A Differential SFBC-OFDM for a DMB System with Multiple Antennas

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ABSTRACT

A differential space-frequency block code - orthogonal frequency division multiplexing (SFBC-OFDM) scheme as a multiple-input multiple-output (MIMO) transmission technique for next-generation digital multimedia broadcasting (DMB) is proposed in this paper. A linear decoding method for differential SFBC, which performs comparably to the ML decoding method, is derived for the cases of two or four transmit antennas. A simple table lookup method is proposed to improve the efficiency of the encoding/decoding process of DSFBC for the case of non-constant modulus constellations. A DMB MIMO channel model, developed by extending the 3GPP MIMO model to fit DMB environments, is used to compare BER performances of differential space block code schemes for various channel environments. Simulation results show that the differential SFBC-16QAM scheme using either four transmit antennas with one receive antenna or two transmit antennas with two receive antennas achieves a performance gain of 12dB than that of the conventional DQPSK scheme, even with a data rate twice faster.

Key Words : Differential SFBC, OFDM, DMB, Linear Decoding, Non-Constant Modulus Constellation

Ⅰ. Introduction

The T-DMB (Terrestrial Digital Multimedia Broadcasting) service, a mobile multimedia broadcasting service based on Eureka-147, has been recently commenced in Korea, giving mobile users access to a relatively high-quality of audio, video and data services in a speedy mobile environment[1]. In this paper, multiple antenna techniques are considered at the transmitter side (receiver side: optional) to obtain diversity gain as well as coding gain for higher-rate data transmission required by the next-generation DMB system. In order to reduce computational overhead for channel estimation and to decrease performance loss in fast-fading channels, a differential space-frequency block code (DSFBC) is employed for the next generation DMB transmission system based on orthogonal frequency division multiplexing (OFDM) transmission. For higher-rate data transmission, an efficient DSFBC decoding method for a non-constant modulus constellation is proposed.

Since [2], the differential space-time coding scheme has been the focus of many papers, mainly because it does not need the channel state information (CSI). Most of these works dealt with the case of constant modulus constellations such as PSK, where the decoder needs to focus only on the phase components of the decoded results for detecting the data symbols, since the amplitude of the transmitted signal matrix can be kept constant in the absence of noise. In [3], a differential space-time coding method is applied with an amplitude/phase shift keying (APSK) constellation consisting of two PSK constellations,

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where the amplitude of the transmitted signal is the same as the previously transmitted one or is switched between two values depending on the value of one bit of the data symbol. In [4]-[6], the differential space-time coding scheme is modified in such a way that it can deal with non-constant modulus constellations. In these approaches, the transmitted signal matrix is commonly divided by the amplitude of the previously transmitted signal matrix so that it will not become too large or too small as it continues to be multiplied by consecutive codeword matrices having non-constant norms. However, the decoding method in [6] still requires computation of the amplitude of the previously transmitted signal matrix for decoding. An efficient method to reduce the computational complexity in the DSFBC encoding/decoding process for the case of non-constant modulus constellations, especially 16-QAM, is proposed in this paper [7]. In addition, a multi-input multi-output (MIMO) channel model for next generation DMB system is used to evaluate the performances of various transmission schemes. [8]-[10].

Ⅱ. DSFBC-OFDM for 16-QAM

There exist several differential coding methods for MIMO transmission systems: differential space-time block code (DSTBC), DSFBC, differential space-time-frequency block code (DSTFBC), and differential space-frequency-time block code (DSFTBC). DSTFBC and DSFTBC commonly map the four elements of each column of the transmitted signal matrix to two sets of two consecutive subcarriers across two successive OFDM symbols; they differ in that DSTFBC applies differential coding across successive OFDM symbols and DSFTBC across consecutive subcarriers. If the channel is frequency-selective and slow-fading, DSTBC has an advantage over DSFBC; if the channel is fast-fading and not so frequency-selective, DSFBC has an advantage over DSTBC. Depending on the characteristic of the channel varying with frequency and time, DSTFBC or DSFTBC can be advantageous over the others. Prioritizing fast mobility of the receivers in the environments of the next generation transmission system, the DSFBC is chosen among the four differential coding schemes.

