비선형 전력증폭기로 인한 CE-CPSK 변조된 DS-CDMA 초기동기 시스템의 성능분석

김 성 철*

Performance Analysis of a CE-CPSK modulated code acquisition system for nonlinear amplified DS-CDMA signal

Seong-Cheol Kim*

요 약

본 논문에서는 송신기의 전력효율을 고려하여 C급 전력증폭기를 사용함으로 인한 증폭기의 비선형성의 영향을 극복하기위해 일정진폭 특성과 연속위상특성을 갖는 CE-CPSK 변조 직접 대역확산 송수신기를 제안하였다. 직접 대역확산 수신기의 초기동기 성능을 평균 동기획득시간, 검출확률을 통해 기존의 BPSK 변조방법과 CE-CPSK변조 방법에 대해 다중사용자환경하에서의 성능을 비교분석하였다. 비선형성을 지닌 채널 환경 하에서 제안한 CE-CPSK변조방식이 기존의 BPSK변조방식에 비해 부대엽 스펙트럼이 상당히 감소되는 것을 알 수 있었으며 코드 동기 획득 성능 또한 우수함을 알 수 있었다.

ABSTRACT

In this paper, to overcome the effect of nonlinear power amplifier, CE-CPSK(Constant Envelope Continuous Phase Shift Keying)modulation method which has the constant envelope and continuous phase characteristics is proposed. The code acquisition performances of the CE-CPSK modulated system and BPSK system are analyzed. And the performance is evaluated in terms of mean acquisition time and detection probability with the output backoff of the amplifier in multiuser environment. Results performed in a channel having a high degree of AM nonlinearity(AM-to-AM conversion), have shown that the CE-CPSK signal, when passed through class C-nonlinear Amplifier, has a spectrum with greatly reduced sidelobes compared to the conventional BPSK signal. The code acquisition performance of a CE-CPSK modulation system is better than that of BPSK system with nonlinear amplified multiuser environments.

키워드

CE-CPSK, detection probability, AM nonlinearity, mean acquisition time, backoff

Ⅰ. INTRODUCTION

After cellular system was introduced to wireless communications, the proliferation of wireless communication is expected to continue and to stimulate service providers to offer a wide range of services. Especially to satisfy increasing demands and capacities, come many portable design challenges that must be met for manufactures to maintain low cost, small size, light weight, high quality cellular terminals. These challenges reside in both the base-band and RF section of the terminal. In particular to the RF section, spectral efficiency is achieved at the expense of requiring a linear

* 우송대학교 철도전기 · 정보통신학부

접수일자 : 2005. 8. 9

transmit power amplifier. But the linear power amplifier cannot meet the low-power consumption requirements. To meet the requirements, as the efficiency increases, so does the distortion(nonlinear characteristics) necessitate the use of compensation schemes. Much work has been accomplished, and continues to be done, in the area of PA linearization techniques[1]-[2]. Meanwhile, band-limited systems have finite rise and falling times, so filtering introduces envelope fluctuations into modulated PSK systems. When these time-varying envelope signals are fed into a nonlinear amplifier, they undergo nonlinear effects such as AM-AM, AM-PM conversion, these introduce the spectral spreading which cause the performance degradation in adjacent channels. To overcome these problems, various constant-envelope modulation techniques have been proposed[3]-[4]. However these techniques have the problem that supplementary signal reducing the envelope variation has DC component, this component appeared line spectrum in the output spectrum. This line spectrum not only increases interference with other system but makes demodulation difficult.

Meanwhile one of the primary functions of the Direct-sequence spread-spectrum(DS/SS) receiver is synchronization of the incoming PN sequence and the locally generated PN sequence. Initially, a coarse alignment of the two PN sequences is produced to within a chip timing offset. This process is referred to as PN acquisition. Because communication can not be taken place before acquisition has been achieved, the quick and effective acquisition scheme is required. Various kinds of acquisition schemes have been proposed and analyzed in the literatures[5]-[7]. Matched filtering is commonly used to synchronize the phase of the local PN sequence with that of the received PN sequence because of its good searching rate capability[8]-[9]. In this paper, a transceiver for DS/SS system with constant envelope continuous phase shift keying(CE-CPSK) modulated scheme is designed and the experimental results and the performance is evaluated. In section II, generation of CE-CPSK signal is described and the spectral characteristics of the CE-CPSK signal is given in section III. PN code acquisition receiver based on DMF and some statistics of the outputs of the correlator for fading channel are shown in section IV. Simulation and the measurement results are compared in section V, and conclusions are drawn in section VI.

