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Performance Investigation of Two Transmit Diversity Schemes with Perfect/Imperfect Channel Evaluation in Wireless Communications

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SUMMARY Transmit diversity, a key technique derived against multi-path mitigation in wireless communication system, is examined and discussed. Especially, we present an approach to investigate perfect/imperfect channel detection when the maximal ratio receiver combined scheme (MRRC) and a simple transmit diversity scheme (STD) are used in the wireless systems, which provide remarkable schemes for diversity transmission over Rayleigh-fading channels using multiple antennas. In order to effectively make use of the transmit diversity techniques, the same approach is extended to process the situation of one transmit antennas and N receive antennas in MRRC scheme ($1 \times N$ MRRC) and two transmit antennas and N receive antennas in STD scheme ($2 \times N$ STD). The effects of perfect/imperfect channel detection and the diversity reception with independent and correlated Rayleigh-fading signals are evaluated and compared carefully.

key words: *transmit diversity, perfect/imperfect channel detection, antenna array, maximal-ratio combining, multi-path mitigation*

1. Introduction

Due to the increasing demand for mobile communications services, the development of highly efficient multiple access scheme [1]–[3], hierarchical structures [4], diversity techniques, smart antenna arrays [5], [6] and interference rejection technologies have been given considerable attention in recent times, especially during the development of IMT-2000 wireless communication systems. With the technology progressing, the goal of IMT-2000 is becoming ever closer. Despite the high capacity offered by code division multiple access (CDMA) technique, the expected demand is likely to outstrip the desired demands. One approach to improve the system performance is the use of spatial processing with antenna array to achieve diversity reception.

In wireless systems, since the received signal is subject to multi-path fading, which severely degrades the signal transmission performance and generates distortion of digital signal symbol. Some auxiliary techniques are necessary to reduce the multi-path fading effects. Theoretically, the most effective technique to mitigate multi-path fading is transmitter power control [5] that can generate the service balance [6]. With this approach, the major problem is the required transmitter dynamic range. For the transmitter to overcome a certain level change generated by fading, it must increase its power by the same level, that will generate the highly interference to all the other users. This is very severe problem in CDMA systems since all the users make use of the same bandwidth. In the other hand, in most cases it is not practical because of the radiation limitations, the cost and the size of the amplifiers.

Another of the most efficient techniques is diversity reception [7]–[9]. Considering most scattering wireless environments, antenna array diversity is practical, effective approach. Therefore, it is a widely applied technique for reducing the effect of multi-path fading. The literature contains many papers [7]–[9] devoted to the effects of diversity reception on the performance of wireless communications, in which two or more antenna arrays are employed at the receivers and/or transmitters to provide a number of independent transmission branches. The motivation of adopting diversity techniques is that when some of branches undergo deep fading, other branches may still have strong signal levels. The spatial separations by antenna arrays need to ensure low correlations of each branches, which depend on the allocated carrier and the structure design of antenna arrays.

References [7] and [8] proposed a delay diversity scheme for the base station simulcasting and later, independently. This scheme was proposed for a single base station in which copies of the same symbol are transmitted through multiple antenna arrays at the different times. Therefore an artificial multi-path distortion of received signal is created. In the receivers, a maximum likelihood sequence estimator (MLSE) or a minimum mean squared error (MMSE) equalizer is adopted to resolve the multi-path distortion and get the diversity gain. Reference [9] is a well known article on the overview of diversity reception that summarized the maximal ratio receiver combined (MRRC) scheme and proposed a simple transmit diversity (STD) scheme.

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In this paper, the author used simulation method and only independent Rayleigh-fading channels and perfect channel evaluation are assumed during the simulation. Although the author discussed the potential effects of imperfect channel evaluation, the practical results have not been given out. STD can improve the signal quality on one side of the link by simple processing across two transmit antennas on the opposite side. In this scheme, there is no feedback required from the receivers to the transmitters. For a fixed level of radiated power per transmit antenna, STD scheme has the same bit error ratio (BER) characteristics as MRRC when the channel evaluation is perfect. In practical, it is impossible to achieve this in the receivers, hence the method we use can be viewed as an extension of Ref. [9] and fill the gap that we focus our interests on the imperfect channel evaluation problem.

From these backgrounds, an analytical approach was proposed in this paper. Not only considering the independent Rayleigh-fading channels, but also the correlated Rayleigh-fading channels. A new transformation method has also been proposed to process the correlated Rayleigh-fading channels in MRRC and STD schemes. On the other aspects, because the imperfect channel evaluation at the receiver is a actual situation and has not been estimated, so in this paper, the imperfect channel evaluation has been defined. In order to effectively make use of the transmit diversity techniques and make them available in diverse wireless environments, such as personal mobile users, mobile stations located on cars or ships and other military mobile stations, the same approach is extended to process the situations of one transmit antennas and N receive antennas in MRRC scheme ($1 \times N$ MRRC) and two transmit antennas and N receive antennas in STD scheme ($2 \times N$ STD).

The remainder of this paper is organized as follows. In Sect. 2, the reviews of MRRC and STD are given. The perfect and imperfect channel evaluations are also defined in this section. Section 3 describes the $M \times N$ transmit diversity (M=1 for MRRC and M=2 for STD), and the derivation of the system BER due to the diversity schemes with independent and correlated Rayleigh-fading signals are also given. The numerical results are given and discussed in detail in Sect. 4. Finally, we conclude this paper.

