Performance of Optimum Diversity Combining and Equalization with Co-Channel Interference in Mobile Radio Channel

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Abstract
This paper presents the statistical and theoretical analysis of Nth-order diversity reception combined with equalization in mobile radio channel. The diversity paths are statistically independent and each is characterized by the sum of multiple delayed and independent Rayleigh-fading beams. In modeling the multi-path radio performance, we take into account co-channel interference (CCI) generated by frequency reuse and additive white Gaussian noise (AWGN). The performance evaluations are made of average error probability and outage probability. The average error rate is determined by using a Monte Carlo simulation for a set of channel parameters such as signal-to-noise ratio (SNR), signal-to-CCI ratio (SIR) and equalization coefficients determined for this channel, etc. A quadrature amplitude modulation (QAM) is used in our signal analysis. Equalization-taps are determined by using minimum mean-squared error (MSE) criterion and it is expressed directly in terms of channel parameters and modulation characteristics of the signal.

1.0 Introduction
For PCS applications, the selection of modulation schemes depends on many factors, including a wide range of operating environments and receiver complexity. The spectrum efficient modulation schemes, such as quadrature amplitude modulation (QAM) and the possibility of utilizing trellis-coded modulation have received much attention. The signal transmission over PCS radio channels is very complex. Unlike the classical AWGN channel, the impulse response of PCS radio channel is time-variant due to multipath propagation and changing environments. In addition, the radio transmission system is limited by the interference due to extensive frequency reuse in the microcell and picocell systems.

The analyses and simulation results of the performance of diversity combining and equalization were presented by Balaban and Salz [2]. However, the CCI was combined with noise source in [2]. In this paper, we present the theoretical analyses and demonstrate the simulation results of diversity combining and equalization with CCI taken into account.

A common measure of system performance is the outage probability for a radio receiver due to transmission impairments. Since radio performance can not actually be measured over varying radio paths, a practical approach is to model the channel by a set of parameters and to evaluate the performance of the radio over this channel. The parameters are varied to fit empirically derived statistical distributions for the radio path so that the outage of the digital radio, if any, can be predicted.

2.0 System Description
A typical radio system is depicted in FIGURE 1. The system consists of several basic building blocks, the evaluation of system performance by purely analytic methods is extremely complex and time consuming. We propose a method in this paper which combines some analytical techniques with Monte Carlo simulation of channel parameters such that the statistical performance of the radio over the channel can be predicted. The block diagram of the transmitter, linear diversity channels, and coherent demodulator with CCI is depicted in FIGURE 2.

2.1 Diversity Channel Model
The complex baseband impulse response of the channel with angular carrier frequency \( \omega_0 \) is represented by \( h_k(t) \). We assume that \( S_n(t) = s_n(t) e^{j\omega_n t} \) is a transmit signal, \( S_n(t) = s_n(t) e^{j\omega_n t} \) is co-channel interference, and \( N_k(t) = N_k(t) e^{j\omega_k t} \) is a complex AWGN in each diversity path. As presented in FIGURE 2, the N received baseband signals are presented as

\[ V_k(t) = S_k(t) \otimes h_k(t) + S_\perp(t) \otimes \alpha_k h_k(t) + N_k(t) \quad (EQ 1) \]

where \( k = 1, \ldots, N \), \( \otimes \) represents convolution and \( \alpha_k \) is the square root of \( \frac{1}{SNR} \) for each diversity path.

The complex baseband QAM modulated signal is

\[ S_n(t) = \sum_n C_n g(t - ntT) \quad (EQ 2) \]

The complex baseband interference generated by the same modulation scheme is

\[ S_\perp(t) = \sum_n D_n g(t - ntT - \tau) \quad (EQ 3) \]

The sequence of complex data symbols \( C_n = c_n + j\eta_n \) and \( D_n = c_n + j\eta_n \) are defined as the signal constellation with \( L^2 \) and the information rate is

\[ R = \frac{\log_2 L^2}{T}, \text{ bits/sec.} \quad (EQ 4) \]

2.2 Optimal Diversity Combiner-Linear Equalizer
In our model, a front-end receiver which presents as a matrix with \( i \) entries of input ports and \( j \) entries of output ports followed by a symbol-by-symbol detector is introduced. The impulse response of matrix filter \( W(t) \) is a \( 1 \times N \) matrix since we are interested in a single output application. The \( 1 \times N \) linear matrix receiver filters are depicted in FIGURE 3.

