Gigabit Distributed Antenna Network And Its Related Wireless Techniques

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Abstract—**For the realization of future wireless networks, gigabit wireless technology which can achieve higher-than-1Gbps data transmission with extremely low transmit power is indispensable. We have been studying the distributed antenna network (DAN) and the frequency-domain wireless signal processing. In DAN, many antennas or clusters of antennas are spatially distributed over a service area and they are connected by means of optical fiber links with DAN signal processing center (SPC). A number of distributed antennas cooperatively serve mobile users using spatial multiplexing, diversity, array or relaying technique. In this paper, the recent research advances of gigabit DAN and its related wireless techniques are introduced.**

Keywords-component; Distributed antenna network, transmit antenna diversity, space-time coding, frequency-domain equalization

I. INTRODUCTION

In early 1980's, cellular mobile communication systems appeared which made "anytime, anywhere" communications possible. Cellular mobile communication systems have evolved from narrowband networks of around 10kbps (1^{st} and 2^{nd} generation systems) to wideband networks of around 10Mbps (3^{rd} generation systems). Along with the advancement of wireless techniques, wireless services have been shifted from simple voice to data and then broadband Internet related data services including video data. Now, the 3^{rd} generation long term evolution (LTE) systems with 100Mbps peak data rate are deployed in some countries [1]. A next step is the development of broadband wireless technology which will be used in the 4^{th} generation systems of a peak data rate of around 1Gbps.

In the future fixed networks, a variety of broadband network services will be available by the so called cloud computing networks. Along with this evolution of fixed networks, wireless networks need to further evolve in order to extend a variety of broadband network services available in the fixed networks to wireless users. Wireless access networks need to be enhanced to provide a gigabit wireless pipe to each user mobile terminal (MT).

However, there are a number of important technical issues to be addressed, e.g., limited bandwidth, frequency-selective fading, limited transmit power. In this paper, some of the above technical issues are discussed and then a distributed antenna network (DAN) is introduced as a promising solution.

II. TECHNICAL ISSUES

The gigabit wireless channels are extremely frequency-selective. The received signal spectrum is severely distorted. An advanced equalization technique is necessary to achieve high quality gigabit wireless data transmissions in a strong frequency-selective channel. A promising equalization technique is frequency-domain equalization (FDE) [2], [3], [4].

In addition to the above, the path loss and shadowing loss cause a severe power loss since the transmit power is limited. The communication range is limited by the uplink (MT-to-BS). With the same MT transmit power as in the present wireless systems (i.e., 3^{rd} generation cellular systems), the communication range of the gigabit wireless networks will significantly shrink and gigabit wireless services may be available near BS only. Therefore, a fundamental change is necessary in wireless access network architecture.

Distributed antenna system or network (DAS or DAN) [5] combined with frequency-domain signal processing has a potential to solve the above problems. For uplink signal transmissions, the single-carrier (SC) transmission is promising since it has a lower peak-to-average power ratio (PAPR) than multi-carrier (MC) transmission (using a transmit power amplifier with the same peak power, SC provides longer communication range than MC). Therefore, we have been investigating the potential of gigabit SC-DAN [5].

III. GIGABIT DAN

One promising solution to solve the problems arising from frequency-selectivity of the channel and limited transmit power is DAN. In DAN, as shown in Fig. 1, the conventional BS is replaced by the signal processing center (SPC) and many antennas or clusters of antennas are spatially distributed around the SPC so that some antennas can always be visible from an MT with a high probability. Antennas or antenna clusters are connected to a SPC by means of optical fiber links or wireless links. A number of distributed antennas cooperate and act as distributed MIMO spatial multiplexing, antenna diversity, antenna array, or relay. Probably the most powerful application is distributed transmit/receive antenna diversity using FDE. The problems can simultaneously be mitigated which result from distance-dependent path loss and shadowing loss as well as the instantaneous signal power variations due to fading.

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For DAN uplink and downlink transmissions, it is desirable to use as many distributed antennas as possible while the number of MT antennas is limited to one or two since there is not enough space to equip too many antennas at an MT. As a promising DAN downlink transmit diversity technique, we have developed a frequency-domain space-time block coded joint transmit/receive diversity (FD-STBC-JTRD) combined with transmit FDE [6]. The FD-STBC-JTRD is an extension of joint transmit diversity/transmit FDE [7] by adding antenna diversity reception and achieves an $N\times M$-th order (full) diversity gain, where N and M represent the number of distributed transmit antennas and that of MT receive antennas, respectively.

