Multicode MIMO Systems With Quaternary LCZ and ZCZ Sequences

Jae-Dong Yang, Xianglan Jin, Kyoung-Young Song, Jong-Seon No, *Member, IEEE*, and Dong-Joon Shin, *Member, IEEE*

***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 between these sequences are within the predetermined correlation zone, and thus, the multiuser 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 numerically shown that the performance of the 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 pairs of binary Hadamard codes.**

***Index Terms*—Hadamard codes, low-correlation zone (LCZ) sequences, multicode multiple-input multiple-output (MIMO), wireless local area network (WLAN), zero-correlation zone (ZCZ) sequences.**

I. Introduction

MULTIPLE-INPUT–multiple-output (MIMO) techniques have widely been used to increase the capacity of wireless communication systems [1]. By utilizing high spatial dimension of 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, i.e., MIMO and multicode techniques, which is called a multicode MIMO system. Multicode MIMO systems are considered to be standard techniques for high-speed downlink packet access systems [3]. In conventional multicode MIMO systems, each transmit antenna uses the same set of spreading codes, and usually, a pair of binary Hadamard codes has been used to make a quaternary spreading code.

Low-correlation zone (LCZ) and zero-correlation zone (ZCZ) sequences [4], [5] have very low autocorrelation and cross-correlation values when the time shifts between the sequences are within the predetermined correlation zone. Therefore, they are suitable for the quasi-synchronous CDMA systems and the multipath resolution for CDMA systems [6], [7]. In this paper, we assume that the delays are much smaller than the data symbol duration (or the period of spreading code) and propose multicode MIMO systems that use quaternary LCZ and ZCZ sequences instead of binary Hadamard codes as spreading codes. Because of 2-D (spatial and code domains) interference, we need 2-D successive interference cancellation (SIC) detection. However, it is shown that for the proposed systems, 1-D (spatial domain) SIC detection shows negligible performance degradation compared with 2-D SIC detection. Also, the bit error probability (BEP) of the proposed systems is theoretically analyzed, which is numerically confirmed, and it is shown that the proposed systems outperform the conventional multicode MIMO systems. Thus, the proposed multicode MIMO system can be used for the high-speed data transmission with multipath resolution in the isolated cell environments, such as hotspots of wireless local area network (WLAN) and indoor wireless communication systems, etc.

This paper is organized as follows: In Section II, the system model of multicode MIMO systems is presented, together with the construction methods of LCZ and ZCZ sequences and their detection scheme. The performance analysis and the numerical results of multicode MIMO systems with LCZ and ZCZ sequences as spreading codes are presented in Section III. Finally, the conclusion is given in Section IV.

II. System Model

The multicode MIMO system is assumed to have N_t transmit antennas and N_r receive antennas, such that each transmit antenna uses the same set of spreading codes $\{c_1(t), c_2(t), \ldots, c_K(t)\}$, where K denotes the number of spreading codes. Fig. 1 shows the multicode MIMO system. We assume that quaternary LCZ and ZCZ sequences are used as spreading codes, that is, complex spreading is used [8]. Spreading codes have the 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.

Manuscript received April 9, 2007; revised September 15, 2007 and October 23, 2007. This work was supported in part by the Ministry of Education and Human Resources Development (MOE), by the Ministry of Commerce, Industry, and Energy (MOCIE), and by the Ministry of Labor (MOLAB), Korea, through the fostering project of the Laboratory of Excellence. The review of this paper was coordinated by Prof. H. Nguyen.

J.-D. Yang, X. Jin, K.-Y. Song, and J.-S. No are with the Institute of New Media and Communications, Department of Electrical Engineering and Computer Science, Seoul National University, Seoul 151-744, Korea (e-mail: yjdong@ccl.snu.ac.kr, xianglan.jin@ccl.snu.ac.kr, sky6174@ccl.snu.ac.kr, jsno@snu.ac.kr).

D.-J. Shin is with the Department of Electronics and Computer Engineering, Hanyang University, Seoul 133-791, Korea (e-mail: djshin@hanyang.ac.kr).

Color versions of one or more of the figures in this paper are available online at http://ieeexplore.ieee.org.

Digital Object Identifier 10.1109/TVT.2007.914059

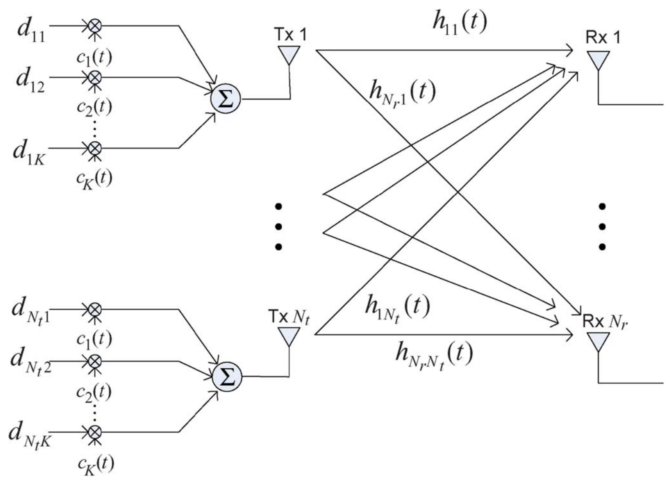

Fig. 1. Multicode MIMO system.