The MSE approach tries to minimize the mean-squared difference between the sequence of transmitted symbols \( C_n \) and sampled output \( V_n(t) \) with the presence of additive noise and interference. With the following assumptions: 1) the stationary infinite and independent data sequences \( C_n \) and \( D_n \) are used. 2)
the CCI is statistical independent, we utilize the expectation at time instant \( \Delta T \) to obtain MSE.

\[
MSE = E[V_0 - C_0^2] = \sigma_c^2 \sum_{k=1}^{N} [f_k(0) \otimes W_k(t) - \hat{W}_k(t)]^2 + \sigma_o^2\sum_{\eta, \eta' \neq 0} \sum_{k=1}^{N} [f_k(0 - nT) \otimes W_k(t)]^2 \\
+ \sigma_o^2 \sum_{\eta, \eta' \neq 0} \int \sum_{k=1}^{N} |W_k(t)|^2 dt + N_o \sum_{k=1}^{N} \int |W_k(t)|^2 dt
\]

\[= MSE[\tilde{W}_1(t), \tilde{W}_2(t), ..., \tilde{W}_N(t)] \quad (\text{EQ 5})
\]

where \( I = 0, V = V(\Delta T) \), the \( \sigma_c^2 = E[C_0^2] \) and \( \sigma_o^2 = E[D_0Q^2] \) are signal and CCI variances, \( N_o \) is the common white noise spectral density, and \( f_k(t) = g(t) \otimes h_k(t) \).

The minimum MSE of the optimal diversity combiner-linear equalizer is

\[
MSE_o = \lim_{(W_1(t), W_2(t), ..., W_N(t))} E[V_0 - C_0^2] = MSE[\hat{W}_1(t), \hat{W}_2(t), ..., \hat{W}_N(t)]
\]

\[= MSE[\hat{W}_1(t), \hat{W}_2(t), ..., \hat{W}_N(t)] \quad (\text{EQ 6})
\]

Let \( \hat{W}_1(t) = W_1(t) + \epsilon_{\hat{W}_1} \) where \( \hat{W}_1(t) \) is the optimal \( W_1(t) \) and use the standard method of the calculus of variations to obtain N equations.

\[
\frac{\partial}{\partial \epsilon_k} (MSE) = \sigma_c^2 \sum_{n} U_n f_k^* (-nT - \tau) + \sigma_o^2 \sum_{n} U_n f_k^* (-nT - \tau) + N_o \tilde{W}_k(\tau), \quad k = 1, ..., N
\]

where \( U_n = \sum_{k=1}^{N} f_k^* (-nT) \otimes \tilde{W}_k(t) \).

\[ \quad (\text{EQ 7})
\]

By defining

\[
\hat{U}_0 = \sigma_c^2 \sum_{n} U_n \quad \text{and} \quad \hat{U}_n = \sum_{\eta, \eta' \neq 0} \left( \frac{1}{N_o} \right) \tilde{W}_k(\tau)
\]

the optimum structure of the kth filter is obtained as

\[
\tilde{W}_k(t) = \sum_{n} \hat{U}_n f_k^* (-nT - \tau). \quad (\text{EQ 8})
\]

For each diversity path, the (EQ 8) shows that the optimum receiving filter is the cascade of a matched filter and a transversal filter with an infinite set of coefficients \( \hat{U}_n \) for T-scaled taps. FIGURE 4. depicts the structure of the optimal diversity combiner-equalizer [1].