2. MRRC, STD, and Imperfect Channel Evaluation

2.1 Characteristics of MRRC and STD

The characteristics of MRRC and STD for two diversity branches, have been investigated in Ref. [9], which are necessary to our extension are briefly recalled in this subsection. As shown in Fig. 1, at a given time, a signal s_0 is sent from only one transmit antenna in MRRC scheme. Like the electrical circuit, the wireless channel may be considered as a transform function and modelled by a complex multiplicative distortion composed a magnitude gain and a phase delay. Because of two receive antennas, referred to antenna 1 and antenna 2 at receiver, there are two diversity branch signals received by the two antennas, respectively. The two transform functions because of the two diversity branches can be expressed as $g_1 = \alpha_1 e^{j\phi_1}$ and $g_2 = \alpha_2 e^{j\phi_2}$. Then receive signals in which noises are added at receive antenna 1 and antenna 2 throughout different diversity branches corresponding to the transmit signal s_0 can be given as [9]

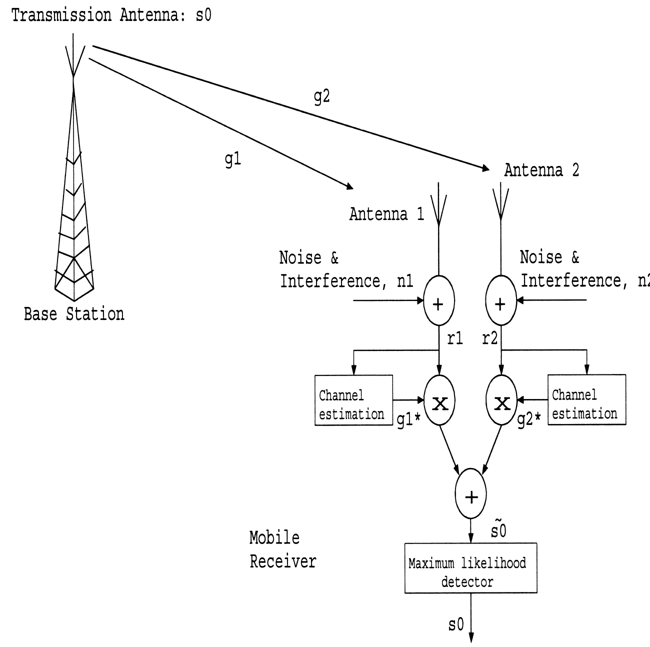

Fig. 1 System configuration of two branch MRRC diversity reception.

$$
\begin{aligned}
r_{1,MRRC} &= g_1 s_0 + n_1 \\
r_{2,MRRC} &= g_2 s_0 + n_2
\end{aligned}
\tag{1}
$$

where n_1 and n_2 are the noises with Gaussian distribution. They are the samples of independent complex Gaussian random variables. In the MRRC scheme with one transmit antenna and two receive antennas, the receive signal combining the two branches is as follows

$$
\begin{aligned}
\tilde{s}_{0,MRRC} &= g_1^* r_{1,MRRC} + g_2^* r_{2,MRRC} \\
&= (\alpha_1^2 + \alpha_2^2) s_0 + g_1^* n_1 + g_2^* n_2
\end{aligned}
\tag{2}
$$

where $*$ is the complex conjugate operation. In order to minimize BER, the maximum likelihood decision detector is considered and it may produce output signal which is a maximum likelihood estimate of s_0. That means to choose $s_0 = 1$ if $Re(\tilde{s}_{0,MRRC}) \geq 0$ and choose $s_0 = -1$, otherwise, when we consider the case of coherent binary phase shift key (BPSK) modulation so that the transmit signal is either $+1$ and -1.

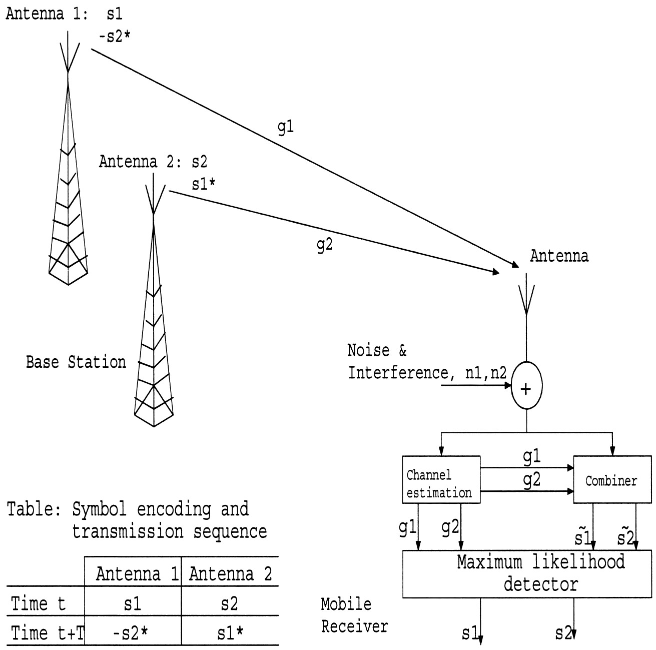

Fig. 2 System configuration of two branch STD diversity reception.

As shown in Fig. 2, STD scheme [9] uses two transmit antennas and one receive antenna that is different with MRRC scheme and may be defined by the following three functions:
(1) The encoding and transmission sequence of information symbols at the transmitter;
(2) The combined scheme at the receiver;
(3) The decision rule for maximum likelihood detection.