The transmitted signal at the nth transmit antenna for one data symbol duration is given as

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

where d_{ni} is the data symbol for the ith spreading code $c_i(t)$ at the nth transmit antenna, and T_c is the chip duration.

For the frequency-selective fading channel, the tapped-delay-line multipath channel model [9] is used. The channel impulse response from the nth transmit antenna to the mth receive antenna with L multipath components is given as

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

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, $\delta(\cdot)$ is the delta function, and τ_l is the 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., $\Omega_l = \Omega_0 e^{-l\zeta}$, $l = 0, 1, \ldots, L-1$, 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

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

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

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

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 $\mathbb{Z}_4$ is isomorphic to the quadrature phase-shift keying (QPSK) symbols, the quaternary LCZ and ZCZ sequences can be used for complex spreading of the QPSK and M-ary quadrature amplitude modulation (MQAM) data symbols. In the next sections, quaternary LCZ and ZCZ sequences are briefly introduced, and the detection scheme is also explained.

A. Quaternary LCZ and ZCZ Sequences

Quaternary LCZ (ZCZ) sequences [4], [5] are sequences that have low (zero) correlation values when the time shifts between the sequences are within the predetermined correlation zone. In this section, we introduce the quaternary LCZ and ZCZ sequences that are used to construct the multicode MIMO systems.

First, we explain a construction method of quaternary LCZ sequences [4]. Let $b(t)$ be a binary m-sequence of period $N = 2^s - 1$. Let e divide s and $M = 2^e - 1$. Then, the quaternary LCZ sequences $c_i(t),\ i = 1, \ldots, M$ can be constructed as

$$\begin{aligned} c_1(t) &= 2b(t) \\ c_{i+1}(t) &= b(t) + 2b(t + iS), \qquad \text{for } i = 1, 2, \ldots, M-1 \end{aligned} \tag{4}$$

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 the 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 $\lfloor S/u \rfloor$, as follows:

$$\text{For } i = 1, 2, \ldots, M \quad \text{and} \quad j = 0, 1, 2, \ldots, u-1$$
$$c_{i+jM}(t) = c_i\left(t + j\left\lfloor \frac{S}{u} \right\rfloor\right) \tag{5}$$

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 the correlation value is 1. Note that even if these sequences are not cyclically distinct, they can be used for multicode MIMO systems, similarly to the cyclically distinct LCZ sequences, and we will also call these sequences LCZ sequences.

Quaternary ZCZ sequences can be constructed using a perfect quaternary sequence, which has perfect autocorrelation property. Let $b(t)$ be a perfect quaternary sequence. Similarly to the LCZ sequences [10], the ZCZ sequences can be iteratively constructed as follows:

$$\begin{aligned} &\text{Initial}: \ c_0^1(t) = b(t), \quad M_0 = 1, \quad N_0 = \text{period of } b(t) \\ &\text{The } i\text{th iteration}: \\ &\quad \text{For } j = 1, 2, \ldots, M_{i-1}, \quad N_i = 2N_{i-1}, \quad M_i = 2M_{i-1} \\ &\quad c_i^j(2t) = c_{i-1}^j(t) \\ &\quad c_i^j(2t+1) = c_{i-1}^j\left(t + \frac{N_0}{2}\right) \\ &\quad c_i^{j+M_{i-1}}(2t) = c_{i-1}^j(t) \\ &\quad c_i^{j+M_{i-1}}(2t+1) = \left[c_{i-1}^j\left(t + \frac{N_0}{2}\right) + 2\right] \bmod 4 \\ &\quad 0 \le t < N_{i-1} \end{aligned} \tag{6}$$

where N_i and M_i are the period and the family size of the ZCZ sequences obtained after the ith iteration, respectively. However, the ZCZ size in (6) is fixed to $2\lfloor (N_0-1)/2 \rfloor$, regardless of the number of iterations. To increase the family size of ZCZ sequences by reducing the ZCZ size, the method (5) for LCZ sequences can be similarly applied.

B. Detection Scheme

Generally, the intersymbol interference in the multicode MIMO systems cannot be ignored, and the interferences in the spatial and code domains are greater than the intersymbol interference. However, since the delay is assumed to be much smaller than the data symbol duration GT_c, the intersymbol interference can be ignored.

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

$$y_{ml'}^k = \int_{l'T_c}^{l'T_c+GT_c} r_m(t) c_k^*(t-l'T_c)dt \tag{7}$$

$$= \sum_{n=1}^{N_t}\sum_{l=0}^{L-1} h_{lmn} d_{nk} R_{kk}(l,l') \tag{8}$$

$$+ \sum_{n=1}^{N_t}\sum_{l=0}^{L-1}\sum_{\substack{i=1\\ i\neq k}}^{K} h_{lmn} d_{ni} R_{ik}(l,l') \tag{9}$$

$$+ n_{ml'}^k \tag{10}$$

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 (7) can be divided into three different signals as

1) sum of the desired signal, the interferences from delayed multipath signals, and signals from other transmit antennas with the same spreading codes in (8);
2) interferences from the signals using different spreading codes from all transmit antennas in (9);
3) additive white Gaussian noise in (10).

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

1) delayed multipath interference, which is resolved by the RAKE receiver;
2) code-domain interference, which is resolved by the code-domain SIC;
3) spatial-domain interference, which is resolved by the spatial-domain SIC.