The FD-STBC-JTRD allows the use of up to M=4 MT receive antennas while requiring only simple addition, subtraction, and conjugate operations at an MT receiver and thus, alleviates the computational complexity problem of MT receivers. Although the number of MT receive antennas is limited to M=4, FD-STBC-JTRD allows the use of an arbitrary number of distributed transmit antennas, N. There are two types of transmit FDE weights: MMSE weight and maximum channel capacity weight.

For the uplink transmissions, frequency-domain space time transmit diversity (FD-STTD) [8] is promising. This is because FD-STTD allows the use of an arbitrary number N of distributed receive antennas although the number M of MT transmit antennas is limited. A combination of FD-STBC-JTRD for downlink and FD-STTD for uplink is suitable for DAN.

In Sect. III-A, we present the signal representation for SC-DAN downlink using FD-STBC-JTRD. Section III-B introduces the transmit FDE weight for FD-STBC-JTRD. The SC-DAN uplink using FD-STTD is described in Sect. III-C. Section III-D discusses the computer simulation results.

The SC signal transmission is a block transmission. In this paper, a block size of N_c symbols is assumed.

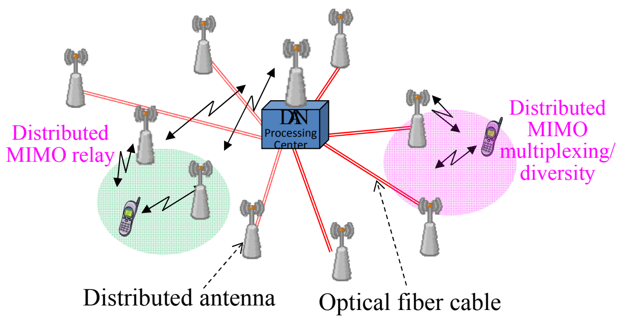

Fig. 1 Distributed antenna network (DAN).

A. FD-STBC-JTRD (downlink)

The FD-STBC-JTRD transmitter/receiver structure using N distributed transmit antennas and M MT receive antennas is illustrated in Fig. 2.

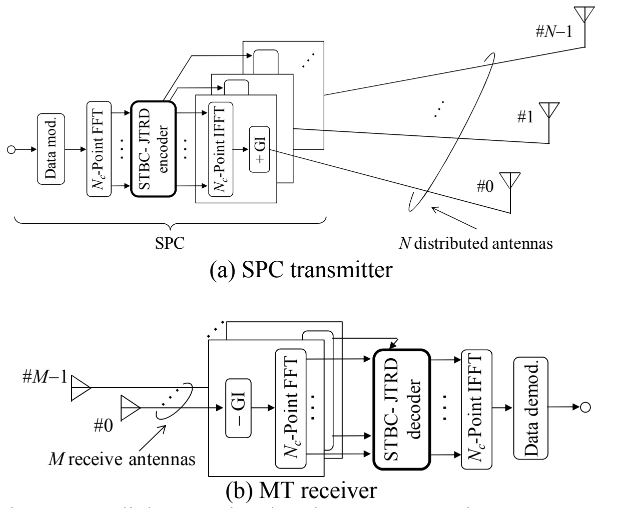

Fig. 2 Downlink transmitter/receiver structure using FD-STBC-JTRD.

At the SPC, a sequence of J blocks of N_c symbols each is transformed by an N_c-point fast Fourier transform (FFT) into a sequence of J frequency-domain signals and then, encoded into N (= the number of distributed transmit antennas) streams of Q coded frequency-domain signal blocks each. FD-STBC-JTRD decoding at an MT receiver needs addition, subtraction, and conjugate operations only. A combination of J and Q and coding rate R are shown in Table 1 for M=1 to 4.

The achievable diversity gain increases with increasing M; however, the coding rate reduces to 3/4 when M=3 and 4. This suggests that the channel capacity or throughput may be maximized at M=2. Below, for the sake of brevity, only the signal representation for M=2 (therefore, J=Q=2) is presented (for M=3 and 4, see [6]). However, the computer simulation results on the channel capacity will be shown for M=1~4 to confirm that M=2 maximizes the channel capacity.