From the above description and system configuration shown in Fig. 2, in order to recover the symbol signals in the combined scheme that are also shown in Eqs. (4) and (5), the symbol encoding and transmission sequence designs are necessary for symbol signal recovery at the receiver. In this STD scheme, at a given symbol period, two symbol signals, denoted by s_1 and s_2, are simultaneously transmitted from the two antennas. In the first symbol period at time t, s_1 and s_2 are transmitted from antennas 1 and 2, respectively. In the time $t+T$, $-s_2^*$ and s_1^* are transmitted from antennas 1 and 2. It is the encoding and transmission sequence of information symbols at the transmitter, which is done in space and time that is often termed as space-time coding [9]. Then, the receive signals at receiver are

$$\begin{aligned} r_{1,STD} &= g_1 s_1 + g_2 s_2 + n_1 \\ r_{2,STD} &= -g_1 s_2^* + g_2 s_1^* + n_2 \end{aligned} \tag{3}$$

According to the combining scheme of these receive signals in Ref. [9], the following two signals are given as

$$\begin{aligned} \tilde{s}_{1,STD} &= g_1^* r_{1,STD} + g_2 r_{2,STD}^* \\ &= (\alpha_1^2 + \alpha_2^2) s_1 + g_1^* n_1 + g_2 n_2^* \end{aligned} \tag{4}$$

and

$$\begin{aligned} \tilde{s}_{2,STD} &= g_2^* r_{1,STD} - g_1 r_{2,STD}^* \\ &= (\alpha_1^2 + \alpha_2^2) s_2 - g_1 n_2^* + g_2^* n_1 \end{aligned} \tag{5}$$

$\tilde{s}_{1,STD}$ and $\tilde{s}_{2,STD}$ combining signals are sent to the maximum likelihood detector in which for these two signals, decision rules are used for recover the original signal. From Eqs. (4) and (5), the combining signals are equivalent to that obtained from two branch MRRC scheme. Only the delay phase are different, that means it do not degrade signal-to-noise ratio (SNR). Please note that all the above combining signals, whether MRRC or STD schemes, are under the perfect evaluation of channel transform functions in the receivers. In practical, it is impossible to ensure this, because of the wireless multi-pathes affected by too many reasons. It will of real important to investigate their effects which will be focused in the following section.

2.2 Perfect/Imperfect Channel Evaluation

All the derivations in the above section are under the consideration of perfect channel evaluation, which is composed of channel gain evaluation and phase evaluation, that means the receiver has perfect knowledge of the each multi-path branch. Although there are many factors that may change the characteristics of the multi-path channel, the channel evaluation error should be minimized as possible even when the channel evaluation detector in receiver could not correctly evaluate the condition of channel. According to the diversity schemes and combining schemes of receive signals, there are two kinds of evaluation errors referred to transmission gain evaluation error and transmission phase evaluation error.

According to the analytical model described in Sect. 2.1, there are two definitions of imperfect channel evaluation. They are the expressions of $\alpha_i e^{j\phi_i} + \beta_i e^{j\theta_i}$ and $(\alpha_i + \beta_i) e^{j(\phi_i + \theta_i)}$. Any one of them can be adopted in the investigation of system performance. In the definition of $\alpha_i e^{j\phi_i} + \beta_i e^{j\theta_i}$, because the fraction of imperfect channel estimation is defined as an independent part, then in all the following formulas, its effects in each expression, such as Eqs. (8) and (9), are the independent. One can see how many effects on the system clearly from these expressions. It is the advantage of this definition. Then in this paper, we adopt this definition as the estimated transform function related with the imperfect channel evaluation shown as follows

$$\hat{g}_i = \alpha_i e^{j\phi_i} + \beta_i e^{j\theta_i}, \quad \text{where} \quad i \in [1, 2...n] \tag{6}$$

where $\hat{g}_i$ is the output of the channel estimator about the i-th channel, in which composed of the exponents of evaluation errors about the transmission gain and phase referred to β_i and θ_i respectively. When β_i and θ_i are equal to zero at the same time, the channel evaluation becomes perfect referred to perfect channel evaluation. The results can be used to obtain BER when $\hat{g}_i$ is drawn according to any arbitrary fading channel.

3. BER Performance Evaluation

3.1 $1 \times N$ Transmit Diversity of MRRC

Based on the derivation process proposed in Sect. 2.1, $1 \times N$ MRRC performance analysis can be conducted by the following steps. When we introduce the imperfect channel evaluation in the derivation, the decision random variable for this MRRC scheme is denoted by $U_{MRRC} = Re(\tilde{s}_{0,MRRC})$. $\tilde{s}_{0,MRRC}$ can be given as

$$\begin{aligned}\tilde{s}_{0,MRRC} &= \sum_{i=1}^{N} \hat{g}_i^* r_{i,MRRC} \\ &= \sum_{i=1}^{N} (\alpha_i e^{j\phi_i} + \beta_i e^{j\theta_i})(\alpha_i e^{j\phi_i} s_0 + n_i) \quad (7)\end{aligned}$$

After some derivation, the mean, $E[U_{MRRC}]$ and the variance, $Var[U_{MRRC}]$ of the decision random variable U_{MRRC} can be expressed as follows

$$E[U_{MRRC}] = \sum_{i=1}^{N} (\alpha_i^2 + \alpha_i\beta_i \cos(\phi_i - \theta_i))s_0 \qquad (8)$$

and

$$Var[U_{MRRC}] = \sum_{i=1}^{N} (\alpha_i \cos\phi_i + \beta_i \cos\theta_i)\sigma_{n_i}^2 \qquad (9)$$

where $\sigma_{n_i}^2$ is the variance of independent complex Gaussian random variable, n_i which represents the noise on the i-th channel branch.