Therefore, in the detection of the multicode MIMO system, 2-D SIC is used to resolve the spatial- and code-domain interferences [11].

To detect data symbols $d_{nk},\ n=1,2,\ldots,N_t$, we collect all finger outputs $y_{ml'}^k$ in (7), and thus, we have

$$\mathbf{y}_k = \mathbf{H}_k\mathbf{d}_k + \sum_{\substack{i=1\\ i\neq k}}^{K} \mathbf{J}_k^i \mathbf{d}_i + \boldsymbol{\eta}_k \tag{11}$$

where

$$\mathbf{y}_k = \begin{bmatrix} y_{10}^k & \cdots & y_{1L-1}^k y_{20}^k & \cdots & y_{N_rL-1}^k \end{bmatrix}^T$$

$$\mathbf{d}_k = \begin{bmatrix} d_{1k} & \cdots & d_{N_tk} \end{bmatrix}^T$$

$$\boldsymbol{\eta}_k = \begin{bmatrix} n_{10}^k & \cdots & n_{1L-1}^k n_{20}^k & \cdots & n_{N_rL-1}^k \end{bmatrix}^T$$

$$\mathbf{H}_k = \begin{bmatrix} \mathbf{R}_{kk} & & \mathbf{0} \\ & \ddots & \\ \mathbf{0} & & \mathbf{R}_{kk} \end{bmatrix} \underbrace{\begin{bmatrix} \mathbf{H}(1) \\ \vdots \\ \mathbf{H}(N_r) \end{bmatrix}}_{\mathcal{H}}$$

$$\mathbf{J}_k^i = \begin{bmatrix} \mathbf{R}_{ik} & & \mathbf{0} \\ & \ddots & \\ \mathbf{0} & & \mathbf{R}_{ik} \end{bmatrix} \begin{bmatrix} \mathbf{H}(1) \\ \vdots \\ \mathbf{H}(N_r) \end{bmatrix}$$

$$\mathbf{R}_{ik} = \begin{bmatrix} R_{ik}(0,0) & \cdots & R_{ik}(L-1,0) \\ \vdots & & \vdots \\ R_{ik}(0,L-1) & \cdots & R_{ik}(L-1,L-1) \end{bmatrix}$$

$$\mathbf{H}(m) = \begin{bmatrix} h_{0m1} & \cdots & h_{0mN_t} \\ \vdots & & \vdots \\ h_{L-1m1} & \cdots & h_{L-1mN_t} \end{bmatrix}.$$

It is easy to check that (11) is similar to the conventional MIMO system. Thus, we can apply the vertical Bell Laboratories layered space–time (V-BLAST) detection scheme [12] to (11) to detect $d_{nk},\ n=1,2,\ldots,N_t$, which uses the spatial-domain SIC. Moreover, we can further improve the performance for detecting $\mathbf{d}_k$ by successively canceling the code-domain interference using the previously detected data symbols $\hat{\mathbf{d}}_1, \hat{\mathbf{d}}_2, \ldots, \hat{\mathbf{d}}_{k-1}$, which is called the code-domain SIC. Here, we do not consider the ordering of codes for the code-domain SIC because each spreading code experiences the same fading channel. Thus, we first detect $\hat{\mathbf{d}}_1$ and remove the interferences from $\hat{\mathbf{d}}_1$, then detect $\hat{\mathbf{d}}_2$ and remove the interferences from $\hat{\mathbf{d}}_2$, and so forth. Then, (11) can be modified into

$$\mathbf{y}'_k = \mathbf{y}_k - \sum_{i=1}^{k-1} \mathbf{J}_k^i \hat{\mathbf{d}}_i = \mathbf{H}_k\mathbf{d}_k + \mathbf{e}_k \tag{12}$$

where

$$\mathbf{e}_k = \sum_{i=1}^{k-1} \mathbf{J}_k^i(\mathbf{d}_i - \hat{\mathbf{d}}_i) + \sum_{i=k+1}^{K} \mathbf{J}_k^i \mathbf{d}_i + \boldsymbol{\eta}_k. \tag{13}$$

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 symbol error probability P_e, we can use the following relation:

$$P_e = E_{\boldsymbol{\mathcal{H}}}[P_{e|\boldsymbol{\mathcal{H}}}] \tag{14}$$

where $E_{\boldsymbol{\mathcal{H}}}[\cdot]$ means the expectation over the channel matrix $\boldsymbol{\mathcal{H}}$, and $P_{e|\boldsymbol{\mathcal{H}}}$ is the conditional symbol error probability, which is defined by

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

To find the conditional symbol error probability of the SIC detection scheme, we should derive the signal-to-interference-and-noise ratio (SINR) and average the symbol error probabilities of data symbols. To simplify the problem, we ignore the error propagation, that is, we assume that the previously detected data symbols are correct, which is a good approximation in the high signal-to-noise ratio (SNR) region.

