# Frequency-domain Space-Time Block Coded-Joint Transmit/Receive Diversity for The Single Carrier Transmission

*Hiromichi TOMEBA*† *Kazuaki TAKEDA*† *and Fumiyuki ADACHI*‡

Dept. of Electrical and Communication Engineering, Graduate School of Engineering, Tohoku University

6-6-05 Aza-Aoba, Aramaki, Aoba-ku, Sendai, 980-8579 Japan

E-mail: †{tomeba, takeda}@mobile.ecei.tohoku.ac.jp, ‡adachi@ecei.tohoku.ac.jp

**Abstract**

Recently, frequency-domain pre-equalization (pre-FDE) transmission has been attracting attention for improving the broadband DS-CDMA (or single carrier (SC)) transmission performance in a frequency-selective fading channel. To further improve the transmission performance, antenna diversity technique is very effective. However, so far the joint use of pre-FDE and the receive diversity has not been fully discussed. Recently, we proposed the space-time block coded-joint transmit/receive diversity (STBC-JTRD) for the narrow-band transmission in a frequency-nonselective fading channel. STBC-JTRD allows an arbitrary number of transmit antennas while limiting the number of receive antennas to 4. In this paper, we extend the STBC-JTRD to the case of frequency-selective fading channel and propose the frequency-domain STBC-JTRD for the SC transmission in a frequency-selective fading channel. The average bit error rate (BER) performance of the SC transmission with frequency-domain STBC-JTRD is evaluated by the computer simulation.

## 1. INTRODUCTION

High speed data services of over 100Mbps are demanded in the next generation mobile communications systems. However, mobile channel is composed of many propagation paths with different time delays, producing severe frequency-selective fading channel, and therefore, the bit error rate (BER) performance of the single-carrier (SC) transmission significantly degrades due to severe inter-symbol inference (ISI) [1-3]. Frequency-domain equalization (FDE) is an effective technique for improving the SC transmission performance in a frequency-selective fading channel [4]. FDE can be applied to DS-CDMA to obtain a good BER performance similar to that of multi-carrier code division multiple access (MC-CDMA) [5, 6]. Recently, frequency-domain pre-equalization (pre-FDE) transmission technique has been proposed for improving the BER performance [7-9]. Recently, we proposed the pre-FDE for DS-CDMA and showed that pre-FDE can achieve a good BER performance similar to the FDE-reception in a frequency-selective fading channel [10].

To further improve the SC transmission performance, the antenna diversity technique is very effective [1-3]. However, so far the joint use of pre-FDE and the receive diversity has not been discussed. Recently, we proposed the space-time block coded-joint transmit/receive diversity (STBC-JTRD), which requires the CSI only at the transmitter side, for narrowband SC transmission in a frequency-nonselective fading channel [11]. An arbitrary number of transmit antennas can be used while the number of receive antennas is limited to 4. In this paper, we extend the STBC-JTRD to the case of frequency-selective fading channel and propose the frequency-domain STBC-JTRD.

The remainder of this paper is organized as follows. Sect. 2 describes the frequency-domain STBC-JTRD for the SC transmission. The pre-equalization weight based on minimum mean square error (MMSE) criterion is derived in Sect. 3. The average BER performance of the SC transmission with frequency-domain STBC-JTRD is evaluated by the computer simulation in a frequency-selective Rayleigh fading channel in Sect. 4. Sect. 5 offers some conclusions.

## 2. FREQUENCY-DOMAIN STBC-JTRD FOR THE SC TRANSMISSION

Figure 1 illustrates the transmitter/receiver structure for the SC transmission using frequency-domain STBC-JTRD with $N_t$ transmit antennas and $N_r$ receive antennas. Throughout the paper, symbol-spaced discrete time representation of the transmitted signal is used. In what follows, without loss of generality, we assume the transmission of one codeword.