If the perfect transmission of each diversity branch is considered and the noise is following the Gaussian distribution, then U_{MRRC} is a linear function of independent Gaussian random variable. From Ref. [10], the BER is $Q(\sqrt{2SNR})$ for BPSK modulation, where SNR is the signal-to-noise ratio. Based on the diversity schemes and the above derivations, $E[U_{MRRC}]$ and $Var[U_{MRRC}]$ are the receive signal power and the noise power, respectively. Then $SNR = \frac{E[U_{MRRC}]}{Var[U_{MRRC}]}$. So one obtains the BER of BPSK for MRRC as [9], [10]

$$P_{BER,MRRC} = Q\left(\sqrt{\frac{2E[U_{MRRC}]}{Var[U_{MRRC}]}}\right) \qquad (10)$$

where,

$$Q(x) = \frac{1}{\sqrt{2\pi}} \int_x^{\infty} e^{-\frac{y^2}{2}} dy \qquad (11)$$

3.2 $2 \times N$ Transmit Diversity of STD

In $2 \times N$ STD scheme, it is possible to provide a diversity order [9] of $2N$ with the two transmit antennas and N receive antennas. According to the explanation of STD in above Sect. 2.1, the receive signals for i-th receive antenna are referred to $r_{i,1}$ and $r_{i,2}$, respectively, which can given as

$$\begin{aligned}r_{i,1} &= g_{i,1}s_1 + g_{i,2}s_2 + n_{i,1} \\ r_{i,2} &= -g_{i,1}s_2^* + g_{i,2}s_1^* + n_{i,2}\end{aligned} \qquad (12)$$

where $g_{i,1}$ and $g_{i,2}$ are the transform functions of signals from antenna 1 and antenna 2 when considering the i-th receive antenna. $n_{i,1}$ and $n_{i,2}$ are the noise on these two branches, respectively.

Recall that in the STD scheme, the two symbol are simultaneously transmitted. When we consider the imperfect channel evaluation described in Sect. 2.2, the combined signals $\tilde{s}_{1,STD}$ and $\tilde{s}_{2,STD}$ are obtained as

$$\begin{aligned}\tilde{s}_{1,STD} &= \sum_{i=1}^{N} \hat{g}_{i,1}^* r_{i,1} + \sum_{i=1}^{N} \hat{g}_{i,2} r_{i,2}^* \\ &= \sum_{i=1}^{N} (\alpha_{i,1}e^{-j\phi_{i,1}} + \beta_{i,1}e^{-j\theta_{i,1}})(\alpha_{i,1}e^{j\phi_{i,1}}s_1 \\ &\quad +\alpha_{i,2}e^{j\phi_{i,2}}s_2 + n_{i,1}) \\ &\quad +\sum_{i=1}^{N} (\alpha_{i,2}e^{j\phi_{i,2}} + \beta_{i,2}e^{j\theta_{i,2}})(-\alpha_{i,1}e^{-j\phi_{i,1}}s_2 \\ &\quad +\alpha_{i,2}e^{-j\phi_{i,2}}s_1 + n_{i,2}^*)\end{aligned} \qquad (13)$$

and

$$\begin{aligned}\tilde{s}_{2,STD} &= -\sum_{i=1}^{N} \hat{g}_{i,1} r_{i,2}^* + \sum_{i=1}^{N} \hat{g}_{i,2}^* r_{i,1} \\ &= -\sum_{i=1}^{N} (\alpha_{i,1}e^{j\phi_{i,1}} + \beta_{i,1}e^{j\theta_{i,1}})(-\alpha_{i,1}e^{-j\phi_{i,1}}s_2 \\ &\quad +\alpha_{i,2}e^{-j\phi_{i,2}}s_1 + n_{i,2}^*) \\ &\quad +\sum_{i=1}^{N} (\alpha_{i,2}e^{-j\phi_{i,2}} + \beta_{i,2}e^{-j\theta_{i,2}})(\alpha_{i,1}e^{j\phi_{i,1}}s_1 \\ &\quad +\alpha_{i,2}e^{j\phi_{i,2}}s_2 + n_{i,1})\end{aligned} \qquad (14)$$

For the symmetry, the performances for the both transmit symbols are the same, so we only need to investigate the BER for symbol signal s_1 or s_2. Here we select s_1 as an sample. By corresponding to the STD scheme, the decision random variable by $U_{STD} = Re(\tilde{s}_{1,STD})$. After some simple formulations, we can obtain the means and variance of U_{STD} as $E[U_{STD}]$ and $Var[U_{STD}]$, respectively. From Eq. (13), $E[U_{STD}]$ and $Var[U_{STD}]$ are given as

$$\begin{aligned}E[U_{STD}] &= \sum_{i=1}^{N} (\alpha_{i,1}^2 + \alpha_{i,2}^2 + \alpha_{i,1}\beta_{i,1}\cos(\phi_{i,1} - \theta_{i,1}) \\ &\quad +\alpha_{i,2}\beta_{i,2}\cos(\phi_{i,2} - \theta_{i,2}))s_1 \\ &\quad +(\alpha_{i,2}\beta_{i,1}\cos(\phi_{i,2} - \phi_{i,1})\end{aligned}$$

$$-\alpha_{i,1}\beta_{i,2}\cos(\phi_{i,1}-\theta_{i,2}))s_2 \quad (15)$$

and

$$Var[U_{STD}] = \sum_{i=1}^{N}(\alpha_{i,1}\cos\phi_{i,1}+\beta_{i,1}\cos\theta_{i,1})\sigma_{n_{i,1}}^2 + (\alpha_{i,2}\cos\phi_{i,2}+\beta_{i,2}\cos\theta_{i,2})\sigma_{n_{i,2}}^2 \quad (16)$$

