-l) + \sum_{j=0, j\neq l}^{N-1} i_\nu'(l) H(j-l), \tag{13}$$

$$S_g(l) = X_g(l) H(0), \tag{14}$$

Where Doppler interference function $H(\cdot)$ is expressed as [4],

$$H(j-l) = \frac{\sin(\pi(j-l+\varepsilon))}{N\sin\left(\frac{\pi}{N}(j-l+\varepsilon)\right)} .\exp\left(i\pi\left(1-\frac{1}{N}\right)(j-l+\varepsilon)\right)$$

Where ϵ is Doppler frequency offset normalized by the subcarrier separation. Therefore the SNIR of OFDM signal on the lth carrier is derived as,

$$\left(\frac{S}{N}\right)_g = \frac{|X_g(l)H(0)|^2}{\sum_{j=0, j\neq l}^{N-1} |X_g'(j)H(j-l)|^2 + \sum_{j=0}^{N-1} |i_\nu'(l)H(j-l)R|^2 + N_g}, \tag{15}$$

Notice that nonlinear inter-modulation interferences by the saturated TWTA propagate even into the demodulation phase as the second term of the denominator of equation (15) through Doppler interference function $H(j-l)$.

Ⅳ. MODULATION SCHEME

Large PAR and frequency shift due to Doppler effects are considered the main disadvantages of OFDM system. Particularly for the OFDM signals transmitted by satellite to the user terminal on the fast vehicles, the effects of those two adverse factors are even aggravated because of power limiting characteristics of the nonlinear TWTA.

The inter-modulation due to the nonlinearity of the TWTA act as a interference by itself and generates another secondary interferences through Doppler interference function expressed in equation (15), so some measures to alleviate the nonlinear inter-modulation interferences should be taken to secure BER performance. In order to diminish the interference power due to inter-modulation, $P(\hat{\gamma}_{th})$ in equation (11) has to be lowered. Obviously, $P(\hat{\gamma}_{th})$ will be lower if the back-off threshold value $\gamma_{th}$is increased. To satisfy link budget of satellite communication with the augmented back-off value, the required maximum power of TWTA should be increased with the same amount of back-up value, which is a redundancy in power of TWTA, to maintain the same effective isotropic radiated power (EIRP). The larger redundancy in TWTA power not only reduces the power efficiency of TWTA itself but also may be an obstacle to design the optimum satellite bus power system.

Referring the related equation (4) and (11), the possible ways to lower the $P(\hat{\gamma}_{th})$ are diminishing the PAR value or enlarging the back-off γ_{th}. Although there are many research works[7-9] to minimize the PAR by controlling constellation, or coding technique, results are not so effective in the respect of system complexity and bandwidth efficiency. Without any expenses in terms of bandwidth and system complexities a more practical and simple approach may be devising some modulation scheme which can lower the value $E\left[|x_s(n)|^2\right]$ and consequently make the available back-off margin γ_{th} larger for the same TWTA. Such a modulation scheme called M-point Zero Padded Phase Shift Keying (M-ZPSK), proposed in [13], may be suitable for this purpose. The signal constellations of M-ZPSK are expressed as,

$$Z_P^M = \left[0,\alpha,\alpha\xi,\cdots,\alpha\xi^{M-2}\right], \tag{16}$$

where $\alpha=\sqrt{M/(M-1)}$, $\xi=\exp(i2\pi/(M-1))$.

To satisfy the primary concern of reducing the average power as well as the compatibility with the other normal M-PSK modulation, which has constant magnitude irrelevant of M, modified M-ZPSK (MM-ZPSK) is proposed as,

$$MZ_P^M = \left[0,1,\xi,\cdots,\xi^{M-2}\right], \tag{17}$$

According to the defined constellations, the minimum Euclidean distance of conventional QPSK, Q-ZPSK and MQ-ZPSK will be as 1.4142, 1.1547 and 1.0 respectively. On the other hand, average power of Q-ZPSK and MQ-ZPSK in reference to the QPSK(0 dB) can be calculated by Parseval's theorem as (-3.01 dB), (-4.25 dB) respectively, so proposed MQ-ZPSK has approximately 4.25 dB less in average power compared to the conventional QPSK modulation so the same value can be used as increased back-off of TWTA. The rationale behind how the MQ-ZPSK can reduce average power is rather obvious since by adopting MQ-ZPSK scheme, one symbol out of four is represented by empty carrier i.e. zero, so the same number of data can be transmitted with three quarter power of the conventional QPSK modulation. As you can see in Table 2 and 3, the mean value of the time domain OFDM signal is reduced by adopting MQ-ZPSK, however the PAR characterized by adding multiple subcarriers are almost same.