Because \( \hat{U}_n \) is a linear function of optimal \( W_k \)'s, we solve the difference equations of \( \hat{U}_n \)'s obtained from (EQ 7) as follows: multiply \( \hat{U}_n \) by \( f_k(t - \tau) \) and integrate, then sum over all the \( k \) terms, the desired difference equations are obtained as

\[
\sum_{k=1}^{N} \sum_{n} r_k(l) \otimes W_k(t) = \sigma_c^2 \sum_{n} U_n \sum_{k=1}^{N} r_k(l - n) + \sigma_o^2 \sum_{n} U_n \sum_{k=1}^{N} r_k(l - n) + N_o U_k, \quad \text{for all } l
\]

where \( r_k(l) = \int \sum_{\tau} f_k(-\tau - IT) f_k^* (-nT - \tau) dt \).

\[ \quad (\text{EQ 9})
\]

By taking Fourier transform of (EQ 9), the coefficients \( U_n \) and \( \hat{U}_n \) are obtained as follows:

\[
U(w) = \frac{\sigma_c^2 \sum_{k=1}^{N} R_k(w)}{\sigma_c^2 \sum_{k=1}^{N} R_k(w) + \sigma_o^2 \sum_{k=1}^{N} R_k(w) + N_o} \quad (\text{EQ 10})
\]

\[
\hat{U}(w) = \frac{\sigma_o^2 \sum_{k=1}^{N} R_k(w)}{\sigma_o^2 \sum_{k=1}^{N} R_k(w) + N_o} \quad (\text{EQ 11})
\]

where \( R_k(w) \) and \( U(w) \) are Fourier transforms of \( r_k(l) \) and \( U_l \).

Now, we have to evaluate the MSE of a system in which \( R_k(w), k = 1, ..., N \) are given, and \( U(w) \) is optimized. Substitute \( W_k(t) \) into (EQ 5) when \( t = 0 \), after algebraic manipulations, the (MSE) is \( \sigma_c^2 \Delta T (1 - U_0) \). Since the \( U(w) \) and \( \hat{U}(w) \) are uniquely determined, the final explicit formula for minimum MSE can be shown to be

\[
MSE_o = \sum_{k=1}^{N} \int_{-T}^{T} \frac{1}{\sigma_o^2 \sum_{k=1}^{N} R_k(w) + N_o} d\omega \quad (\text{EQ 12})
\]

The expression of the MSE is achieved by optimizing the receiving filter for a given diversity channel and a given shaping filter. Note that (MSE) is a random variable.

2.3 Optimal Decision Feedback Equalizer

The optimal combiner-decision feedback equalizer is depicted in FIGURE 5. The matched filter is followed by the anticausal feedforward filter (anticausal taps) [4]. The sampled output signal of optimal combiner-decision feedback equalizer in the absence of noise and CCI is expressed as

\[
V_{os} (\Delta T) = \sum_{l=0}^{N} q_l c_{l-n} = C_0 q_{0} + \sum_{\eta > 0} q_{\eta} c_{\eta-n} + \sum_{\eta < 0} q_{\eta} c_{\eta-n}
\]

where \( q_l = \sum_{k=1}^{N} g(t) \otimes f_k(t) \otimes W_k(t) \)

and \( C_{l-n} \) is error free for \( n > 0 \).

\[ \quad (\text{EQ 13})
\]

The desired equations of coefficients \( U_n \)'s are obtained without post-cursive terms.
\[
\sigma^2_c \sum_{k=1}^{N} r_k(l) = \sigma^2_c \sum_{n=0}^{\infty} U_n \sum_{k=1}^{N} r_k(l-n) \\
+ \sigma^2 \sum_{n=0}^{\infty} U_n \sum_{k=1}^{N} r_k(l-n) + N_o U_l, \\
(l = 0, -1, ..., -\infty). \tag{EQ 14}
\]

(EQ 14) is a set of Wiener-Hopf equations. To solve the equations, some techniques are used, the detail derivation without CCI can be found on [2] and [3]. With the present of CCI, we summarize the results of our derivations as follows:

Let \( Q_l = \sum_{k=1}^{N} r_k(l) \) and define the infinite sequence of numbers
\[
M_l = Q_l + \frac{\sigma^2}{\sigma^2_c} Q_l + N_o \delta_{l0} \text{ and } M_l = M_l^+ \otimes M_l^- \text{ where } M_l^+ = 0, \\l <0 \text{ and } M_l^- = 0, l > 0.
\]

