Two-Valued Integrand Coding Transmission of Multi-Rate Quasi-Synchronous CDMA Signals Convolved by Real-Valued Self-Orthogonal Finite-Length Sequences

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Abstract: The next generation mobile communication system is desired to transmit multimedia information at multi-rate. Two-valued integrand codes are applied to the realization of the multi-rate quasi-synchronous CDMA system using real-valued self-orthogonal finite-length sequences with zero correlation zone. Each user data of multi-level and multi-interval and a common synchronizing sequence are convolved with the user sequence and converted to two-valued integrand coding signal. The signal passes through low pass filters in a transmitter and a receiver and is changed to a real-valued signal. In the receiver, the desired data are detected by the correlation processing with the respective code. This transmission system suppresses distortion on amplitude limitation and quantization, and interchannel and intersymbol interferences. Two-valued integrand coding contributes to easy amplification and modulation.

Key–Words: multi-rate transmission, two-valued integrand code, self-orthogonal finite-length sequence, zero correlation zone, convolution, amplitude distortion, quasi-synchronous CDMA.

1 Introduction

The next generation mobile communication system is desired to transmit multimedia information at multi-rate. Code division multiple access (CDMA) system and orthogonal frequency division multiplexing (OFDM) system are candidates for the next generation system. Multi-rate CDMA systems using multiple codes of the same length[1] and using variable-length codes at the same chip rate[2] are proposed. The both systems require reference codes more than one in a receiver and reduce user member. Multi-rate OFDM system allotting many sub-carriers is proposed[3]. This system accompanies troublesome sub-carrier allocation.

Basically, the OFDM system has the higher spectral efficiency and the more tolerance for multi-path than the CDMA system, but is caused to intercarrier interferences leading to intersymbol in interferences and interchannel interferences by peak distortion and synchronization error.

Quasi-synchronous CDMA system using zero-correlation-zone (ZCZ) sequences has the tolerance for the synchronization error and the applicability to improve the multi-path. Tanada developed the real-valued self-orthogonal finite-length sequences of zero correlation zone, which have zero sidelobe autocorrelation functions except at both shift ends and zero crosscorrelation functions in a limited shift range[4],[5],[6],[7].

In this paper, two-valued integrand codes[8] are applied to the realization of the multi-rate quasi-synchronous CDMA system using real-vlaued self-orthogonal finite-length sequences with zero correlation zone[9], this system is an expansion of the convolutional coding data transmission system using real-valued finite-length sequences for a single user[10],[11]. Data rate is controlled by multi-level and multi-interval data, which is convolved with the distinctive sequences of each user, and the convolved real-valued signal is converted to two-valued signal by integrand coding. The two-valued signal is changed to real-valued signal through the band limited channel. The desired data are detected by the correlation between the real-valued signal and the respective code. Numerical experiments show the suppression of the distortions on amplitude limitation and quantization and the interchannel and intersymbol interferences in a certain shift range.

2 Orthogonal Set of Real-Valued Finite-Length PN Sequences

This section summarizes an orthogonal set of real-valued finite-length pseudonoise (PN) sequences to be applied to a multi-rate quasi-synchronous CDMA system[5],[6],[7]. The sequence is called a self-orthogonal or shift-orthogonal finite-length sequence, also called Huffman sequence, since its shifted sequences are orthogonal within a limited shift range.

An aperiodic autocorrelation function of the self-orthogonal finite-length sequence $\{a_{M,\ell,i}\}$ of length M, member ℓ and ordinal i is given by

$$\begin{aligned}\rho_{M,\ell,\ell,i'} &= \frac{1}{M}\sum_{i=0}^{M-1} a_{M,\ell,i}a_{M,\ell,i-i'} \\ &= \begin{cases} 1 & ;i'=0 \\ \varepsilon_{M-1} & ;i'=\pm(M-1) \\ 0 & ;elsewhere \end{cases} \end{aligned} \quad (1)$$

where $a_{M,\ell,i} = 0$ for $i < 0$ and $i > M-1$, i' is shift, and ε_{M-1} is a shift-end correlation value. The sequence is replaced by an impulse train with weight $a_{M,\ell,i}$ at every time-chip interval T_c

$$a_{M,\ell}(t) = \sum_{i=0}^{M-1} a_{M,\ell,i}\delta(t-iT_c) \quad (2)$$

and its Fourier transform

$$\begin{aligned}A_{M,\ell}(f) &= \int_{-\infty}^{+\infty} a_{M,\ell}(t)\cdot e^{-j2\pi ft}dt \\ &= \sum_{i=0}^{M-1} a_{M,\ell,i}Z^{-i} \end{aligned} \quad (3)$$

where $\delta(t)$ is Dirac's delta function of time t, and $Z = e^{j2\pi fT_c}$, and f is frequency.