### 2.1. Encoding

At the transmitter, a sequence of data modulated symbols $\{d(i);\ i=0\sim(GN_c-1)\}$ is divided into a sequence of $G$ information symbol blocks with $N_c$ symbols each. The $g$-th symbol block $\{s_g(t);\ t=0\sim(N_c-1)\}$, $g=0\sim(G-1)$ is expressed as

$$s_g(t)=d(t+gN_c)\,. \tag{1}$$

$N_c$-point fast Fourier transform (FFT) is applied to decompose the $g$-th symbol block into $N_c$ frequency components $\{S_g(k);k=0\sim(N_c-1)\}$ as

$$S_g(k)=\sum_{t=0}^{N_c-1}s_g(t)\exp\left(-j2\pi t\frac{k}{N_c}\right)\quad. \tag{2}$$

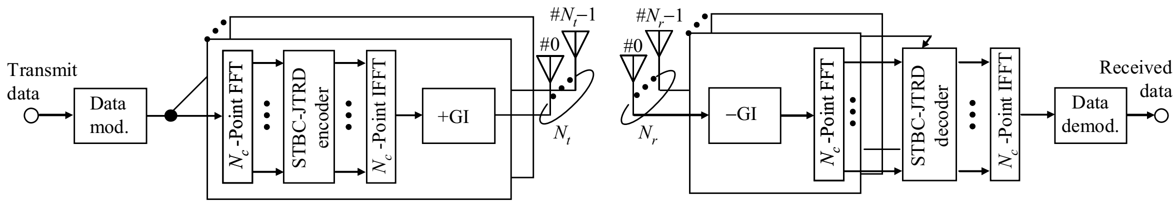


Fig. 1 Transmitter/receiver structure for the SC transmission with frequency-domain STBC-JTRD.

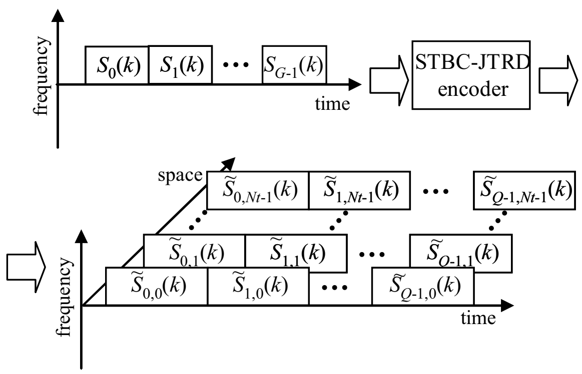


Fig. 2 STBC-JTRD encoding for the $k$-th frequency components.

Table 1 $G$, $Q$ and $R$ for $N_r$=2~4

| No. of receive antennas, $N_r$ | No. of information symbol blocks in a codeword, $G$ | No. of coded symbol blocks in a codeword, $Q$ | Coding rate $R$ |
|---|---|---|---|
| 2 | 2 | 2 | 1 |
| 3 | 3 | 4 | 3/4 |
| 4 | 3 | 4 | 3/4 |

The resultant $G$ consecutive $k$-th frequency components $\{S_0(k),..., S_g(k),.., S_{G-1}(k)\}$ are encoded into $N_t$ parallel codewords (see Fig. 2); the $n$-th codeword consists of $Q$ consective frequency components $\{\widetilde{S}_{0,n}(k),...,\widetilde{S}_{q,n}(k),...,\widetilde{S}_{Q-1,n}(k)\}$ and is transmitted from the $n$-th transmit antenna after performing $N_c$-point inverse FFT (IFFT). Table 1 shows the number $G$ of information symbol in a codeword, the length $Q$ of a codeword, and the coding rate $R$ for $N_r$ (the number of receive antennas) =2, 3 and 4. Frequency-domain encoding for $N_r$=2~4 is expressed as follows:

*2.1.1 $N_r$=2 (G=Q=2)*

$$\begin{pmatrix}\widetilde{S}_{0,n}(k)\\ \widetilde{S}_{1,n}(k)\end{pmatrix}=C_2\begin{pmatrix}S_0(k)w^*_{0,n}(k)+S_1(k)w^*_{1,n}(k)\\ S^*_0(k)w^*_{1,n}(k)-S^*_1(k)w^*_{0,n}(k)\end{pmatrix}. \qquad (3)$$