Hence, we have \( Q_l = U_l \otimes M_l^+ \otimes M_l^- \), \((l < 0)\). The results of \( U(w) \), \( \hat{U}(w) \) and \( \hat{W}(\tau) \) are obtained as
\[
\frac{1}{\sigma^2_c} \frac{N_o}{M_l(M(w))} U(w) = \frac{1}{\sigma^2_c} \frac{1}{1 + \frac{\sigma^2}{\sigma^2_c}} \frac{N_o}{M_l(M(w))} . \tag{EQ 15}
\]

Again, we define \( \hat{W}_k(\tau) = \sum_{n_1} U_{n_1} \delta_{l-n_1} \) as it's defined in (EQ 8), the optimum structure of the \( k \)th filter is obtained as
\[
\hat{W}_k(\tau) = \sum_{n_1} \hat{U}_{n_1} \delta_{l-n_1} . \tag{EQ 16}
\]

For each diversity path, (EQ 16) shows that the optimum receiving filter is the cascade of a matched filter and a transversal filter with one-side infinite set of coefficients \( \hat{U}_n \) for T-spaced taps.

\[
\hat{U}(w) = \frac{1}{\sigma^2_c} \frac{N_o}{M_l(M(w))} \left( 1 + \frac{\sigma^2}{\sigma^2_c} \right) U(w) = \frac{1}{\sigma^2_c} \frac{N_o}{M_l(M(w))} . \tag{EQ 17}
\]

Like the derivation approach outlined in [3], the \( MSE_{od} = \sigma^2_c (1 - U_0) \) with CCI is obtained by the same manipulation.

\[
MSE_{od} = \sigma^2_c \left( \frac{\pi}{T} \int_{-\pi}^{\pi} \frac{\sigma^2_c \sum_{k=1}^{N} R_k(w)}{\sigma^2_c \sum_{k=1}^{N} R_k(w) + N_o} \, dw \right) . \tag{EQ 18}
\]

2.4 Matched Filter

To obtain an explicit formula of the MSE for a matched filter which ideally reduces the ISI, we simply discard the ISI from (EQ 10). Only \( U_0 \) is interested since the \( U_n = 0 \) when \( n < 0 \).
\[ V_{0i} = b_0 U_0^* + \sum_{n: (n \neq 0)} b_n U_{n\pi}^* + \sum_{n: (n \neq 0)} a_n U_{n\eta}^* + \sum_{n: (n \neq 0)} a_n U_{n\chi}^* + v_{0i} \]  
\[ \text{where the } U_0 \text{ is real in the case of } W_k's \text{ are comprised of a matched filter and followed by taps.} \]

Suppose that the detection threshold is placed in the middle of adjacent elements for each real-part and imaginary-part. The detection threshold is in the set of \([-L-2, \ldots, -2, 0, 2, \ldots, L-2]\) when \(a_n, b_n, \gamma_n, \text{ and } \chi_n \) are in the set of \([-L-1, \ldots, -3, 1, 1, \ldots, L-1]\). The estimated error rate is

\[ p_e = P_e \left[ v_{0i} \geq U_0 - \sum_{n: (n \neq 0)} (a_n U_{n\pi}^* + b_n U_{n\eta}^*) - \sum_{n: (n \neq 0)} (\gamma_n a_n U_{n\pi}^* + \chi_n a_n U_{n\eta}^*) \right] \]

\[ = P_e \left[ v_{0i} \geq U_0 - \sum_{n: (n \neq 0)} (b_n U_{n\pi}^* + a_n U_{n\eta}^*) - \sum_{n: (n \neq 0)} (\gamma_n a_n U_{n\pi}^* + \chi_n a_n U_{n\eta}^*) \right] \]  
\[ \text{Equation (26)} \]

The second equality holds since \(v_{0i}\) and \(v_{0i}\) are identically distributed Gaussian random variables and intersymbol interference and cross-symbol interference have identical probability distribution.

