A Simple Transmit Diversity Technique for Wireless Communications

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***Abstract*— This paper presents a simple two-branch transmit diversity scheme. Using two transmit antennas and one receive antenna the scheme provides the same diversity order as maximal-ratio receiver combining (MRRC) with one transmit antenna, and two receive antennas. It is also shown that the scheme may easily be generalized to two transmit antennas and M receive antennas to provide a diversity order of $2M$. The new scheme does not require any bandwidth expansion any feedback from the receiver to the transmitter and its computation complexity is similar to MRRC.**

***Index Terms*—Antenna array processing, baseband processing, diversity, estimation and detection, fade mitigation, maximal-ratio combining, Rayleigh fading, smart antennas, space block coding, space–time coding, transmit diversity, wireless communications.**

I. Introduction

THE NEXT-generation wireless systems are required to have high voice quality as compared to current cellular mobile radio standards and provide high bit rate data services (up to 2 Mbits/s). At the same time, the remote units are supposed to be small lightweight pocket communicators. Furthermore, they are to operate reliably in different types of environments: macro, micro, and picocellular; urban, suburban, and rural; indoor and outdoor. In other words, the next generation systems are supposed to have better quality and coverage, be more power and bandwidth efficient, and be deployed in diverse environments. Yet the services must remain affordable for widespread market acceptance. Inevitably, the new pocket communicators must remain relatively simple. Fortunately, however, the economy of scale may allow more complex base stations. In fact, it appears that base station complexity may be the only plausible trade space for achieving the requirements of next generation wireless systems.

The fundamental phenomenon which makes reliable wireless transmission difficult is time-varying multipath fading [1]. It is this phenomenon which makes tetherless transmission a challenge when compared to fiber, coaxial cable, line-of-sight microwave or even satellite transmissions.

Increasing the quality or reducing the effective error rate in a multipath fading channel is extremely difficult. In additive white Gaussian noise (AWGN), using typical modulation and coding schemes, reducing the effective bit error rate (BER) from 10^{-2} to 10^{-3} may require only 1- or 2-dB higher signal-to-noise ratio (SNR). Achieving the same in a multipath fading environment, however, may require up to 10 dB improvement in SNR. The improvement in SNR may not be achieved by higher transmit power or additional bandwidth, as it is contrary to the requirements of next generation systems. It is therefore crucial to effectively combat or reduce the effect of fading at both the remote units and the base stations, without additional power or any sacrifice in bandwidth.

Theoretically, the most effective technique to mitigate multipath fading in a wireless channel is transmitter power control. If channel conditions as experienced by the receiver on one side of the link are known at the transmitter on the other side, the transmitter can predistort the signal in order to overcome the effect of the channel at the receiver. There are two fundamental problems with this approach. The major problem is the required transmitter dynamic range. For the transmitter to overcome a certain level of fading, it must increase its power by that same level, which in most cases is not practical because of radiation power limitations and the size and cost of the amplifiers. The second problem is that the transmitter does not have any knowledge of the channel experienced by the receiver except in systems where the uplink (remote to base) and downlink (base to remote) transmissions are carried over the same frequency. Hence, the channel information has to be fed back from the receiver to the transmitter, which results in throughput degradation and considerable added complexity to both the transmitter and the receiver. Moreover, in some applications there may not be a link to feed back the channel information.

Other effective techniques are time and frequency diversity. Time interleaving, together with error correction coding, can provide diversity improvement. The same holds for spread spectrum. However, time interleaving results in large delays when the channel is slowly varying. Equivalently, spread spectrum techniques are ineffective when the coherence bandwidth of the channel is larger than the spreading bandwidth or, equivalently, where there is relatively small delay spread in the channel.