If the perfect transmission is assumed and the noise is following the Gaussian distribution, then U_{STD} is a linear function of independent Gaussian random variable that is relative with transmit symbol s_1 and s_2. As we considered and select the maximum of $E[U_{STD}]$ and minimum of $E[U_{STD}]$, the average BER of BPSK for STD scheme for s_1 can be given as [10], [11]

$$P_{BER,STD} = \frac{1}{2}\left\{Q\left(\sqrt{\frac{2(a+b)}{Var[U_{STD}]}}\right) + Q\left(\sqrt{\frac{2(a-b)}{Var[U_{STD}]}}\right)\right\} \quad (17)$$

where,

$$a = \sum_{i=1}^{N}(\alpha_{i,1}^2+\alpha_{i,2}^2+\alpha_{i,1}\beta_{i,1}\cos(\phi_{i,1}-\theta_{i,1}) + \alpha_{i,2}\beta_{i,2}\cos(\phi_{i,2}-\theta_{i,2}) \quad (18)$$

and

$$b = \sum_{i=1}^{N}(\alpha_{i,2}\beta_{i,1}\cos(\phi_{i,2}-\phi_{i,1}) - \alpha_{i,1}\beta_{i,2}\cos(\phi_{i,1}-\theta_{i,2}) \quad (19)$$

3.3 Independent Rayleigh-Fading Channels

As described in the above section, if the transmit branches are perfect and independent, the average BER's of the diversity schemes can be directly calculated by Eqs. (10) and (17). In practice, if the Rayleigh-fading branches are considered and the branches are also independent, we have to consider the effects of fading. In this case, the average BER's of the diversity schemes can be computed by averaging Eqs. (10) and (17) respectively over the probability distributions for the transform functions.

According to Eqs. (7) and (13), because of the small imperfect channel evaluation in practice, the combined signal, whether it is MRRC or STD schemes, has the approximate formulation. We select MRRC as an example, there is the form of $C\sum_{i=1}^{N}\alpha_i^2$ where C is an constant. Since the Rayleigh-fading channels are considered, that means α_i is a Rayleigh distribution random variable with $E[\alpha_i] = 1.253\sigma_i$ (σ_i^2 is the variance of α_i). If we consider the total noise as a fixed value N_{noise}, then the pdf of SNR_i (is also a random variable), referred to $\gamma_i = \alpha_i^2/N_{noise}$ ($E[\alpha_i^2] = 2\sigma_i^2$)is given as [9], [10]

$$f(\gamma_i) = \frac{1}{\Gamma_i}e^{-\frac{\gamma_i}{\Gamma_i}} \quad (20)$$

where Γ_i is the mean square value of $\alpha_i^2/N_{noise}, i = 1, 2...N$. The derivation of the pdf of total SNR, symbolized as γ is performed in following step. The pdf of γ is obtained by taking the inverse Laplace transform of the product of the Laplace transforms of $f(\gamma_1), f(\gamma_2), ...f(\gamma_N)$. The pdf of γ is given by [10]

$$f(\gamma) = \frac{\Gamma_1^{N-2}}{(\Gamma_1-\Gamma_2)...(\Gamma_1-\Gamma_N)}e^{-\frac{\gamma}{\Gamma_1}} + ... + \frac{\Gamma_N^{N-2}}{(\Gamma_N-\Gamma_1)...(\Gamma_N-\Gamma_{N-1})}e^{-\frac{\gamma}{\Gamma_N}} \quad (21)$$

3.4 Correlated Rayleigh-Fading Channels

In the above section, all the formulations are analyzed assuming that the receive signals are independent Rayleigh-fading signals. In practice, such an assumption is valid only if the diversity antennas are sufficiently separated by a distance, generally larger than a half wavelength. As the size of the hand phone tend to smaller, the separated space available for the diversity antennas mounted on the hand phone is limited so that the receive signals become correlated [11]. The correlation is also affected by the surrounding environments of antenna location. Antennas at the base station can be located on a variety of structures, including rooftops, sides of buildings, sides of water towers and inside offices operating through windows. Each structure will influence the scattering patterns and propagation delay across the antenna array. Then, it is difficult to assume independent Rayleigh-fading channels in actual situations whether the diversity reception scheme is MRRC or STD. Therefore, investigation and quantifying the effects of correlation on diversity receive are of interest for system designers and researchers [11]–[13]. In this section, the transformation technique is introduced to convert two correlated Rayleigh-fading signals into two independent Rayleigh-fading signals, so the performance analysis of diversity scheme with independent Rayleigh-fading signals, formulated in the above section, can be adopted to investigate the diversity scheme with correlated Rayleigh-fading signals [12], [13]. In order to be understood easily and simplification, we select $2 \times N$ STD scheme as a typical example to be formulated.

At first, let's assume that the noise components from receive antennas are independent and the receive signals at different antennas are independent Rayleigh-fading signals. Only the receive signals at the same antenna are correlated Rayleigh-fading signals with correlation coefficient, ρ. We can have the transform functions $g_{i,1}$ and $g_{i,2}$ as

$$g_{i,1} = [X_{i,1} + jY_{i,1}] + n_{i,1}$$
$$g_{i,2} = [X_{i,2} + jY_{i,2}] + n_{i,2} \quad (22)$$

where $X_{i,1}$, $X_{i,2}$, $Y_{i,1}$, $Y_{i,2}$ are zero mean Gaussian random variables with variance of $\sigma_{i,1}^2$ and $\sigma_{i,2}^2$. Here, generally assumed $\sigma_{i,1}^2 = \sigma_{i,2}^2 = \sigma^2$ for simplicity. According the above assumptions, we have [12], [13]

$$E[X_{i,1}Y_{i,1}] = 0, i = 1, 2, ...N$$
$$C[X_{i,1}X_{i,2}] = \rho\sigma^2$$
$$C[Y_{i,1}Y_{i,2}] = \rho\sigma^2 \quad (23)$$