For positive ε_{M-1} and odd M, we have the spectrum solution of the sequence $\{a_{M,\ell,i}\}$ as

$$\begin{aligned}A_{M,\ell}(f) &= \sqrt{M\varepsilon_{M-1}}K_{M,\ell} \\ &\quad\times\prod_{m=1}^{\frac{M-1}{2}}\{Z^{-2}+2\gamma_{M,\ell,m}Z^{-1} \\ &\quad\times\cos\tfrac{(2m-1)\pi}{M-1}+\gamma_{M,\ell,m}^2\} \end{aligned} \quad (4)$$

where $\gamma_{M,\ell,m} = \alpha_M \;\; or \;\; \beta_M$, $\beta_M = 1/\alpha_M$ and

$$K_{M,\ell} = 1/\prod_{m=0}^{\frac{M-1}{2}}\gamma_{M,\ell,m} \quad (5)$$

$$\alpha_M = \left(\frac{1+\sqrt{1-4|\varepsilon_{M-1}|^2}}{2|\varepsilon_{M-1}|}\right)^{\frac{1}{M-1}}. \quad (6)$$

For negative ε'_{M-1} and odd M, we have the spectrum solution of the sequence $\{a'_{M,\ell,i}\}$ as

$$\begin{aligned}A'_{M,\ell}(f) &= -\sqrt{M|\varepsilon'_{M-1}|}K'_{M,\ell} \\ &\quad\times(Z^{-1}-\gamma'_{M,\ell,0})(Z^{-1}+\gamma'_{M,\ell,\frac{M-1}{2}}) \\ &\quad\times\prod_{m=1}^{\frac{M-3}{2}}\{Z^{-2}-2\gamma'_{M,\ell,m}Z^{-1} \\ &\quad\times\cos\tfrac{2m\pi}{M-1}+\gamma'^2_{M,\ell,m}\} \end{aligned} \quad (7)$$

$$K'_{M,\ell} = 1/\{\sqrt{\gamma'_{M,\ell,0}\gamma'_{M,\ell,\frac{M-1}{2}}}\prod_{m=1}^{\frac{M-3}{2}}\gamma'_{M,\ell,m}\}, \quad (8)$$

where the parameters with the mark $'$ are defined as similarly as those without the mark $'$.

We can synthesize the sequence $\{a'_{M_0,\lambda,i}\}$ of length $M_0 = 2M-1$ and shift-end negative value ε'_{M_0-1} from the sequence $\{a_{M,\ell,i}\}$ of odd length M and shift-end positive value ε_{M-1} and the sequence $\{a'_{M,\ell',i}\}$ of odd length M and shift-end negative value $\varepsilon'_{M-1} = -\varepsilon_{M-1}$. From Eqs.(4) and (7), we obtain the following spectrum of the sequence $\{a'_{M_0,\lambda,i}\}$ as

$$A'_{M_0,\lambda}(f) = K_s\cdot A_{M,\ell}(f)\cdot A'_{M,\ell'}(f) \quad (9)$$

where $\alpha'_{M_0} = \alpha'_M = \alpha_M$ and

$$K_s = \sqrt{M_0|\varepsilon'_{M_0-1}|}/(M|\varepsilon_{M-1}|) \quad (10)$$

$$\varepsilon'_{M_0-1} = -|\varepsilon_{M-1}|^2/(1-2|\varepsilon_{M-1}|^2)\,. \quad (11)$$

For the orthogonal set of the suppressed amplitude distinct sequences $\{a'_{M,\ell'_n,i-n}\}$ of $M = 2^{\nu+1}+1$, $\nu = 3,4,5,\cdots$, we have the spectrum set

$$\begin{aligned}& A'_{M,\ell'_n}(f)Z^{-n} \\ &= -\sqrt{M|\varepsilon'_{M-1}|}F(\gamma'^2_{M,\lambda_0,0},Z^{-2})G(\gamma'_{M,\lambda_0,1},Z^{-1}) \\ &\quad\times\prod_{m=2}^{\nu-1}G(\gamma'^{2^{m-1}}_{M,\ell'_n,m},Z^{-2^{m-1}})Z^{-n} \end{aligned} \quad (12)$$

where

$$\begin{aligned}& F(\gamma'_{M,\lambda_0,0},Z^{-1}) \\ &= (Z^{-1}-\gamma'_{M,\lambda_0,0})(Z^{-1}+\gamma'^{-1}_{M,\lambda_0,0}) \\ &= Z^{-2}-(\gamma'_{M,\lambda_0,0}-\gamma'^{-1}_{M,\lambda_0,0})Z^{-1}-1\,, \end{aligned} \quad (13)$$

$$\begin{aligned}& G(\gamma'_{M,\lambda_0,\nu},Z^{-1}) \\ &= Z^{-4}+\sqrt{2}(\gamma'_{M,\lambda_0,\nu}-\gamma'^{-1}_{M,\lambda_0,\nu})Z^{-3} \\ &\quad+(\gamma'_{M,\lambda_0,\nu}-\gamma'^{-1}_{M,\lambda_0,\nu})^2Z^{-2} \\ &\quad-\sqrt{2}(\gamma'_{M,\lambda_0,\nu}-\gamma'^{-1}_{M,\lambda_0,\nu})Z^{-1}+1 \end{aligned} \quad (14)$$

