MULTICODE MIMO SYSTEMS WITH QUATERNARY LCZ AND ZCZ SEQUENCES

Jae-Dong Yang, Xianglan Jin, Kyoung-Young Song, Department of EECS, Seoul National University, Seoul, Korea, Email: {yjdong, xianglan.jin, sky6174}@ccl.snu.ac.kr

Jong-Seon No, Department of EECS, Seoul National University, Seoul, Korea, Email: jsno@snu.ac.kr

Dong-Joon Shin, Division of ECE, Hanyang University, Seoul, Korea, Email: djshin@hanyang.ac.kr *

Abstract

In this paper, we propose multicode multiple-input multiple-output (MIMO) systems with quaternary low correlation zone (LCZ) and zero correlation zone (ZCZ) sequences as spreading codes. Quaternary LCZ and ZCZ sequences have very low correlation values when the time shifts of correlation function are within the predetermined correlation zone and thus the multi-user or multipath interference can be substantially reduced when the delay is within a few chips. The bit error probability of the proposed systems is theoretically analyzed, which is numerically confirmed. It is also numerially shown that the performance of multicode MIMO systems with quaternary LCZ and ZCZ sequences is better than that of the conventional multicode MIMO systems with quaternary spreading codes constructed from a pair of binary Hadamard codes.

I. Introduction

Multiple-input multiple-output (MIMO) system is a very attractive technique for increasing the capacity of wireless communication systems [1]. By utilizing high spatial dimension by multiple antennas, high spectral efficiency can be achieved in the wireless communication systems. In code division multiple access (CDMA) systems, higher data rate communication can be achieved by using multicode channelization [2]. To accommodate the demand for various high data rate services, we can construct a system by combining these two techniques, that is, multiple antennas and multicode schemes, which is called a multicode MIMO system. Multicode MIMO systems are considered as a standard technique for high speed downlink packet access (HSDPA) systems [3]. In the multicode MIMO systems, each transmit antenna uses the same set of spreading codes and usually, a pair of binary Hadamard codes have been used as a quaternary spreading code.

If the channel is frequency selective, there are delayed multipaths that make an irreducible error floor. Power allocation technique is proposed to improve the performance [4], but accurate feedback information is needed.

Low correlation zone (LCZ) and zero correlation zone (ZCZ) sequences [5, 6] have very low autocorrelation and cross-correlation values when their time shifts are within the predetermined correlation zone. Therefore, they are suitable for the quasi-synchronous CDMA systems and multipath resolution for CDMA systems [7]. In this paper, we assume that the delays are much smaller than the symbol duration (or the period of spreading code) and propose multicode MIMO systems which use quaternary LCZ and ZCZ sequences instead of binary Hadamard codes as spreading codes. Thus, the proposed multicode MIMO system can be used for the high speed data transmission with multipath resolution in the isolated cell environments. Because of two-dimensional (spatial and code domains) interference, we need two-dimensional successive interference cancellation (SIC) detection. However, it is shown that in our systems only one-dimensional (spatial domain) SIC detection shows negligible performance degradation compared with two-dimensional SIC detection. The bit error probability of the proposed systems is theoretically analyzed, which is numerically confirmed.

II. System Model

We assume the multicode MIMO system with N_t transmit antennas and N_r receive antennas, such that each transmit antenna uses the same set of spreading codes $\{c_1(t), c_2(t), \cdots, c_K(t)\}$, where K denotes the number of spreading codes. Fig. 1 shows the multicode MIMO system. We assume that the quadrature phase shift keying (QPSK) modulation is used and quaternary LCZ and ZCZ sequences are used as spreading codes, that is, complex spreading is used. Spreading codes have a period G, which corresponds to the processing gain. A data stream is demultiplexed into N_t groups and each group is partitioned into K streams of data symbols. The transmitted signal at the nth transmit antenna for one data symbol duration is given as

$$x_n(t) = \sum_{i=1}^{K} d_{ni}\ c_i(t),\ n = 1, 2, \cdots, N_t,\ 0 \le t \le GT_c \quad (1)$$

where d_{ni} is a data symbol at the nth transmit antenna for the ith spreading code $c_i(t)$ and T_c is a chip duration. Since delay is assume to be much smaller than the symbol duration, the intersymbol interference can be ignored. Note that, generally the intersymbol interference in the multicode MIMO systems cannot be ignored. Thus interferences in the spatial and code domains are greater than the intersymbol interference.