Here, we consider the DSFBC-OFDM transmission scheme with four transmit antennas and two receive antennas. In the DSFBC scheme [4]-[6], the transmitted signal matrix S_i at the i-th frequency block is generated by

$$S_{i+1} = \widetilde{S}_i X_{i+1} \tag{1}$$

with $S_0 = I_{N_T}$ (an $N_T \times N_T$ identity matrix). X_{i+1} is the space frequency block codeword at the $(i+1)$ -th frequency block. Here, $\widetilde{S}_i$ is defined as

$$\widetilde{S}_i = \frac{1}{a_i} S_i \tag{2}$$

where S_i is divided by its amplitude a_i, i.e.,

$$a_i = \sqrt[N_T]{|S_i|} = \sqrt[N_T]{|X_i|} = \sqrt{\frac{1}{K}\sum_{l=0}^{K-1} |x_{K(i-1)+l}|^2} \tag{3}$$

for normalization. Note that $|S_i|$ is the determinant of S_i, K the number of data symbols carried in the DSFBC encoded signal matrix S_i on each frequency block, and N_T, the number of transmit antennas, equals the dimension(size) of S_i.

Similar to the STBC proposed by Alamouti [11], the space-time block codeword matrix for transmitting two data symbols through two antennas on one frequency block of two subcarriers is given by

$$X_{i+1} = \frac{1}{\sqrt{2}} \begin{bmatrix} x_{2i} & -x^*_{2i+1} \\ x_{2i+1} & x^*_{2i+1} \end{bmatrix} \tag{4}$$

Also, the codeword matrix for transmitting three data symbols through four antennas on one frequency block of four subcarriers is given by

$$X_{i+1} = \frac{1}{\sqrt{3}} \times \begin{bmatrix} x_{3i} & -x^*_{3i+1} & \frac{x^*_{3i+2}}{\sqrt{2}} & \frac{x^*_{3i+2}}{\sqrt{2}} \\ x_{3i+1} & x^*_{3i} & \frac{x^*_{3i+2}}{\sqrt{2}} & -\frac{x^*_{3i+2}}{\sqrt{2}} \\ \frac{x_{3i+2}}{\sqrt{2}} & \frac{x_{3i+2}}{\sqrt{2}} & -x_{3i,R}+jx_{3i+1,I} & x_{3i+1,R}+jx_{3i,I} \\ \frac{x_{3i+2}}{\sqrt{2}} & -\frac{x_{3i+2}}{\sqrt{2}} & -x_{3i+1,R}+jx_{3i,I} & -x_{3i,R}-jx_{3i+1,I} \end{bmatrix} \tag{5}$$

where $x_i = x_{i,R} + jx_{i,I}$. Here, the normalization process of the transmitted signal matrix S_i (by dividing it by a_i) is needed to prevent it from becoming too large or too small.

Let the data symbol x_i's be chosen from the rectangular 16-QAM constellation. Then, depending on whether (4) or (5) is used as the codeword matrix, the amplitude of S_i i.e. a_i, belongs to one of the following two sets:

$$\left\{\sqrt{\frac{4}{20}},\sqrt{\frac{12}{20}},\sqrt{\frac{20}{20}},\sqrt{\frac{28}{20}},\sqrt{\frac{36}{20}}\right\} \quad \left\{\sqrt{\frac{6}{30}},\sqrt{\frac{14}{30}},\sqrt{\frac{22}{30}},\sqrt{\frac{30}{30}},\sqrt{\frac{38}{30}},\sqrt{\frac{46}{30}},\sqrt{\frac{54}{30}}\right\} \tag{6}$$

Here, each row of S_i is transmitted through one of N_T different transmit antennas while each column of S_i is transmitted on one of N_T consecutive subcarriers. Thus, the signal received by the m-th receive antenna through a channel with the frequency response $\mathbf{h}_m=\left[h_1^m,\cdots,h_{N_T}^m\right]$ can be described as

$$\begin{aligned}\mathrm{R}_{i+1}^m &= \mathbf{h}_m S_{i+1}+\mathrm{W}_{i+1}^m \\ &= \frac{1}{a_i}\mathbf{h}_m S_i X_{i+1}+\mathrm{W}_{i+1}^m\end{aligned} \tag{7}$$

Here, h_l^m is the channel gain from the l-th transmit antenna to the m-th receive antenna, the W_{i+1}^m noise, and

$$\begin{aligned}\mathrm{R}_i^m &= \left[\mathrm{R}_{i,1}^m,\cdots,\mathrm{R}_{i,N_T}^m\right] \\ &= \left[R(m,N_T i+1),\cdots,R(m,N_T(i+1))\right]\end{aligned} \tag{8}$$

where $R(m,k)$ is the FFT output of $r(m,n)$. Here, $r(m,n)$ is the signal received by the m-th receive antenna at the n-th sampling time, and $R(m,k)$ is the frequency component of the k-th subcarrier at the m-th antenna.