In most scattering environments, antenna diversity is a practical, effective and, hence, a widely applied technique for reducing the effect of multipath fading [1]. The classical approach is to use multiple antennas at the receiver and perform combining or selection and switching in order to improve the quality of the received signal. The major problem with using the receive diversity approach is the cost, size, and power of the remote units. The use of multiple antennas and radio frequency (RF) chains (or selection and switching circuits) makes the remote units larger and more expensive. As a result, diversity techniques have almost exclusively been

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applied to base stations to improve their reception quality. A base station often serves hundreds to thousands of remote units. It is therefore more economical to add equipment to base stations rather than the remote units. For this reason, transmit diversity schemes are very attractive. For instance, one antenna and one transmit chain may be added to a base station to improve the reception quality of all the remote units in that base station's coverage area.[1] The alternative is to add more antennas and receivers to all the remote units. The first solution is definitely more economical.

Recently, some interesting approaches for transmit diversity have been suggested. A delay diversity scheme was proposed by Wittneben [2], [3] for base station simulcasting and later, independently, a similar scheme was suggested by Seshadri and Winters [4], [5] for a single base station in which copies of the same symbol are transmitted through multiple antennas at different times, hence creating an artificial multipath distortion. A maximum likelihood sequence estimator (MLSE) or a minimum mean squared error (MMSE) equalizer is then used to resolve multipath distortion and obtain diversity gain. Another interesting approach is space–time trellis coding, introduced in [6], where symbols are encoded according to the antennas through which they are simultaneously transmitted and are decoded using a maximum likelihood decoder. This scheme is very effective, as it combines the benefits of forward error correction (FEC) coding and diversity transmission to provide considerable performance gains. The cost for this scheme is additional processing, which increases exponentially as a function of bandwidth efficiency (bits/s/Hz) and the required diversity order. Therefore, for some applications it may not be practical or cost-effective.

The technique proposed in this paper is a simple transmit diversity scheme which improves the signal quality at the receiver on one side of the link by simple processing across two transmit antennas on the opposite side. The obtained diversity order is equal to applying maximal-ratio receiver combining (MRRC) with two antennas at the receiver. The scheme may easily be generalized to two transmit antennas and M receive antennas to provide a diversity order of $2M$. This is done without any feedback from the receiver to the transmitter and with small computation complexity. The scheme requires no bandwidth expansion, as redundancy is applied in space across multiple antennas, not in time or frequency.

The new transmit diversity scheme can improve the error performance, data rate, or capacity of wireless communications systems. The decreased sensitivity to fading may allow the use of higher level modulation schemes to increase the effective data rate, or smaller reuse factors in a multicell environment to increase system capacity. The scheme may also be used to increase the range or the coverage area of wireless systems. In other words, the new scheme is effective in all of the applications where system capacity is limited by multipath fading and, hence, may be a simple and cost-effective way to address the market demands for quality and efficiency without a complete redesign of existing systems. Furthermore, the scheme seems to be a superb candidate for next-generation wireless systems, as it effectively reduces the effect of fading at the remote units using multiple transmit antennas at the base stations.

[1] In fact, many cellular base stations already have two receive antennas for receive diversity. The same antennas may be used for transmit diversity.

In Section II, the classical maximal ratio receive diversity combining is discussed and simple mathematical descriptions are given. In Section III, the new two-branch transmit diversity schemes with one and with two receive antennas are discussed. In Section IV, the bit-error performance of the new scheme with coherent binary phase-shift keying (BPSK) modulation is presented and is compared with MRRC. There are cost and performance differences between the practical implementations of the proposed scheme and the classical MRRC. These differences are discussed in detail in Section V.

II. Classical Maximal-Ratio Receive Combining (MRRC) Scheme

Fig. 1 shows the baseband representation of the classical two-branch MRRC.

At a given time, a signal s_0 is sent from the transmitter. The channel including the effects of the transmit chain, the airlink, and the receive chain may be modeled by a complex multiplicative distortion composed of a magnitude response and a phase response. The channel between the transmit antenna and the receive antenna zero is denoted by $\boldsymbol{h_0}$ and between the transmit antenna and the receive antenna one is denoted by $\boldsymbol{h_1}$ where

$$\begin{aligned}\boldsymbol{h_0} &= \alpha_0 e^{j\theta_0}\\ \boldsymbol{h_1} &= \alpha_1 e^{j\theta_1}.\end{aligned}\tag{1}$$

Noise and interference are added at the two receivers. The resulting received baseband signals are

$$\begin{aligned}\boldsymbol{r_0} &= \boldsymbol{h_0}\boldsymbol{s_0} + \boldsymbol{n_0}\\ \boldsymbol{r_1} &= \boldsymbol{h_1}\boldsymbol{s_0} + \boldsymbol{n_1}\end{aligned}\tag{2}$$

where $\boldsymbol{n_0}$ and $\boldsymbol{n_1}$ represent complex noise and interference.