*2.1.2 $N_r$=3 (G=3, Q=4)*

$$\begin{pmatrix}\widetilde{S}_{0,n}(k)\\ \widetilde{S}_{1,n}(k)\\ \widetilde{S}_{2,n}(k)\\ \widetilde{S}_{3,n}(k)\end{pmatrix}=C_3\begin{pmatrix}S_0(k)w^*_{0,n}(k)+S_1(k)w^*_{1,n}(k)+S_2(k)w^*_{2,n}(k)\\ S^*_0(k)w^*_{1,n}(k)-S^*_1(k)w^*_{0,n}(k)\\ S^*_0(k)w^*_{2,n}(k)-S^*_2(k)w^*_{0,n}(k)\\ S^*_1(k)w^*_{2,n}(k)-S^*_2(k)w^*_{1,n}(k)\end{pmatrix}. \qquad (4)$$

*2.1.3 $N_r$=4 (G=3, Q=4)*

$$\begin{pmatrix}\widetilde{S}_{0,n}(k)\\ \widetilde{S}_{1,n}(k)\\ \widetilde{S}_{2,n}(k)\\ \widetilde{S}_{3,n}(k)\end{pmatrix}=C_4\begin{pmatrix}S_0(k)w^*_{0,n}(k)+S_1(k)w^*_{1,n}(k)+S_2(k)w^*_{2,n}(k)\\ S^*_0(k)w^*_{1,n}(k)-S^*_1(k)w^*_{0,n}(k)+S_2(k)w^*_{3,n}(k)\\ S^*_0(k)w^*_{2,n}(k)-S_1(k)w^*_{3,n}(k)-S^*_2(k)w^*_{0,n}(k)\\ S_0(k)w^*_{3,n}(k)+S^*_1(k)w^*_{2,n}(k)-S^*_2(k)w^*_{1,n}(k)\end{pmatrix}. \qquad (5)$$

In Eqs. (3)~(5), $\{w_{m,n}(k); k=0\sim(N_c-1)\}$, $m=0\sim(N_r-1)$ and $n=0\sim(N_t-1)$, are the MMSE pre-equalization weights, which will be derived in Sect. 3. $C_{Nr}$ is the power normalization factor, given by

$$C_{N_r}=\sqrt{\frac{N_c}{\sum_{m=0}^{N_r-1}\sum_{n=0}^{N_t-1}\sum_{k=0}^{N_c-1}|w_{m,n}(k)|^2}}, \qquad (6)$$

which acts as the transmit power control factor.

The $n$-th time-domain codeword $\{\widetilde{s}_{q,n}(t)$; $t=0\sim(N_c-1)\}$, $n=0\sim(N_t-1)$ and $q=0\sim(Q-1)$, is obtained by applying $N_c$-point IFFT to $\{\widetilde{S}_{q,n}(k)$; $k=0\sim(N_c-1)\}$ as

$$\widetilde{s}_{q,n}(t)=\sqrt{\frac{2E_s}{T_s}}\left\{\frac{1}{N_c}\sum_{k=0}^{N_c-1}\widetilde{S}_{q,n}(k)\exp\left(j2\pi k\frac{t}{N_c}\right)\right\}, \qquad (7)$$

where $E_s$ and $T_s$ denote the transmit symbol energy and the symbol period, respectively. After inserting a cyclic prefix of $N_g$ symbols into the guard interval (GI), $N_t$ codewords of $(N_c+N_g)$ symbols each are transmitted from the $N_t$ transmit antennas.

