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DS-CDMA 통신시스템 기반에서의 새로운 SNR 성능평가

정계현, 김성수
충북대학교 전기전자컴퓨터공학부

The Advanced SNR Performance analysis for DS-CDMA Communication System

Jeong Kehon, Kim Sung-Soo
School of Electrical and Computer Engineering, Chungbuk National University

Abstract - This paper proposes the new signal-to-noise ratio(SNR) that is different from the conventional signal-to-noise ratio(SNR) in direct-sequence code-division multiple-access communication system employing offset quadrature phase-shift keying(OQPSK) and using a chip waveform. The conventional SNR value is so different from a real SNR value. Therefore, we propose a new SNR equation approximated to real SNR. The multiple-access interference(MAI) in DS-CDMA communication system has an effect on SNR performance and MAI is concerned with the correlation functions of the chip waveform. For this reason, we considered all possible correlations of chip waveforms. In conclusion, the SNR value of proposed method is enclosed to the real SNR value.

1. INTRODUCTION

In this paper, we propose the new signal-to-noise ratio(SNR) that is different from the conventional SNR in direct-sequence code-division multiple-access(DS-CDMA) communication system employing offset quadrature phase-shift keying(OQPSK) and using a chip waveform. The multiple-access interference(MAI) in DS-CDMA communication system has an effect on SNR performance and MAI is concerned with the correlation functions of the chip waveform. Therefore, we will consider some concerned equations and terminologies referred upper.

The rest of the paper is organized as follows. In Section 2.1, the system and signal models used are discussed. In Section 2.2, a conventional MAI and SNR are discussed. In Section 2.3, we propose a new SNR equation. And final conclusions are presented in Section 3.

2. BASIC EQUATIONS AND A NEW SNR METHOD

2.1 SYSTEM AND SIGNAL MODEL

We employ offset quadrature phase-shift keying(OQPSK) in DS-CDMA communication system. We briefly review this model here. Assuming K active users with transmitted signals of the generalized OQPSK format, the information bits of user k, $1 \le k \le K$, are split into inphase and quadrature streams

$$b_k^I(t) = \sum_{l=-\infty}^{\infty} b_{k,l}^I u(t-lT) \qquad (1)$$

$$b_k^Q(t) = \sum_{l=-\infty}^{\infty} b_{k,l}^Q u(t-lT) \qquad (2)$$

where $u(t)$ is the unit pulse on the bit interval [0, T], and data bits $b_k^I(t)$ and $b_k^Q(t)$ are binary random variables $\in \{+1, -1\}$ with equal probability. The data streams are multiplied by the spectrum-spreading signals

$$a_k^I(t) = \sum_{l=-\infty}^{\infty} a_{k,m}^I \psi_c(t-mT_c) \qquad (3)$$

$$a_k^Q(t) = \sum_{l=-\infty}^{\infty} a_{k,m}^Q \psi_c(t-mT_c) \qquad (4)$$

where $a_k^I(t)$ and $a_k^Q(t)$ are two distinct signature sequences modeled as random, periodic, with independent, equally likely chip symbols $a_{k,m}^I$, $a_{k,m}^Q$ $\in \{+1, -1\}$. We assume that each bit is coded with N chips(i.e. $T = NT_c$). The chip waveform $\psi_c(t)$ is time-limited to the interval $[0, T_c]$ and normalized to have energy T_c, so that

$$\int_0^{T_c} \psi_c^2(t)dt = T_c. \qquad (5)$$

The two spreading signals are modulated onto two carriers in quadrature, with a time offset of half a chip period $T_c/2$ introduced in the cosine branch. the resulting inphase and quadrature signals are

$$s_k^I(t) = a_k^I(t-T_c/2)b_k^I(t-T_c/2)\cos(w_c t + \theta_k) \qquad (6)$$

$$s_k^Q(t) = a_k^Q(t)b_k^Q(t)\sin(w_c t + \theta_k) \qquad (7)$$

where θ_k is the carrier phase angle and w_c is angular frequency. The resulting transmitted signal from the k th user is obtained by the sum

$$s_k(t) = s_k^I(t) = s_k^Q(t). \qquad (8)$$

At the front end of the receiver, the received signal $r(t)$ is the sum of delayed versions of the transmitted signals of all active users, with additive thermal noise

$$r(t) = \sqrt{2P}\sum_{k=1}^{K} a_k^I(t-\frac{T_c}{2}-\tau_k)b_k^I(t-\frac{T_c}{2}-\tau_k)\cos(w_c t+\phi_k) + \sqrt{2P}\sum_{k=1}^{K} a_k^Q(t-\tau_k)b_k^Q(t-\tau_k)\sin(w_c t+\phi_k) + n(t) \qquad (9)$$

where $n(t)$ is white Gaussian process with PDS

$N_0/2$ modeling the thermal noise, P is the received signal power in each inphase and quadrature component, assumed to assumed to be the same for all users. τ_k is the signal time delay, and ϕ_k is cumulated phase equal to $\theta_k - w_c\tau_k$. If perfect synchronization were assumed, we could set $\tau_k = \phi_k = 0$. The desired receiver decision statistic is then given by

$$Z_1^Q(t) = \int_{-T}^{t} r(t)sin(w_c\tau)d\tau. \quad (10)$$

The decision statistic is sampled at multiples of the bit duration T, followed by a zero-threshold decision device that outputs a symbol $b_{1,i-1}^{*Q} = 1$ if $Z_1^Q(iT) > 0$ and a symbol $b_{1,i-1}^{*Q} = -1$ otherwise [1][2].