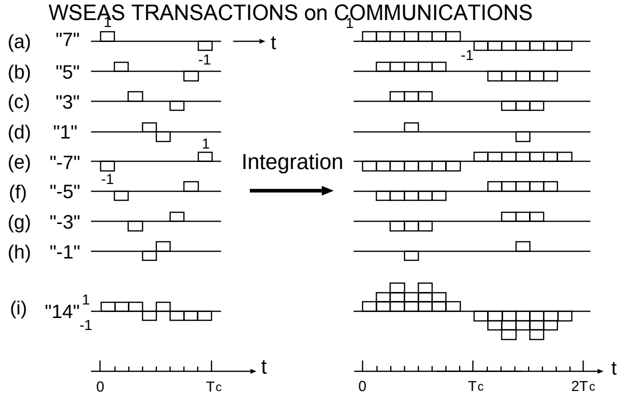

Figure 1: Composition of integrand code and its integration.

and $n = 0, 1, \cdots, \frac{M-1}{8} - 1$. The $\frac{M-1}{8}$ sequences are orthogonal at the shift $i' = 0 \ mod \ \frac{M-1}{8}$. From Eq.(12), for $M = 33$, we have the spectrum set for the sequence set $\{a'_{33,\ell'_0,i}\}$, $\{a'_{33,\ell'_1,i-1}\}$, $\{a'_{33,\ell'_2,i-2}\}$ and $\{a'_{33,\ell'_3,i-3}\}$,

$$\begin{aligned} A'_{33,\ell'_0}(f) &= -\sqrt{33|\varepsilon_{32}|}F(\alpha_{33}^2, Z^{-2}) \\ &\times G(\alpha_{33}, Z^{-1})G(\alpha_{33}^2, Z^{-2}) \\ &\times G(\alpha_{33}^4, Z^{-4}) \end{aligned} \quad (15)$$

$$\begin{aligned} A'_{33,\ell'_1}(f)Z^{-1} &= -\sqrt{33|\varepsilon_{32}|}F(\alpha_{33}^2, Z^{-2}) \\ &\times G(\alpha_{33}, Z^{-1})G(\beta_{33}^2, Z^{-2}) \\ &\times G(\alpha_{33}^4, Z^{-4})Z^{-1} \end{aligned} \quad (16)$$

$$\begin{aligned} A'_{33,\ell'_2}(f)Z^{-2} &= -\sqrt{33|\varepsilon_{32}|}F(\alpha_{33}^2, Z^{-2}) \\ &\times G(\alpha_{33}, Z^{-1})G(\alpha_{33}^2, Z^{-2}) \\ &\times G(\beta_{33}^4, Z^{-4})Z^{-2} \end{aligned} \quad (17)$$

$$\begin{aligned} A'_{33,\ell'_3}(f)Z^{-3} &= -\sqrt{33|\varepsilon_{32}|}F(\alpha_{33}^2, Z^{-2}) \\ &\times G(\alpha_{33}, Z^{-1})G(\beta_{33}^2, Z^{-2}) \\ &\times G(\beta_{33}^4, Z^{-4})Z^{-3} \end{aligned} \quad (18)$$

where $\gamma_{M,\lambda_0,0} = \alpha_{33}$, $\gamma_{M,\lambda_0,1} = \alpha_{33}$, and the number n in $A'_{33,\ell'_n}(f)Z^{-n}$, $n = 0, 1, 2, 3$, is defined by the binary notation as $n = (1\ 0)_2 = 2$ corresponding to the reversed product $G(\beta_{33}^4, Z^{-4})G(\alpha_{33}^2, Z^{-2})$ in Eq.(17). The set of $\{a'_{33,\ell'_0,i}\}$ and $\{a'_{33,\ell'_2,i-2}\}$ is orthogonal at the shifts $i' = 0 \pm 1 \ mod \ 4$ and is the sequence set with the zero correlation zone. Similarly, we have a set of $\{a'_{65,\ell'_0,i}\}$, $\{a'_{65,\ell'_2,i-2}\}$, $\{a'_{65,\ell'_4,i-4}\}$ and $\{a'_{65,\ell'_6,i-6}\}$ of length $M = 65$, which is orthogonal at the shifts $i' = 0 \pm 1 \ mod \ 8$.

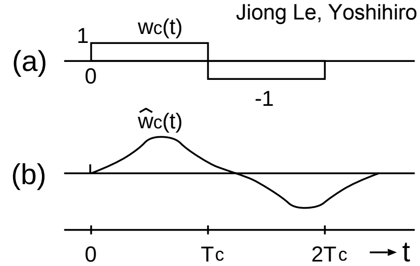

Figure 2: Chip waveforms.

Table 1: Positive values of two-valued integrand code (n=16).