Assuming $\boldsymbol{n_0}$ and $\boldsymbol{n_1}$ are Gaussian distributed, the maximum likelihood decision rule at the receiver for these received signals is to choose signal $\boldsymbol{s_i}$ if and only if (iff)

$$\begin{aligned}&d^2(\boldsymbol{r_0}, \boldsymbol{h_0}\boldsymbol{s_i}) + d^2(\boldsymbol{r_1}, \boldsymbol{h_1}\boldsymbol{s_i}) \le d^2(\boldsymbol{r_0}, \boldsymbol{h_0}\boldsymbol{s_k})\\ &\quad + d^2(\boldsymbol{r_1}, \boldsymbol{h_1}\boldsymbol{s_k}), \qquad \forall \boldsymbol{i} \ne \boldsymbol{k}\end{aligned}\tag{3}$$

where $d^2(\mathbf{x}, \mathbf{y})$ is the squared Euclidean distance between signals $\boldsymbol{x}$ and $\boldsymbol{y}$ calculated by the following expression:

$$d^2(\boldsymbol{x}, \boldsymbol{y}) = (\boldsymbol{x} - \boldsymbol{y})(\boldsymbol{x}^* - \boldsymbol{y}^*).\tag{4}$$

The receiver combining scheme for two-branch MRRC is as follows:

$$\begin{aligned}\tilde{\boldsymbol{s}}_0 &= \boldsymbol{h}_0^*\boldsymbol{r_0} + \boldsymbol{h}_1^*\boldsymbol{r_1}\\ &= \boldsymbol{h}_0^*(\boldsymbol{h_0}\boldsymbol{s_0} + \boldsymbol{n_0}) + \boldsymbol{h}_1^*(\boldsymbol{h_1}\boldsymbol{s_0} + \boldsymbol{n_1})\\ &= (\alpha_0^2 + \alpha_1^2)\boldsymbol{s_0} + \boldsymbol{h}_0^*\boldsymbol{n_0} + \boldsymbol{h}_1^*\boldsymbol{n_1}.\end{aligned}\tag{5}$$

Expanding (3) and using (4) and (5) we get

choose $\boldsymbol{s_i}$ iff

$$\begin{aligned}&(\alpha_0^2 + \alpha_1^2)|\boldsymbol{s_i}|^2 - \tilde{\boldsymbol{s}}_0\boldsymbol{s}_i^* - \tilde{\boldsymbol{s}}_0^*\boldsymbol{s_i}\\ &\quad \le (\alpha_0^2 + \alpha_1^2)|\boldsymbol{s_k}|^2 - \tilde{\boldsymbol{s}}_0\boldsymbol{s}_k^* - \tilde{\boldsymbol{s}}_0^*\boldsymbol{s_k}, \qquad \forall \boldsymbol{i} \ne \boldsymbol{k}\end{aligned}\tag{6}$$