Under the assumption that the frequency responses of adjacent subcarriers are identical, the received signal can be written as

$$\begin{aligned}\mathrm{R}_{i+1}^m &= \frac{1}{a_i}\left(\mathrm{R}_i^m - W_{i+1}^m\right)X_{i+1}+W_{i+1}^m \\ &= \frac{1}{a_i}\mathrm{R}_i^m X_{i+1} - \frac{1}{a_i}W_i^m X_{i+1} + W_{i+1}^m\end{aligned} \tag{9}$$

As suggested in [4], the near-optimal differential decoder with the dependence of the noise on the transmitted signal neglected is given by

$$\begin{aligned}\hat{\mathbf{x}}_{i+1} &= \left[\hat{x}_{Ki}\cdots\hat{x}_{K(i+1)-1}\right] \\ &= \underset{\mathbf{x}_{i+1}}{argmin}\sum_{m=1}^{N_R}\left\|\mathrm{R}_{i+1}^m-\frac{1}{a_i}\mathrm{R}_i^m X_{i+1}\right\|^2 \\ &= \underset{\mathbf{x}_{i+1}}{argmin}\sum_{m=1}^{N_R}\left\{-2Re\left[\mathrm{R}_i^m\frac{1}{a_i}X_{i+1}\mathrm{R}_{i+1}^{m*}\right]\right. \\ &\qquad \left.+\mathrm{R}_i^m\frac{1}{a_i}X_{i+1}\frac{1}{a_i}X_{i+1}^*\mathrm{R}_i^{m*}\right\}\end{aligned} \tag{10}$$

where $*$ denotes the complex conjugate transpose and N_R the number of receive antennas. Both of the two codeword matrices (4) and (5) can be expressed as

$$X_{i+1}=\frac{1}{\sqrt{K}}\sum_{n=0}^{K-1}\left(A_n x_{Ki+n,R}+jB_n x_{Ki+n,I}\right) \tag{11}$$

where the matrices $\{A_n,B_n|n=0:K-1\}$ constitute an amicable orthogonal design [12] in the sense that they satisfy the following conditions:

$$\left.\begin{aligned}A_m^*A_m&=I_{N_T}\\ B_m^*B_m&=I_{N_T}\\ A_m^*B_n&=B_n^*A_m\end{aligned}\right\}\forall\, m,n \quad \left.\begin{aligned}A_m^*A_n&=-A_n^*A_m\\ B_m^*B_n&=-B_n^*B_m\end{aligned}\right\}\forall\, m\neq n \tag{12}$$

where I_{N_T} is an $N_T\times N_T$ identity matrix.

Thus, the cost function of the above decoder to be minimized can be rewritten as

$$\begin{aligned}&J(x_{Ki+n,R},x_{Ki+n,I}) \\ &=-\frac{1}{a_i\sqrt{K}}\left[\sum_{m=1}^{N_R}\sum_{n=0}^{K-1}\left\{\left(\mathrm{R}_i^m A_n \mathrm{R}_{i+1}^{m*}+\mathrm{R}_i^{m*}A_n\mathrm{R}_{i+1}^m\right)\right.\right. \\ &\quad \left.\left.\cdot\, x_{Ki+n,R}+j\left(\mathrm{R}_i^m B_n\mathrm{R}_{i+1}^{m*}-\mathrm{R}_i^{m*}B_n^*\mathrm{R}_{i+1}^m\right)x_{Ki+n,I}\right\}\right] \\ &\quad +\frac{1}{a_i^2K}\left\{\sum_{n=0}^{K-1}\left(x_{Ki+n,R}^2+x_{Ki+n,I}^2\right)\right\}\left\{\sum_{m=1}^{N_R}\sum_{n=1}^{N_T}|\mathrm{R}_{i,n}^m|^2\right\}\end{aligned} \tag{13}$$