Value	Code elements	Value	Code elements
0	+ - + - - + - +	16	+ + - + + - + +
1	- + + + - - - -	17	+ + - + + + - -
2	- + + + - - - +	18	+ + - + + + - +
3	+ - - + + - + +	19	+ + + - + + - -
4	+ - - + + + - -	20	+ + - + + + + -
5	+ - - + + + - +	21	+ + - + + + + +
6	+ - + - + + - -	22	+ + + - + + + -
7	+ - + - + + - +	23	+ + + - + + + +
8	+ - + + - + - -	24	+ + + + - + + -
9	+ - + - + + + -	25	+ + + + - + + +
10	+ - + - + + + +	26	+ + + + + - + -
11	+ - + + - + + -	27	+ + + + + - + +
12	+ - + + - + + +	28	+ + + + + + - -
13	+ - + + + - + -	29	+ + + + + + - +
14	+ - + + + - + +	31	+ + + + + + + -
15	+ - + + + + - -	32	+ + + + + + + +

3 Two-Valued Integrand Codes for Transmission of Real-Valued Signals

This section explains two-valued integrand codes for the transmission of real-valued signals, because two-valued signals are easily amplified and modulated. Real-valued signals are converted to integer signals by quantization. Fig.1 illustrates the composition of a two-valued integrand code and its integration. One chip time duration T_c is divided into 8 time slots and the 8-point moving average is used instead of an integration. (a), (b), (c), (d), (e), (f), (g) and (h) in Fig.1 are the code components which produce the respective front area $7, 5, 3, 1, -7, -5, -3$ and -1 after the moving average, and Fig.1(i) is the code which produces the front area $14 = 7 + 5 + 3 - 1$ after the moving average. Generally, for the time slots of even division number n_b, the value of the front area is given by

$$N_{b,i} = \sum_{k=0}^{n_b/2-1} (2k+1) b_{i,k} \quad (19)$$

where $b_{i,k} \in \{-1,+1\}$. The two-valued integrand code makes the values $N_{b,i}/2 = 0, \pm1, \pm2, \cdots, \pm n_b^2/8$ except a pair values $\pm(n_b^2/8-2)$ that are mapped into $\pm(n_b^2/8-1)$ or $\pm(n_b^2/8-3)$ so as to reduce the error which decreases as the division number increases. The integrand codes pass through low pass filters in a transmitter and a receiver, and become sine-like pulse. Each integrand code has some combinations for a given value, and is selected to the optinum combination so that every sine-like pulse might resemble. Table 1 shows the positive values and the optimized front half code elements of two-valued integrand codes with $n_b = 16$ [8].

The waveform of the two-valued integrand code for the maximum value is shown in Fig.2(a). At the low pass filter output in the receiver, the waveform is changed to a sine-like pulse in Fig.2(b) as well as those of the other codes, where a waveform weighted to $w_c(t)$ is regarded to be transmitted for each code [8]. A signal constructed by the integrand code and a sequence $\{a_{M,\ell,i}\}$ is equivalently expressed as

$$x(t) = \sum_{i=0}^{M-1} a_{M,\ell,i} \cdot w_c(t - iT_c) \quad (20)$$

and a signal at the low pass filter output in the receiver is given by

$$\hat{x}(t) = \sum_{i=0}^{M-1} a_{M,\ell,i} \cdot \hat{w}_c(t - iT_c) \quad . \quad (21)$$

4 Multi-Rate Quasi-Synchronous CDMA System

Fig.3 illustrates a quasi-synchronous CDMA system by convolution between multi-level and multi-interval data and real-valued self-orthogonal finite-length sequences with zero correlation zone $-1 \le i' \le 1$, and Fig.4 shows signal allotment of the system[9],[10],[11],[12]. Data of each user are arranged to those of multi-level and multi-interval according to the demand for data rate and data reliability. In Fig.4(a), the arranged data for user 0 and a synchronizing sequence $\{w_0 a_{M,\ell,i}\}$ are allotted , where w_0 is the weight for the balance with the data power. A train of p-valued data $d_{0,k} \in \{-V_{p-1}, \cdots, -V_{2\nu-1}, \cdots, -V_1, V_1, \cdots, V_{2\nu-1}, \cdots, V_{p-1}\}$, $\nu = 1, 2, \cdots, p/2$, $k = 0, 1, \cdots, K_0(N-1)-1$ for user 0 is allotted at the front data interval $N(M-1)T_c$, where p-valued data with uniform distribution and average power 1 takes the value

$$V_{2\nu-1} = (2\nu-1)\sqrt{3/(p^2-1)} \quad (22)$$

and K_0 is the number of data for user 0 during $(M-1)T_c$. Convolving the signals of Fig.4(a) with the sequence $\{a'_{M,\ell'_0,i}\}$ for user 0 makes the signals of Fig.4(c). Here, the signal $x_{0d,i}$ and its average power P_{0d} in the data interval are given by

$$x_{0d,i} = \sum_{k=0}^{K_0(N-1)-1} d_{0,k} a'_{M,\ell'_0,i-k\mu} \quad (23)$$

$$P_{0d} = \frac{1}{N(M-1)} \sum_{i=0}^{N(M-1)-1} x_{0d,i}^2 \cong K_0 \quad (24)$$

where $\mu = (M-1)/K_0$, and the synchronizing signal $x_{0s,i}$ and its average power P_{0s} in the synchronizing interval are given by

$$x_{0s,i} = \sum_{i'=0}^{M-1} w_0 a_{M,\ell,i-i'+2(M-1)} a'_{M,\ell'_0,i'} = \frac{w_0}{K_s} a'_{2M-1,\lambda'_0,i+2(M-1)} \quad (25)$$

