Two Signaling Schemes for Improving the Error Performance of Frequency-Division-Duplex (FDD) Transmission Systems Using Transmitter Antenna Diversity

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Abstract

We propose two signaling schemes that exploit the availability of multiple (N) antennas at the transmitter to provide diversity benefit to the receiver. In the first scheme, a channel code of length N and minimum Hamming distance $d_{\min} \leq N$ is used to encode a group of K information bits. Channel code symbol c_i is transmitted with the i^{th} antenna. At the receiver, a maximum likelihood decoder for the channel code provides a diversity of $d_{\min}$ as long as each transmitted code symbol is subjected to independent fading. The second scheme introduces deliberate resolvable multi-path distortion by transmitting the data bearing signal with antenna 1, and $N-1$ delayed versions of it with antennas 2 through N. The delays are unique to each antenna and are chosen to be multiples of the symbol interval. At the receiver, a maximum likelihood sequence estimator resolves the multi-path in an optimal manner to realize a diversity benefit of N.

1. Introduction

Here we propose two techniques for providing diversity benefit to a mobile using base station antenna array for a FDD (or TDD with fast channel variations) based approach to transmission. Ideally, we would like to maintain the transmission efficiency to be at the same value as that of the baseline modulation scheme (that gets no diversity benefit). This is achieved in the first technique by using coded modulation schemes and in the second technique by creating intentional multi-path distortion. In the former situation the code symbols of a codeword are transmitted in an orthogonal manner using different antennas. The received codeword is decoded using the maximum likelihood receiver. In the latter situation, a maximum likelihood sequence estimator (MLSE) or a minimum mean squared error (MMSE) equalizer is used to completely resolve the multi-path distortion. It has been shown through simulations and analysis [1,2] that MLSE and MMSE based receivers provide diversity gain. The second scheme is backward compatible with IS-54 (North America) and GSM (Europe) digital cellular systems. In both cases, the maximum diversity benefit is upper bounded by the number of antenna elements at the base. At the completion of this work, it was brought to the attention of the authors that a similar scheme has been proposed in [3] using multiple base-station simulcasting. This scheme uses different FIR filters at each base-station to filter the modulation symbols before transmission. The FIR filters are chosen so that diversity benefit is obtained at the receiver. The technique proposed here is a special case of [3] with appropriate choice of FIR filters. Reference [3] considers only the MLSE scheme with background noise being white Gaussian. Here, we consider the performance of MMSE as well as MLSE equalizers.

2. Diversity Benefit Using Channel Coding

In order to provide different copies of the transmitted signal at the receiver, the proposed scheme transmits the same signal at different instants of time. At each instant a different antenna is used for transmission. The receiver now gets N different copies of the transmitted signal where each copy is subject to a fade that is statistically independent from those of the other copies. These copies are then combined to obtain the diversity advantage. It can be noticed that what is being proposed is a repetition code whose length is equal to the number of antenna elements. However the transmission (bandwidth) efficiency goes down by a factor of N. In order to increase the bandwidth and power efficiency, we propose to use combined modulation and coding techniques. This is by now well understood for fading channels [4] and these codes can be directly applied here.

2.1. Channel Model

We suppose that the overall channel is made up of N channels, each undergoing independent slow (static) Rayleigh fading. The channel impulse response for the i^{th} channel is given by

$$h_i(t) = z_i \delta(t)\, e^{j w_o t}, \quad 1 \leq i \leq N, \tag{1}$$

where w_o is the angular carrier frequency and z_i is the static complex fade value whose phase is a random variable that is uniformly distributed over $(-\pi, \pi)$ and whose magnitude is Rayleigh distributed. The transmitted signal from the i^{th} antenna is given by

$$\tilde{s}_i(t) = \text{Re}\, (s_i(t)\, e^{j w_o t}) , \tag{2}$$

where

$$s_i(t) = \sum_n c_{in}\, p(t - nT) . \tag{3}$$

c_{in} is the i^{th} complex (M-ary data) symbol in a codeword $\mathbf{c}_n$ which is generated by encoding the n^{th} group of k information bits into N channel symbols using a channel code with a minimum Hamming distance $d_{\min} \leq N$, $\mathbf{c}_n = (c_{1n}, c_{2n}, \ldots, c_{nN})$, $N \geq 2$. The transmitter impulse response is $p(t)$.