Setting the derivative of this cost function with respect to $x_{Ki+n,R}$ and $x_{Ki+n,I}$ to zero, we obtain

$$\begin{aligned}\hat{x}_{Ki+n} &= \hat{x}_{Ki+n,R}+j\hat{x}_{Ki+n,I} \\ &= \frac{\sqrt{K}a_i}{2\sum_{m=1}^{N_R}\sum_{n=1}^{N_T}|\mathrm{R}_{i,n}^m|^2}\sum_{m=1}^{N_R}\sum_{n=0}^{K-1}\left\{\left(\mathrm{R}_i^m A_n\mathrm{R}_{i+1}^{m*}\right.\right. \\ &\quad \left.\left.+\mathrm{R}_i^{m*}A_n\mathrm{R}_{i+1}^m\right)-\left(\mathrm{R}_i^m B_n\mathrm{R}_{i+1}^{m*}-\mathrm{R}_i^{m*}B_n\mathrm{R}_{i+1}^m\right)\right\}\end{aligned} \tag{14}$$

This equation leads to the following decoding formulas:

$$\begin{aligned}\hat{\mathbf{x}}_{i+1} &= \begin{bmatrix}\hat{x}_{2i}\\ \hat{x}_{2i+1}\end{bmatrix} \\ &= \frac{\sqrt{2}\,\hat{a}_i}{\sum_{m=1}^{N_R}\left(|\mathrm{R}_{i,1}^m|^2+|\mathrm{R}_{i,2}^m|^2\right)}\sum_{m=1}^{N_R}\begin{bmatrix}\mathrm{R}_{i,1}^{m*}\mathrm{R}_{i+1,1}^m+\mathrm{R}_{i,2}^m\mathrm{R}_{i+1,2}^{m*}\\ \mathrm{R}_{i,2}^{m*}\mathrm{R}_{i+1,1}^m-\mathrm{R}_{i,1}^m\mathrm{R}_{i+1,2}^{m*}\end{bmatrix}\end{aligned} \tag{15}$$

and

$$\hat{\mathbf{x}}_{i+1} = \begin{bmatrix} \hat{x}_{3i} \\ \hat{x}_{3i+1} \\ \hat{x}_{3i+2} \end{bmatrix} = \frac{\sqrt{3}\,\hat{a}_i}{\sum_{m=1}^{N_R}\sum_{n=1}^{N_T}|\mathrm{R}_{i,n}^m|^2} \cdot \sum_{m=1}^{N_R} \begin{bmatrix} \psi_0(m) \\ \psi_1(m) \\ \psi_2(m) \end{bmatrix} \quad (16)$$

for the case of two and four transmit antennas, respectively. Here, $\mathrm{R}_{i,n}^m$ denotes the n-th component of the i-th frequency block received by the m-th receive antenna. Unlike the method proposed in [8], this decoding process does not require any channel power estimation.

In the proposed method, possible values of a_i, as in (6), are stored so that the DSFBC encoder can choose the value corresponding to the set of K data symbols $\{x_{K(i-1)} \cdots x_{Ki-1}\}$, which was encoded at the previous frequency block. Also, the DSFBC decoder can choose the value corresponding to the set of estimates of the data symbol $\{\hat{x}_{K(i-1)} \cdots \hat{x}_{Ki-1}\}$, which is obtained from the decoded results at the previous frequency block. While the direct use of (3) for computing a_i or $\hat{a}_i$ requires $2K+1$ real multiplications, $K-1$ real addition(s), and one square root per frequency block, this table lookup method needs no more computation than $2K$ comparisons for choosing an appropriate value of a_i or $\hat{a}_i$.

Ⅲ. A Channel Model for DMB MIMO

Since the existing COST207 channel model[10] is a single-input single-output (SISO) model, it cannot be used for a performance analysis of MIMO transmission schemes. In order to evaluate and compare the performances of the various DMB MIMO transmission schemes discussed in this paper, we develop a channel model for DMB MIMO by modifying the parameters of the 3GPP MIMO channel model [8]-[9] such that it can match the DMB MIMO environments. In the DMB MIMO channel model, the PDP's between the transmit antennas and receive antennas are set to those of the COST207 channel model. The distance among receive antennas is set to 0.5λ, and the distance among transmit antennas is 4λ. A reduced typical urban (RTU) model, CH1, has a uniform power azimuth spectrum (PAS) and an angle of arrival (AoA) of 22.5° for the receiver, and a Laplacian PAS with azimuth spread (AS)= 5° and AoA of 20° for the transmitter. Another RTU model, CH2, has a Laplacian PAS with AS=35° and AoA of 56° for the receiver and the same parameters for the transmitter as CH1. A reduced bad urban (RBU) model, CH3, is the same as CH2, except that the receiver has a Laplacian PAS with AS=22° and the transmitter has an AoA of 50°. Velocity of 3~120km/h is considered.