Via the same TWTA, if the conventional QPSK/OFDM is transmitting without back-off from the saturation, the available back-off margin for the MQ-ZPSK will be around 4.25dB. Therefore the probability of saturation for MQ-ZPSK is reduced accordingly depending on the equation (4).

V. DOPPLER SHIFT COMPENSATION

To clarify the exact effects of the Doppler shift on the received data symbol, any other interferences and AWGN are excluded by assuming $P(\check{\gamma}_{th})\cong 1, P(\hat{\gamma}_{th})\cong 0$ and $N_g \cong 0$. Then equation (12) can be simplified and redefined as

$$X(l) \cong X_g'(l), \tag{18}$$

The received signal on l-th subcarrier under the effects of Doppler shift channel can be expressed as [4],

$$Y(l) = X(l)H(0) + \sum_{j=0, l\neq j}^{N-1} X(j)H(j-l), \tag{19}$$

Since the significant interferences to the l-th subcarrier are coming from the neighboring few subcarriers, equation (19) can be approximated with two adjacent terms in both sides,

$$\begin{aligned} Y(l) &\cong X(l)H(0)+X(l-1)H(-1)+X(l+1)H(+1) \\ &+X(l-2)H(-2)+X(l+2)H(+2), \end{aligned} \tag{20}$$

To simplify the equation (20) further, the first and second terms of $H(j-l)$ are manipulated using basic trigonometric equations and approximated as,

$$\frac{\sin(\pi(j-l+\varepsilon))}{N\sin\left(\frac{\pi}{N}(j-l+\varepsilon)\right)} \cong \frac{\sin\pi\varepsilon}{N\sin\left\{\frac{\pi}{N}(j-l)\right\}} \cong \frac{\varepsilon}{j-l}$$
$$(j-l\neq 0), \tag{21}$$

$$\exp(i\pi(1-1/N)(j-l+\varepsilon)) \cong \exp(i\pi(l-k))\cdot\exp(i\pi\varepsilon), \tag{22}$$

After substituting equation (21), (22) into (20) and dividing both sides by $X(l)$, equation (20) becomes as,

$$\frac{Y(l)}{X(l)} \cong \exp(i\pi\varepsilon)$$
$$+\left\{\frac{-X(l-1)-X(l+1)}{X(l)}\right\}\varepsilon\exp(i\pi\varepsilon)$$
$$+\left\{\frac{X(l-2)+X(l+2)}{X(l)}\right\}\frac{\varepsilon}{2}\exp(i\pi\varepsilon), \tag{23}$$

Two composite random sequences are defined as,

$$Z_1=\{(-X(l-1)-X(l+1))/X(l)\}, \tag{24}$$
$$Z_2=\{(-X(l-2)+X(l+1))/X(l)\}, \tag{25}$$

Since $E[Z_1]=0, E[Z_2]=0$, the expectation value of equation (23) will be

$$E[Y(k)/X(K)] \cong \exp(i\pi\varepsilon), \tag{26}$$

Therefore the center of constellation of $Y(l)$ will be shifted $(\pi\varepsilon)$ radian from that of $X(l)$ on complex unit circle and the dispersiveness of $Y(l)$ is determined by the variance of another composite random sequence defined as $Z_1\varepsilon\exp(i\pi\varepsilon)+Z_2(\varepsilon/2)\exp(i\pi\varepsilon)$, that is also proportional to ε and $\exp(i\pi\varepsilon)$.