First, we assume that QPSK modulation is used. To calculate the BEP of the proposed multicode MIMO system, the SINR of the input signal to the detector should be derived. Let SINR_{ni} be the SINR corresponding to d_{ni}. It is well known that the BEP of QPSK modulation is given as

$$P_b = Q\left(\sqrt{\text{SINR}}\right) \tag{15}$$

and for the proposed system, the conditional BEP of the QPSK symbol corresponding to d_{ni} is given as

$$P_{b|\boldsymbol{\mathcal{H}}} = Q\left(\sqrt{\text{SINR}_{ni}}\right).$$

Since we assume that there is no error propagation, (13) can be rewritten as

$$\mathbf{e}_k = \sum_{i=k+1}^{K} \mathbf{J}_k^i \mathbf{d}_i + \boldsymbol{\eta}_k. \tag{16}$$

If L is smaller than the LCZ and ZCZ sizes, $R_{ab}(l,l')$ is much smaller than the period G of the 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, (12) can be modified as

$$\mathbf{y}_k(j) = \mathbf{H}_k(j)\mathbf{d}_k(j) + \mathbf{e}_k, \qquad j = 0, 1, \ldots, N_t - 1$$

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 [12]. Let $\mathbf{H}_k(j)^\dagger$ be the Moore–Penrose pseudoinverse of $\mathbf{H}_k(j)$. According to the zero-forcing criterion [12], we perform the following procedure:

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

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

$$\begin{aligned} E\left[\left(\mathbf{H}_k(j)^\dagger \mathbf{e}_k\right)\left(\mathbf{H}_k(j)^\dagger \mathbf{e}_k\right)^H\right] &\\ = \mathbf{H}_k(j)^\dagger E\left[\mathbf{e}_k\mathbf{e}_k^H\right]\left(\mathbf{H}_k(j)^\dagger\right)^H & \end{aligned} \tag{17}$$

where $(\cdot)^H$ denotes the transpose and complex conjugate. From (16), we have

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

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

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

where $\sigma^2 = E[|n(t)|^2]$, and $\mathbf{R}_{kk}$ is the autocorrelation matrix of the spreading code $c_k(t)$ that is defined in (11). If quaternary LCZ or ZCZ sequences are used, $\mathbf{R}_{kk}$ becomes similar to the scaled 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

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

Therefore, we have

$$E\left[\mathbf{e}_k\mathbf{e}_k^H\right] \simeq \sum_{i=k+1}^{K}\mathbf{J}_k^i\left(\mathbf{J}_k^i\right)^H + \sigma^2 G\mathbf{I}. \tag{18}$$