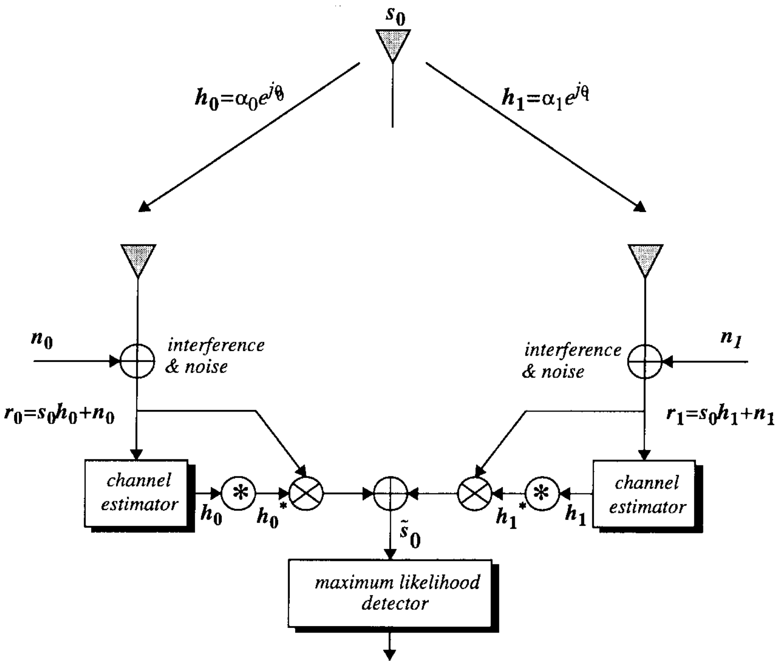

Fig. 1. Two-branch MRRC.

or equivalently

choose $\boldsymbol{s_i}$ iff

$$\begin{aligned}&(\alpha_0^2+\alpha_1^2-1)|\boldsymbol{s_i}|^2+d^2(\tilde{\boldsymbol{s}}_0,\boldsymbol{s_i})\\ &\quad\le(\alpha_0^2+\alpha_1^2-1)|\boldsymbol{s_k}|^2+d^2(\tilde{\boldsymbol{s}}_0,\boldsymbol{s_k}),\qquad\forall \boldsymbol{i}\neq\boldsymbol{k}.\end{aligned}\tag{7}$$

For PSK signals (equal energy constellations)

$$|\boldsymbol{s_i}|^2=|\boldsymbol{s_k}|^2=E_s,\qquad\forall\,\boldsymbol{i},\boldsymbol{k}\tag{8}$$

where E_s is the energy of the signal. Therefore, for PSK signals, the decision rule in (7) may be simplified to

choose $\boldsymbol{s_i}$ iff

$$d^2(\tilde{\boldsymbol{s}}_0,\boldsymbol{s_i})\le d^2(\tilde{\boldsymbol{s}}_0,\boldsymbol{s_k}),\qquad\forall\boldsymbol{i}\neq\boldsymbol{k}.\tag{9}$$

The maximal-ratio combiner may then construct the signal $\tilde{\boldsymbol{s}}_0$, as shown in Fig. 1, so that the maximum likelihood detector may produce $\hat{\boldsymbol{s}}_0$, which is a maximum likelihood estimate of $\boldsymbol{s}_0$.

III. The New Transmit Diversity Scheme

A. Two-Branch Transmit Diversity with One Receiver

Fig. 2 shows the baseband representation of the new two-branch transmit diversity scheme.

The scheme uses two transmit antennas and one receive antenna and may be defined by the following three functions:

- the encoding and transmission sequence of information symbols at the transmitter;
- the combining scheme at the receiver;
- the decision rule for maximum likelihood detection.

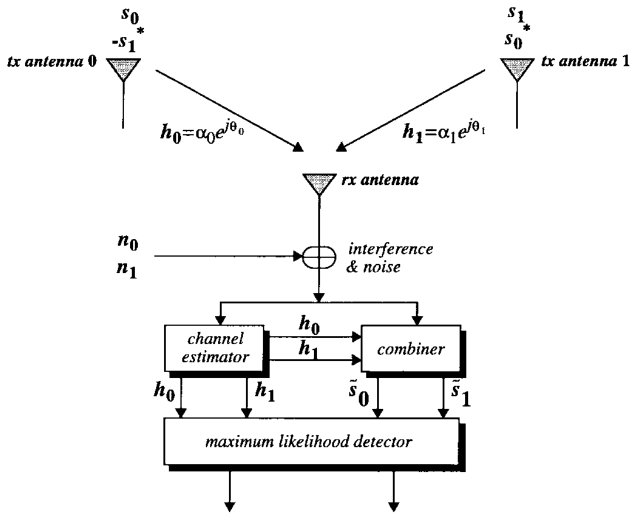

Fig. 2. The new two-branch transmit diversity scheme with one receiver.