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The received signal corresponding to the i^{th} transmitted signal is

$$r_i(t) = \mathrm{Re}\left(s_i(t) * h_i(t)\right) + \mathrm{Re}\left(n_i(t)\ e^{jw_o t}\right) \qquad (4)$$

where $*$ denotes the convolution operator and $n_i(t)\ e^{jw_o t}$ is additive co-channel interference (modeled as white Gaussian) plus any other source of noise which is also modeled as white Gaussian.

Assuming ideal coherent demodulation, for the i^{th} transmitted signal, the output of the receive filter (matched to the baseband square root Nyquist pulse $p(t)$) sampled at the ideal timing instant $t = jT$ is given by

$$r_i(jT) = \sum |z_i| c_{ij} + w_{ij}\ . \qquad (5)$$

The maximum likelihood decoder for recovering the j^{th} group of information bits forms the decision statistic

$$\Lambda(\hat{c}) = \sum_k |z_k| c_{kj} \hat{c}_{kj} + \sum_k w_{kj} c_{kj} \qquad (6)$$

where $\hat{c}$ is one of the channel codewords that could have been transmitted. The correlation is performed for every possible codeword.

The codeword $\hat{c}$ with the highest correlation is chosen as the transmitted codeword which in turn yields the decoded j^{th} group of k information bits.

2.2. Code Constructions (No. of antennas = 2)

The codes constructed in this subsection assume that the number of antennas at the base is equal to 2. The baseline scheme is uncoded 4-PSK for the purpose of error rate comparison.

Example 1 (1 bit per symbol): A repetition code of length 2, using the QPSK symbols is used to realize a rate $R = 1$ code. Symbol 1 is transmitted with antenna 1 and symbol 2 with antenna 2. The channel is assumed to be a static time-selective fading channel. After demodulation, the demodulated 4 dimensional signal is correlated with each of the four codewords and the one with the highest correlation is the decoded data. The time diversity of this code is 2.

Example 2 (1.5 bits/symbol): This code of length 2, $d_{\min} = 2$ and product distance 2 is formed from a natural binary encoded 8-PSK constellation. The code consists of 8 4-dimensional codewords $\mathbf{C} = \{(0,0), (1,5), (2,2), (3,7), (4,4), (5,1), (6,6), (7,3)\}$. A distinct pair of codewords differ in at least two positions and the minimum product distance is 2 and is the product of (for example) $d^2(0,1)$ and $d^2(0,5)$. Three information bits are conveyed over two intervals and hence the rate is 1.5 bits/symbol.

Example 3 (2.0 bits/symbol): In order to achieve $d_{\min} = 2$ and stay with the constraint that the block length of the code= 2, it is necessary to have at least 16 codewords. Hence, 16-PSK is the smallest constellation with which we can get diversity benefit of 2 and maintain the bandwidth efficiency. In general, with N antennas and with $d_{\min} = N$, in order to maintain bandwidth efficiency, the minimum constellation expansion factor is 2^N. The 4D-16 PSK code is $\mathbf{C} = \{(0,0), (2,2), (4,4), (6,6), (8,8), (10,10), (12,12), (14,14), (1,7), (3,9), (5,11), (7,13), (9,15), (11,1), (13,3), (15,5)\}$.

Some performance improvement can be obtained by Gray coding the information data so that the 4-D signal points that are separated by a higher product distance are also subjected to a higher number of information bit errors. The minimum product distance is $(0.587)^2$.

2.3. Simulation Results

Figure 1 shows the performance of the codes that were constructed in section 2.3 on a time-selective Rayleigh fading channel. The channel is assumed to be static over the duration of a burst which is 200 code symbols. A new fade value is generated for each of the two channels for every burst. It can be seen from figure 1 that at a bit error rate of 10^{-3}, about 8 dB gain in SNR is obtained at a bit rate of 2 bits/symbol over uncoded 4-PSK. This gain is smaller than if one were to use diversity reception at the receiver or have perfect adaptive retransmission. Thus, a penalty is paid for an open-loop transmission versus an ideal closed-loop transmission. By backing off on the rate, the loss in diversity gain can be recovered. At 1.5 bits/symbol, the gain is 12 dB, and at 1 bit/symbol, the gain is about 16 dB over an uncoded 4-PSK system. The large expansion in the signal constellation size in order to maintain the bandwidth efficiency makes this scheme somewhat inefficient as the number of antennas at the base becomes large.