$$P_{0s} = \frac{w_0^2}{K_s^2} \quad . \quad (26)$$

For user 1, the synchronizing sequence $\{w_1 a_{M,\ell,i}\}$ and a train of q-valued data $d_{1,k} \in \{-V_{q-1}, \cdots, -V_1, V_1, \cdots, V_{q-1}\}$, $k = 0, 1, \cdots, K_1(N-1)-1$, in Fig.4(b) make the signal in Fig.4(d) through the convolution with $\{a'_{M,\ell'_2,i-2}\}$. The similar signals are allotted for the other users. The heights of the data signal $x_{0d,i}$ and the synchronizing signal $x_{0s,i}$ present approximately Gaussian distributions. If we adjust the signal power of each user n to $\sigma_{nx}^2 = P_{nd} = P_{ns}$, $n = 0, 1, 2, \cdots$, then the distribution of the height x of the signals $x_{nd,i}$ and $x_{ns,i}$, $n = 0, 1, 2, \cdots$, is approximately represented by

$$q_n(x) = \frac{1}{\sqrt{2\pi}\sigma_{nx}} e^{-\frac{x^2}{2\sigma_{nx}^2}} \quad (27)$$

where $\sigma_{nx}^2 = K_n$. When $|\varepsilon_{M-1}| = 1/M$, we obtain $K_s \cong \sqrt{2/M}$ and $w_n = K_s\sqrt{K_n} \cong \sqrt{2K_n/M}$. The signal $x_{n,i}$ of user n is limited between the levels $-r$ and r, and quantized to the integer signal $\hat{x}_{n,i}$ between $-A$ and A as follows:

$$\hat{x}_{n,i} \cong \frac{1}{K_{nc}} x_{n,i} \;;\; -r < x_{n,i} < r \quad (28)$$

$$\hat{x}_{n,i} = \begin{cases} A & ;\; r \le x_{n,i} \\ -A & ;\; x_{n,i} \le -r \end{cases} \quad (29)$$

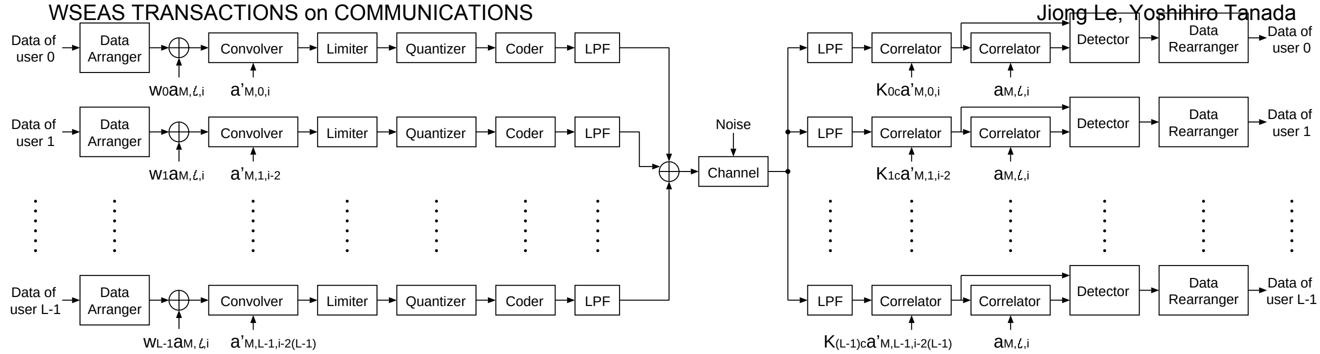

Figure 3: Model of Multi-Rate Quasi-Synchronous CDMA System.

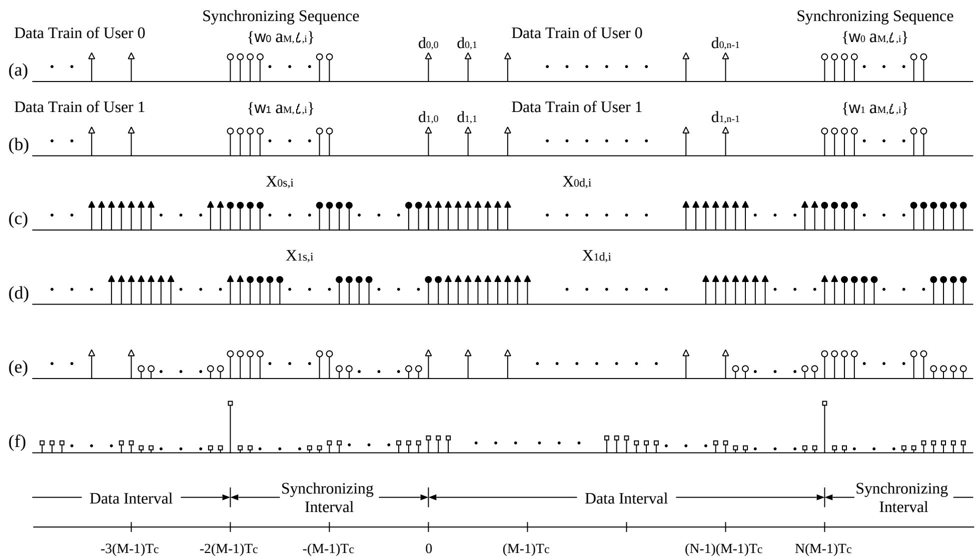

Figure 4: Signal allotment of Multi-Rate Quasi-Synchronous CDMA System.