1) The Encoding and Transmission Sequence: At a given symbol period, two signals are simultaneously transmitted from the two antennas. The signal transmitted from antenna zero is denoted by $\boldsymbol{s}_0$ and from antenna one by $\boldsymbol{s}_1$. During the next symbol period signal $(-\boldsymbol{s}_1^*)$ is transmitted from antenna zero, and signal $\boldsymbol{s}_0^*$ is transmitted from antenna one where $*$ is the complex conjugate operation. This sequence is shown in Table I.

In Table I, the encoding is done in space and time (space–time coding). The encoding, however, may also be done in space and frequency. Instead of two adjacent symbol periods, two adjacent carriers may be used (space–frequency coding).

TABLE I
THE ENCODING AND TRANSMISSION SEQUENCE FOR THE TWO-BRANCH TRANSMIT DIVERSITY SCHEME

	antenna 0	antenna 1
time t	s_0	s_1
time $t+T$	$-s_1^*$	s_0^*

The channel at time t may be modeled by a complex multiplicative distortion $h_0(t)$ for transmit antenna zero and $h_1(t)$ for transmit antenna one. Assuming that fading is constant across two consecutive symbols, we can write

$$\begin{aligned} h_0(t) &= h_0(t+T) = h_0 = \alpha_0 e^{j\theta_0} \\ h_1(t) &= h_1(t+T) = h_1 = \alpha_1 e^{j\theta_1} \end{aligned} \tag{10}$$

where T is the symbol duration. The received signals can then be expressed as

$$\begin{aligned} r_0 &= r(t) = h_0 s_0 + h_1 s_1 + n_0 \\ r_1 &= r(t+T) = -h_0 s_1^* + h_1 s_0^* + n_1 \end{aligned} \tag{11}$$

where r_0 and r_1 are the received signals at time t and $t+T$ and n_0 and n_1 are complex random variables representing receiver noise and interference.

2) The Combining Scheme: The combiner shown in Fig. 2 builds the following two combined signals that are sent to the maximum likelihood detector:

$$\begin{aligned} \tilde{s}_0 &= h_0^* r_0 + h_1 r_1^* \\ \tilde{s}_1 &= h_1^* r_0 - h_0 r_1^*. \end{aligned} \tag{12}$$

It is important to note that this combining scheme is different from the MRRC in (5). Substituting (10) and (11) into (12) we get

$$\begin{aligned} \tilde{s}_0 &= (\alpha_0^2 + \alpha_1^2) s_0 + h_0^* n_0 + h_1 n_1^* \\ \tilde{s}_1 &= (\alpha_0^2 + \alpha_1^2) s_1 - h_0 n_1^* + h*_1 n_0. \end{aligned} \tag{13}$$

3) The Maximum Likelihood Decision Rule: These combined signals are then sent to the maximum likelihood detector which, for each of the signals s_0 and s_1, uses the decision rule expressed in (7) or (9) for PSK signals.

The resulting combined signals in (13) are equivalent to that obtained from two-branch MRRC in (5). The only difference is phase rotations on the noise components which do not degrade the effective SNR. Therefore, the resulting diversity order from the new two-branch transmit diversity scheme with one receiver is equal to that of two-branch MRRC.

B. Two-Branch Transmit Diversity with M Receivers

There may be applications where a higher order of diversity is needed and multiple receive antennas at the remote units are feasible. In such cases, it is possible to provide a diversity order of $2M$ with two transmit and M receive antennas. For illustration, we discuss the special case of two transmit and two receive antennas in detail. The generalization to M receive antennas is trivial.