Since the noise components are assumed as independent and uncorrelated with each other, that is

$$C[n_{i,1}n_{i,2}] = C[n_{i,1}X_{i,2}] = C[n_{i,2}Y_{i,2}] = 0 \quad (24)$$

In order to convert the correlated $g_{i,1}$ and $g_{i,2}$ into independent $g'_{i,1}$ and $g'_{i,2}$. Let's use the following transformation matrix Λ as

$$\Lambda = \begin{bmatrix} \frac{\sqrt{2}}{2} & \frac{\sqrt{2}}{2} \\ -\frac{\sqrt{2}}{2} & \frac{\sqrt{2}}{2} \end{bmatrix} \quad (25)$$

to convert $g_{i,1}$ and $g_{i,2}$ by

$$\begin{bmatrix} g'_{i,1} \\ g'_{i,2} \end{bmatrix} = \Lambda \begin{bmatrix} g_{i,1} \\ g_{i,2} \end{bmatrix} \quad (26)$$

Based on Eq. (26), $g'_{i,1}$ and $g'_{i,2}$ can be obtained by

$$g'_{i,1} = X'_{i,1} + jY'_{i,1} + \frac{\sqrt{2}}{2}(n_{i,2} + n_{i,1})$$
$$g'_{i,2} = X'_{i,2} + jY'_{i,2} + \frac{\sqrt{2}}{2}(n_{i,2} - n_{i,1}) \quad (27)$$

where

$$X'_{i,1} = \frac{\sqrt{2}}{2}X_{i,1} + \frac{\sqrt{2}}{2}X_{i,2}$$
$$X'_{i,2} = -\frac{\sqrt{2}}{2}X_{i,1} + \frac{\sqrt{2}}{2}X_{i,2}$$
$$Y'_{i,1} = \frac{\sqrt{2}}{2}Y_{i,1} + \frac{\sqrt{2}}{2}Y_{i,2}$$
$$Y'_{i,2} = -\frac{\sqrt{2}}{2}Y_{i,1} + \frac{\sqrt{2}}{2}Y_{i,2} \quad (28)$$

Since $X_{i,1}$, $X_{i,2}$, $Y_{i,1}$ and $Y_{i,2}$ are Gaussian random variables with zero means, so $X'_{i,1}$, $X'_{i,2}$, $Y'_{i,1}$ and $Y'_{i,2}$ are also Gaussian random variables with zero means. With Eqs. (23) and (28), the correlation between them can be calculated by

$$C[X'_{i,1}X'_{i,2}] = C[Y'_{i,1}Y'_{i,2}] = 0 \quad (29)$$

Therefore, $X'_{i,1}$, $X'_{i,2}$, $Y'_{i,1}$ and $Y'_{i,2}$ are mutually independent Gaussian random variables. From Eq. (28), we have the variances of $g'_{i,1}$ and $g'_{i,2}$ as

$$E[X'_{i,1}] = E[Y'_{i,1}] = \sigma^2(1 + \rho) \quad (30)$$

$$E[X'_{i,2}] = E[Y'_{i,2}] = \sigma^2(1 - \rho) \quad (31)$$

Then based on the above transformation, the independent Rayleigh-fading signals can be converted from correlated Rayleigh-fading signals. All the analysis in Sects. 3.1, 3.2 and 3.3 are used to process the diversity scheme when the diversity reception with correlated Rayleigh-fading signals is considered.

4. Numerical Results and Discussions

For presentation of the numerical results, two system situations are considered in our investigation as: (1) The channel gains and the channel evaluation errors are fixed with values selected randomly (no fading case). The link channels are only considered as Gaussian channels, therefore BER's for MRRC and STD can be calculated by Eqs. (10) and (17), respectively. (2) Channels are considered as Rayleigh-fading channels, then BER's of MRRC and STD can be computed by averaging Eqs. (10) and (17) respectively over the probability distributions for the channel gains. For the two cases, the numerical results are given as follows and compared with these presented in Refs. [9] and [10].

Diversity gain is a function of many parameters, including the modulation scheme and FEC coding. Figures 3, 4 and 5 show the BER's of un-coded BPSK for MRRC and STD in case 1. Figure 3 shows the BER as a function of SNR per bit (dB) in which we randomly selected the channel gains in the diversity. The diversity reception, wether it is MRRC and STD, can combined the each channel gain to improve the average received SNR, therefore BER can be also improved. According to our numerical results, MRRC and STD show the same performance in the perfect channel evaluation. It also shows when in no diversity (1 Tx, 1 Rx), our calculation results is nearly identical to these denoted in Ref. [10].

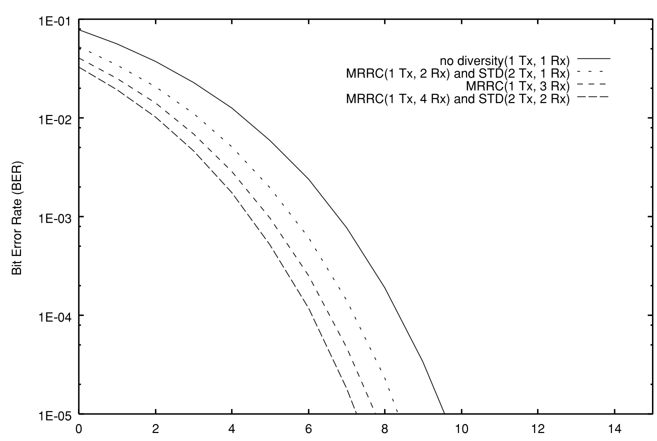

Fig. 3 BER performance of MRRC and STD diversity schemes with perfect channel evaluation in Gaussian channels (no fading case).