The channel impulse response from the nth transmit antenna to the mth receive antenna with L multipath components is given as

$$h_{mn}(t) = \sum_{l=0}^{L-1} h_{lmn}\delta(t - \tau_l) \quad (2)$$

*This work was supported by the ITRC program of the MIC, Korea, and by the MOE, the MOCIE, and the MOLAB, Korea, through the fostering project of the Lab of Excellency.

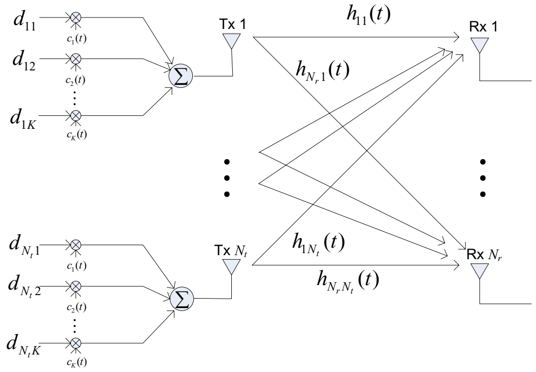


Figure 1: Multicode MIMO system.

where h_{lmn} is the channel coefficient of the lth multipath from the nth transmit antenna to the mth receive antenna with zero mean complex Gaussian distribution, $\delta(\cdot)$ a delta function, and τ_l a time delay of the lth multipath. We assume $\tau_l = lT_c$ with $L \ll G$ and the exponential multipath intensity profile so that the powers of channel coefficients $\Omega_l = E[|h_{lmn}|^2]$ are exponentially distributed, i.e., $\Omega_l = \Omega_0 e^{-l\zeta}, \quad l = 0, 1, \cdots, L-1$, and $\sum_{l=0}^{L-1} \Omega_l = 1$, where ζ is the rate of the exponential decay.

The received signal at the mth receive antenna at time t is given as

$$r_m(t) = \sum_{n=1}^{N_t} h_{mn}(t) * x_n(t) + n_m(t), \;\; m = 1, 2, \cdots, N_r \quad (3)$$

where $*$ denotes the convolution and $n_m(t)$ an additive white Gaussian noise. Using (1) and (2), (3) can be rewritten as

$$r_m(t) = \sum_{n=1}^{N_t} \sum_{l=0}^{L-1} h_{lmn} \sum_{i=1}^{K} d_{ni} c_i(t - lT_c) + n_m(t). \quad (4)$$

In this paper, we propose and analyze the multicode MIMO system using quaternary LCZ and ZCZ sequences as spreading codes, which consist of elements in $\mathbb{Z}_4 = \{0, 1, 2, 3\}$. Since QPSK modulation is used to transmit symbols in $\mathbb{Z}_4$, the quaternary LCZ and ZCZ sequences can be easily used for complex spreading. In the next subsections, quaternary LCZ and ZCZ sequences are briefly introduced and the detection scheme is explained.

A. Quaternary LCZ and ZCZ Sequences

Quaternary LCZ (ZCZ) sequences [5, 6] are sequences which have low (zero) correlation values when their time shifts are within the predetermined correlation zone. In this subsection, we introduce the quaternary LCZ and ZCZ sequences which are used to construct the multicode MIMO systems.