Ⅱ. GENERATION OF CE-CPSK SIGNALS

As in the case of OQPSK modulation, smooth phase transitions can be achieved if the phase-vector rotated with constant amplitude through the $90°$ phase state to $180°$. This implies that a source of $90°$ shifted component is necessary to achieve the desired result. This component is only needed during the data transition, and ideally its amplitude should be of the following form.

$$Y(t) = \pm\sqrt{A^2 - X^2} \tag{1}$$

where $X(t)$ is the filtered input data to the multiplier of the in-phase channel, $Y(t)$ is the supplementary signal for constant amplitude generation, and A is the peak amplitude of $|X(t)|$. When $X(t)$ takes the peak value, i.e., $X(t) = \pm A$, then $Y(t) = 0$ which is to say that in the zero and $180°$ phase states, we do not need the $90°$ phase component. During the transition, as $X(t)$ becomes zero, $Y(t)$ becomes $\pm A$ which creates a phase vector rotation passing through $90°$. The block diagram of the CE-CPSK Modulator is shown in Fig. 1. The output can be written as

$$Z(t) = X(t)\cos\omega_0 t + Y(t)\sin\omega_0 t \tag{2}$$

equation (2) can be written in the form of

$$Z(t) = B(t)\cos(\omega_0 t - \phi(t)) \tag{3}$$

where

$$B(t) = \sqrt{X^2(t) + Y^2(t)} \tag{4}$$

and

$$\phi(t) = \tan^{-1}\frac{Y(t)}{X(t)},\ 0 \le \phi \le \pi \tag{5}$$

As in(3) and(4), to have a constant envelope we should keep the value of (4) constant. To meet this requirement, data signals be shaped sinusoidly and supplementary signal should be generated by using the shaped signal. i.e

$$X(t) = A\cos w_a t,\quad Y(t) = A\sin w_a t \quad (w_a << \omega_0) \tag{6}$$

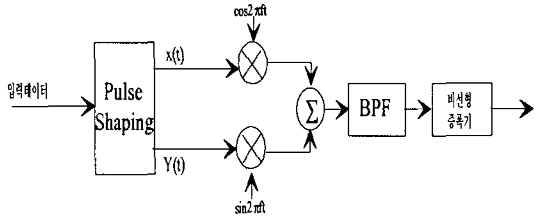

Fig. 1. The block diagram of CE-CPSK modulator

Ⅲ. SPECTRAL CHARACTERISTICS OF THE CE-CPSK SIGNAL

The derivation of the power spectrum of the modulated Z(t) signal is fairy complex, as evidenced by the correlation between the X(t) and Y(t) signals. In oder to have some insight into the shape of the spectrum, the power spectra of the X(t) and Y(t) baseband signals are first calculated. From these spectra, the power spectrum of the out put signal Z(t) can be estimated.

1. The PSD of the Baseband Signal X(t)

Assume that the input signal is random and is described by

$$S(t) = \sum_{n=-\infty}^{\infty} a_n g(t - nT_s) \tag{7}$$

where

$$a_n = \begin{cases} +1 & with \quad prob\ .1/2 \\ -1 & with \quad prob\ .1/2 \end{cases}$$

and g(t) is a rectangular pulse of duration Ts. The power spectrum of s(t) is given by

$$S_s(f) = 2f_s|G(f)|^2 = \frac{2}{f_s}\left|\frac{A\sin \pi f/f_s}{\pi f/f_s}\right|^2 \tag{8}$$

where G(f) is the Fourier transform of g(t), and $f_s = 1/T_s$ is the signaling frequency. The power spectrum of x(t) is

$$S_x(f) = |H(f)|^2 S_s(f) = \frac{2}{f_s}|H(f)|^2\left|\frac{A\sin \pi f/f_s}{\pi f/f_s}\right|^2 \tag{9}$$

where H(f) is the transfer function of the premodulation transversal filter.