Table 1. J, Q, and coding rate R

No. of transmit antennas	No. of receive antennas	J	Q	Coding rate
Arbitrary	1	1	1	1
	2	2	2	1
	3	3	4	3/4
	4	3	4	3/4

When M=2, a pair of N_c-symbol blocks (J=2) to be transmitted are transformed by N_c-point FFT into a pair of frequency-domain signals $\{D_0(k), D_1(k); k=0\sim N_c-1\}$ (Q=2), which is represented by $\mathbf{D}(k)=[D_0(k)\quad D_1(k)]^T$. $\mathbf{D}(k)$ is then encoded into N pairs of frequency-domain coded signal blocks as

$$\mathbf{S}_{stbc-jtrd}(k)=\begin{bmatrix} S_{0,0}(k) & S_{1,0}(k) \\ S_{0,1}(k) & S_{1,1}(k) \\ \vdots & \vdots \\ S_{0,N-1}(k) & S_{1,N-1}(k) \end{bmatrix}=\mathbf{W}(k)\mathbf{D}_{stbc-jtrd}(k), \qquad (1)$$

where $S_{q,n}(k)$, q=0~1, is the k-th frequency component at the n-th transmit antenna. In Eq. (1), $\mathbf{D}_{stbc-jtrd}(k)$ is the encoding matrix of size 2×2 for M=2, given by

$$\mathbf{D}_{stbc-jtrd}(k)=\begin{bmatrix} D_0(k) & -D_1^*(k) \\ D_1(k) & D_0^*(k) \end{bmatrix}. \qquad (2)$$

$\mathbf{W}(k)$ is the transmit FDE weight matrix of size 2×N, given as

$$\mathbf{W}(k)=A(k)\mathbf{H}^H(k), \qquad (3)$$

where

$$\mathbf{H}(k)=\begin{bmatrix} H_{0,0}(k) & H_{0,1}(k) & \cdots & H_{0,N-1}(k) \\ H_{1,0}(k) & H_{1,1}(k) & \cdots & H_{1,N-1}(k) \end{bmatrix} \qquad (4)$$

and $()^H$ denotes the Hermitian transpose. In Eq. (3), $A(k)$ is introduced to keep the transmit power intact after the encoding and will be discussed in Sect. III-B. $H_{m,n}(k)$ in Eq. (4) is the channel gain between the n-th distributed transmit antenna and the m-th MT receive antenna. An N_c-point inverse FFT (IFFT) is applied to $\{\mathbf{S}(k); k=0\sim N_c-1\}$ to obtain N pairs of N_c-symbol blocks to be transmitted from N distributed antennas.

At the MT receiver, a superposition of N pairs of coded N_c-symbol blocks is received by M=2 antennas. The received signals are transformed by N_c-point FFT into frequency-domain signals, $\{R_{m,0}(k)$ and $R_{m,1}(k)$; m=0~1$\}$, which can be expressed using the matrix form as

$$\mathbf{R}(k)=\begin{bmatrix} R_{0.0}(k) & R_{0,1}(k) \\ R_{1.0}(k) & R_{1,1}(k) \end{bmatrix} =\sqrt{\frac{2E_s}{T_s}}\mathbf{H}(k)\mathbf{S}_{stbc-jtrd}(k)+\mathbf{N}(k), \qquad (5)$$

where

$$\mathbf{N}(k)=\begin{bmatrix} N_{0,0}(k) & N_{0,1}(k) \\ N_{1,0}(k) & N_{1,1}(k) \end{bmatrix} \qquad (6)$$

is the noise matrix with $\{N_{m,q}(k)$; m=0~1, q=0~1$\}$ being independent zero-mean complex Gaussian variables having variance $2N_0/T_s$, where N_0 denotes the single-sided power spectrum density of AWGN and T_s denotes the data symbol duration.