First, we explain a construction method of quaternary LCZ sequences [5]. Let $b(t)$ be a binary m-sequence of period $N = 2^s - 1$. Let $e|s$ and $M = 2^e - 1$. Then, LCZ sequences $c_i(t), i = 1, \cdots, M$, can be constructed as

$$\begin{aligned} c_1(t) &= 2b(t) & (5)\\ c_{i+1}(t) &= b(t) + 2b\left(t + iS\right), \text{ for } i = 1, 2, \cdots, M-1 \end{aligned}$$

where M is the family size of LCZ sequences and the LCZ size $S = N/M$. Note that if the delay between two LCZ sequences is less than S, the magnitude of correlation value becomes 1.

It is easy to check that the family size of these LCZ sequences can be increased u times by reducing the LCZ size to $\lfloor S/u \rfloor$, that is,

$$\begin{aligned} &\text{For } i = 1, 2, \cdots, M, \text{ and } j = 0, 1, 2, \cdots, u-1 \\ &c_{i+jM}(t) = c_i\left(t + j\left\lfloor \frac{S}{u} \right\rfloor\right) \quad (6) \end{aligned}$$

where $\lfloor S/u \rfloor$ denotes the largest integer less than or equal to S/u. Clearly, if the delay between these sequences is less than $\lfloor S/u \rfloor$, the magnitude of correlation value is 1. Note that even if these sequences are not cyclically distinct, they can be used for multicode MIMO systems as other LCZ sequences and we will call these sequences as LCZ sequences.

Quaternary ZCZ seqeunces can be constructed using a quaternary perfect sequence which has perfect autocorrelation property [6]. Let $b(t)$ be a perfect sequence. Then the ZCZ sequences can be iteratively constructed as follows.

$$\begin{aligned} &\text{Initial}: c_0^1(t) = b(t),\ M_0 = 1,\ N_0 = \text{period of } b(t) \\ &\text{The } i\text{th iteration}: \\ &\quad \text{For } j = 1, 2, \cdots, M_{i-1},\ N_i = 2N_{i-1},\ M_i = 2M_{i-1}, \\ &\qquad c_i^j(2t) = c_{i-1}^j(t), \\ &\qquad c_i^j(2t+1) = c_{i-1}^j(t + \tfrac{N_0}{2}), \\ &\qquad c_i^{j+M_{i-1}}(2t) = c_{i-1}^j(t), \qquad (7)\\ &\qquad c_i^{j+M_{i-1}}(2t+1) = [c_{i-1}^j(t + \tfrac{N_0}{2}) + 2] \bmod 4, \\ &\qquad 0 \le t < N_{i-1} \end{aligned}$$

where N_i is the period of the ZCZ sequences after the ith iteration and M_i the family size of the ZCZ sequences after the ith iteration.

B. Detection Scheme

We assume that there are L different delayed multipath signals and RAKE receiver with L fingers. The correlator output of the l'th finger at the mth receive antenna for the data symbols spreaded with $c_k(t)$ is given by

$$\begin{aligned} y_{ml'}^k &= \int_{l'T_c}^{l'T_c + GT_c} r_m(t) c_k^*(t - l'T_c) dt & (8)\\ &= \sum_{n=1}^{N_t} \sum_{l=0}^{L-1} h_{lmn} d_{nk} R_{kk}(l, l') & (9)\\ &+ \sum_{n=1}^{N_t} \sum_{l=0}^{L-1} \sum_{i \neq k}^{K} h_{lmn} d_{ni} R_{ik}(l, l') & (10)\\ &+ n_{ml'}^k & (11) \end{aligned}$$

where $R_{ab}(l, l') = \int_{l'T_c}^{l'T_c+GT_c} c_a(t - lT_c) c_b^*(t - l'T_c) dt$ and $n_{ml'}^k = \int_{l'T_c}^{l'T_c+GT_c} n_m(t) c_k^*(t - l'T_c) dt$. Equation (8) can be divided into three different signals as:

i) Sum of the desired signal and the interferences from delayed multipath signals and signals from other transmit antennas with the same spreading codes in (9);

ii) Interferences from signals using different spreading codes from all transmit antennas in (10);
iii) Additive white Gaussian noise in (11).