### 2.2. Decoding

A superposition of $N_t$ codewords is received via a frequency-selective fading channel by $N_r$ receive antennas. We assume a symbol-spaced $L$-path frequency-selective block fading channel. The channel impulse response $h_{m,n}(t)$ between the $n$-th transmit antenna and the $m$-th receive antenna is expressed as

$$h_{m,n}(t)=\sum_{l=0}^{L-1}h_{m,n,l}\delta(t-\tau_l), \qquad (8)$$

where $h_{m,n,l}$ and $\tau_l$ are the complex-valued path gain and time delay of the $l$-th propagation path. The $q$-th symbol block $r_{q,m}(t)$, in a codeword, received by the $m$-th receive antenna, can be expressed as

$$r_{q,m}(t)=\sum_{n=0}^{N_t-1}\sum_{l=0}^{L-1}h_{m,n,l}\widetilde{s}_{q,n}(t-\tau_l)+\eta_{q,m}(t), \qquad (9)$$

where $\eta_{q,m}(t)$ is the additive white Gaussian noise (AWGN) process with a zero mean and a variance $2N_0/T_s$ with $N_0$ being the single-sided power spectrum density.

After the removal of the GI, $\{r_{q,m}(t);t=0\sim(N_c-1)\}$ is decomposed by $N_c$-point FFT into $N_c$ frequency components:

$$R_{q,m}(k)=\sum_{t=0}^{N_c-1} r_{q,m}(t)\exp\left(-j2\pi t\frac{k}{N_c}\right). \qquad (10)$$

Substituting Eq. (9) into Eq. (10), the $k$-th frequency component $R_{q,m}(k)$ of the $q$-th symbol block is given by

$$R_{q,m}(k)=\sqrt{\frac{2E_s}{T_s}}\sum_{n=0}^{N_t-1} H_{m,n}(k)\widetilde{S}_{q,n}(k)+\Pi_{q,m}(k) \quad ,(11)$$

where

$$\begin{cases} H_{m,n}(k)=\sum_{l=0}^{L-1} h_{m,n,l}\exp\left(-j2\pi k\frac{\tau_l}{N_c}\right) \\ \Pi_{q,m}(k)=\sum_{t=0}^{N_c-1}\eta_{q,m}(t)\left(-j2\pi k\frac{t}{N_c}\right) \end{cases}. \qquad (12)$$

Frequency-domain STBC-JTRD decoding is carried out on $\{R_{q,m}(k);q=0\sim(Q-1)\}$ as follows:

*2.2.1 $N_r$=2 (G=Q=2)*

$$\begin{pmatrix}\hat{S}_0(k)\\ \hat{S}_1(k)\end{pmatrix}=\begin{pmatrix}R_{0,0}(k)+R^*_{1,1}(k)\\ R_{0,1}(k)-R^*_{1,0}(k)\end{pmatrix}. \qquad (13)$$

*2.2.2 $N_r$=3 (G=3, Q=4)*

$$\begin{pmatrix}\hat{S}_0(k)\\ \hat{S}_1(k)\\ \hat{S}_2(k)\end{pmatrix}=\begin{pmatrix}R_{0,0}(k)+R^*_{1,1}(k)+R^*_{2,2}(k)\\ R_{0,1}(k)-R^*_{1,0}(k)+R^*_{3,2}(k)\\ R_{0,2}(k)-R^*_{2,0}(k)-R^*_{3,1}(k)\end{pmatrix}. \qquad (14)$$

*2.2.3 $N_r$=4 (G=3, Q=4)*

$$\begin{pmatrix}\hat{S}_0(k)\\ \hat{S}_1(k)\\ \hat{S}_2(k)\end{pmatrix}=\begin{pmatrix}R_{0,0}(k)+R^*_{1,1}(k)+R^*_{2,2}(k)+R_{3,3}(k)\\ R_{0,1}(k)-R^*_{1,0}(k)-R_{2,3}(k)+R^*_{3,2}(k)\\ R_{0,2}(k)+R_{1,3}(k)-R^*_{2,0}(k)-R^*_{3,1}(k)\end{pmatrix}. \qquad (15)$$

The STBC-JTRD decoding requires the addition/subtraction and conjugate operations, but requires no channel state information (CSI). Then, $N_c$-point IFFT is applied to transform $\{\hat{S}_g(k);k=0\sim(N_c-1)\}$ into the time-domain symbol block as

$$\hat{s}_g(t)=\frac{1}{N_c}\sum_{k=0}^{N_c-1}\hat{S}_g(k)\exp\left(j2\pi k\frac{t}{N_c}\right). \qquad (16)$$

Finally, the data demodulation is carried out.