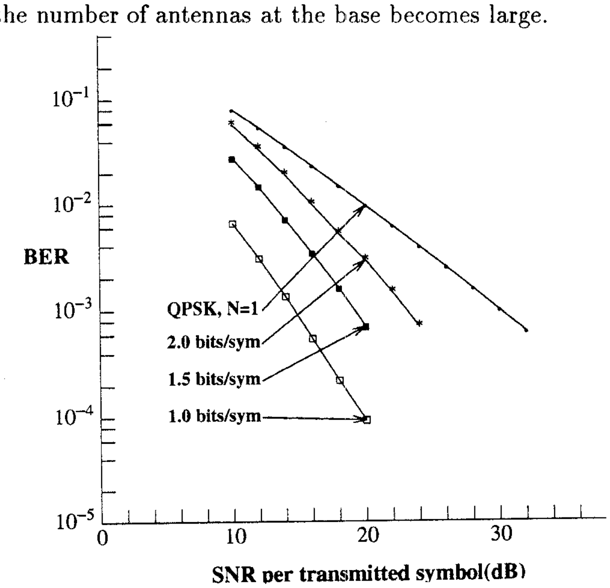

Figure 1

Simulated BER for three 4-dimensional Block Coded PSK Modulation schemes on a Quasi-Static Fading Channel with the proposed approach using 2 antennas.

3. Diversity Benefit Using Intentional Multi-Path Distortion

In scheme 1, example 1, the N channel symbols of the repetition code were transmitted from N different antennas at non-overlapping time instants. The effect of this is that the bandwidth efficiency is decreased by a factor of N. The subsequent examples 2 and 3 used coded modulation techniques to partially or completely offset the loss.

However the SNR gain is reduced due to the use of these techniques. In this section, we introduce intentional multipath distortion to compensate for the bandwidth loss. This is done by transmitting the data bearing signal $s(t)$ and $N-1$ replicas of it simultaneously from the N antennas. In order to get the diversity advantage, the overall channel is made frequency selective by introducing a delay of T between successive transmissions. Thus the transmitted signal from the i^{th} antenna, $i = 1, \ldots, N$, is $s(t-(i-1)T)$. A minimum mean squared error (MMSE) decision feedback equalizer (DFE) or a maximum likelihood sequence estimator (MLSE) is used to resolve the multi-path distortion and thus obtain diversity benefit that is due to the frequency selective channel.

3.1. Transmission Scheme

The data bearing signal

$$\tilde{s}_1(t) = \text{Re}\left[\left(\sum_n (a_n + jb_n)\, p(t-nT)\right) e^{jw_o t}\right] \tag{7}$$

and its $N-1$ delayed versions $\tilde{s}_2(t) = \tilde{s}_1(t-T), \ldots, \tilde{s}_N(t) = \tilde{s}_1\,(t-(N-1)T)$ are transmitted using antenna 1, antenna 2, ... and antenna N respectively. Here $a_n + jb_n$ is the complex 4-QAM data with a_n, $b_n = \pm 1$, and $p(t)$ is the transmit pulse which for example is a square root Nyquist Pulse.

Each transmitted signal $\tilde{s}_i(t)$ is subjected to a multiplicative distortion $\tilde{z}_i(t)$ which is given by $\tilde{z}_i(t) = z_i\, e^{jw_o t}$ where $z_i = x_i + jy_i$ with x_i and y_i being Gaussian random variates with zero mean and $E[x_i^2] = E[y_i^2] = \frac{1}{2}$.