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

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

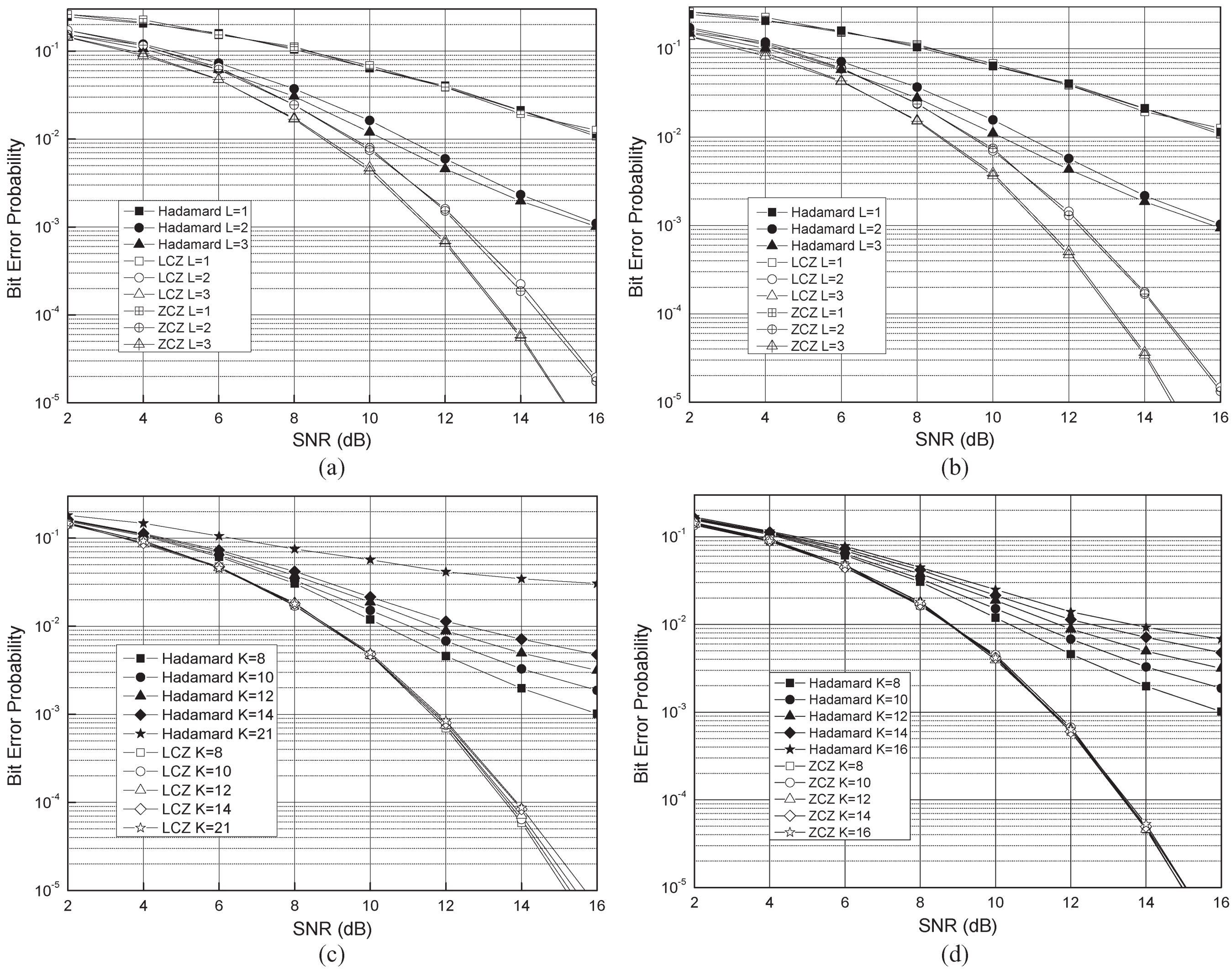

Fig. 2. Performance comparison of 4 × 4 multicode MIMO systems with QPSK using 2-D SIC. (a) Various L values with $K = 8$ under exponential power profile. (b) Various L values with $K = 8$ under uniform power profile. (c) Various K values with $L = 3$ for LCZ sequences under exponential power profile. (d) Various K values with $L = 3$ for ZCZ sequences under exponential power profile.

where $[\cdot]_{p(j)}$ denotes the $p(j)$th diagonal element. Using (17) and (18), the SINR of the data symbol $d_{p(j)k}$ for detection can be derived as

$$\frac{1}{\mathrm{SINR}_{p(j)k}} \simeq \min_{p(j)} \left[\left[\mathbf{H}_k(j)^\dagger \left(\sum_{i=k+1}^{K} \mathbf{J}_k^i \left(\mathbf{J}_k^i\right)^H \right) \left(\mathbf{H}_k(j)^\dagger\right)^H \right]_{p(j)} + \sigma^2 G \left[\mathbf{H}_k(j)^\dagger \left(\mathbf{H}_k(j)^\dagger\right)^H \right]_{p(j)} \right]. \tag{19}$$

From (14), (15), and (19), the average BEP can be derived as

$$P_b \simeq E_{\mathcal{H}} \left[\frac{1}{KN_t} \sum_{j,k} Q\left(\sqrt{\mathrm{SINR}_{p(j)k}} \right) \right]. \tag{20}$$

Using (20) and the Monte Carlo method, the average BEP of the proposed multicode MIMO systems using QPSK modulation can be evaluated. For other modulations such as MQAM and M-ary phase-shift keying, the same analysis can straightforwardly be applied, that is, we only need to replace the bit error function in (15) by the bit error function corresponding to the modulation.

For the simulation, we assume that the processing gain G is 64, there are multipath signals up to $L = 3$, four transmit and four receive antennas are used, and the exponential multipath intensity profile with $\zeta = 0.5$. Quaternary LCZ sequences with period 63, family size $M = 7$, and LCZ size 9 are generated by (4) using a binary m-sequence with period 63. To increase the family size of quaternary LCZ sequences, (5) is used with $u = 3$ to generate 21 LCZ sequences. To make $G = 64$, one zero symbol is padded to each quaternary LCZ sequence. Quaternary ZCZ sequences with ZCZ size 6 are generated by (6) using a perfect quaternary 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. To double the family size of quaternary ZCZ sequences, (5) is used with $u = 2$. The conventional multicode MIMO system is assumed to use the complex spreading by quaternary spreading codes, each of which is constructed