Now, applying the Chernoff bound to estimate the upper bound of equation (Equation (26))

\[ P_e \leq \exp \left\{ -\frac{1}{2} \frac{U_0^2}{\sigma_v^2} - \frac{1}{2} \sum_{n: (n \neq 0)} \left| U_n^* \right|^2 + \frac{1}{2} \sum_{n: (n \neq 0)} \left| U_n \right|^2 \right\} \]

\[ = \exp \left\{ -\frac{MSE_{e}}{\sigma_v^2} - \frac{1}{2} \sum_{n: (n \neq 0)} \left| U_n^* \right|^2 + \frac{MSE_{e}}{\sigma_v^2} \right\} \]

\[ = \exp \left\{ -\frac{MSE_{e}}{\sigma_v^2} \right\} \]  
\[ \text{Equation (27)} \]

where \(E(a^2) = \sigma_a^2\), \(E(y^2) = \sigma_y^2\), and \(E(v_{0i})^2 = \sigma_{v_{0i}}^2\).

The mean-squared error from equation (22) is

\[ MSE = E\left| V_0 - C_0 \right|^2 = \sigma_v^2 \left( 1 - U_0 \right)^2 + \sum_{n: (n \neq 0)} \left| U_n^* \right|^2 + \sum_{n: (n \neq 0)} \left| U_n \right|^2 + E\left| v_{0i} \right|^2 \]

\[ \text{Equation (28)} \]

Finally, we define the outage probability for a given error probability threshold \( P_{th} \) as

\[ P_e = P_e \left( P_e > P_{th} \right) \]  
\[ \text{Equation (29)} \]

3.0 Mobile Radio Transmission Channel Model

The transmission characteristic of a mobile radio channel is modeled by frequency-selective fading of both amplitude and group delay [15]. It is assumed that the channel is time invariant over some period of time in order to establish a tractable mathematical framework. The complex envelope impulse response of the model with \( M \) received Rayleigh-fading beams is

\[ f(t) = \sum_{m=1}^{M} \rho_m (r_m + j\beta_m) \delta(t - \tau_m) \]  
\[ \text{Equation (30)} \]

where each of the \( M \) beams is characterized by \( \gamma_m, \beta_m \), and \( \tau_m \).

The \( \gamma_m \) and \( \beta_m \) are independent zero-mean Gaussian random variables, \( E(\gamma_m^2) = E(\beta_m^2) = \sigma^2 \).

The overall channel impulse response of \( N \) diversity paths is

\[ h(t) = \sum_{k=1}^{N} \sum_{m=1}^{M} \rho_{km} (r_{km} + j\beta_{km}) \delta(t - \tau_{km}) \]  
\[ \text{Equation (31)} \]

We use \( M=2 \) in our numerical computation. Consider the output signal and output CCI on the \( k \)th diversity path when the input signal and CCI signal are defined in (Equation 2) and (Equation 3). The output desired signal is

\[ V_{sk}(t) = \sum_{n} C_n (A_k g(t - nT) + B_k g(t - nT - \tau_k)) \]  
\[ \text{Equation (32)} \]

and the output CCI signal

\[ V_{sk}(t) = \sum_{n} D_n (Y_k g(t - nT) + Y_k g(t - nT - \tau_k)) \]  
\[ \text{Equation (33)} \]

where \( g(t) \) is raised-cosine shape function, the relative delay spreads, \( \tau_k = \tau_0 - \tau_2 \) and \( \tau_k = \tau_1 + \tau_2 \) and

\[ A_k = \rho_{k1} (\gamma_{k1} + j \beta_{k1}), B_k = \rho_{k2} (\gamma_{k2} + j \beta_{k2}), \]

\[ Y_k = \rho_{k1} (\gamma_{k1} + j \beta_{k1}), Y_k = \rho_{k2} (\gamma_{k2} + j \beta_{k2}). \]

For calibration purpose, the average received desired signal power and CCI signal power on the \( k \)th diversity path are presented as follows:

\[ P_{sk}(A_k, B_k) = \lim_{\lambda \to \infty} E \left\{ \frac{1}{2\lambda^T} \int_{-\lambda T}^{\lambda T} \left| V_{sk}(t)^2 \right| \right\} \]

\[ = \frac{\delta_k^2}{\pi} \lambda \int_{-\lambda T}^{\lambda T} \left| G(w) \right|^2 \left| A_k + B_k e^{j\omega s} \right|^2 dw. \]  
\[ \text{Equation (34)} \]

\( E \{ \} \) denotes expectation w.r.t. the symbols \( C_n \) and \( E|\gamma|^2 = \sigma_y^2 \).