### 2.3. Decoding without FFT/IFFT

The frequency-domain STBC-JTRD decoding presented in Sect. 2.2 requires $N_c$-point FFT/IFFT operations to obtain $\{R_{q,m}(k)\}$ and transform them back to time-domain symbol blocks. An equivalent time-domain decoding that requires no FFT/IFFT operations is possible. $N_c$-point IFFT of $\{R^*_{q,m}(k);k=0\sim(N_c-1)\}$ can be expressed as

$$\frac{1}{N_c}\sum_{k=0}^{N_c-1} R^*_{q,m}(k)\left(j2\pi k\frac{t}{N_c}\right)=r^*_{q,m}((N_c-t)\bmod N_c). \qquad (17)$$

Hence, frequency-domain STBC-JTRD decoding and IFFT of Eqs. (13) and (16) can be replaced by

$$\begin{pmatrix}\hat{s}_0(t)\\ \hat{s}_1(t)\end{pmatrix}=\begin{pmatrix}r_{0,0}(t)+r^*_{1,1}((N_c-t)\bmod N_c)\\ r_{0,1}(t)-r^*_{1,0}((N_c-t)\bmod N_c)\end{pmatrix}$$

for $N_r$=2. (18)

Equivalent time-domain decoding for $N_r$=3 and 4 are omitted for the sake of brevity.

### 2.4. Comparison with STTD

The well-known space-time transmit diversity (STTD) [12-16] can also be applied to the SC transmission with MMSE-FDE [6]. This is called frequency-domain STTD in this paper. Frequency-domain STBC-JTRD is compared with STTD in Table 2. STBC-JTRD requires the CSI at the transmitter while STTD requires it at the receiver. In STBC-JTRD, an arbitrary number of transmit antennas can be used while the number $N_r$ of receive antennas is limited to 4. This is advantageous for the downlink (base-to-mobile) applications since more antennas can be implemented at a base station than at a mobile receiver. When $N_r$=2, the coding rate is $R$=1. However, when $N_r$=3 and 4, the coding rate reduces to $R$=3/4. On the other hand, if STTD is used, only $N_t$=2 antennas can be equipped at the base station for coding rate $R$=1 while an arbitrary number of receiver antennas can be implemented at a mobile station (note that although more than 2 ($N_t$>2) transmit antennas can be used at the base station, the coding rate reduces to $R$=3/4~5/8, [14-16]).

Table 2 Comparison of STBC-JTRD and STTD

| Diversity scheme | No. of transmit antennas, $N_t$ | No. of receive antennas, $N_r$ | CSI required at | Coding rate $R$ |
|---|---|---|---|---|
| STBC -JTRD | Arbitrary | 2 | Transmitter side | 1 |
| | | 3 | | 3/4 |
| | | 4 | | 3/4 |
| STTD | 2 | Arbitrary | Receiver side | 1 |
| | 3 | | | 3/4 |
| | 4 | | | 3/4 |
| | 5 | | | 2/3 |
| | 6 | | | 2/3 |
| | 7 | | | 5/8 |

## 3. MMSE PRE-EQUALIZATION WEIGHT

We will derive the MMSE weight that minimizes the mean square error (MSE) between the transmit signal and the received signal. Since the frequency-domain STBC-JTRD encoding alters the transmitted signal spectrum shape in a frequency-selective fading channel, the signal-to-noise power ratio (SNR) is unproportional to the MSE. Therefore, in this paper, we introduce the relative equalization error $e(k)$ defined as

$$e(k)=\frac{\hat{S}_g(k)-C_{N_r}\sqrt{2E_s/T_s}S_g(k)}{C_{N_r}\sqrt{2E_s/T_s}\sqrt{E[|S_g(k)|^2]}}\quad . \tag{19}$$