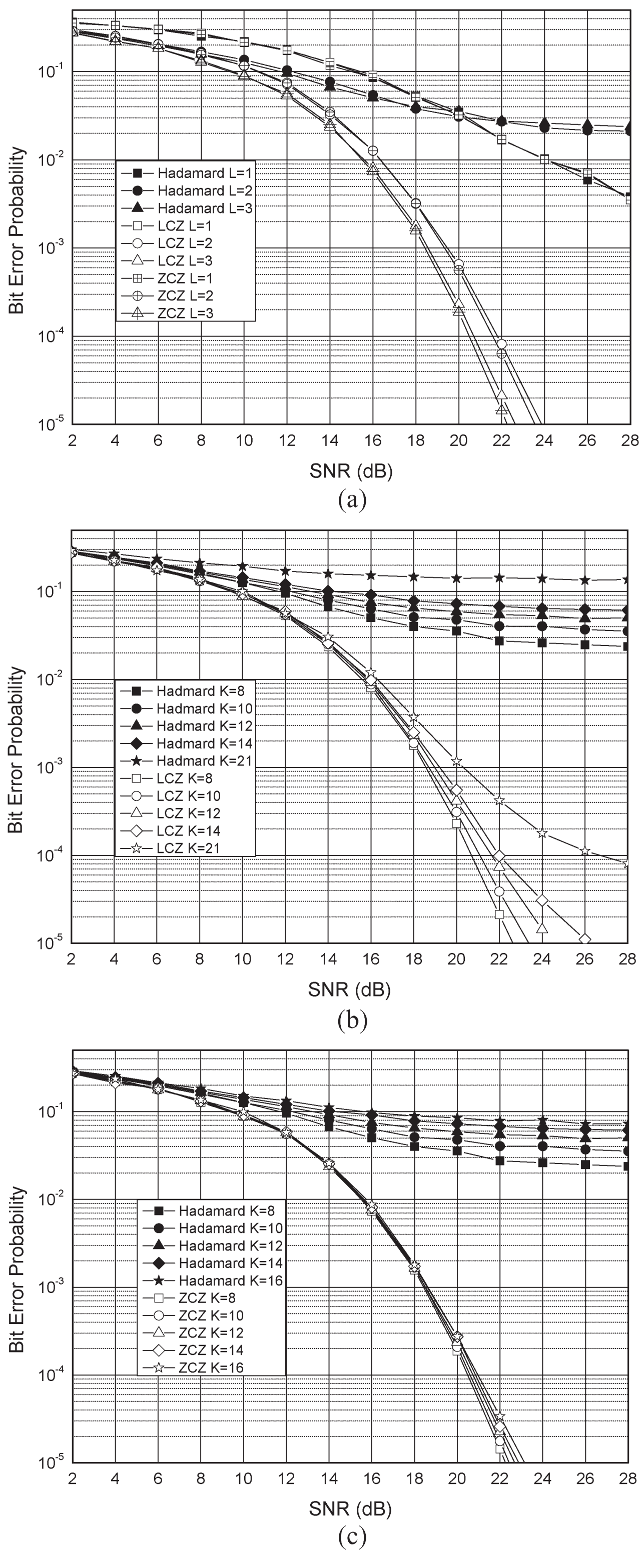

Fig. 3. Performance comparison of 4 × 4 multicode MIMO systems with 16QAM using 2-D SIC. (a) Various L values with $K = 8$. (b) Various K values with $L = 3$ for LCZ sequences. (c) Various K values with $L = 3$ for ZCZ sequences.

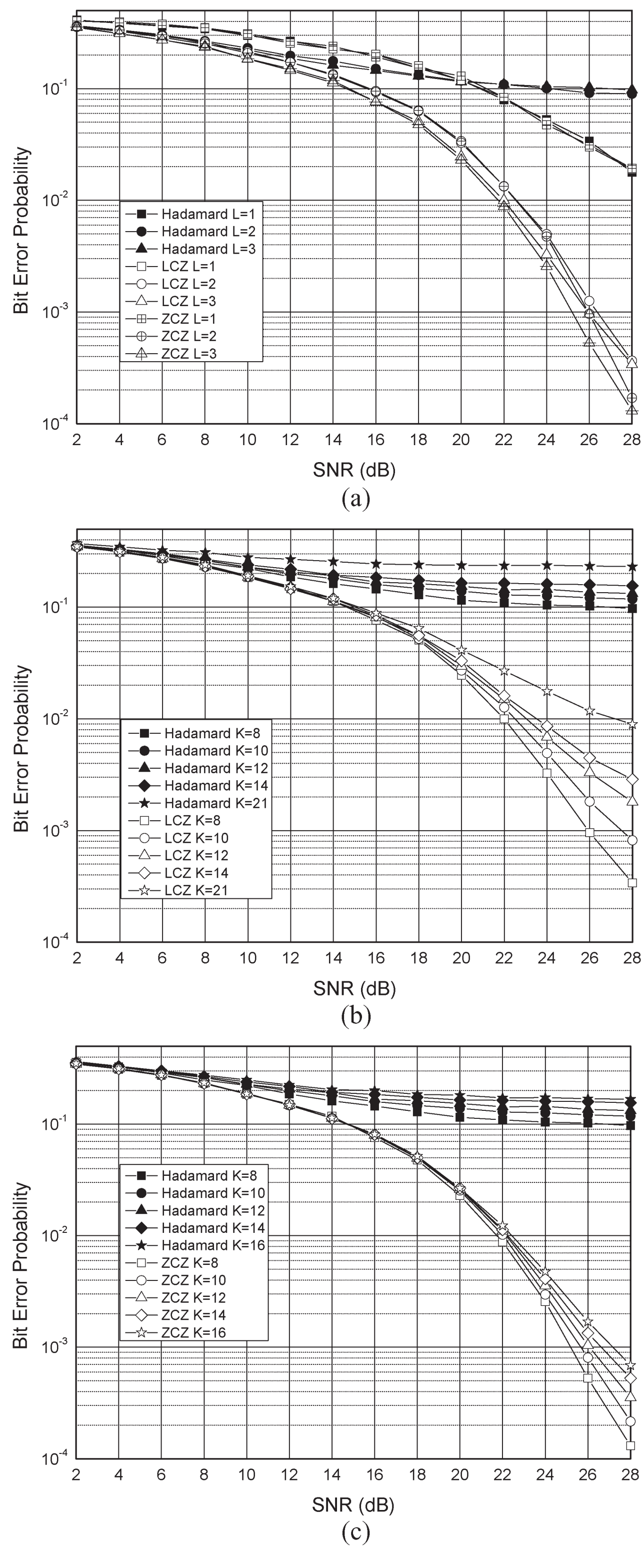

Fig. 4. Performance comparison of 4 × 4 multicode MIMO systems with 64QAM using 2-D SIC. (a) Various L values with $K = 8$. (b) Various K values with $L = 3$ for LCZ sequences. (c) Various K values with $L = 3$ for ZCZ sequences.

from a pair of binary Hadamard codes of length 64. Also, the scrambling code in [2] is used to randomize the mutual interference of Hadamard codes. Each transmit antenna uses K spreading codes. When LCZ sequences are used as spreading codes, we consider $K = 8$, 10, 12, 14, and 21. When ZCZ sequences are used as spreading codes, we consider $K = 8$, 10, 12, 14, and 16.