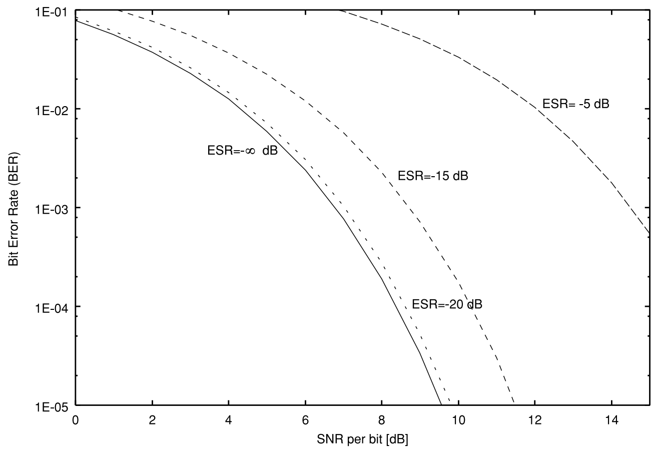

Fig. 4 BER performance in no diversity schemes with perfect/imperfect channel evaluation in Gaussian channel (no fading case).

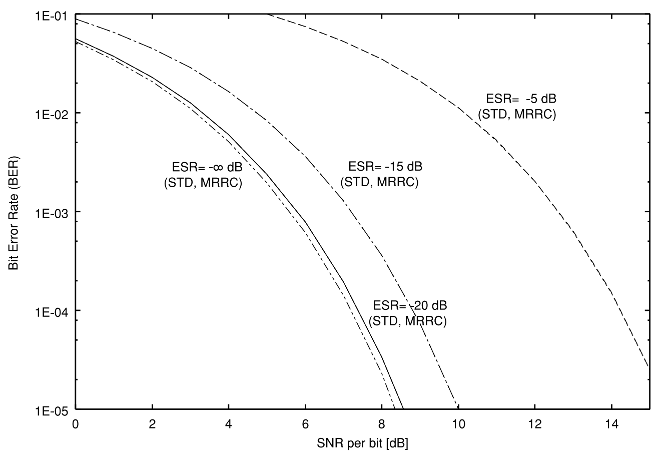

Fig. 5 BER performance of MRRC (1Tx 2Rx) and STD (2Tx 1Rx) diversity schemes with perfect/imperfect channel evaluation in Gaussian channel (no fading case).

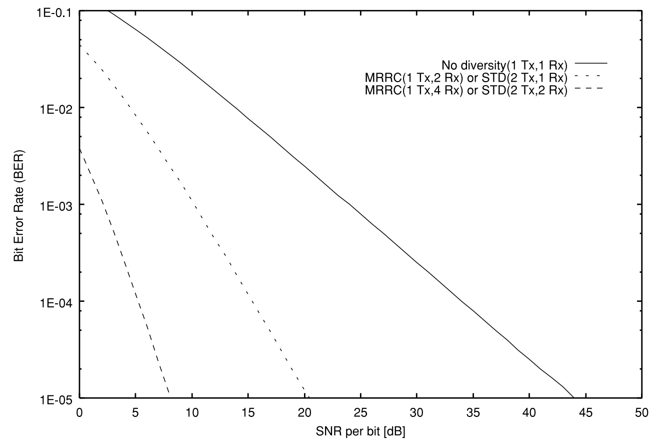

Fig. 6 BER performance of MRRC and STD diversity schemes with perfect channel evaluation in independent Rayleigh-fading channel.

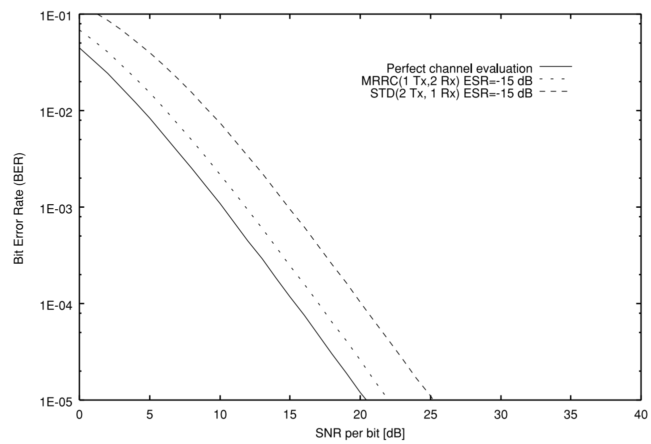

Fig. 7 BER performance of MRRC (1Tx 2Rx) and STD (2Tx 1Rx) diversity schemes with perfect/imperfect channel evaluation in independent Rayleigh-fading channel.

The imperfect channel estimation ESR (dB) is defined as $10\log(\frac{\beta}{\alpha})$ (see Eq. (6) in Sect. 2.2). For simply calculation, the imperfect channel estimations in each signal branch are assumed as the same. Based on the definition, Figs. 4 and 5 show the effects of ESR on the diversity scheme for case 1. BER's of MRRC and STD are plotted as a function of SNR for ESR$=-\infty$ (perfect channel evaluation), -20 dB, -15 dB and -5 dB. We see BER will be increased rapidly with the increase of ESR. On the other hand, the numerical results also show that STD has the same BER as the standard MRRC in case 1 whether the channel evaluations are perfect or imperfect.