where $r = K_{nc}A$, and K_{nc} is a coefficient so that the power of the approximated signal $K_{nc}\hat{x}_{n,i}$ might nearly equal the power of the amplitude-limited signal in the quantization input. The height distribution of the amplitude-limited real-valued signal $x'_{n,i}$ is given by

$$q_{nr}(x') = \begin{cases} q_n(x') & ; \ -r < x < r \\ Q_0\tilde{\delta}(x' - r) & ; \ r \leq x \\ Q_0\tilde{\delta}(x' + r) & ; \ x \leq -r \end{cases} \quad (30)$$

where $\tilde{\delta}(x)$ is Dirac's delta function of x and

$$\begin{aligned} Q_0 &= \int_r^{+\infty} q_n(x')dx' \\ &= \frac{1}{2}erfc(\frac{r}{\sqrt{2}\sigma_{nx}}) \quad . \end{aligned} \quad (31)$$

The power of the amplitude-limited signal is calculated as

$$\begin{aligned} P_{nr} &= \int_{-\infty}^{+\infty} x'^2 q_{nr}(x')dx' \\ &= r^2 - (r^2 - \sigma_{nx}^2)erf(\frac{r}{\sqrt{2}\sigma_{nx}}) \\ &\quad -\sqrt{\frac{2}{\pi}} r\sigma_{nx} e^{-\frac{r^2}{2\sigma_{nx}^2}} \quad , \end{aligned} \quad (32)$$

and an efficiency of signal transform is given by

$$\begin{aligned} \eta &= \frac{\sqrt{P_{nr}}}{\sigma_{nx}} \\ &\cong \begin{cases} 0.718 & ; \ r = \sigma_{nx} \\ 0.960 & ; \ r = 2\sigma_{nx} \\ 0.998 & ; \ r = 3\sigma_{nx} \quad . \end{cases} \end{aligned} \quad (33)$$

A train of the integer signal $\hat{x}_{n,i}$ is converted to two-valued signal by two-valued integrand code of $A =$

$n_b^2/8$. The two-valued signal is amplified and transmitted through a low pass filter with impulse response $h_s(t)$. The received signal through a low pass filter with impulse response $h_r(t)$ is processed by a correlator with the reference code from $\{rd'_{M,\ell'_n,i-2n}\}$, to produce the signal shown in Fig.4(e), which is processed by another correlator with the reference code from $\{a_{M,\ell,i}\}$ to give the synchronizing pulse as shown in Fig.4(f). We can detect the data $d_{0,k}$ in Fig.4(e) by the synchronizing pulse in Fig.4(f). The multi-level and multi-interval data are rearranged to the original data array.

The distortion based on amplitude limitation and quantization is treated as an error of signal. The relation between the integer signal $\hat{x}_{n,i}$ and the real-valued signal $x_{n,i}$ is expressed as

$$K_{nc}\hat{x}_{n,i} = \eta\, x_{n,i} + \triangle x_{n,i} \quad (34)$$

where $\triangle x_{n,i}$ is an error with mean value 0 and variance σ_{nD}^2 from the real-valued signal $x_{n,i}$ and $\triangle x_{n,i}/K_{nc}$ is the error at the quantizer output. The transmittances of the quantization and the integrand coding are $1/K_{nc}(= A/r)$ and $1/A$, respectively. When the correlator output signal takes the peak magnitude at $t = t_0 + \mu T_c$ in the absence of noise and interchannel interference, the correlator output between the equivalent correlator input $(\eta x_{nd,i} + \triangle x_{nd,i})K_{Sin}/(K_{nc}A)$ and the reference sequence $\{ra'_{M,\ell'_n,i-2n}\}$ in the discrete model is given by

$$\begin{aligned}\tilde{x}_{nd,i'} &= \frac{K_{Sout}}{M}\sum_i\{\frac{\eta x_{nd,i+i'} + \triangle x_{nd,i+i'}}{K_{nc}A}\} \\ &\quad \times ra'_{M,\ell'_n,i-2n} \\ &= K_{Sout}(\eta \sum_{k=0}^{K_n(N-1)-1} d_{n,k}\cdot\rho_{M,\ell'_n,\ell'_n,i'-k\mu} \\ &\quad + \frac{1}{M}\sum_{i=0}^{M-1}\triangle x_{nd,i+i'}\cdot a'_{M,\ell'_n,i-2n}) \quad (35)\end{aligned}$$

where K_{Sin} and K_{Sout} are the coefficients based on low-pass filtering and correlation processing. In Eq.(35), the first term provides the output data signal power $\eta^2 K_{Sout}^2$, and the second term provides the output distortion $\sigma_{nD}^2 K_{Sout}^2/M$. The input data signal power is given by

$$\begin{aligned}S_{in} &= E[(\eta x_{nd,i}K_{Sin}/r)^2] \\ &= \frac{\eta^2 K_n K_{Sin}^2}{r^2} \quad (36)\end{aligned}$$

where $E[\,\cdot\,]$ denotes expectation.