Based on the result of statistical estimation as in equation (26), Doppler effects compensation can be performed in frequency domain as well as time domain. Frequency domain compensation, classified as post-FFT compensation technique, can be done by the equation as follows:

$$Y_c(l)=\frac{Y(l)}{E[Y(l)/X(l)]} \cong \frac{Y(l)}{\exp(i\pi\varepsilon)}, \tag{27}$$

With this compensation, the center of constellation of $Y(l)$ move to the nominal center of $X(l)$ however the dispersiveness of $Y(l)$, still dominated by $Z_1\varepsilon+Z_2(\varepsilon/2)$, an not be corrected. The other compensation technique performed in time domain (Pre-FFT) as in equation (28) is able to correct not only the biased constellation center but also the dispersiveness of constellation.

$$y_c(l)=y(l)\bullet\{IFFT[E[Y(l)/X(l)]^{-1}]\}$$
$$\cong y(l)\bullet\{IFFT[\exp(-i\pi\varepsilon)]\}, \tag{28}$$

Naturally this time domain approach shows better performance than frequency domain since Doppler frequency shift is fundamentally corrected in time domain before FFT resulting no bias and dispersiveness of constellation of received symbol. By multiplying $[\exp(-i\pi\varepsilon)]$ to the time domain signal, the shifted centers of each subcarrier are moved back to the nominal carrier frequency so all the other receiver blocks are not affected functionally.

Since the proposed method relies on the statistical data of the received symbol loaded on the subcarriers, it is reasonable to delay one frame to apply this proposed method. Considering the WiBro system simulated, the channel bandwidth occupied by subcarriers are 8.75 MHz therefore time duration of one frame is approximately 0.11 μsec. The potential velocity change of even the fastest airplane and express railroad train during that time span is negligible so the estimated Doppler shift parameter of the previous frame can be used to adjust current frame symbol data without any delay.

Ⅵ. PROS AND CONS OF MQ-ZPSK FOR SATELLITE MOBILE OFDM

Basic concept of MQ-ZPSK is using the null point as one of the constellations to transmit symbols. By doing this, the transmitting average power is saved but Euclidean distance between the constellations is sacrificed which is critical for reliable symbol detection in noisy channel. So in AWGN channel, obviously the conventional QPSK shows better BER performance than MQ-ZPSK. However if mobile OFDM signal has to be relayed by power limiting nonlinear satellite transponder thus the nonlinear interferences of the conventional QPSK are greater than that of MQ-ZPSK then the BER performances of MQ-ZPSK are expected better than those of QPSK. For the same PAR, the higher the average power is the higher the probability of signal will be in the non linear region. By Parseval's theorem the average power of MQ-ZPSK 4.25 dB less than that of normal QPSK. That means the back-off margin of MQ-ZPSK is 4.25 dB greater than that of QPSK. To make decision of preference between MQ-ZPSK and normal QPSK in both non-linear interference and AWGN environment, primary factor dominating BER performance should be clarified. Which one is more harmful to BER performance, non-linear interference or AWGN? If the nonlinear interferences are severe compared to AWGN then MQ-ZPSK are preferable but QPSK is preferable for the opposite case.

Doppler shift compensation method proposed in previous section can be applied effectively to MQ-ZPSK and QPSK as well. Besides the advantage of MQ-ZPSK over QPSK in the respect of average power, MQ-ZPSK has robustness to ICI due to Doppler shift compared to QPSK since a quarter of the adjacent subcarriers are likely emptied which otherwise bring ISI to the neighboring subcarrier according to the equation (20).

In fact the inter-modulation interference due to nonlinearity of TWTA and ISI from Doppler shift are independent to each other, nevertheless by adopting MQ-ZPSK the nonlinear interference and ISI are inter-related over MQ-ZPSK.

Ⅶ. SIMULATION RESULTS AND DISCUSSION

Computer simulations are performed based on specification of WiBro (Wireless Broadband) system parameters of which are summarized in table 1.

The statistical values of QPSK and MQ-ZPSK signals used in simulation are summarized in table 2.

The mean power difference of MQ-ZPSK and QPSK modulation is measured statistically as 4.3 dB which is almost same as the analytical calculation based on Parseval's theorem. However the PAR has no difference therefore Complementary Cumulative Probability (CCP) of the signals for both modulations are almost same if the threshold value $\bar{\gamma}$ from the mean is set equal as can be seen table 3. Since the average power of MQ-ZPSK is 4.3 dB lower than that of QPSK, therefore the probability that some signal is located in nonlinear region of TWTA for MQ-ZPSK is lower accordingly.