Then, there are three different types of interferences, which can be resolved by using the following methods:

i) Delayed multipath interference, which is resolved by the RAKE receiver;
ii) Code domain interference, which is resolved by the code domain SIC;
iii) Spatial domain interference, which is resolved by the spatial domain SIC.

Therefore, for the multicode MIMO system two-dimensional SIC for the detection is used to resolve the spatial and code domain interferences [4].

For detecting data symbols $d_{nk}, n = 1, 2, \cdots, N_t$, we collect all finger outputs $y^k_{ml'}$ in (8) and thus we have

$$\mathbf{y}_k = \mathbf{H}_k \mathbf{d}_k + \sum_{i \neq k}^{K} \mathbf{J}^i_k \mathbf{d}_i + \boldsymbol{\eta}_k \tag{12}$$

where

$$\begin{aligned}
\mathbf{y}_k &= [y^k_{10} \cdots y^k_{1L-1} y^k_{20} \cdots y^k_{N_r L-1}]^T, \mathbf{d}_k = [d_{1k} \cdots d_{N_t k}]^T, \\
\boldsymbol{\eta}_k &= [n^k_{10} \cdots n^k_{1L-1} n^k_{20} \cdots n^k_{N_r L-1}]^T, \\
\mathbf{H}_k &= \begin{bmatrix} \mathbf{R}_{kk} & & \mathbf{0} \\ & \ddots & \\ \mathbf{0} & & \mathbf{R}_{kk} \end{bmatrix} \underbrace{\begin{bmatrix} \mathbf{H}(1) \\ \vdots \\ \mathbf{H}(N_r) \end{bmatrix}}_{\boldsymbol{\mathcal{H}}}, \\
\mathbf{J}^i_k &= \begin{bmatrix} \mathbf{R}_{ik} & & \mathbf{0} \\ & \ddots & \\ \mathbf{0} & & \mathbf{R}_{ik} \end{bmatrix} \begin{bmatrix} \mathbf{H}(1) \\ \vdots \\ \mathbf{H}(N_r) \end{bmatrix}, \\
\mathbf{R}_{ik} &= \begin{bmatrix} R_{ik}(0,0) & \cdots & R_{ik}(L-1,0) \\ \vdots & & \vdots \\ R_{ik}(0,L-1) & \cdots & R_{ik}(L-1,L-1) \end{bmatrix}, \text{and} \\
\mathbf{H}(m) &= \begin{bmatrix} h_{0m1} & \cdots & h_{0mN_t} \\ \vdots & & \vdots \\ h_{L-1m1} & \cdots & h_{L-1mN_t} \end{bmatrix}.
\end{aligned}$$

It is easy to check that (12) is similar to the conventional MIMO system. Thus, we can apply V-BLAST detection scheme [8] to (12) for detecting $d_{nk}, n = 1, 2, \cdots, N_t$, which uses the spatial domain SIC. Moreover, we can further improve the performance for detecting $\mathbf{d}_k$ by successively cancelling the code domain interference using the previously detected data symbols $\hat{\mathbf{d}}_1, \hat{\mathbf{d}}_2, \cdots, \hat{\mathbf{d}}_{k-1}$, which is called the code domain SIC. Here, we do not consider the ordering for code domain SIC. Then (12) can be modified into

$$\mathbf{y}'_k = \mathbf{y}_k - \sum_{i=1}^{k-1} \mathbf{J}^i_k \hat{\mathbf{d}}_i = \mathbf{H}_k \mathbf{d}_k + \mathbf{e}_k \tag{13}$$

where

$$\mathbf{e}_k = \sum_{i=1}^{k-1} \mathbf{J}^i_k (\mathbf{d}_i - \hat{\mathbf{d}}_i) + \sum_{i=k+1}^{K} \mathbf{J}^i_k \mathbf{d}_i + \boldsymbol{\eta}_k. \tag{14}$$

In the proposed multicode MIMO systems, quaternary LCZ and ZCZ sequences are used as spreading codes and we already assume that the maximum time delay $(L-1)T_c$ is assumed to be less than their correlation zone size. Therefore, we can only perform the spatial domain SIC in the receiver without performance degradation, which is shown in the next section.