The average received desired signal power w.r.t. channel characteristics is obtained as

\[ P_{sk} = E \left\{ P_{sk}(A_k, B_k) \right\} = \frac{\sigma_y^2}{\lambda T} E \left\{ A_k + B_k \right\} = \frac{\sigma_y^2}{\lambda T} (\sigma_{A_k}^2 + \sigma_{B_k}^2) \]  
\[ \text{Equation (35)} \]

where the \( \tau_0 = 0 \) in case of flat fading.

By applying the same manipulation to the received CCI signal, the average received CCI signal power can be obtained as

\[ P_{sk} = E \left\{ P_{sk}(Y_k, Z_k) \right\} = \frac{\sigma_y^2}{\lambda T} E \left\{ Y_k + Z_k \right\} = \frac{\sigma_y^2}{\lambda T} (\sigma_{Y_k}^2 + \sigma_{Z_k}^2) \]  
\[ \text{Equation (36)} \]

where the \( \tau_1 = 0 \) in case of flat fading.

The noise power is defined as

\[ P_{nk} = N_{\text{oa}} 2W = \frac{N_{\text{oa}}}{T}. \]  
\[ \text{Equation (37)} \]
Finally, we define the signal-to-noise ratio (SNR) and signal-to-CCI ratio (SIR) for kth diversity path which we shall use in subsequent derivations and numerical evaluations. In defining the SNR and SIR, we normalized (\(\sigma^2_{A_k}, \sigma^2_{B_k}\)) to (\(\sigma^2_A, \sigma^2_B\)) and (\(\sigma^2_{Y_k}, \sigma^2_{Z_k}\)) to (\(\sigma^2, \sigma^2\)).

With this manipulation, the SNR on the kth diversity path becomes

\[
P_k = \frac{P_{sk}}{P_{ik}} = \frac{\sigma_c^2}{N_o} \left( \frac{\sigma^2_A + \sigma^2_B}{\sigma^2} \right) \quad (\text{EQ 38})
\]

and the SIR on the kth diversity path becomes

\[
\eta_k = \frac{P_{sk}}{P_{ik}} = \frac{\sigma^2_c}{\sigma^2} \left( \frac{\sigma^2_A + \sigma^2_B}{\sigma^2} \right) \quad (\text{EQ 39})
\]

where \(\lambda_k, B_k, \nu_k, Y_k, Z_k\) are i.i.d., zero-mean, complex Gaussian random variables with common variance, \(\sigma^2=1/2\).

4.0 Simulations Results

In our numerical evaluation, we use the Quasi-analytic Monte Carlo simulation method. Random sets of channel variables are drawn from known probabilities and averages. We evaluate the formulas by using numerical method. This operation is repeated and the result is recorded in a data file for each draw until the values fall into the confidence interval.

The selected set of simulation results of average error probabilities and outage probabilities with a variety of channel parameters are demonstrated in the following sections. The Shannon capacity and attainable data rate of our channel model were presented in [6]. We shall exhibit the numerical results of a system operating in a 30 kHz bandwidth, 45 kbit/sec data rate, and using 35% filter rolloff.

4.1 Noise Dominated Channel (\(\eta_k>60\text{dB}, 15\text{dB}<\eta_k<30\text{dB}\))

For comparison, we summarize the numerical results of noise dominated channel in Rayleigh flat-fading (\(\tau_k=0, \tau_j=0\)) condition as follows:

- All equalizers achieve the matched filter bound due to flat fading, delay spread \(\tau_k=\tau_j=0\).
- The slope in dual space diversity gain is doubled compared to non-diversity. The result agrees with the well known result of N-channel combining in a flat fading environment.

\[
P_e = \left( \frac{\sigma^2_c}{\rho} \right)^N \quad (\text{EQ 40})
\]

Also, our numerical results agree closely with the results presented in [2].

4.2 CCI Dominated Channel (\(\eta_k>60\text{dB}, 15\text{dB}<\eta_k<30\text{dB}\))

For CCI dominated channels, noise is considered relatively very small. The system performance degradation is dominated by the impairments of ISI caused by delay spread and CCI.