As in equation (9), Sx(f) has spectral nulls at the signaling frequency and its integral multiples.

2. The PSD of thesupplementary signal Y(t)

The supplementary signal Y(t) refers to (1). To simplify the calculation, it is modified as

$$Y(t) = \sqrt{A^2 - X(t)^2} \tag{10}$$

Considering (7), the X(t) signal can be expressed as

$$X(t) = \sum_{n=-\infty}^{\infty} a_n v(t - nT_s) \tag{11}$$

where

$$v(t) = g(t) * h(t) \tag{12}$$

h(t) is the impulse response of the transversal filter and the symbol * denotes convolution. It is assume that the transversal filter round off the edges of the pulse shape g(t) to a sine shape i. e.,

$$v(t) = \begin{cases} A\sin\dfrac{\pi t}{2aT_s}, & 0 < t < aT_s \\ A\quad , & aT_s < t < (1 - aT_s) \\ A\sin\dfrac{\pi(T_s - t)}{2aT_s}, & (1-a)T_s < t < T_s \end{cases} \tag{13}$$

$$(0 < a < 1/2)$$

Then using (11), (12), and (13), Y(t) can be represented by

$$Y(t) = \sum_{n=-\infty}^{\infty} Y_n(t) \tag{14}$$

where for the interval $(2n-1)T_s/2 < t < (2n+1)T_s/2$

$$Y_n(t)=\begin{cases} p(t-nT_s)=\sqrt{A^2-v(t-nT_s)} \\ \qquad \cong A-|v(t-nT_s)| & with\ prob\ q=0.5 \\ 0 & with\ prob\ q=0.5 \end{cases} \tag{15}$$

and

$$p(t)=\begin{cases} 0 & -T_s/2<t<-aT_s \\ A(1+\sin\dfrac{\pi t}{2aT_s}), & -aT_s<t<0 \\ A(1-\sin\dfrac{\pi t}{2aT_s}) & 0<t<aT_s \\ 0 & aT_s<t<T_s/2 \end{cases} \tag{16}$$

The power spectrum of Y(t) can be written as

$$S_y(f)=2f_s q(1-q)|p(f)|^2+q^2 f_s^2 [p(0)]^2\delta(f)+2q^2 f_s^2\sum_{m=1}^{\infty}|p(mf_s)|^2\delta(f-mf_s) \tag{17}$$

where p(f) is the Fourier transform of p(t) and is given by

$$p(f)=\int_{-\infty}^{\infty}p(t)e^{-j2\pi ft}dt=\frac{\dfrac{Af_s}{2a}}{\left(\dfrac{\pi f_s}{2a}\right)^2-(2\pi f)^2}\left[\pi\frac{\sin(2a\pi f/f_s)}{2a\pi f/f_s}-2\right] \tag{18}$$

Substituting (18) into (19) gives the power spectrum of the signal Y(t), i. e.,

$$\begin{aligned} S_y(f)&=2f_s q(1-q)\left[\frac{Af_s/2a}{(\pi f_s/2a)^2-(2\pi f)^2}\right]^2\bullet\left[\pi\frac{\sin(2a\pi f/f_s)}{2a\pi f/f_s}-2\right] \\ &+q^2(2aA\frac{\pi-2}{k})^2\delta(f)+8q^2\left(\frac{aA}{\pi}\right)^2\sum_{m=1}^{\infty}\left[\frac{1}{1+(4am)^2}\right]^2 \\ &\bullet\left|\frac{\sin 2a\pi m}{2a\pi m}-2\right|^2\delta(f-mf_s) \end{aligned} \tag{19}$$

The first bracket in (19) indicates the continuous part of the spectrum, while the second bracket gives the discrete part, which consists of spectral lines at zero and integral multiples of the signaling frequency. This spectral line is due to the DC component of the Y(t) signal. To simplify calculation, we assumed like (11). In this case the Y(t) signal can have only positive values during any transition of the X(t) signal hence, it has a positive DC average component. But in the proposed method, the DC component can be removed by letting the Y(t) signal alternate between positive and negative values during the transitions of the X(t) signal to make the polarities of the complementary signal equally probable. In this case, the DC component of Y(t) signal would be zero, and therefore no spectral line would appear at zero frequency in Sy(f). This process is designed through the digital logic circuit.