Fig. 2(a) compares the BEP of the proposed multicode MIMO systems and the conventional multicode MIMO system when the number of transmit antennas is four, $K = 8$,

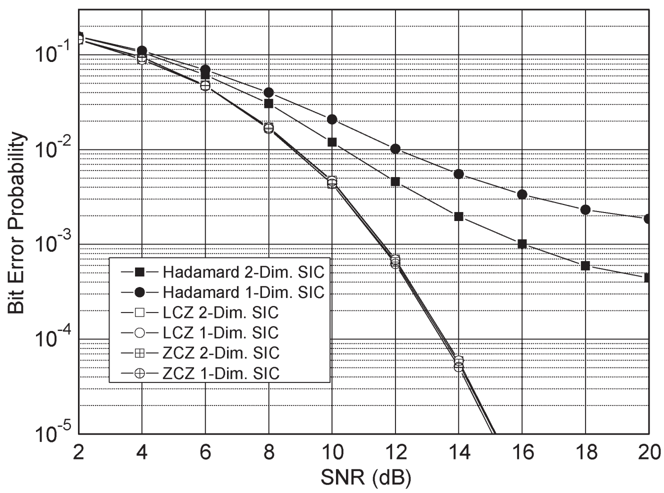

Fig. 5. Performance comparison of 2-D and 1-D SICs for 4 × 4 multicode MIMO systems with QPSK when $K = 8$ and $L = 3$.

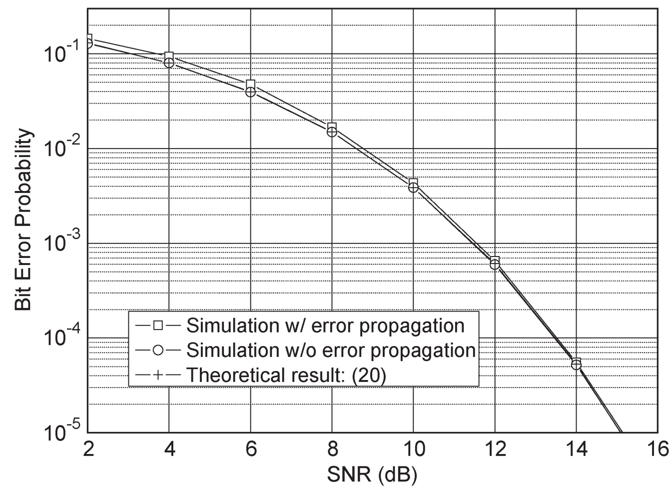

Fig. 6. Comparison of theoretical and simulation results when ZCZ sequences are used for 4 × 4 multicode MIMO systems with $K = 8$ and $L = 3$.

and QPSK modulation is used. When $L = 1$, i.e., there is no delayed multipath signal, these systems show almost identical performance. If $L > 1$, there is an irreducible error floor due to the code-domain interference in the conventional multicode MIMO system. However, the proposed multicode MIMO systems show no error floor, and their performance becomes better than that of the conventional multicode MIMO system as L increases. Fig. 2(b) compares the BEP for various L under uniform power profile, and it shows a similar trend as for the exponential power profile. Thus, we assume the exponential power profile for the remaining simulations. Fig. 2(c) and (d) compares the BEP for the various K values of spreading codes when QPSK modulation is used and $L = 3$. As K increases, the performance of the conventional multicode MIMO system becomes worse, but the proposed multicode MIMO systems show negligible performance degradation due to the good correlation property of quaternary LCZ and ZCZ sequences.

Fig. 3(a) compares the BEP of the proposed multicode MIMO system and the conventional multicode MIMO system when the number of transmit antennas is four, $K = 8$, and 16QAM is used. Similarly to Fig. 2(a), when $L = 1$, these three systems show almost the identical BEP. If $L > 1$, the proposed multicode MIMO systems have much better BEP than the conventional multicode MIMO system. Fig. 3(b) and (c) compares the BEP of the proposed multicode MIMO systems and the conventional multicode MIMO system as K increases. We observed that as K increases, the conventional multicode MIMO system shows worse error floor performance, but the proposed systems have much better error floor performance. In particular, when up to 16 ZCZ sequences are used, there is no error floor above the BEP $= 10^{-5}$. Fig. 4 compares the BEP of the proposed multicode MIMO systems and the conventional multicode MIMO system when 64QAM is used. It shows similar trends as in Fig. 3.

Fig. 5 compares the performance of 2-D and 1-D SICs for the multicode MIMO systems. As expected, there is a negligible performance loss in the proposed multicode MIMO systems, but there is a significant performance loss in the conventional multicode MIMO system. Thus, 1-D SIC detection is sufficient for the proposed multicode MIMO systems, and therefore, the detection complexity can be reduced.

Fig. 6 shows the theoretical and simulation results for the proposed multicode MIMO system with quaternary ZCZ sequences. Since we assume that there is no error propagation when we derive the BEP, there is a small gap between the theoretical performance and simulation results in the low SNR region. When we simulate the proposed multicode MIMO system by assuming a genie detector, the theoretical and simulation results match well over the whole SNR region.