- Flat Fading (\(\tau_k=0, \tau_j=0\))

Figure 6 exhibits the error probability obtained in Rayleigh flat-fading. The logarithm of average error probability versus SIR is linear when \(\mu\) is measured in decibels. The results described in section 4.1 is maintained with modification on average error probability equation (EQ 40) as

\[
P_e = \alpha_\eta \left( \frac{\sigma^2_c}{\eta} \right)^N - \beta_\eta
\]

where \(\alpha_\eta>0\) and \(N>\beta_\eta>0\) when the SIR is set to a equivalent SNR, \((\eta=\rho)\). The values of \(\alpha_\eta\) and \(\beta_\eta\) depend on the characteristics of CCI. In our case, \(\alpha_\eta>1\) and \(N>\beta_\eta>0\) because a independent CCI is used in the simulation.

- Arbitrary \(\tau_k, \tau_j\) and Filter Rolloff of Desired and CCI Signal

The logarithm of average error probability are plotted versus SIR when arbitrary relative beam delay=0.6T and filter rolloff factor \(\beta=0.35\) in Figure 7. Figures 8 and 9 show the influences of SIR and multipath delay spread on the performance. We evaluate average error probability when the SNR and SIR are in 15-30 dB range.

4.2.1 Outage Probability Numerical Results

The outage probability in our radio channel is presented by using criterion given in (EQ 29). We present the outage probability in the range of \(10^{-1}-10^{-2}\) based on the error probability threshold of \(P_{th}=10^{-4}\).

Figure 10 depicts the outage probability versus SNR for non-diversity and dual diversity when SIR takes three different values.

5.0 Conclusion

This paper has presented the analyses of system performance when diversity reception and different equalization techniques are used over frequency-selective fading channels. The performance evaluations were done on the average error probability and outage probability due to AWGN, CCI, filter rolloff, and multipath delay spread.

In our numerical results, the performance of a 4-QAM (QPSK) system operating in a 30 kHz bandwidth, 45 kbit/sec data rate, and using 35% filter rolloff were presented as follows:

1. The degradation due to co-channel interference in the channel capacity is more severe than by equivalent additive white Gaussian noise. This degradation is 3dB without diversity and 5dB with dual diversity.

2. The dual diversity gain on average error probability depends on the CCI level. It is of the order-of magnitude in noise dominated channel and of the one order-of magnitude in CCI dominated channel once the multi-path delay spread exceeds the 0.6T.

3. The average error probability and outage probability are irreducible by increasing the SNR once the SIR is less than 20dB.

4. The performance of decision feedback equalizer is uniformly superior to the linear equalizer and closely follows the matched filter bound with the presumption of no error propagation. Both equalizers take advantage of the multi-path delay spread of the desired signal and CCI signal to eliminate the ISI and CCI.

Further study is to investigate the performance of finite-tap equalizer in combating the ISI and CCI and the interference control schemes, such as adaptive power control, intra-cell hand-off algorithm, and dynamic channel assignment.
REFERENCE


FIGURE 2. Transmitter, Linear Diversity Channels and Coherent Demodulator with Co-Channel Interference

\[
\sum_n C_n e^{j\omega t} (t - nT) \rightarrow V_1(t) \rightarrow \tilde{V}_1(t) \rightarrow V_1(t)
\]

\[
\sum_n D_n e^{j\omega t} (t - nT) \rightarrow \tilde{V}_2(t) \rightarrow V_2(t)
\]

\[
\sum_n N_n e^{j\omega t} (t - nT) \rightarrow \tilde{V}_N(t) \rightarrow V_N(t)
\]

FIGURE 3. IXN Linear Matrix Receiver Filter

\[
V_1(t) \rightarrow W_1(t) \rightarrow V_o(t) \rightarrow V_o(nT) \quad t_o = nT
\]

\[
MSE = |V_o(nT) - C_n|^2
\]

FIGURE 4. Optimal Combiner-Linear Equalizer Structure

\[
V_1(t) \rightarrow f_1^*(t) \rightarrow V_2(t) \rightarrow f_2^*(t) \rightarrow \ldots \rightarrow V_N(t) \rightarrow f_N^*(t) \rightarrow V_o(nT)
\]

N-order Space Diversity

Linear Equalizer