Table 1. DMB MIMO Channel Model

		CH1	CH2	CH3
No. of paths		6	6	6
PDP		RTU	RTU	RBU
Doppler spectrum		Classical	Classical	Classical
Speed (km/h)		3/40/120	3/40/120	3/40/120
Rx	Topology	0.5λ spacing	0.5λ spacing	0.5λ spacing
	PAS	Uniform over 360	Laplacian, AS=35	Laplacian, AS=22
	DoM (deg)	0	22.5	22.5
	AoA (deg)	22.5	56.0	56.0
Tx	Topology	4λ spacing	4λ spacing	4λ spacing
	PAS	Laplacian, AS=5	Laplacian, AS=5	Laplacian, AS=5
	AoA (deg)	20	20	20

The correlations among the antennas at the transmitter or receiver are determined depending on the DMB MIMO channel model parameters, such as AS, AoA, and PAS type, listed in Table 1. Among the three models, CH1 has the smallest correlation and CH3 has the largest correlation, all in the space domain. Note that the correlation at the receiver side is negligible for CH1 due to the uniform PAS over 360°, signifying a full receive diversity gain. On the other hand, for

CH3, the correlation is about 0.7 due to a smaller AS, implying that significant receive diversity gain cannot be expected.

Ⅳ. Simulation

In this section, performances of the proposed method are evaluated by computer simulation with the DMB MIMO channel discussed in Section III. First, the proposed decoding method is compared with the ML decoding method for the DSFBC-OFDM system with an input of 16QAM. From the simulation results with CH1 shown in Fig. 1, it can be seen that the BER performance of the proposed decoding method is comparable to that of the ML decoding method.

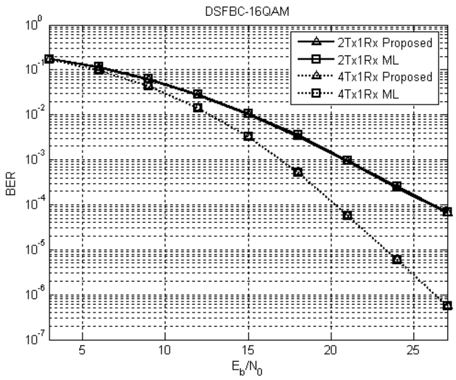

Fig. 1. Performance comparison between the proposed decoding method and ML decoding method (CH1)

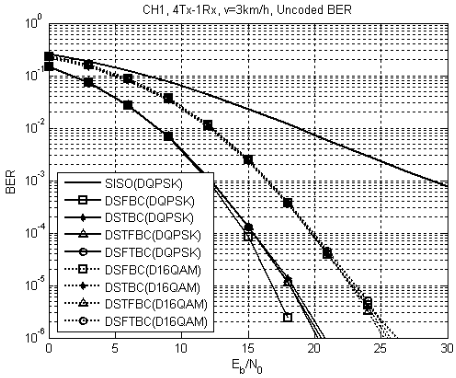

Fig. 2. BER performances of various differential space block codes at 3km/h (CH1, 4Tx-1Rx)

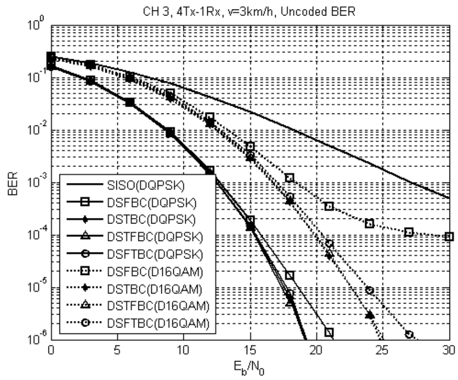

Fig. 3. BER performances of various differential space block codes at 3km/h (CH3, 4Tx-1Rx)