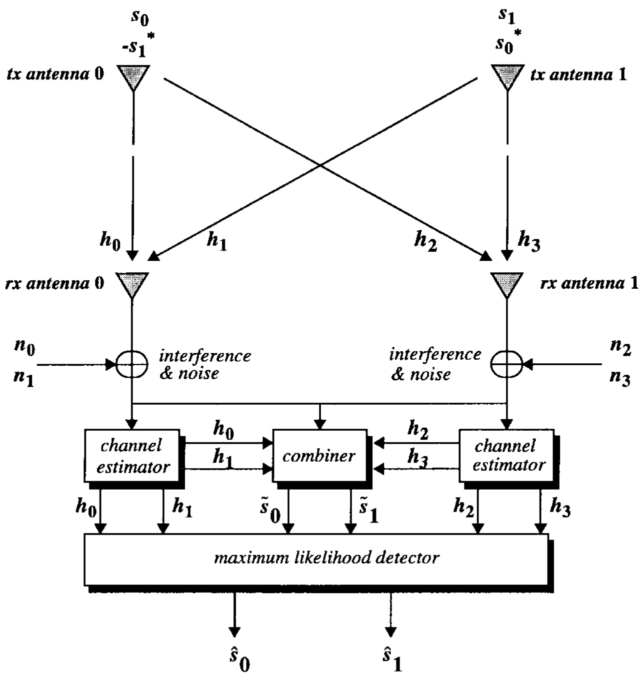

Fig. 3. The new two-branch transmit diversity scheme with two receivers.

TABLE II
THE DEFINITION OF CHANNELS BETWEEN THE TRANSMIT AND RECEIVE ANTENNAS

	rx antenna 0	rx antenna 1
tx antenna 0	h_0	h_2
tx antenna 1	h_1	h_3

TABLE III
THE NOTATION FOR THE RECEIVED SIGNALS AT THE TWO RECEIVE ANTENNAS

	rx antenna 0	rx antenna 1
time t	r_0	r_2
time $t+T$	r_1	r_3

Fig. 3 shows the baseband representation of the new scheme with two transmit and two receive antennas.

The encoding and transmission sequence of the information symbols for this configuration is identical to the case of a single receiver, shown in Table I. Table II defines the channels between the transmit and receive antennas, and Table III defines the notation for the received signal at the two receive antennas.

Where

$$\begin{aligned} r_0 &= h_0 s_0 + h_1 s_1 + n_0 \\ r_1 &= -h_0 s_1^* + h_1 s_0^* + n_1 \\ r_2 &= h_2 s_0 + h_3 s_1 + n_2 \\ r_3 &= -h_2 s_1^* + h_3 s_0^* + n_3 \end{aligned} \tag{14}$$

n_0, n_1, n_2, and n_3 are complex random variables representing receiver thermal noise and interference. The combiner in Fig. 3 builds the following two signals that are sent to the maximum

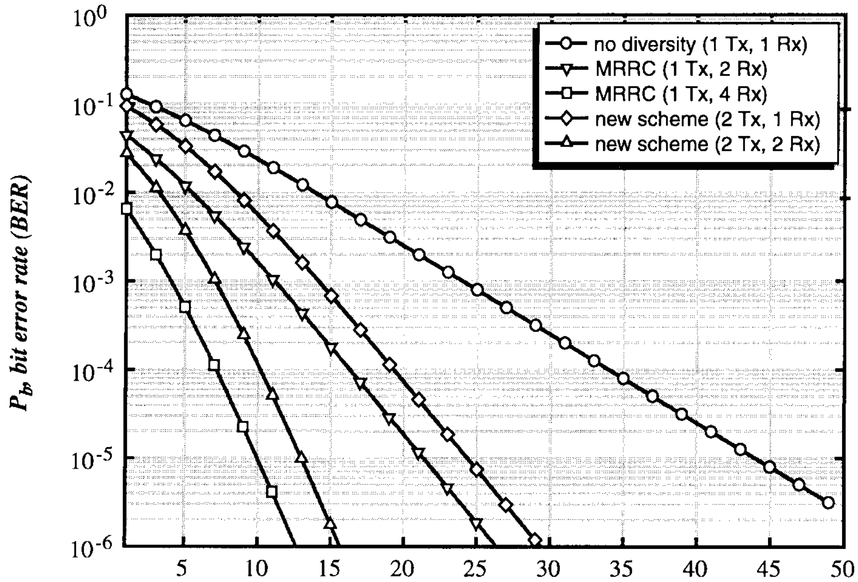

Fig. 4. The BER performance comparison of coherent BPSK with MRRC and two-branch transmit diversity in Rayleigh fading.