Table 1. System parameters of Simulation

Parameters	Value
FFT Size	1024
Nominal Channel Bandwidth	8.75 MHz
Sampling Frequency	10 MHz
Subcarrier Spacing	9.7656 KHz
Number of Sub Carriers	865
Carrier Frequency	2.3 GHz

Table 2. Statistical Signal Parameters during one Frame

	Mean(dB)	Max(dB)	PAR(dB)
QPSK	33.3	42.1	8.7
MQ-ZPSK	29.0	37.9	8.9

Table 3. Complementary Cumulative Probability

CCP	$\bar{\gamma}$ =3	$\bar{\gamma}$ =7	$\bar{\gamma}$ =11	$\bar{\gamma}$ =13(dB)
QPSK	0.35	0.2	0.09	0.04
MQ-ZPSK	0.35	0.21	0.09	0.03

For the sake of simulation of non-linear TWTA, the typical nonlinear gain curve is fitted by seventh order odd polynomial i.e., $\nu_o = \alpha_1\nu_i + \alpha_3\nu_i^3 + \alpha_5\nu_i^5 + \alpha_7\nu_i^7$ by least square fitting algorithm, where α_1=1.9817, α_3 =-1.3638, a_5 =0.4507 and α_7=-0.0524. The resultant non-linear curve modeled by those polynomials is shown in Figure 1 as dotted line.

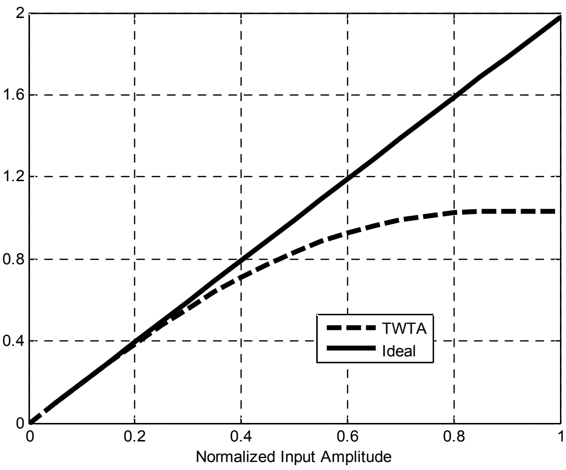

Figure 1. Non linear input/output curve of TWTA

To show the advantage of MQ-ZPSK over QPSK in terms of saving average power, simulations for MQ-ZPSK and QPSK are performed without back-off. In the figure 1, the threshold input voltage of the nonlinearity can be assumed 0.4 so the QPSK signal is to be driven at the