III. Performance Analysis And Numerical Results

For the performance analysis of the proposed multicode MIMO system, we assume that the channel and interference matrices are already known. To find the average error probability P_e, we can use the following relation

$$P_e = E_{\boldsymbol{\mathcal{H}}}[P_{e|\boldsymbol{\mathcal{H}}}] \tag{15}$$

where $P_{e|\boldsymbol{\mathcal{H}}}$ is the conditional error probability defined by

$$P_{e|\boldsymbol{\mathcal{H}}} = \frac{1}{KN_t} \sum_{n,i} P(d_{ni} \neq \hat{d}_{ni} | \boldsymbol{\mathcal{H}}) \tag{16}$$

and $E_{\boldsymbol{\mathcal{H}}}[\cdot]$ means the expectation over the channel matrix $\boldsymbol{\mathcal{H}}$. To find the conditional error probability of the SIC detection scheme, we should derive the signal to interference and noise ratio (SINR) and average the error probabilities of data symbols.

To simplify the problem, we ignore the error propagation, i.e., we assume that the previously detected data symbols are correct. In order to derive the error probability of the proposed multicode MIMO system, the SINR of the input signal of the detector should be derived. Let $SINR_{ni}$ be the SINR corresponding to d_{ni}. It is well known that the error probability of QPSK modulation is given as

$$P_e = Q(\sqrt{SINR}) \tag{17}$$

and for the proposed system, the conditional error probability of d_{ni} is given as $P(d_{ni} \neq \hat{d}_{ni} | \boldsymbol{\mathcal{H}}) = Q(\sqrt{SINR_{ni}})$.

Since we assume that there is no error propagation, (14) can be rewritten as

$$\mathbf{e}_k = \sum_{i=k+1}^{K} \mathbf{J}^i_k \mathbf{d}_i + \boldsymbol{\eta}_k. \tag{18}$$

If L is smaller than the LCZ and ZCZ size, $R_{ab}(l, l')$ is much smaller than the period of sequence when $a \neq b$ or $l \neq l'$ and thus in the proposed system, the code domain interference level is very low. If the spatial domain SIC is performed, (13) can be modified as

$$\mathbf{y}_k(j) = \mathbf{H}_k(j) \mathbf{d}_k(j) + \mathbf{e}_k, \quad j = 0, 1, \cdots, N_t - 1 \tag{19}$$

where j is the number of previously detected data symbols, $\mathbf{y}_k(j)$ is the j spatial domain interference cancelled signal, and $\mathbf{H}_k(j)$ is the corresponding deflated channel matrix [8]. Let $\mathbf{H}_k(j)^\dagger$ be the Moore-Penrose pseudo-inverse of $\mathbf{H}_k(j)$. According to the zero-forcing criterion [8], we perform the following procedure

$$\mathbf{H}_k(j)^\dagger \mathbf{y}_k(j) = \mathbf{d}_k(j) + \mathbf{H}_k(j)^\dagger \mathbf{e}_k. \tag{20}$$

Since we can assume that the data symbols have unit power, we only need to calculate the interference and noise power as

$$E\left[\left(\mathbf{H}_k(j)^\dagger \mathbf{e}_k\right)\left(\mathbf{H}_k(j)^\dagger \mathbf{e}_k\right)^H\right] = \mathbf{H}_k(j)^\dagger E\left[\mathbf{e}_k\mathbf{e}_k^H\right]\left(\mathbf{H}_k(j)^\dagger\right)^H \tag{21}$$

where $(\cdot)^H$ denotes the transpose and complex conjugate. From (18), we have

$$E\left[\mathbf{e}_k\mathbf{e}_k^H\right] = E\left[\left(\sum_{i=k+1}^{K}\mathbf{J}_k^i\mathbf{d}_i + \boldsymbol{\eta}_k\right)\left(\sum_{i=k+1}^{K}\mathbf{J}_k^i\mathbf{d}_i + \boldsymbol{\eta}_k\right)^H\right]$$
$$= E\left[\sum_{a=k+1}^{K}\sum_{b=k+1}^{K}\mathbf{J}_k^a\mathbf{d}_a\mathbf{d}_b^H(\mathbf{J}_k^b)^H\right] + E\left[\boldsymbol{\eta}_k\boldsymbol{\eta}_k^H\right].$$