Case 2 can really present the diversity reception of wireless transmission channels in which considering the Rayleigh-fading. In order to verify our approach, we investigated the same situation shown in Ref. [9] that BER of MRRC and STD with perfect channel evaluation in independent Rayleigh-fading channel. After compared our calculation results shown in Fig. 6 with the simulation results in Fig. 4 of Ref. [9], our results are approximatively identical each other and the differences between them are below 0.5 dB. The results also show that the diversity reception really improve the quality of signal transmissions and mitigate the multi-path effects, such as Rayleigh-fading [9].

As stated in Ref. [9], there are many factors that may degrade the performance of systems such as mismatched interpolation coefficients and quantization effects. Because of the time variance of the channel, the estimation errors of channels are the dominant reasons to affect the performance. Figure 7 shows the BER performance of MRRC (1 Tx, 2 Rx) and STD (2 Tx, 1 Rx) with ESR$=-15$ dB in the independent Rayleigh-fading channel. According to the results, STD is more sensitive to channel estimation errors than MRRC scheme in Rayleigh-fading channel. BER of STD is near 2.5 dB worse than that of MRRC. Therefore from practical implementation aspects, the system design of MRRC and

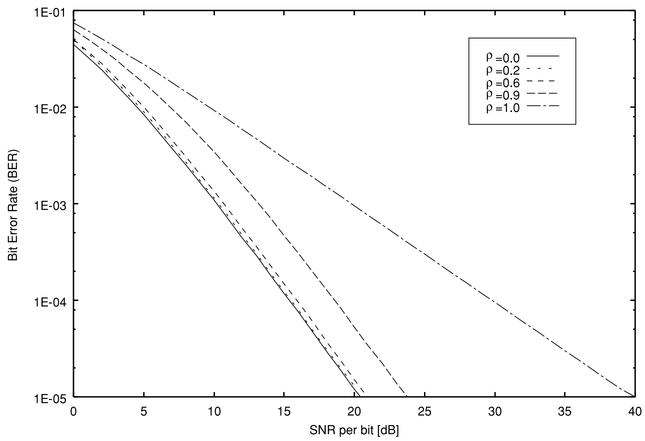

Fig. 8 BER performance of STD (2Tx 1Rx) with perfect channel evaluation in correlated Rayleigh-fading channel (ESR=$-\infty$).

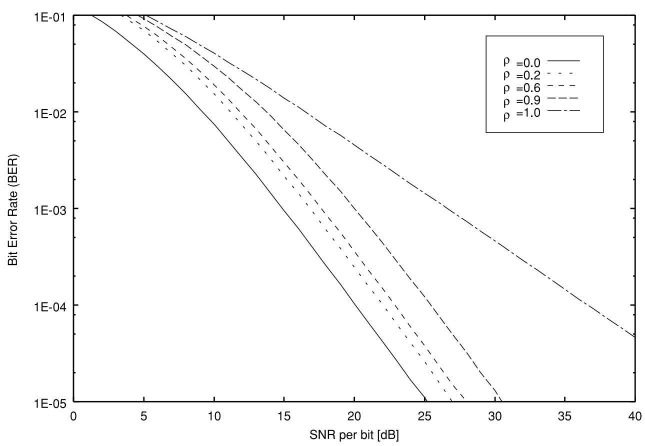

Fig. 9 BER performance of STD (2Tx 1Rx) with imperfect channel evaluation in correlated Rayleigh-fading channel (ESR=-15 dB).

STD schemes may differ according to practical reasons. From this opinion, although STD is an attractive technique for small mobile handsets and for multi-path mitigation compared with MRRC, it becomes necessary to investigate its practical implementation possibility in various shadow, fading channels such as shadow with log-normal distribution, Rican fading or Nakagami fading channels that will become our future works.

Figures 8 and 9 show the numerical results of STD diversity scheme with correlated Rayleigh-fading signals. Especially when the channel estimation is imperfect (ESR=-15 dB), the results show that the STD scheme's sensitivity is more worse when the receive signals are correlated with the parameter ρ. We can conclude that STD diversity scheme with correlated Rayleigh-fading signals can improve the BER performance when ρ is smaller, especially the channel estimations are the perfect at the same time. With the increase of the correlation coefficient ρ, the degradation rate of BER performance is increased because of the correlation.

5. Conclusions

A recently proposed STD scheme in Ref. [9] allows implementation of diversity reception with multiple antennas at the receiver that is really attractive technique for small mobile handsets and for multi-path mitigation. An theoretical approach was proposed to investigate the BER's performance of MRRC and STD with perfect/Imperfect channel evaluation. The numerical results show that STD has the same BER as the standard MRRC whether the channel evaluations are perfect or imperfect in no fading case, but in Rayleigh-fading case, STD scheme is more sensitive to the channel estimation errors than that of MRRC, near 2.5 dB worse. Because of this reason, the practical implementation of the STD scheme should be carefully considered and it is necessary to investigate its performances in various wireless mediums such as other kinds of shadow with log-normal distribution, Rican fading or Nakagami fading channels.

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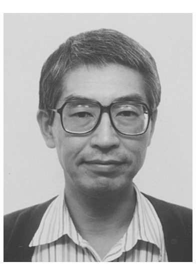

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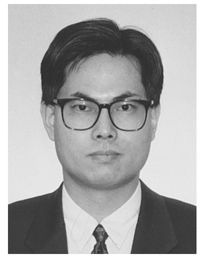

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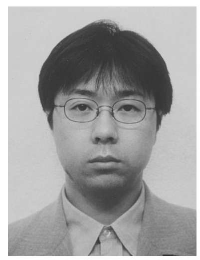

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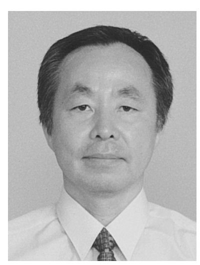

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