3. The PSD of the Resultant CE-CPSK signal Z(t)

The CE-CPSK signal is given by (2). The power spectra of the Z(t) signal can be expressed by

$$S_z(f)=S_x(f)+S_y(f+f_0)+CPSD\ of\ I\ \text{and}\ Q \tag{20}$$

The first term in (20) denotes the useful energy which contains the information, and the last two terms indicate the required energy (much less than the first term, because of the low power contribution of the Q-channel) for smoothing the 180 phase transition of the I-channel RF signal. The overall shape of the output power spectra is much the same as the shape of $S_x(f+f_0)$. The proposed modified-BPSK [2] includes the DC component of Y(t) signal caused discreet spectral lines at f_0 and $f_0+mf_s(m=1,2,\cdots)$. However, in this paper, the DC component of Y(t) signal would be zero, and therefore no spectral line would appear at the carrier frequency in the $S_z(f)$.

Ⅳ. Code-Acquisition System

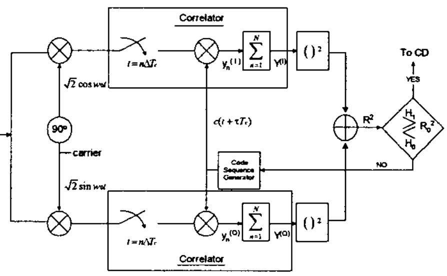

Fig. 2. A typical digital matched filter detector

A typical structure for DS/SS PN code acquisition system based DMF(Digital Matched Filter) correlator is shown in Fig. 2. The received signal is given by

$$r(t)=\sqrt{2P}\sum_{k=1}^{K}\gamma_k(t)Z_k(t+\Delta_k T_c)+n_w(t) \tag{21}$$

Where K is the number of users, $\gamma_k(t)$ is the time-variant channel coefficients, Δ_k is the time delay of the channel, and $n_w(t)$ is additive white Gaussian noise(AWGN) with one-side power spectral density of N_o.

We assume that during the acquisition process, no data are sent. The inphase and the quadrature components obtained by integrating over N chips in the correlator of the reference user(user1) are denoted by $Y^{(I)}$ and $Y^{(Q)}$, and are given by

$$Y^{(I)} = D^{(I)} + N^{(I)} + L_I^{(I)} + NL_I^{(I)} \tag{22a}$$

and

$$Y^{(Q)} = D^{(Q)} + N^{(Q)} + L_I^{(Q)} + NL_I^{(Q)} \tag{22b}$$

Where D is the desired component of the first user and is given by

$$D^{(I)} = \gamma_1\sqrt{P/2}R_c(\delta_1)\cos\phi_1, \qquad \delta_1 = \left|(\Delta_1 - \hat{\Delta}_1)T_c\right| \tag{23a}$$

$$D^{(Q)} = \gamma_1\sqrt{P/2}R_c(\delta_1)\sin\phi_1, \qquad \delta_1 = \left|(\Delta_1 - \hat{\Delta}_1)T_c\right| \tag{23b}$$

where

$$R_c(\delta) = \frac{1}{T_p}\int_0^{T_p} a_1(t+\Delta_1 T_c)a_1(t+\hat{\Delta}_1 T_c)dt$$
$$= \begin{cases} 1-\delta & \delta<1 \quad (H_1 : correct phase state) \\ 0 & \delta>1 \quad (H_0 : all other states) \end{cases} \tag{24}$$

where $\hat{\Delta} T_c$ is the phase offset of the locally generate code in the receiver. Linear other user interferences are represented as

$$L_I^{(I)} = \sqrt{P/2}\sum_{k=2}^{K}\gamma_k \cos\phi_k Z_k, \tag{25a}$$

$$L_I^{(Q)} = \sqrt{P/2}\sum_{k=2}^{K}\gamma_k \sin\phi_k Z_k, \tag{25b}$$

Where Z_k is partial correlation function between the desired signal and kth user, is given by [10]