It is easy to check that the covariance matrix of the noise term is given as

$$E[\boldsymbol{\eta}_k\boldsymbol{\eta}_k^H] = \sigma^2\begin{bmatrix}\mathbf{R}_{kk} & & \mathbf{0}\\ & \ddots & \\ \mathbf{0} & & \mathbf{R}_{kk}\end{bmatrix} \simeq \sigma^2 G\mathbf{I}$$

where $\sigma^2 = E[|n(t)|^2]$ and $\mathbf{R}_{kk}$ is an autocorrelation matrix of the spreading code $c_k(t)$ defined in (12). If quaternary LCZ or ZCZ sequences are used, $\mathbf{R}_{kk}$ becomes similar to the identity matrix because the off-diagonal terms of $\mathbf{R}_{kk}$ are much smaller than the identical diagonal terms. The covariance matrix of the interference term can be rewritten as

$$E\left[\sum_{a=k+1}^{K}\sum_{b=k+1}^{K}\mathbf{J}_k^a\mathbf{d}_a\mathbf{d}_b^H(\mathbf{J}_k^b)^H\right]$$
$$= \sum_{a=k+1}^{K}\sum_{b=k+1}^{K}\mathbf{J}_k^a E\left[\mathbf{d}_a\mathbf{d}_b^H\right](\mathbf{J}_k^b)^H$$
$$= \sum_{i=k+1}^{K}\mathbf{J}_k^i E\left[\mathbf{d}_i\mathbf{d}_i^H\right](\mathbf{J}_k^i)^H = \sum_{i=k+1}^{K}\mathbf{J}_k^i(\mathbf{J}_k^i)^H.$$

Therefore, we have

$$E\left[\mathbf{e}_k\mathbf{e}_k^H\right] \simeq \sum_{i=k+1}^{K}\mathbf{J}_k^i(\mathbf{J}_k^i)^H + \sigma^2 G\mathbf{I}. \tag{22}$$

The ordering of data symbol detection in the spatial domain SIC should be determined by using the SINR, that is, the data symbol with the largest SINR should be detected first. In other words, we should detect the data symbol with the minimum interference and noise power, which corresponds to

$$\min_{p(j)}\left\{\left[\mathbf{H}_k(j)^\dagger\left(\sum_{i=k+1}^{K}\mathbf{J}_k^i(\mathbf{J}_k^i)^H\right)(\mathbf{H}_k(j)^\dagger)^H\right]_{p(j)} + \sigma^2 G\left[\mathbf{H}_k(j)^\dagger(\mathbf{H}_k(j)^\dagger)^H\right]_{p(j)}\right\}$$

where $[\cdot]_{p(j)}$ denotes the $p(j)$-th diagonal element. Using (21) and (22), SINR of the data symbol $d_{p(j)k}$ for detection can be derived as

$$SINR_{p(j)k} \simeq \left[\min_{p(j)}\left[\left[\mathbf{H}_k(j)^\dagger\left(\sum_{i=k+1}^{K}\mathbf{J}_k^i(\mathbf{J}_k^i)^H\right)(\mathbf{H}_k(j)^\dagger)^H\right]_{p(j)} + \sigma^2 G\left[\mathbf{H}_k(j)^\dagger(\mathbf{H}_k(j)^\dagger)^H\right]_{p(j)}\right]\right]^{-1}. \tag{23}$$

From (15), (17), and (23), the average error probability can be derived as

$$P_e \simeq E_{\boldsymbol{\mathcal{H}}}\left[\frac{1}{KN_t}\sum_{j,k} Q\left(\sqrt{SINR_{p(j)k}}\right)\right]. \tag{24}$$