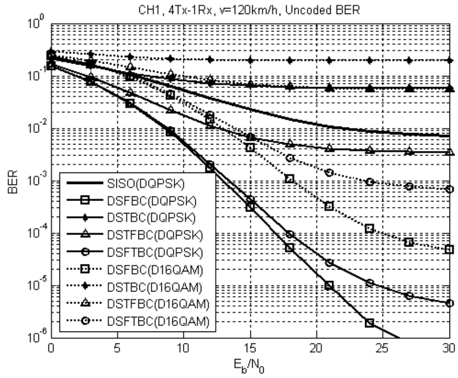

Fig. 4. BER performances of various differential space block codes at 120km/h (CH1, 4Tx-1Rx)

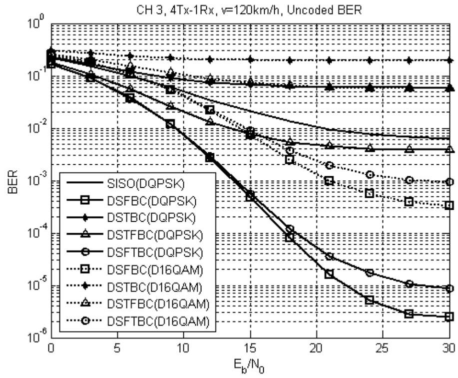

Fig. 5. BER performances of various differential space block codes at 120km/h (CH3, 4Tx-1Rx)

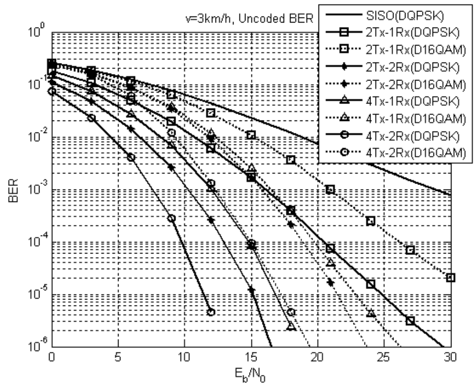

Fig. 6. BER performances of DSFBC-OFDM when the number of antennas and modulation order vary (CH1)

Next, the performances of differential space block coding methods are compared by using the DMB MIMO channels with four transmit antennas and one receive antenna. As shown in Fig. 2, there is not a large difference in BER performance among the differential MIMO schemes for CH1 with a large coherence bandwidth. For CH3 with a smaller coherence bandwidth (Fig. 3), the performances of the differential encoding schemes along the frequency domain, such as DSFBC, worsen as the modulation order increases. Note, however, that the BER performance of the DSFBC scheme is always better than that of the SISO case. Figures 4~5 show that the performances of the differential encoding schemes along the time domain, such as DSTBC, worsen as the speed increases. Error floor phenomena can be observed in the case of DSTBC since the channel varies during two successive OFDM symbol periods. Note that DSFBC works well at high speed (120km/h). Fig. 6 shows the BER performances of the DSFBC-OFDM when input signals of different modulation order are applied to the DMB MIMO channel with a different number of transmit/receive antennas at a speed of 3km/h. From Fig. 6, it can be seen that DSFBC-QPSK with 2Tx-1Rx, DSFBC-16QAM with 2Tx-2Rx, and DSFBC-16QAM with 4Tx-1Rx achieve an Eb/N0 gain of 12dB over the conventional SISO-DQPSK at a BER of 10^{-3}. Note that this gain is achieved even with a data rate twice faster than the conventional case for the last two DSFBC-16QAM cases. Also, DSFBC-16QAM with 4Tx-2Rx has an Eb/N0 gain of 17dB over the conventional SISO-DQPSK, even with twice faster data rate.

V. Conclusion

The linear DSFBC decoding method for non-constant modulus constellations was derived in this paper for the cases of two or four transmit antennas. In addition, a table lookup method was proposed so as to reduce the computational complexity in the encoding/decoding process in DSTBC or DSFBC for the case of a 16-QAM constellation. Among the several differential coding schemes applicable to MIMO transmission, we have chosen the DSFBC scheme for next generation DMB transmission. This choice is strongly supported by various simulation results with the DMB MIMO channel model developed in this paper. It was also shown by simulation that the DSFBC-16QAM with either four transmit antennas and one receive antenna or two transmit antennas and two receive antennas achieves a performance gain of 12dB than the case of the conventional SISO-DQPSK, even with a data rate twice faster.

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