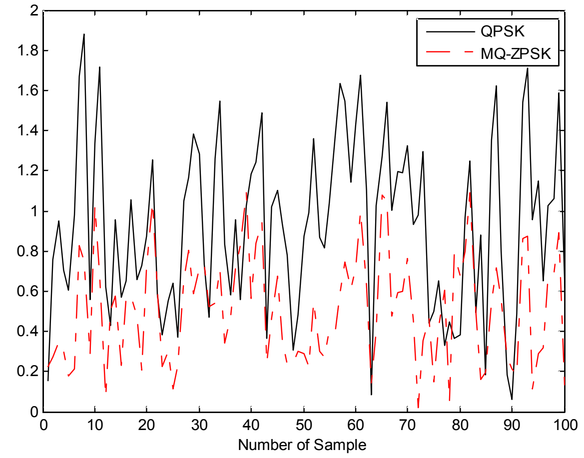

Figure 2. Normalized Amplitudes of QPSK and MQ-ZPSK

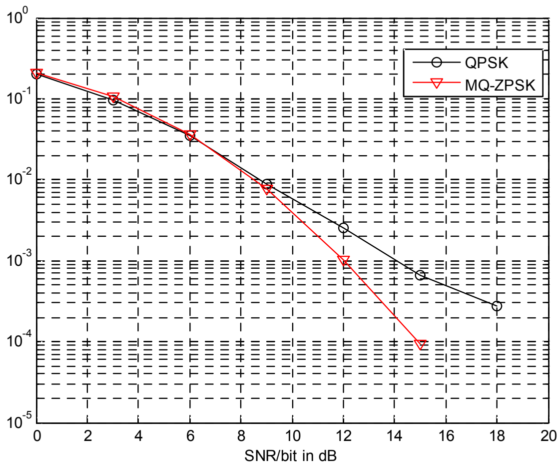

Figure 3. BER performance of MQ-ZPSK and QPSK without back-off

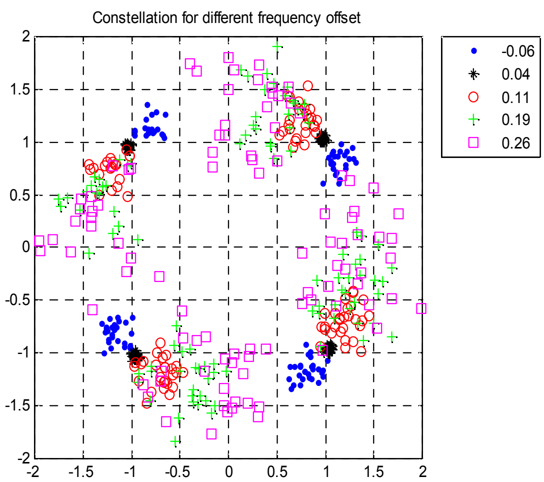

Figure 4. Constellation shift for different ε

average input voltage 0.4, while the MQ-ZPSK is driven at 4.3 dB less average power which is equivalent 0.2436 in input average voltage. The typical time domain signal both QPSK and MQ-ZPSK are shown in Figure 2.

AS expected, the BER performance of MQ-ZPSK is better that of QPSK as can be seen in Figure 3.

Figure 4 shows constellation figures for different frequency offset values in a very low AWGN environment (SNR = 40dB). Angular rotation and the dispersiveness of the constellations are very proportional to frequency offset value ε, which is set to -0.06, 0.04, 0.11, 0.19 and 0.26. The previous arguments raised by (23) and (26) are certified by the simulation results of Figure 4.

Figure 5 shows the compensated constellation by frequency domain and time domain compensation methods for QPSK signal. The result of 'NONE' shows the constellation without compensation. By frequency domain compensation method, designate by 'FREQ', only the constellation center is corrected wile both constellation center and the dispersiveness are corrected by the time domain method designated by 'TIME'.

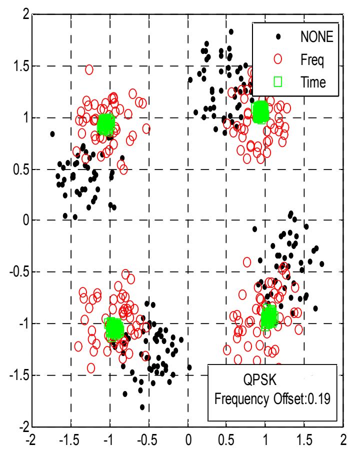

Figure 5. Compensated Constellations (QPSK)

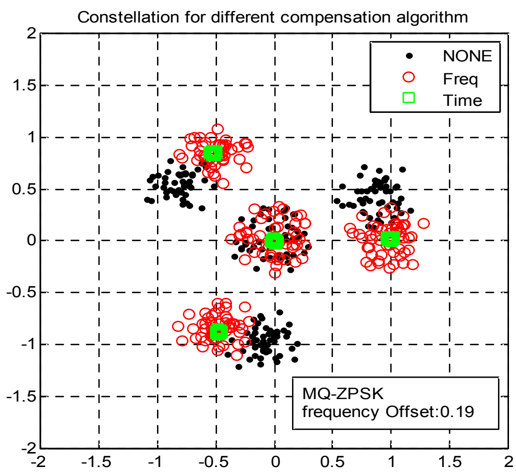

Figure 6. Compensated Constellations(MQ-ZPSK)

Figure 6 shows the same simulation results for MQ-ZPSK that is identical to the results of Q-ZPSK except the position of the constellations.

Figure 7 shows BER performance for QPSK and MQ-ZPSK in different Doppler frequencies without any compensations. When Doppler frequency shifts are given as 500 Hz and 800 Hz, MQ-ZPSK is superior to QPSK, on the contrary for 200 Hz of Doppler shift, QPSK has better performance. This contradiction is explained by the fact that the relative interference intensity from the adjacent carriers versus AWGN becomes stronger as the Doppler frequency shift increases.