$$Z_k = \frac{1}{N}\left[(1-\delta_k)\sum_{n=0}^{N-1} a_{1,n}a_{k,n-\eta_k} + \delta_k\sum_{n=0}^{N-1} a_{1,n-1}a_{k,n-\eta_k}\right] \tag{26}$$

where η_k is integer, $0 \le \delta_k < 1$ since $a_{1,n}$ and $a_{k,n}$ are independent random variables taking on +1 and -1 values with equal probability, based on the central limit theorem, Z_k has a Gaussian distribution when $N >> 1$ with mean $E[Z_k] = 0$ and variance $\sigma^2{}_{zk} = E[Z^2{}_k] \approx 2T_p/3N$.

Due to the random phase ϕ_k uniformly distributed over $[0, 2\pi]$, each term L_I is independent of any other terms, therefore L_I is a Gaussian random variable with mean 0, and variance is given by

$$\sigma^2{}_{L_I} = \frac{T^2{}_p P}{2}\sum_{k=2}^{K}\gamma_k\frac{1}{3N} \tag{27}$$

Analytical study of the nonlinear other user interferences are complex. In this paper, the amount of spectral restoration and the degradations caused by this are evaluated by computer simulations. And the background noise is AWGN with mean 0 and variance $N_0/4T_p$. The received signal can be modeled as Gaussian random process by central limit theorem. From this theorem, received signals into the I-channel and the Q-channel can be assumed as Gaussian distributed variables, and amplitudes can be Rayleigh distributed. Assuming that no timing error exists in the correlation procedure, the autocorrelation value after DMF correlator is used for the detection threshold region such as H_1 region(true acquisition) and H_0 (false alarm). the output of the correalator is defined by

$$V = Y^{(I)^2} + Y^{(Q)^2} \tag{28}$$

Its squared mean E is represented by

$$E^2 = \frac{NP}{2}\gamma_1 R_c{}^2(\delta_1) \tag{29}$$

and the variance of $Y^{(I)}$ is given by

$$\sigma^2{}_{1I} = \sigma^2{}_{L_I} + \sigma^2{}_{NL_I} + \sigma^2{}_{N_I}$$
$$= \frac{T_p{}^2 P}{2}\sum_{k=2}^{K}\gamma_k\frac{1}{3N} + \sigma^2{}_{NL_I} + \frac{N_0}{4T_p} \tag{30}$$

The acquisition process is modeled using state transition diagram shown [6]. For worst case code phase offset over code uncertainty region, mean acquisition time is obtained by

$$\overline{T}_{ACQ} = \left(\frac{P_{ACQ}}{P_{D_1}}\right)\tau_d\{(K+1)P_{D_1} + \overline{T}_M P_{ACQ}$$
$$+ (v-1)\overline{T}_0[P_{D_1} + P_{ACQ}(1-P_{D_1})]/(1-P_{FA_1})\} \quad (31)$$

where $\overline{T_0}$ is mean time per H_0 normalized by τ_d, $\overline{T}_M$ is miss detection mean time. And each is represented by

$$\overline{T}_0 \equiv \tau_d^{-1}H_0(1) = (1-P_{FA_0}) + (K+1)P_{FA_0}(1-P_{FA_1}) \quad (32)$$

$$\overline{T}_M \equiv \tau_d^{-1}H_M(1) = (1-P_{D_0}) + (K+1)P_{D_0}(1-P_{D_1}) \quad (33)$$

And P_{D_0} is detection probability in search mode at cell H_1, P_{D_1} is detection probability in verification mode at cell H_1. P_{FA_0} is false alarm probability in search mode, and P_{FA_1} is false alarm probability in verification mode and is represented by

$$P_{FA_0} = \int_\theta^\infty f_{H_0}(V)dV = \exp^{-\theta/2\sigma_1^2} \quad (34)$$

$$P_{D_0} = \int_\theta^\infty f_{H_1}(V)dV = \int_{\theta/2\sigma_1^2}^\infty e^{-(x+\lambda)} I_0(2\sqrt{\lambda x})dx \quad (35)$$

$$P_{D_1} = \sum_{n=B}^{A}\binom{A}{n} P^n{}_{D_0}(1-P_{D_0})^{A-n} \quad (36)$$

$$P_{FA_1} = \sum_{n=B}^{A}\binom{A}{n} P^n{}_{FA_0}(1-P_{FA_0})^{A-n} \quad (37)$$

where I_0 is 0 order modified Bessel function and

$$x = \frac{V}{2\sigma_1^2}, \quad \lambda = \frac{E^2}{2\sigma_1^2} \quad (38)$$