Since the same data symbol is spread over all frequencies, the BER performance achievable with frequency-domain STBC-JTRD depends on the sum of mean square relative equalization errors given as

$$e^2=\sum_{k=0}^{N_c-1}E[|e(k)|^2]\quad . \tag{20}$$

We can find the MMSE pre-equalization weight $w_{m,n}(k)$ that minimizes Eq. (20), i.e.,

$$\frac{\partial e^2}{\partial w_{m,n}(k)}=0 \tag{21}$$

for all $k$=0~($N_c$−1), $m$=0~($N_r$−1) and $n$=0~($N_t$−1). After some manipulations, the following MMSE pre-equalization weight is obtained:

$$w_{m,n}(k)=\frac{H_{m,n}(k)}{\frac{1}{N_r}\sum_{n=0}^{N_t-1}\sum_{m=0}^{N_r-1}|H_{m,n}(k)|^2+\left(\frac{E_s}{N_0}\right)^{-1}}\ . \tag{22}$$

## 4. COMPUTER SIMULATION

Table 3 summarizes the simulation conditions. We assume $N_c$=256, $N_g$=32, and a symbol-spaced $L$=16-path frequency-selective block Rayleigh fading channel with uniform power delay profile (i.e., the ensemble average of $|h_{m,n,l}|^2$ is $E[|h_{m,n,l}|^2]=1/L$ for all $m$, $n$, $l$). The ideal channel estimation is assumed. For comparison, we also evaluate, by the computer simulation, the average BER performance with frequency-domain STTD.

Table 3 Simulation condition

| | | |
|---|---|---|
| Data modulation | | QPSK |
| Transmitter | No. of FFT points | $N_c$=256 |
| | Guard interval | $N_g$=32 |
| | No. of transmit antennas | $N_t$=1, 2, 3, 4 |
| | Pre-equalization weight | MMSE |
| Channel model | No. of paths | $L$=16 |
| | Power delay profile | Uniform |
| | Time delay | $\tau_l=l$, $l$=0~$L$-1 |
| Receiver | No. of receive antennas | $N_r$=1, 2, 3, 4 |
| Channel estimation | | Ideal |

### 4.1. Average BER performance

Figure 3 shows the average BER performance achievable with frequency-domain STBC-JTRD as a function of the transmit $E_b/N_0$ (=0.5($E_s/N_0$)(1+$N_g/N_c$)) with $N_r$ as a parameter for $N_t$=1. It is seen from Fig. 3 that the BER performance is significantly improved by increasing $N_r$. This is because, for a large number of receiver antennas, the equivalent channel gain varies less in the frequency-domain (i.e., the channel becomes close to a frequency-nonselective channel), and hence the inter-symbol interference (ISI) can be sufficiently suppressed. The reduction of the required $E_b/N_0$ for a BER=$10^{-4}$ is as much as 6 dB when $N_r$ is increased from 1 to 4.

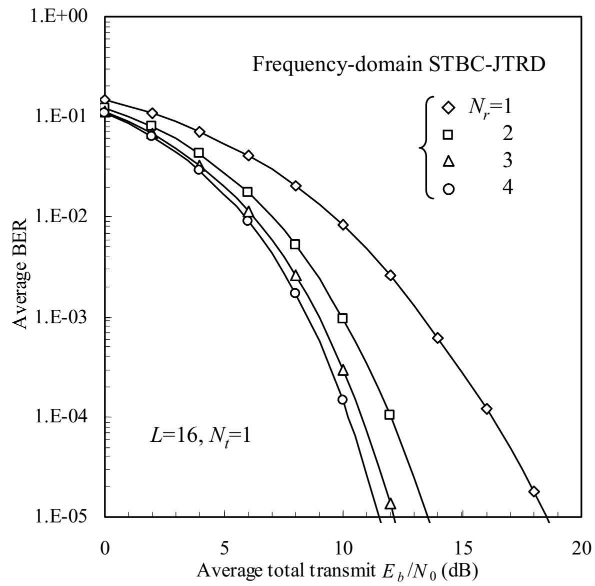


Fig. 3 Average BER performance with $N_r$ as a parameter.