likelihood detector:

$$\begin{aligned}\tilde{s}_0 &= h_0^* r_0 + h_1 r_1^* + h_2^* r_2 + h_3 r_3^* \\ \tilde{s}_1 &= h_1^* r_0 - h_0 r_1^* + h_3^* r_2 - h_2 r_3^*. \end{aligned} \tag{15}$$

Substituting the appropriate equations we have

$$\begin{aligned}\tilde{s}_0 &= (\alpha_0^2 + \alpha_1^2 + \alpha_2^2 + \alpha_3^2) s_0 + h_0^* n_0 + h_1 n_1^* \\ &\quad + h_2^* n_2 + h_3 n_3^* \\ \tilde{s}_1 &= (\alpha_0^2 + \alpha_1^2 + \alpha_2^2 + \alpha_3^2) s_1 - h_0 n_1^* + h_1^* n_0 \\ &\quad - h_2 n_3^* + h_3^* n_2. \end{aligned} \tag{16}$$

These combined signals are then sent to the maximum likelihood decoder which for signal s_0 uses the decision criteria expressed in (17) or (18) for PSK signals.

Choose s_i iff

$$\begin{aligned}&(\alpha_0^2 + \alpha_1^2 + \alpha_2^2 + \alpha_3^2 - 1)|s_i|^2 + d^2(\tilde{s}_0, s_i) \\ &\quad \le (\alpha_0^2 + \alpha_1^2 + \alpha_2^2 + \alpha_3^2 - 1|s_k|^2 + d^2(\tilde{s}_0, s_k). \end{aligned} \tag{17}$$

Choose s_i iff

$$d^2(\tilde{s}_0, s_i) \le d^2(\tilde{s}_0, s_k), \qquad \forall i \ne k. \tag{18}$$

Similarly, for s_1, using the decision rule is to choose signal s_i iff

$$\begin{aligned}&(\alpha_0^2 + \alpha_1^2 + \alpha_2^2 + \alpha_3^2 - 1)|s_i|^2 + d^2(\tilde{s}_1, s_i) \\ &\quad \le (\alpha_0^2 + \alpha_1^2 + \alpha_2^2 + \alpha_3^2 - 1)|s_k|^2 + d^2(\tilde{s}_1, s_k) \end{aligned} \tag{19}$$

or, for PSK signals,

choose s_i iff

$$d^2(\tilde{s}_1, s_i) \le d^2(\tilde{s}_1, s_k), \qquad \forall i \ne k. \tag{20}$$

The combined signals in (16) are equivalent to that of four-branch MRRC, not shown in the paper. Therefore, the resulting diversity order from the new two-branch transmit diversity scheme with two receivers is equal to that of the four-branch MRRC scheme.

It is interesting to note that the combined signals from the two receive antennas are the simple addition of the combined signals from each receive antenna, i.e., the combining scheme is identical to the case with a single receive antenna. We may hence conclude that, using two transmit and M receive antennas, we can use the combiner for each receive antenna and then simply add the combined signals from all the receive antennas to obtain the same diversity order as $2M$-branch MRRC. In other words, using two antennas at the transmitter, the scheme doubles the diversity order of systems with one transmit and multiple receive antennas.

An interesting configuration may be to employ two antennas at each side of the link, with a transmitter and receiver chain connected to each antenna to obtain a diversity order of four at both sides of the link.

IV. Error Performance Simulations

The diversity gain is a function of many parameters, including the modulation scheme and FEC coding. Fig. 4 shows the BER performance of uncoded coherent BPSK for MRRC and the new transmit diversity scheme in Rayleigh fading.

It is assumed that the total transmit power from the two antennas for the new scheme is the same as the transmit power from the single transmit antenna for MRRC. It is also assumed that the amplitudes of fading from each transmit antenna to each receive antenna are mutually uncorrelated Rayleigh distributed and that the average signal powers at each receive antenna from each transmit antenna are the same. Further, we assume that the receiver has perfect knowledge of the channel.

Although the assumptions in the simulations may seem highly unrealistic, they provide reference performance curves for comparison with known techniques. An important issue is

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