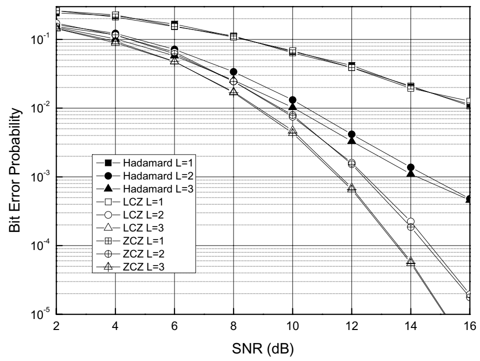
Using (24) and the Monte Carlo method, the average bit error probability (BEP) of the proposed multicode MIMO systems with quaternary LCZ and ZCZ sequences can be evaluated.

We assume that the processing gain G is 64 and there are multipath signals upto $L = 3$, and use 4 transmit antennas and the exponential multipath intensity profile with $\zeta = 0.5$. Each transmit antenna uses 8 spreading codes, $K = 8$, when ZCZ sequences are used, and upto 21 spreading codes, $K = 21$, when LCZ sequences are used. More specifically, quaternary LCZ sequences with period 63 and the family size $M = 7$ are generated by (5) using a binary m-sequence with period 63. To increase the family size of quaternary LCZ sequences, (6) is used with $u = 3$ to generate 21 LCZ sequences. To make $G = 64$, one zero is padded to each quaternary LCZ sequence. Quaternary ZCZ sequences are generated by (7) using a quaternary perfect sequence with period 8, i.e., $\{0, 0, 1, 2, 0, 2, 1, 0\}$. To make $G = 64$ and $M = 8$, three iterations are performed. The conventional multicode MIMO systems assume to use the complex spreading by quaternary spreading codes constructed from a pair of binary Hadamard codes of length 64.

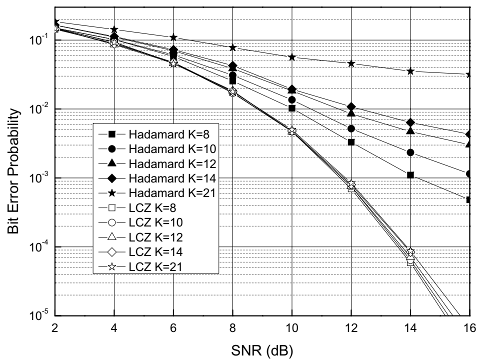
Fig. 2(a) compares the BEP of the proposed multicode MIMO system and the conventional multicode MIMO system using binary Hadamard codes when the number of transmit antennas is four and $K = 8$. When $L = 1$, i.e., there is no delayed multipath signal, three systems show the almost identical performance. If $L > 1$, there is an irreducible error floor due to the code domain interference in the conventional multicode MIMO system with binary Hadamard codes. But the proposed multicode MIMO system shows no error floor and its performance becomes better than that of the conventional system as L becomes lager. Fig. 2(b) compares the BEP for $L = 3$ and the various number K of spreading codes. As K becomes larger, the performance of the multicode MIMO systems with binary Hadamard codes becomes worse, but the proposed system shows very small performance degradation due to the correlation property of quaternary LCZ sequences.

Fig. 3 compares the performance of two-dimensional SIC and one-dimensional SIC. As expected, there is negligible performance loss in the proposed systems, but there is significant performance loss in the multicode MIMO system with binary Hadamard codes. Thus, one-dimensional SIC detection is sufficient for the proposed multicode MIMO systems and the detection complexity can be reduced.

Fig. 4 shows the theoretical and simulation results for the multicode MIMO system with quaternary ZCZ sequences.



(a) Various L values with $K = 8$



(b) Various K values with $L = 3$

Figure 2: Performance comparison of 4×4 multicode MIMO systems with QPSK using two-dimensional SIC.