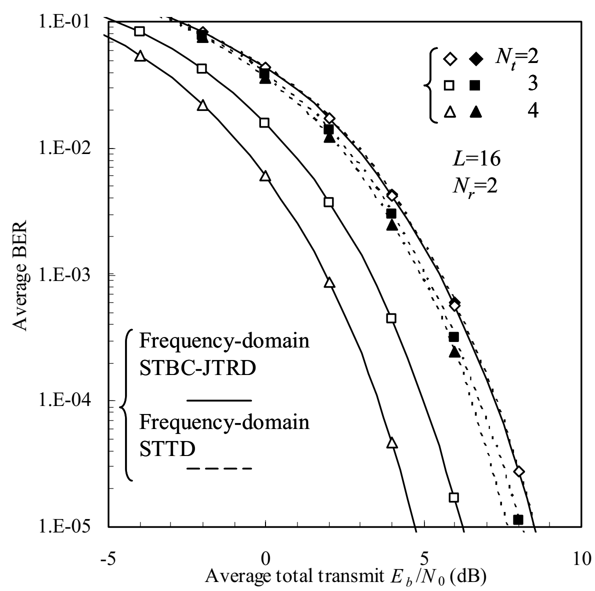


Fig. 4 Comparison of frequency-domain STBC-JTRD and STTD with $N_t$ as a parameter.

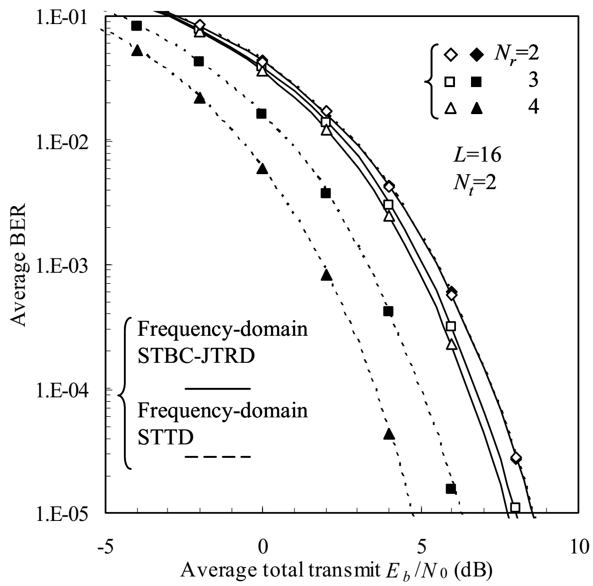


Fig. 5 Comparison of frequency-domain STBC-JTRD and STTD with $N_r$ as a parameter.

### 4.2. Comparison of frequency-domain STBC-JTRD and frequency-domain STTD

Figure 4 compares the average BER performances of frequency-domain STBC-JTRD and frequency-domain STTD with $N_t$ and $N_r$ as parameters. It is seen from Fig. 4 that the performance improvement of frequency-domain STBC-JTRD by increasing $N_t$ is larger than frequency-domain STTD, since the received SNR of frequency-domain STTD is reduced by a factor of $1/N_t$ [11] than frequency-domain STBC-JTRD using $N_t$ transmit antennas. Note that when $N_t$>2, the coding-rate is reduced to $R$=3/4 for STTD, but it is kept to $R$=1 for frequency-domain STBC-JTRD. On the other hand, it can be seen from Fig. 5 that when $N_r$ is increased, the BER performance of frequency-domain STTD is significantly improved, while that of frequency-domain STBC-JTRD is only slightly improved. Thereby, the use of frequency-domain STBC-JTRD is advantageous for the downlink applications where the number of antennas at a mobile terminal is limited due to available space limitation while a relationally large number of antennas can be implemented at a base station.

It should be noted that two-antenna frequency-domain STTD is a good option for the uplink applications since an arbitrary number of receiver antennas can be implemented at a base station without loss of the transmission efficiency.