IV. Conclusion

We proposed and analyzed the multicode MIMO systems using quaternary LCZ and ZCZ sequences as spreading codes instead of binary Hadamard codes. In the high SNR region, the analytical result of BEP matches well with the numerical result, whereas in the low SNR region, there is a small gap due to the no error propagation assumption. Since LCZ and ZCZ sequences have very good correlation property within the predetermined correlation zone, the proposed multicode MIMO systems outperform the conventional multicode MIMO system using binary Hadamard codes if the delay is limited within the LCZ or ZCZ. The throughput of the proposed multicode MIMO system can also be improved via increasing the number of spreading codes without performance degradation. Thus, the proposed multicode MIMO systems can be used as a hotspot solution in the WLAN to support high data rate. Since the indoor wireless communication systems experience multipaths with a few chip delay in the isolated area, the proposed systems can also be effectively used.

Acknowledgment

The authors would like to thank the anonymous referees and the Associate Editor for their valuable suggestions, which have improved this paper.

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, no. 3, 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] 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.

[5] H. Torii, M. Nakamura, and N. Suehiro, "A new class of zero-correlation zone sequences," *IEEE Trans. Inf. Theory*, vol. 50, no. 3, pp. 559–565, Mar. 2004.

[6] R. De Gaudenzi, C. Elia, and R. Viola, "Bandlimited quasi-synchronous CDMA: A novel satellite access technique for mobile and personal communication system," *IEEE J. Sel. Areas Commun.*, vol. 10, no. 2, pp. 328–343, Feb. 1992.

[7] B. Long, P. Zhang, and J. Hu, "A generalized QS-CDMA system and the design of new spreading codes," *IEEE Trans. Veh. Technol.*, vol. 47, no. 4, pp. 1268–1275, Nov. 1998.

[8] V. K. N. Lau, "On the analysis of peak-to-average ratio (PAR) for IS95 and CDMA2000 systems," *IEEE Trans. Veh. Technol.*, vol. 49, no. 6, pp. 2174–2188, Nov. 2000.

[9] J. G. Proakis, *Digital Communications*, 3rd ed. New York: McGraw-Hill, 1995.

[10] Y.-S. Kim, J.-W. Jang, J.-S. No, and H. Chung, "New design of low correlation zone sequence sets," *IEEE Trans. Inf. Theory*, vol. 52, no. 10, pp. 4607–4616, Oct. 2006.

[11] C.-S. Park and K.-B. Lee, "Transmit power allocation for successive interference cancellation in multicode MIMO systems," in *Proc. IEEE ICC*, May 2005, pp. 2431–2436.

[12] 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.

Jae-Dong Yang received the B.S. degree in electronic engineering in 2002 from Seoul National University, Seoul, Korea, where he is currently working toward the Ph.D. degree, through the M.S./Ph.D. joint course, within the Department of Electrical Engineering and Computer Science.

His research interests include space–time codes, MIMO, coding theory, information theory, and wireless communication systems.

Xianglan Jin received the B.S. and M.S. degrees in telecommunication engineering from Beijing University of Posts and Telecommunications, Beijing, China, in 1999 and 2002, respectively. She is currently working toward the Ph.D. degree with the Department of Electrical Engineering and Computer Science, Seoul National University, Seoul, Korea.

She was an Engineer with Beijing Samsung Telecommunications R&D Center from April 2002 to June 2004. Her research interests include MIMO, space–time codes, and information theory.

Kyoung-Young Song received the B.S. degree in electrical engineering (major) and mathematics (minor) from Korea University, Seoul, Korea, in 2004. He is currently working toward the Ph.D. degree, through the M.S./Ph.D. joint course, within the Department of Electrical Engineering and Computer Science, Seoul National University.

His research interests include algebraic coding theory, cryptography, and wireless communication systems such as MIMO communication systems, space–time coding, etc.

Jong-Seon No (S'86–M'88) received the B.S. and M.S.E.E. degrees in electronics engineering from Seoul National University, Seoul, Korea, in 1981 and 1984, respectively, and the Ph.D. degree in electrical engineering from the University of Southern California, Los Angeles, in 1988.

He was a Senior MTS with Hughes Network Systems, Germantown, MD, from February 1988 to July 1990. He was an Associate Professor with the Department of Electronic Engineering, Konkuk University, Seoul, from September 1990 to July 1999. He joined the faculty of the Department of Electrical Engineering and Computer Science, Seoul National University, in August 1999, where he is currently a Professor. His research interests include error-correcting codes, sequences, cryptography, space–time codes, LDPC codes, and wireless communication systems.

Dong-Joon Shin (S'96–M'99) received the B.S. degree in electronics engineering from Seoul National University, Seoul, Korea, the M.S. degree in electrical engineering from Northwestern University, Evanston, IL, and the Ph.D. degree in electrical engineering from the University of Southern California, Los Angeles.

From 1999 to 2000, he was a member of Technical Staff with the Wireless Network Division and Satellite Network Division, Hughes Network Systems, Germantown, MD. Since September 2000, he has been an Associate Professor with the Department of Electronics and Computer Engineering, Hanyang University, Seoul. His current research interests include digital communication systems, error-correcting codes, sequences, and discrete mathematics.