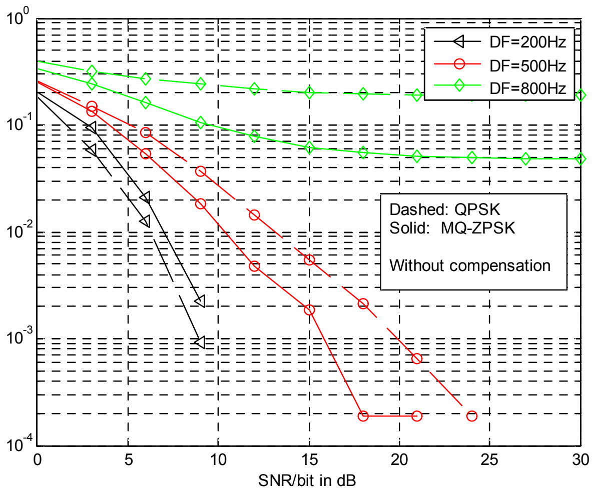

Figure 7. BER Performance for Modulations

In other words, when the Doppler frequency shifts are great, the effects of reduced interference due to the 25% emptiness of MQ-ZPSK subcarriers overrule the effects of diminished Euclidean distance of MQ-ZPSK in BER performance. On the contrary, for the case of smaller

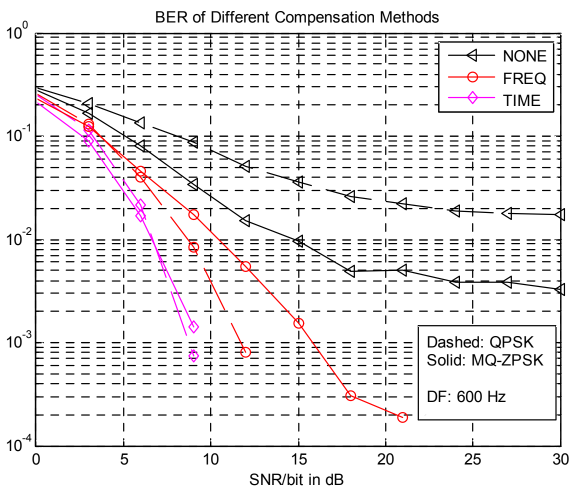

Figure 8. BER Performance with Compensations

Doppler frequency shift, effects of the diminished Euclidean distance becomes dominant so the reversion in BER performance occurs in the range of lower Doppler frequency shift.

Figure 8 shows enhancements of BER performance for two different modulation schemes in relatively severe Doppler frequency shifting environment. The BER performance graph without any compensation is unacceptably bounded at Doppler frequency shift of 600 Hz. However as can be noticed, frequency domain compensation method, and time domain compensation method are very effective to both QPSK and MQ-ZPSK and among two compensations, time domain method shows better performance by compensating the distorted constellation in terms of shift as well as the dispersiveness.

Ⅷ. CONCLUSION

Providing broadband OFDM satellite internet satellite services via nonlinear satellite transponder for high speed vehicles, the both high PAR and Doppler frequency offset problems of OFDM may be obstacles to secure reliable BER performances. In this paper to alleviate inter-modulation interferences because of the nonlinearity of TWTA, MQ-ZPSK is proposed which can provide 4.3 dB back-off effects compared to conventional QPSK.

A novel Doppler effects compensation schemes based on the constellation estimation are proposed which are applicable both QPSK and MQ-ZPSK. The compensation methods are classified into time domain method and frequency domain method whether the compensation techniques are applied before FFT or after FFT respectively. The time domain compensation shows better performance because both rotation and dispersiveness of the received constellations are corrected.

Unlike the other methods suggested to combat PAR and Doppler shift problems of OFDM, the proposed methods in this paper can be implemented without expenses in terms of system complexities and frequency bandwidth.

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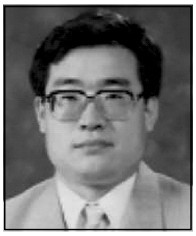

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