## 5. CONCLUSION

In this paper, we proposed the frequency-domain space-time block coded-joint transmit/receive diversity (frequency-domain STBC-JTRD), which requires CSI only at a transmitter side. An arbitrary number of transmit antennas can be used without reducing the coding rate when 2 receive antennas are used; however the coding rate reduces to 3/4 when 3 or 4 receive antennas are used. Frequency-domain STBC-JTRD is suitable for the downlink (base-to-mobile) applications since most of the antennas can be implemented at the base station for the given total number of antennas. The average BER performance of frequency-domain STBC-JTRD in a frequency-selective fading channel was evaluated by the computer simulation. It was shown that the BER performance improvement by increasing the transmit antennas of frequency-domain STBC-JTRD is larger than frequency-domain STTD.

## REFERENCES

[1] W.C., Jakes Jr, Ed, *Microwave mobile communications*, Wiley, New York, 1974.

[2] J.G. Proakis, *Digital communications*, 2nd ed., McGraw-Hill, 1995.

[3] Y. Akaiwa, *Introduction to digital mobile communication*, Wiley, Newyork, 1997.

[4] D. Falconer, S. L. Ariyavisitakul, A. Benyamin-Seeyar, and B. Eidson, "Frequency domain equalization for single-carrier broadband wireless systems," IEEE Commun. Mag., Vol. 40, pp. 58-66, Apr. 2002.

[5] F. Adachi, D. Garg, S. Takaoka, and K. Takeda, "Broadband CDMA techniques," IEEE Wireless Commun., Vol. 12, No. 2, pp. 8-18, Apr. 2005.

[6] K. Takeda and F. Adachi, "MMSE frequency-domain equalization combined with space-time transmit diversity and antenna received diversity for DS-CDMA," Proc. IEEE VTC'04 spring, May. 2004.

[7] I. Cosovic, M. Schnell, and A. Springer, "On the performance of different channel pre-compensation techniques for uplink time division duplex MC-CDMA," Proc. IEEE VTC'03 fall, Oct. 2003.

[8] S. Abe, S. Takaoka, H. Tomeba and, F. Adachi, "Frequency-domain pre-equalization for MC-CDMA/TDD uplink and its bit error rate analysis," IEICE Trans. Commun., Vol. E89-B, No. 1, pp. 162-173, Jan. 2006.

[9] Lai-U Choi and Ross D. Murch, "Frequncy domain pre-equalization with transmit diversity for MISO broadband wireless communications," Proc. IEEE VTC'02 Fall, Oct. 2002.

[10] H. Tomeba, K. Takeda, and F. Adachi, "Frequency-domain pre-equalization for multi-code DS-CDMA mobile radio," Proc. IEEE VTS 2nd APWCS, pp. 184-188, Aug. 2005.

[11] H. Tomeba, K. Takeda, and F. Adachi, "Space-time block coded transmit/receive diversity," Proc. IEEE VTC05 fall, Dallas, USA, Sept. 25-28, 2005.

[12] S. M. Alamouti, "A simple transmit diversity technique for wireless communications," IEEE J. Select. Areas. Commun, Vol. 16, No. 8, pp. 1451-1458, Oct. 1998.

[13] V. Tarokh, H. Jafarkhani, and A. R. Calderbank, "Space-time block coding for wireless communications: performance results," IEEE J. Select. Areas. Commun, Vol. 17, No. 3, pp. 451-460, Mar. 1999.

[14] V. Tarokh, H. Jafarkhani, and A. R. Calderbank, "Space-time block codes from orthogonal designs," IEEE Trans. Inform. Theory, Vol. 45, No. 5, pp. 1456-1467, July 1999.

[15] X.-B. Liang, "A high-rate orthogonal space-time block code," IEEE Commun., Lett., Vol. 7, No. 5, pp. 222-223, May 2003.

[16] W. Su, X. G. Xia, and K. J. R Liu, "A systematic design of high-rate complex orthogonal space-time block codes," IEEE Commun., Lett., Vol. 8, No. 6, pp. 380-382, June 2004.