If there is an additive white Gaussian noise with power $N_{in} = \sigma_{in}^2$ in the channel, the input and the output signal-to-noise ratios for signal level $V_{2\nu-1}$ are given by

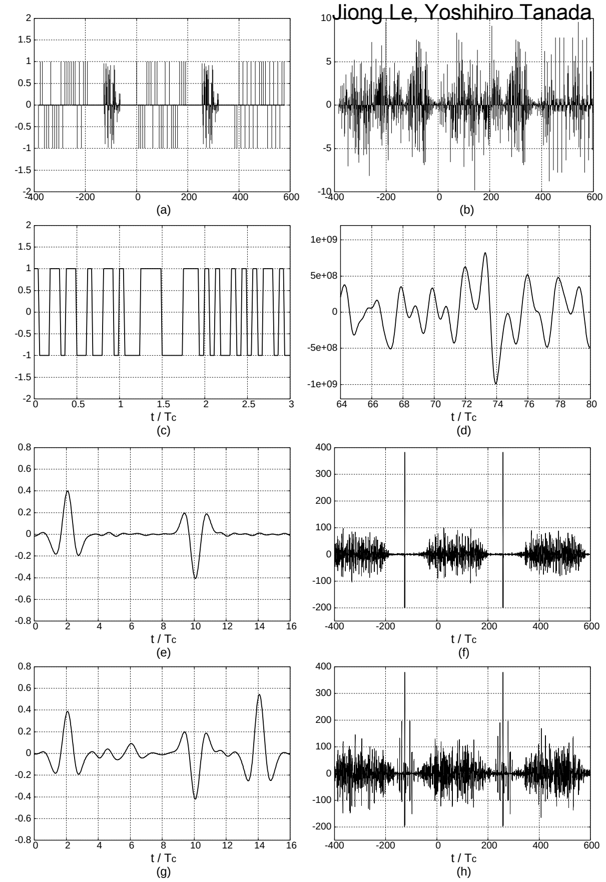

Figure 5: Signals in the experimental system.

$$SNR_{in} = \frac{S_{in}}{N_{in}} = \frac{\eta^2 K_n K_{Sin}^2}{r^2\sigma_{in}^2} \quad (37)$$

$$\begin{aligned}SNR_{out} &= \frac{S_{out,2\nu-1}}{N_{out}} \\ &= \frac{\eta^2\cdot V_{2\nu-1}^2}{\sigma_{nD}^2/M + K_{Nout}^2\sigma_{in}^2/M} \quad (38)\end{aligned}$$

where K_{Nout} is the coefficient based on low-pass filtering and correlation processing. If there are interchannel interference and the above noise, the expression corresponding to SNR_{out} has an additional term of interchannel interference in the denominator of Eq.(38). However, the magnitude of the interfernece is small in the permissible time shift range of this quasi-synchronous CDMA system. The values of the coefficients K_{Sin}, K_{Sout} and K_{Nout} are obtained by specifying the impulse responses $h_s(t)$ and $h_r(t)$[8].

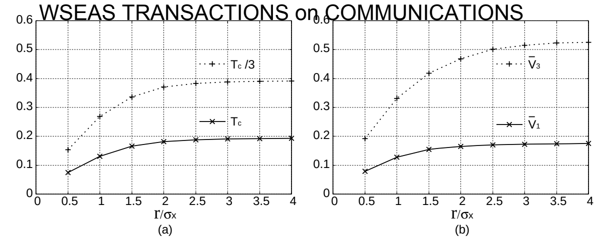

Figure 6: Correlator data output levels. (a) Two-valued data. (b) Four-valued data.

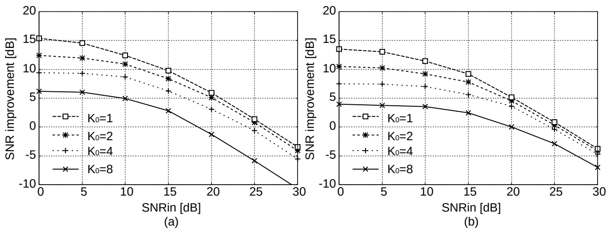

Figure 7: SNR improvement by correlation processing. (a) Moving average of $T_c/3$. (b) Moving average of T_c.

5 Numerical Experiments

We examine the proposed CDMA system using a set of sequences with the zero correlation zone $-1 \leq i' \leq 1$ by numerical experiments, where $M = 65$, $N = 4$, $K_n = 1, 2, \cdots, 8$ and $p = 2, 4$. We use two-valued integrand codes with $n_b = 16$ ($A = 32$) and the orthogonal sequences $\{d_{65,\ell'_n,i-2n}; n = 0, 1, 2, 3\}$, $\varepsilon'_{64} = -1/65$, and the common sequence $\{a_{65,\ell,i}\}$, $\varepsilon_{64} = 1/65$, where $|a'_{65,\ell'_n,i}|_{max} \cong 2.7463$, $|\rho_{65,\ell_n,\ell'_n,i'}|_{max} \cong 0.5268$, $|a_{65,\ell,i}|_{max} \cong 1.9979$.