V. SIMULATION AND EXPERIMENTAL MEASUREMENT RESULTS

1. CE-CPSK transmitter

The design is made comparison between conventional BPSK and CE-CPSK modulator. The measured power spectra of Z(t)(input to the power amplifier) and output are shown in Fig. 3. Also simulated BPSK spectra are shown in Fig.4.

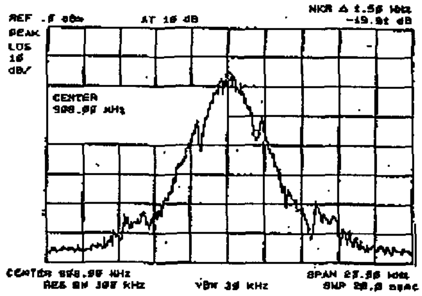

Fig 3.(a) Power spectrum at power amplifier input

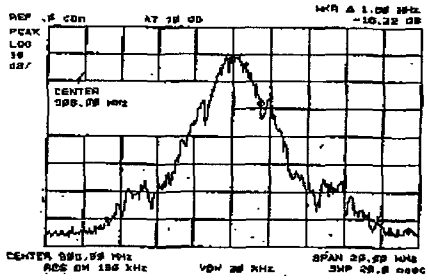

Fig.3.(b) Power spectrum at power amplifier output

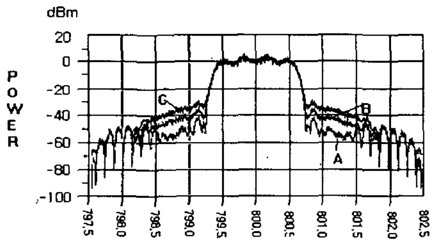

Fig. 4. Power spectrum density with output backoff (BPSK). (A:backoff 8dB, B: backoff 0dB, C: backoff 4 dB)

As can be seen, the nonlinear amplifier removes the effect of filtering on a BPSK signal, but it has a minimal effect on the CE-CPSK signal.

2. DS/SS PN code acquisition receiver based on DMF

The mean acquisition time normalized to uncertainty time and detection probability is analyzed for a nonlinear amplifier on multiuser environment. For the analysis, the following parameters are considered : the signal to interference(S/I), the number of correlation register(N), number of user's(K), and amplifier output back off. In Fig5, the signal to interference ratio of both BPSK and CE-CPSK system operated in the nonlinear region (output backoff -4dB) are compared. The performance degradation due to amplifier nonlinearity is much more severe in the BPSK system than proposed CE-CPSK system as the number of users increases.

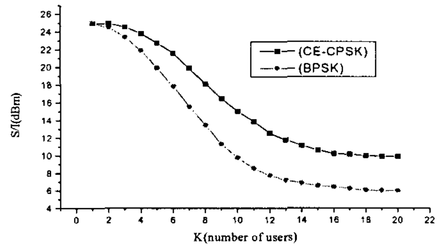

Fig. 5. Signal to interference power ratio with the number of use when the amplifier is operated in the nonlinear region

The effect of nonlinearity on the detection probability in both of the BPSK and CE-CPSK system are shown in Fig.6 with the correlation window N. The results show that the correlation window is not critical in the performance when the users are rare. As other users increase, using a larger value of N reduces the performance degradation.