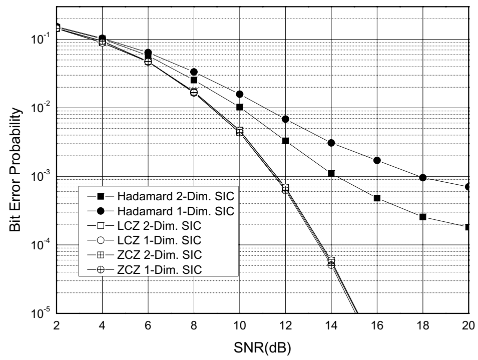


Figure 3: Performance comparision of two-dimensional SIC and one-dimensional SIC for 4×4 multicode MIMO systems with QPSK when $K = 8$ and $L = 3$.

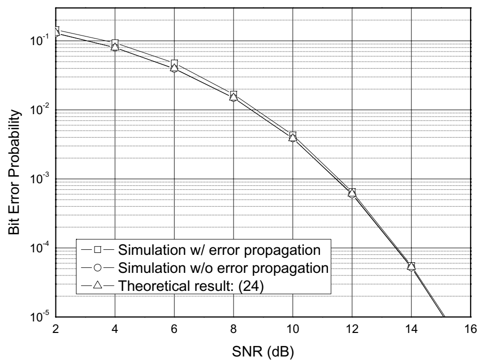


Figure 4: Comparison of theoretical and simulation results when ZCZ sequences are used for 4×4 multicode MIMO systems with $K = 8$ and $L = 3$.

Since we assume that there is no error propagation when we derive the BEP, there is a gap between the theoretical performance and simulation result in low SNR region. When we simulate the proposed system by assuming a genie detector, the theoretical and simulation results match well over the whole SNR region.

IV. Conclusion

We proposed and analyzed multicode MIMO systems using quaternary LCZ and ZCZ sequences instead of binary Hadamard codes. Since LCZ and ZCZ sequences have very good correlation property within the predetermined correlation zone, the proposed systems outperform the conventional multicode MIMO system with binary Hadamard codes if the delay is limited within the correlation zone. The throughput of the proposed system can also be improved via increasing the number of spreading codes without performance degradation. The theoretical analysis matches well with the numerical result in high SNR region, and in low SNR region there is a small gap due to the no error propagation assumption.

References

[1] G. J. Foschini and M. J. Gans, "On limits of wireless communications in a fading environment when using multiple antennas," *Wireless Pers. Commun.*, vol. 6, pp. 311–335, Mar. 1998.

[2] "3G TS 25.213 version 3.0.0, Spreading and Modulation (FDD)," 3GPP TSG-RAN, 1999–10.

[3] "Physical layer aspects of UTRA High Speed Downlink Packet Access," Tech. Rep. 3GPP TR 25.848 v4.0.0, 2001–3.

[4] C.-S. Park and K.-B. Lee, "Transmit power allocation for succesive interference cancellation in multicode MIMO systems," *in Proc. IEEE ICC2005*, May 2005.

[5] S.-H. Kim, J.-W. Jang, J.-S. No, and H. Chung, "New constructions of quaternary low correlation zone sequences," *IEEE Trans. Inf. Theory*, vol. 51, no. 4, pp. 1469–1477, Apr. 2005.

[6] H. Torii, M. Nakamura, and N. Suehiro, "A new class of zero-correlation zone sequences," *IEEE Trans. Inf. Theory*, vol. 50(3), pp. 559–565, Mar. 2004.

[7] R. De Gaudenzi, C. Elia, and R. Viola, "Bandlimited quasisynchronous CDMA: A novel satellite access technique for mobile and personal communication system," *IEEE J. Sel. Area Commun.,* vol. 10, pp. 328–343, Feb. 1992.

[8] P. W. Wolniansky, G. J. Foschini, G. D. Golden, and R. A. Valenzuela, "V-BLAST: An architecture for realizing very high data rates over the rich-scattering wireless channel," in *Proc. ISSSE*, Pisa, Italy, Sep.–Oct. 1998, pp. 295–300.