FD-STBC-JTRD decoding for M=2 is carried out to obtain the frequency-domain signal vector $\hat{\mathbf{D}}(k)=[\hat{D}_0(k) \quad \hat{D}_1(k)]^T$ corresponding to the transmitted signal vector $\mathbf{D}(k)=[D_0(k) \quad D_1(k)]^T$ as

$$\hat{\mathbf{D}}(k)=\begin{bmatrix} R_{0,0}(k)+R_{1,1}^*(k) \\ R_{0,1}(k)-R_{1,0}^*(k) \end{bmatrix} =\sqrt{\frac{2E_s}{T_s}}A(k)\sum_{m=0}^{1}\sum_{n=0}^{N-1}\left|H_{m,n}(k)\right|^2\mathbf{D}(k) +\begin{bmatrix} N_{0,0}(k)+N_{1,1}^*(k) \\ N_{0,1}(k)-N_{1,0}^*(k) \end{bmatrix}. \qquad (7)$$

Finally, N_c-point IFFT is applied to transform $\{\hat{\mathbf{D}}(k)$; k=0~N_c−1$\}$ into a pair of soft decision N_c-symbol blocks (J=2).

It can be clearly seen from Eq. (7) that the downlink FD-STBC-JTRD using N distributed transmit antennas and M MT receive antennas can achieve an N×M(=2)-th order (full) diversity gain.

In the following, we consider a SC-DAN with N distributed antennas and M MT antennas.

B. Transmit FDE weight for FD-STBC-JTRD (downlink)

Two types of transmit FDE weights are presented: joint water-filling and maximal ratio transmission (WF-MRT) weight [9] and minimum mean square error (MMSE) weight [6].

i) Joint WF-MRT weight

The joint WF-MRT weight is the one which maximizes the channel capacity. The problem formulation of channel capacity maximization under the transmit power constraint can be written from Eq. (7) as

$$\max_{\{A(k)\}} C=\frac{1}{N_c}\sum_{k=0}^{N_c-1}\log_2\left(1+\frac{E_s}{N_0}\left|A_{stbc-jtrd}^{wf-mrt}(k)\right|^2\left(\sum_{m=0}^{M-1}\sum_{n=0}^{N-1}\left|H_{m,n}(k)\right|^2\right)^2\right)$$
$$\text{s.t.} \quad \sum_{k=0}^{N_c-1}\left(\left|A_{stbc-jtrd}^{wf-mrt}(k)\right|^2\sum_{m=0}^{M-1}\sum_{n=0}^{N-1}\left|H_{m,n}(k)\right|^2\right)=N_c. \qquad (8)$$

The solution of Eq. (8) is denoted by $\left\{A_{stbc-jtrd}^{wf-mrt}(k); k=0\sim N_c-1\right\}$ and is given by

$$A_{stbc-jtrd}^{wf-mrt}(k)=\frac{1}{\sqrt{\sum_{m=0}^{M-1}\sum_{n=0}^{N-1}\left|H_{m,n}(k)\right|^2}} \times\left[\max\left\{\varphi_{2D}-\frac{(E_s/2N_0)^{-1}}{\sum_{m=0}^{M-1}\sum_{n=0}^{N-1}\left|H_{m,n}(k)\right|^2},0\right\}\right]^{\frac{1}{2}}, \qquad (9)$$

where φ_{2D} is set so as to satisfy the transmit power constraint shown in Eq. (8). Substituting Eq. (9) into Eq. (3) gives the joint WF-MRT weight. The channel capacity C (bps/Hz) for the given channel realization is given as

$$C=\frac{1}{N_c}\sum_{k=0}^{N_c-1}\log_2\left(\begin{array}{l}1+\frac{E_s}{N_0}\sum_{m=0}^{M-1}\sum_{n=0}^{N-1}\left|H_{m,n}(k)\right|^2\times\\ \left[\max\left\{\varphi_{2D}-\frac{(E_s/2N_0)^{-1}}{\sum_{m=0}^{M-1}\sum_{n=0}^{N-1}\left|H_{m,n}(k)\right|^2},0\right\}\right]\end{array}\right). \quad (10)$$

The channel capacity distribution of SC-DAN downlink using FD-STBC-JTRD with joint WF-MRT weight is evaluated by Monte-Carlo numerical computation method. The channel condition is as follows. Distributed antennas are equally distantly located in the area and N antennas closest from an MT are selected for downlink transmissions. The MT equipped with M antennas is randomly generated in the area. The channel is assumed to follow an L=16-path frequency-selective Rayleigh fading, a log-normally distributed shadowing having standard deviation σ=7.0 dB, and a distance dependent path loss having path loss exponent α= 3.5.