2.2 A CONVENTIONAL MAI AND SNR METHOD

We have already investigated the signal model and decision statistic in DS-CDMA communication system. From (10), the general SNR expression can be obtained as

$$SNR = \frac{[E[Z_1^Q(T)]]^2}{var[Z_1^Q(T)]} \quad (11)$$

where the numerator part of (11) means signal energy and the denominator part of it means the sum of noise energy and multiple-access interference(MAI) energy. Signal energy is

$$[E[Z_1^Q(T)]]^2 = E_b \quad (12)$$

where E_b denotes the energy per bit. The denominator part of (11) is rewritten as

$$var[Z_1^Q(T)] = \frac{N_0}{2} + 2(K-1)E_b\frac{M_c}{N} \quad (13)$$

where $N_0/2$ is energy of noise, K is the number of users, N is the number of chip waveform per bit, and M_c is the normalized mean-squared partial chip correlation

$$M_c = \frac{1}{T_c^3}\int_0^{Tc} R_{\psi_c}^2(s)ds = \frac{1}{T_c^3}\int_0^{Tc} R_{\psi_c}^{*2}(s)ds \quad (14)$$

which is obtained from the continuous-time partial autocorrelation functions given by

$$R_{\psi_c}(s) = \int_0^{s} \psi_c(t)\psi_c(t+T_c-s)dt \quad (15)$$

$$R_{\psi_c}^{*}(s) = \int_s^{T_c} \psi_c(t)\psi_c(t-s)dt. \quad (16)$$

From (12) to (16), we can rewrite (11) as

$$SNR = \frac{E_b}{\frac{N_0}{2} + 2(K-1)E_b\frac{M_c}{N}} \quad . \quad (17)$$

This equation (17) is a conventional SNR [2][3].

2.3 A NEW SNR EQUATION

When the conventional SNR equation (17) is used, the SNR performance is better than real SNR performance. This result is caused by missing some interferences. In Section 2.2, while we result in (17), we have passed over statistic probability. When chip waveforms are transmitting continuously, if $\psi_c(t)$ were a present chip waveform, $\psi_c(t+T_c)$ would be situated on the front of $\psi_c(t)$, and $\psi_c(t-T_c)$ would be situated on the behind of $\psi_c(t)$. By the way, the conventional SNR method in Section 2.2 is satisfied with the hypothesis that a present chip waveform $\psi_c(t)$ is correlated with both $\psi_c(t+T_c)$ and $\psi_c(t-T_c)$. But, actually this hypothesis is not true. Thus, we must change this hypothesis. When we investigate possible cases about correlation of a present chip waveform, there are four cases which have the same happening probability($Pr = 1/4$).

These four cases are following. The <case 1> is when $\psi_c(t+T_c)$ and $\psi_c(t)$ are only correlated, the <case 2> is when $\psi_c(t)$ and $\psi_c(t-T_c)$ are only correlated, the <case 3> is when $\psi_c(t)$ is correlated with both $\psi_c(t+T_c)$ and $\psi_c(t-T_c)$, and the last <case 4> is when $\psi_c(t)$ is un-correlated with both $\psi_c(t+T_c)$ and $\psi_c(t-T_c)$ [4].

Now, we can calculate the SNR for each case. We define that SNR_k is the SNR of the <case k>. For calculating SNR of <case k>, we can use (15), (16), (14), (13), and (17), successively.

For the <case 1>, (15) is not zero and (16) is zero. Thus, the normalized mean-squared partial chip correlation is M_c. The denominator part of SNR_1 is $var[Z_1^Q(T)] = \frac{N_0}{2} + (K-1)E_b\frac{M_c}{N}$. Therefore,

$$SNR_1 = \frac{E_b}{\frac{N_0}{2} + (K-1)E_b\frac{M_c}{N}}. \quad (18)$$

Appling the same rule to the <case 2>, the normalized mean-squared partial chip correlation is the same as the <case 1> M_c. Therefore,

$$SNR_2 = \frac{E_b}{\frac{N_0}{2} + (K-1)E_b\frac{M_c}{N}}. \quad (19)$$

And the <case 3> is identical with the conventional SNR method. Therefore, the normalized mean-squared partial chip correlation is $2M_c$ and

$$SNR_3 = \frac{E_b}{\frac{N_0}{2} + 2(K-1)E_b\frac{M_c}{N}}. \quad (20)$$

At the rest, all of the partial autocorrelation functions of the <case 4> are zero. In the same reason, the normalized mean-squared partial chip correlation is zero. Thus, form (13), $var[Z_1^Q(T)] = \frac{N_0}{2}$. Therefore,

$$SNR_4 = \frac{E_b}{N_0/2}. \quad (21)$$

We have already known that the probabilities of the each case is same as 1/4. In conclusion, we can calculate the average SNR from (18), (19), (20), and (21) that is

$$SNR_{av} = 1/4\sum_{k=1}^{4} SNR_k$$

$$= \frac{2E_b}{2N_0 + 4(K-1)E_b\frac{M_c}{N}} \qquad (22)$$

$$+ \frac{E_b}{2N_0 + 8(K-1)E_b\frac{M_c}{N}} + \frac{E_b}{2N_0}$$

The average SNR_{av} of (22) is the proposed new SNR. this SNR_{av} is more close to the real SNR value than the conventional SNR (17).

3. CONCLUSION

This paper proposed the new SNR method that is different from the conventional SNR method in DS-CDMA communication systems employing OQPSK and using a chip waveforms. In spite of great efforts, we could not formulate a SNR equation that is identical to a real SNR. This result may be caused by impossibility to consider all constraints of a real system. But, the proposed SNR(SNR_{av}) is more approximated to the real SNR than the conventional SNR. In the future, it will be necessary to make simulation for visual objects.

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