A low pass filter in the transmitter is replaced by the moving averager of $T_c/3$ or T_c, and a low pass filter in the receiver is by the 6-stage connection of the moving averager of $T_c/3$. Fig.5 shows signals of the transmitter and receiver on user 0 in the experimental CDMA system, where $K_0 = 8$, $p = 2$, and the moving averager of T_c is used for the low pass filter in the transmitter, and there exists no additive noise. Figs.5(a), (b), (c), (d), (e), and (f) show a train of data and synchronizing sequence, its convolved signal, the integrand coding signal, the correlator input signal, the correlator output signal for data and that for synchronizing pulse, respectively. Figs.5(g) and (h) show the correlator output signal for data and that for synchronizing pulse when the signal of user 1 is mixed. Fig.6(a) shows the correlator data ouput levels for $p = 2$ when the moving average durations are $T_c/3$ and T_c in the transmitter. Fig.6(b) shows the correlator data output levels for $p = 4$ when the moving average duration is $T_c/3$ in the transmitter. Figs.7(a) and (b) show the SNR improvements for $p = 4$ and $\nu = 1$ when the moving average durations are $T_c/3$ and T_c in the transmitter. These results explain that the higher cut-off low pass filter in the transmitter gives the output data with the higher level. Fig.8 shows the correlation output distortion for $p = 2$

$$D = 20log_{10}\frac{\sigma_D}{\bar{V}_1} \tag{39}$$

where σ_D is the standard deviation of the output distortion and $\bar{V}_1$ is the mean absolute value of the output data. In Figs.8(a) and (b), the moving averager of $T_c/3$ and T_c, respectively, are used in the transmitter. This result explains that the amplitude limitation of $r = 3\sigma_x$ is sufficient and the low pass filter of lower cut-off suppresses the distortion to lower level. Fig.9 shows the correlation output interference for $p = 2$

$$I = 20log_{10}\frac{\sigma_I}{\bar{V}_1} \tag{40}$$

where there is no noise, the signals of users 1, 2 and 3 are inphase, the signal of user 0 is shifted in the time range of $-T_c \leq \delta T_c \leq T_c$, σ_I is the standard deviation of interference and $\bar{V}_1$ is the mean absolute value of the output data. In Figs.9(a) and (b), the moving averager of $T_c/3$ and T_c, respectively, are used in the transmitter. This result shows that interference suppression in a certain shift range and the low pass filter of higher cut-off suppresses the interference in the wider shift range due to narrower pulse width.

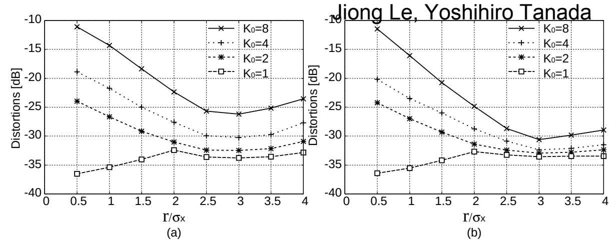

Figure 8: Output distortion (p=2). (a) Moving average of $T_c/3$. (b) Moving average of T_c.

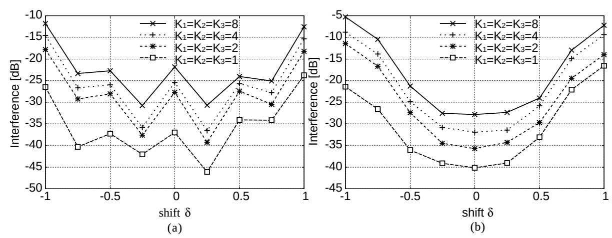

Figure 9: Interference suppression (p=2). (a) Moving average of $T_c/3$. (b) Moving average of T_c.

6 Conclusion

Two-valued integrand codes are applied to the realization of the multi-rate quasi-synchronous CDMA system using real-valued self-orthogonal finite-length sequences with ZCZ. Data rate is controlled by the multi-level and multi-interval data, which is convolved with the distinctive real-valued self-orthogonal finite-length sequence with ZCZ. This convolution ensures the suppression of noise and interchannel interference in the channel. The convolved real-valued signal is converted to two-valued integrand coding signal for easy amplification and modulation. The integrand coding is the technique to use much fewer clock pulses than the pulse-width-modulation. The two-valued signal is changed to real-valued signal through the low pass filters in the channel. The correlation between the real-valued signal and the user code yields the sharp output signal for the discrimination. The proposed system suppresses distortion on amplitude limitation and quantization, intersymbol interference and interchannel interference. The system is useful for the multimedia data transmission for the demand of data rate and data reliability. Furthermore, it only need one spreading code for each user.

For the future works, there exist the practical subjects; the evaluation of the system performances in the communication environment of near and far distances, multi-path, moving stations and so on by numerical experiments; the evaluation of the performances of the real communication system of physical channels and hardware devices by the practical experiments; the comparison of the performances between this system and the other systems. This system, after verifying the significance of the performances, is to be applied to cellular mobile communication, wireless LAN, power-line LAN, ad hoc network etc.

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