Fig. 3 plots the 10% outage capacity, below which the channel capacity falls with 10% probability, for the normalized transmit E_s/N_0=10 dB (i.e., the transmit power is the one which provides the received E_s/N_0=10 dB at the adjacent antenna location). It can be seen from the figure that M=2 maximizes the channel capacity. This is a consequence of trade-off between diversity gain and coding rate; as M increases, the diversity gain increases, but the coding rate R reduces to 3/4 when M=3 (see Table 1).

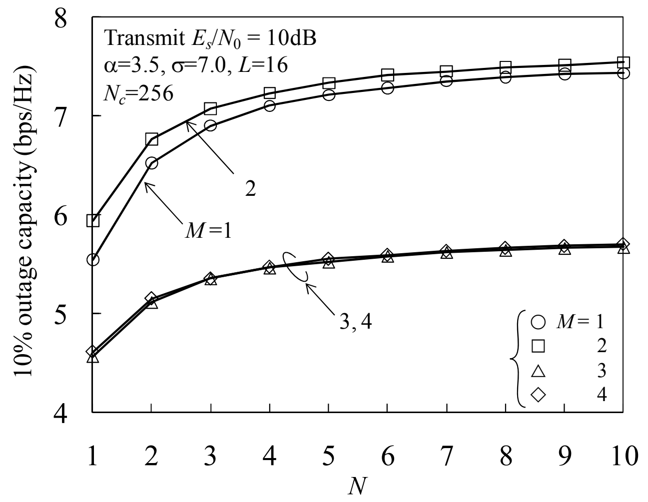

Fig. 3 10% outage channel capacity of SC-DAN downlink using FD-STBC-JTRD with joint WF-MRT weight.

ii) MMSE weight

The MMSE weight is the one which minimizes the mean square error (MSE) between $\hat{\mathbf{D}}(k)=[D_0(k)\quad D_1(k)]^T$ and $\mathbf{D}(k)=[D_0(k)\quad D_1(k)]^T$. Similar to Eq. (8), finding the MMSE weight can be formulated as

$$\begin{array}{l}\min_{\{A(k)\}}\quad e=\sum_{k=0}^{N_c-1}\sum_{m=0}^{M-1}E[\left|\hat{D}_m(k)-D_m(k)\right|^2]\\ \text{s.t.}\quad \sum_{k=0}^{N_c-1}\left(\left|G\cdot A_{stbc-jtrd}^{mmse}(k)\right|^2\sum_{m=0}^{M-1}\sum_{n=0}^{N-1}\left|H_{m,n}(k)\right|^2\right)=N_c\end{array}. \quad (11)$$

By solving Eq. (11), $\{A_{stbc-jtrd}^{mmse}(k);k=0\sim N_c-1\}$ is given as [7]

$$A_{stbc-jtrd}^{mmse}(k)=\frac{1}{\frac{1}{M}\sum_{m=0}^{M-1}\sum_{n=0}^{N-1}\left|H_{m,n}(k)\right|^2+\left(\frac{E_s}{N_0}\right)^{-1}} \quad (12)$$

and

$$G=\left(\frac{1}{N_c}\sum_{k=0}^{N_c-1}\{A_{stbc-jtrd}^{mmse}(k)\}^2\sum_{m=0}^{M-1}\sum_{n=0}^{N-1}\left|H_{m,n}(k)\right|^2\right)^{-1/2}. \quad (13)$$

Substituting Eq. (12) into Eq. (3) gives the MMSE weight.

C. Receive FDE weight for FD-STTD (uplink)

FD-STBC-JTRD is suitable for the downlink transmission. For the uplink transmission, FD-STTD [8] is suitable since FD-STTD allows the use of an arbitrary number N of distributed receive antennas, although the number M of MT transmit antennas is limited. A pair of N_c-symbol block (subscripts "0" and "1" representing even and odd blocks, respectively) are transformed by N_c -point FFT into frequency-domain signals represented by $\mathbf{D}(k)=[D_0(k)\quad D_1(k)]^T$. The FD-STTD encoding is expressed using the matrix form as [8]

$$\mathbf{D}_{sttd}(k)=\sqrt{\frac{1}{2}}\begin{pmatrix}D_0(k) & D_1(k)\\ -D_1^*(k) & D_0^*(k)\end{pmatrix}. \quad (14)$$