Assuming ideal coherent demodulation, the received signal is given by

$$\begin{aligned} r(t) &= \sum_i z_i \left(\sum_n (a_n + jb_n)\, p(t-(i-1)T-nT)\right) \\ &= \sum_n (a_n + jb_n)\left[\sum_i z_i p(t-Ii-1)T-nT)\right] \end{aligned} \tag{8}$$

Thus the overall baseband channel up to the receiver front end is given by

$$g_{\mathbf{z}}(t) = \sum_{i=1}^{N} z_i\; p(t-(i-1)T) \tag{9}$$

3.2. Maximum Likelihood Sequence Estimator (MLSE)

The maximum likelihood receiver finds that complex 4-QAM sequence $\hat{\mathbf{c}} = (\hat{c}_o, \ldots, \hat{c}_n)$, $\hat{c}_i = \hat{a}_i + j\hat{b}_i$, a_i, $b_i = \pm 1$, such that

$$\hat{\mathbf{c}} = \arg\min_{\hat{\mathbf{c}}} \int \left| r(t) - \sum_n \hat{c}_n g_z(t-nT)\right|^2 dt\ . \tag{10}$$

It is well known that this can be accomplished using the Viterbi algorithm.

3.3. Lower Bound on the Error Probability

The error performance of this receiver can be lower bounded by the matched filter bound [2]. The matched filter bound assumes that all the past and future data symbols have been decoded correctly and that a decision has to be made about a transmitted complex signal $a_o + jb_o$.

The probability of a bit error conditioned on a particular fade realization $\mathbf{z} = (z_1, \ldots, z_N)$ is given by

$$P_b(\mathbf{z}) = \frac{1}{2}\,\text{erfc}\left(\sqrt{\frac{E_s(\mathbf{z})}{2N_o}}\right) \tag{11}$$

where $E_s(\mathbf{z})$ is the energy in the pulse $g_z(t)$ which is given by

$$\begin{aligned} E_s(\mathbf{z}) &= \int |g_z(t)|^2\, dt \\ &= E_p\left[|z_1|^2 + |z_2|^2 + \ldots + |z_N|^2\right]\ , \end{aligned} \tag{12}$$

where E_p is the energy in the transmitter pulse $p(t)$.

The conditional bit error probability is now averaged over the fade statistics to give the average error probability as

$$\overline{P}_b = \left(\frac{1-\mu}{2}\right)^N \sum_{k=0}^{N-1} \binom{N-1+k}{k} \left(\frac{1+\mu}{2}\right)^k \tag{13}$$

where

$$\mu = \sqrt{\frac{\text{SNR}}{1+\text{SNR}}}\ , \tag{14}$$

and

$$\text{SNR} = \frac{E_p}{N_o}\ .$$

Thus a diversity benefit of N is obtained.

Figure 2 shows the performance of the MLSE with $N = 2$ antennas.

3.4. Optimum Linear Equalization and Decision Feedback Equalization

As an alternative to MLSE, linear equalization (LE) or MMSE-DFE can be used to tradeoff complexity with performance. Here we consider the performance of the optimum LE and DFE assuming an infinite number of taps in both cases. With LE and $N = 2$ and $M \geq 1$, bounds on the average BER with optimum combining and equalization are presented in [5].

For LE and $N \geq 2$, we used Monte Carlo simulation to determine the average BER. The transmission channel between the N transmit and one receive antenna is given by $C(\omega) = \sum_{i=1}^{N} z_i e^{-j2\omega(i-1)T}$, where the z_i's are independent, complex Gaussian random variables. As shown in [5], an exponentially tight upper bound on the average BER with coherent detection of QPSK is given by

$$P_e = e^{-\frac{1}{2\,\text{MSE}}}\ , \tag{15}$$

where the mean-square error, MSE, is given by

$$\text{MSE} = T/(2\pi)\int_{-\pi/T}^{\pi/T} \left[1 + \rho \mid C(\omega) \mid^2\right]^{-1} d\omega\ , \tag{16}$$

where ρ is the received signal-to-noise ratio and we have assumed for simplicity that the transmitted signal spectrum