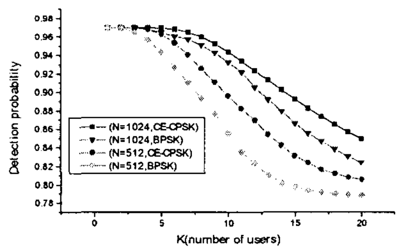

Fig.6. Detection probability with the number of use when the amplifier is operated in the nonlinear region

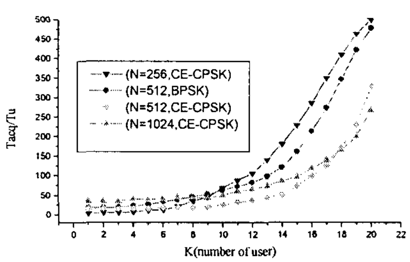

Fig. 7. Acquisition time with the number of use when the amplifier is operated in the nonlinear region

Fig.7 shows the acquisition time normalized to uncertainty time as a function of K which is the number of users. When there is a small other user interference, using N=256 gives the best performance. But the difference in the performance is not critical. As the users increase, the situation is changes. Using a larger value of N reduces the mean acquisition time.

Ⅵ. CONCLUSION

CE-CPSK(Constant Envelope Continuous Phase Shift Keying) modulated DS/SS transceiver is proposed. The performance of the proposed transceiver is evaluated in the wireless communication channel environment. When time-varying envelope signals are fed into a nonlinear amplifier such as class C amplifier, it undergoes nonlinear effects which introduce the spectral spreading or spectral regeneration, resulting to the performance degradation in adjacent channels. To overcome these problems, CE-CPSK modulation method which has the constant envelope and continuous phase characteristics is proposed. Simulation and experimental examination has been carried out. We also analyzed and compared the code-acquisition performance of CE-CPSK system with that of conventional BPSK system in the nonlinear amplified system with the correlation window. In the case of small K, the correlation window N=256 is large enough to achieve a small variance of the decision variables so that code acquisition can be achieved correctly. As a result, using a smaller Nvalue speed up the acquisition process. As the user increase, a longer correlation interval is necessary to ensure a high detection probability and, therefor, to keep a small value of the mean acquisition time. As a result, a larger N value results in better acquisition when the other user interference is severe. And comparing BPSK with CE-CPSK system, it is evident that the interference results from the nonlinearity of the amplifier play a dominant role in the performance.

REFERENCE

[1] Y.Nagata,"Linear amplification technique for digital mobile communications," Proc. IEEE Vehic. Technol. Comm., pp.159-164.1989

[2] A.A.MSaleh and J.Salz,"Adaptive linearization of power amplifiers in digital radio systems," Bell Syst.Tech.J., vol62 no.4 pp.1019-1033. April 1983

[3] C. Andren, "PSK sidebands reduced by premodulation filtering," Microwave Journal, vol.21pp.69-73 Jan. 1978

[4] E.T. Dickerson and W.B. Warren, "Generation of constant-envelope-signals," IEEE Trans.Commun. vol. COM-30 no.12 pp2554-2556 Dec.1982

[5] G.L.Turin, "Introduction to Spread Spectrum Antimultipath Techniques and Their Application to Urban Digital Radio," Proceeding of the IEEE. Vol.68,no.3 Mach 1980

[6] A.Polydors and C.Weber, "A Unified approach to serial search spread spectrum code acquisition Part I: General Theory," IEEE Trans. Comm., vol.COM-32, pp.542-549, May 1984.

[7] A.Polydors and C.Weber, "A Unified approach to serial search spread spectrum code acquisition Part II: A Matched filter Receiver," IEEE Trans. Comm., vol. COM-32, pp.550-560, May 1984.

[8] K.Simon, K.Omura, A.Scholtz and K.Levitt, Spread Spectrum Communications, vol.III, Computer Science Press,1985

[9] A. J. Viterbi, CDMA : Principles of Spread Spectrum Communication, Addison Wesley,1995

[10] R.L.Peterson,andR.E.Zeimer, Introduction to spread spectrum communications. Englewood Cliffs, NJ:Prentice Hall, 1995.

저자소개

김 성 철(Seong-cheol Kim)

1987년 고려대학교 전자공학과 학사
1989년 고려대학교 전자공학과 석사
1989년～1994년: 삼성전자(주) 근무
1997년 고려대학교 전자공학과 박사
1997년～현재 : 우송대학교 철도전기정보통신학부 부교수
※ 관심분야 : CDMA 이동통신, 통신이론, 차세대 광대역 이동 멀티미디어 시스템