Two pairs of FD-STTD codeword are transmitted from M=2 antennas and received by N distributed antennas, followed by N_c-point FFT. N pairs of the received frequency-domain signals, $\{R_{n,0}(k)$ and $R_{n,1}(k)$; n=0~N–1}, are expressed using the matrix form as

$$\begin{aligned}\mathbf{R}_{sttd}(k)&=\begin{bmatrix}R_{0,0}(k) & R_{0,1}(k)\\ R_{1,0}(k) & R_{1,1}(k)\\ \vdots & \vdots\\ R_{N-1,0}(k) & R_{N-1,1}(k)\end{bmatrix}\\ &=\sqrt{\frac{2E_s}{T_s}}\mathbf{H}^{\mathrm{T}}(k)\mathbf{D}_{sttd}(k)+\mathbf{N}(k)\end{aligned}. \quad (15)$$

where

$$\mathbf{N}(k)=\begin{bmatrix}N_{0,0}(k) & N_{0,1}(k)\\ N_{1,0}(k) & N_{1,1}(k)\\ \vdots & \vdots\\ N_{N-1,0}(k) & N_{N-1,1}(k)\end{bmatrix}. \quad (16)$$

is the noise matrix.

After the FD-STTD decoding, the frequency-domain signal vector $\hat{\mathbf{D}}(k) = [\hat{D}_0(k) \quad \hat{D}_1(k)]^T$ associated with $\mathbf{D}(k)=[D_0(k) \quad D_1(k)]^T$ is obtained as

$$\begin{aligned}\hat{\mathbf{D}}(k) &= \begin{bmatrix} \sum_{n=0}^{N-1} R_{n,0}(k)W_{n,0}^*(k) + \sum_{n=0}^{N-1} R_{n,0}^*(k)W_{n,1}(k) \\ \sum_{n=0}^{N-1} R_{n,0}(k)W_{n,1}^*(k) - \sum_{n=0}^{N-1} R_{n,1}^*(k)W_{n,0}(k) \end{bmatrix} \\ &= \sqrt{\frac{2E_s}{T_s}} A(k) \sum_{m=0}^{1}\sum_{n=0}^{N-1} \left|H_{m,n}(k)\right|^2 \mathbf{D}(k) \\ &+ \begin{bmatrix} \sum_{n=0}^{N-1} N_{n,0}(k)W_{n,0}^*(k) + \sum_{n=0}^{N-1} N_{n,0}^*(k)W_{n,1}(k) \\ \sum_{n=0}^{N-1} N_{n,0}(k)W_{n,1}^*(k) - \sum_{n=0}^{N-1} N_{n,1}^*(k)W_{n,0}(k) \end{bmatrix}\end{aligned} \quad . \quad (17)$$

Similar to the downlink case, the MMSE weight matrix that minimizes the MSE between $\hat{\mathbf{D}}(k)$ and $\mathbf{D}(k)$ is given, for the case of M MT antennas, as [8]

$$\begin{aligned}\mathbf{W}_{sttd}^{mmse}(k) &= A_{sttd}^{mmse}(k)\mathbf{H}^*(k) \\ &= \frac{\mathbf{H}^*(k)}{\frac{1}{M}\sum_{m=0}^{M-1}\sum_{n=0}^{N-1}\left|H_{m,n}(k)\right|^2 + \left(\frac{E_s}{N_0}\right)^{-1}},\end{aligned} \quad (18)$$

where

$$A_{sttd}^{mmse}(k) = \frac{1}{\frac{1}{M}\sum_{m=0}^{M-1}\sum_{n=0}^{N-1}\left|H_{m,n}(k)\right|^2 + \left(\frac{E_s}{N_0}\right)^{-1}}. \quad (19)$$

D. Uplink/downlink performance comparison

We compare the uplink and downlink BER performances. We consider FD-STBC-JTRD using joint WF-MRT weight for the uplink transmission and FD-STTD using MMSE weight for the downlink transmission. N arbitrary distributed antennas and M=2 MT antennas are assumed.

The complementary cumulative distribution function (CCDF) of the measured downlink and uplink bit error rates (BERs) are plotted in Fig.4 for E_s/N_0=5 dB. It can be clearly seen from the figure that well balanced downlink and uplink transmissions can be achieved. By increasing N (the number of distributed transmit/receive antennas), the probability that the received signal power drops due to path loss, shadowing loss, and frequency-selective fading can be more reduced and hence, the BER can be significantly reduced.