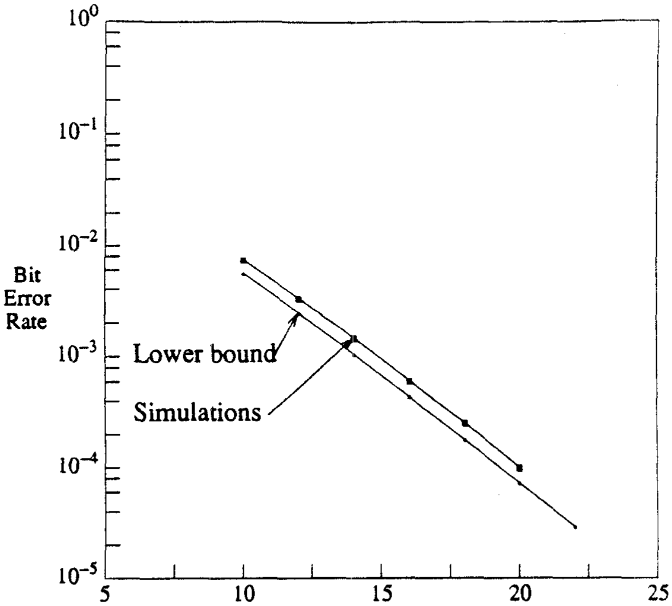

Figure 2

Simulated BER for a MLSE with QPSK on a 2-Ray static fading channel. Ideal knowledge of the channel is assumed.

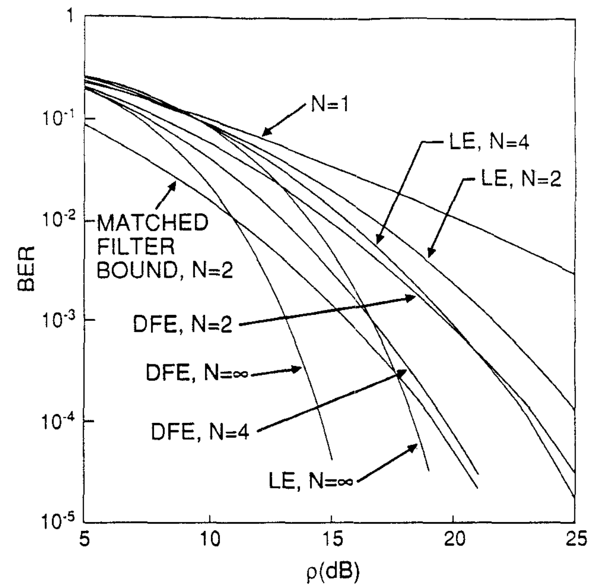

Figure 3

Simulated BER for LE and DFE with QPSK and N transmit antennas with noise only.

is zero outside of a $2\pi/T$ bandwidth. As in [5], we used Monte Carlo simulation to derive 1000 transmission channels $C(\omega)$ and numerically calculated the average BER. The BER averaged over these 1000 channels is shown in Figure 3 for $N = 2$ and 4.

For DFE, we used Monte Carlo simulation as above, with the following modification. The average BER with DFE is given by (15) with the MSE given by

$$\text{MSE} = \exp\left\{-T/(2\pi)\int_{-\pi/T}^{\pi/T} \ln\left[1+\rho \mid C(\omega)\mid^2\right]\, d\omega\right\}. \tag{17}$$

Results for DFE with $N = 2$ and 4 are also shown in Figure 3.

Figure 3 shows the average BER versus the SNR (ρ) for LE and DFE with $N = 2$ and 4. Results for the matched filter bound with $N = 2$ and the performance bounds for DFE and LE are also shown. For $N = 2$, the DFE is 4 dB worse than the matched filter bound and LE is 2 dB worse than DFE. Note that, from Figure 2, MLSE was only 0.6 dB worse than the matched filter bound. For $N = 4$ at a 10^{-2} BER, performance with LE improves by 1 dB over $N = 2$ and with DFE improves by 2 dB over $N = 2$. Also, with both LE and DFE, $N = 4$ improves performance over $N = 2$ by about half (in dB) of the maximum performance improvement possible as $N \to \infty$.

4. Conclusions

We have proposed two signaling schemes that makes use of multiple transmitter antennas to provide diversity benefit to a receiver that is normally equipped with only one antenna as in cellular radio. The proposed schemes have been developed for frequency division duplex (FDD) schemes, such as cellular radio systems. Unlike time division duplex (TDD) schemes, we assume that the channel characteristics in both directions are independent and no prior knowledge of the downlink channel is assumed at the mobile.

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