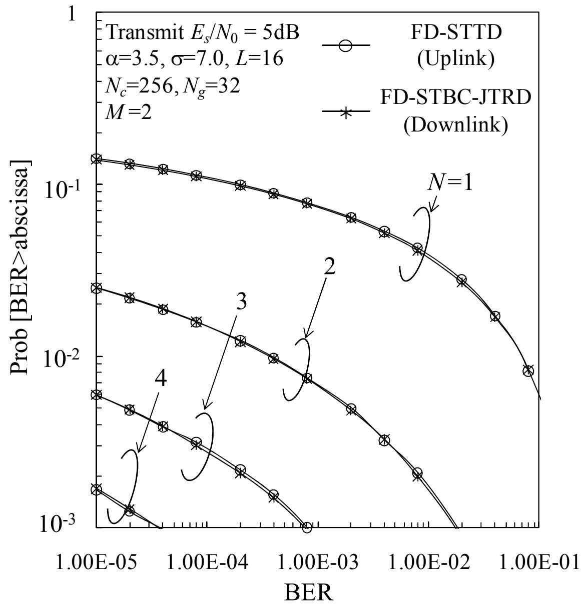

Fig. 4 Uplink/downlink performance comparison.

IV. RELATED WIRELESS TECHNIQUES

For gigabit wireless transmissions, development of some equalization techniques is necessary. FDE is an attractive equalization technique which can be combined with transmit diversity, receive diversity, beamforming, cooperative relaying, hybrid automatic repeat request (HARQ) combining, etc. FD-STBC-JTRD and FD-STTD introduced in Sect.III are a combination of transmit/receive diversity and one-tap FDE. Below, one of recent advances in frequency-domain equalization is introduced.

Recently, we successfully applied maximum likelihood detection employing QR decomposition and M-algorithm (QRM-MLD) [10] to the reception of cyclic prefix (CP) inserted SC block signals transmitted over a frequency-selective fading channel and proposed a new SC block detection called SC frequency-domain QRM-ML block detection (QRM-MLBD) [11]. SC frequency-domain QRM-MLBD is a powerful signal detection scheme combined with FDE; however, it requires higher computational complexity than conventional one-tap MMSE-FDE. The receiver structure of SC frequency-domain QRM-MLBD for a CP inserted SC (CP-SC) block transmission is illustrated in Fig. 5.

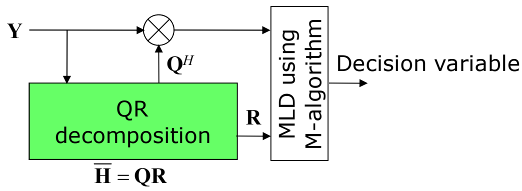

Fig. 5 SC frequency-domain QRM-MLBD.

Assuming the single-input single output (SISO) transmissions, the frequency-domain received signal vector $\mathbf{Y}=[Y(0),\ldots,Y(k),...,Y(N_c-1)]^T$ is expressed as

$$\mathbf{Y}=\sqrt{\frac{2E_s}{T_s}}\mathbf{HFd}+\mathbf{N}=\sqrt{\frac{2E_s}{T_s}}\overline{\mathbf{H}}\mathbf{d}+\mathbf{N}\,, \qquad (20)$$

where $\mathbf{F}$ is the DFT matrix of size $N_c\times N_c$, $\mathbf{H}=diag[H(0),\ldots,H(k),\ldots,H(N_c-1)]$ is the frequency-domain channel matrix, and $\mathbf{N}=[N(0),...,N(k),...,N(N_c-1)]^T$ is the noise vector.

QRM-MLBD consists of two steps: QR decomposition and M-algorithm. First, the QR decomposition is applied to the equivalent channel matrix $\overline{\mathbf{H}}$ to obtain $\overline{\mathbf{H}}=\mathbf{QR}$, where $\mathbf{Q}$ is an $N_c\times N_c$ unitary matrix and $\mathbf{R}$ is an $N_c\times N_c$ upper triangular matrix. The transformed frequency-domain received signal $\hat{\mathbf{Y}}=[\hat{Y}(0),...,\hat{Y}(n),...,\hat{Y}(N_c-1)]^T$ is obtained as

$$\hat{\mathbf{Y}}=\mathbf{Q}^H\mathbf{Y}=\sqrt{\frac{2E_s}{T_s}}\mathbf{Rd}+\mathbf{Q}^H\mathbf{N}\,. \qquad (21)$$

It can be understood from Eq. (21) that the MLD can be converted to the successive tree search problem and the computational complexity can be reduced by introducing the M-algorithm into the successive tree search. SC frequency-domain QRM-MLBD can achieve the BER performance close to the matched filter (MF) bound with significantly reduced computational complexity compared to the MLD.

However, the CP-SC block transmission requires a fairly large number of surviving paths in the M-algorithm and therefore, its computational complexity is still high compared to conventional one-tap MMSE-FDE. To overcome this problem, we suggested using a known training sequence (TS) aided SC (TA-SC) transmissions, in which the known TS in the previous block acts as the CP in the present block, as shown in Fig. 6 [12]. The known TS is exploited in the M-algorithm to reduce the number of surviving paths.

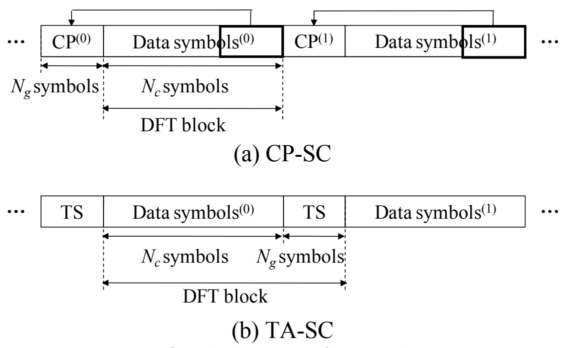

Fig. 6 CP-SC and TA-SC.

We compare CP-SC and TA-SC. We assume a block transmission of N_c=64 symbols. The channel is assumed to be a symbol-spaced L=16-path frequency-selective block Rayleigh fading channel having uniform power delay profile. Ideal channel estimation is assumed. The average BER performance of SC frequency-domain QRM-MLBD is plotted in Fig. 7. Also plotted are the BER performance achievable by one-tap MMSE-FDE and the theoretical MF bound. As the number P of surviving paths in the M-algorithm increases, the BER performance improves and approaches the MF bound.

When TA-SC is used, the required number of surviving paths in the M-algorithm is greatly reduced while achieving almost the same BER performance as CP-SC. In case of 16QAM, the overall complexity of frequency-domain QRM-MLBD for TA-SC reduces to about 7.4% of that for CP-SC.

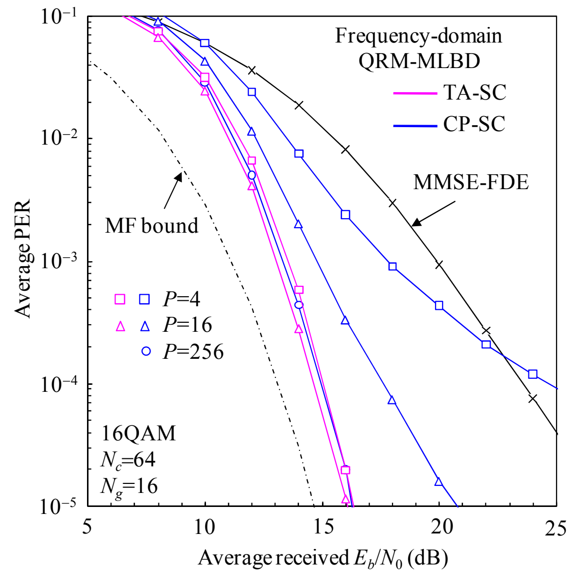

Fig. 7 BER performance comparison between CP-SC and TA-SC.

V. CONCLUSIONS

DAN is a promising wireless technology to realize gigabit wireless data transmissions. In this paper, we have introduced gigabit DAN combined with SC frequency-domain signal processing. Distributed transmit/received diversity can solve the problems arising from severe channel selectivity and limited transmit power. It is desirable to use as many distributed antennas as possible to achieve higher diversity gain while limiting the number of MT antennas to one or two so as to alleviate the complexity problem of MT. It was shown that the balanced downlink/uplink performance can be achieved by using FD-STBC-JTRD for the downlink while FD-STTD for the uplink. We also presented an advanced SC block detection called frequency-domain QRM-MLBD. This can be extended to MIMO multiplexing to significantly improve the transmission performance compared to the MMSE spatial filtering [13] and further to distributed MIMO multiplexing in SC-DAN.

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