A Co-Channel Interference Cancellation Technique Using Orthogonal Convolutional Codes on Multipath Rayleigh Fading Channel

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Abstract—This paper proposes and evaluates a new co-channel interference cancellation technique that utilizes orthogonal convolutional codes on a multipath Rayleigh fading channel. In a spread spectrum multiple access environment, co-channel interference (CCI) limits the performance of the communication link. To remove this interference, several CCI cancellation techniques have been proposed, including the technique that does not require the receiver to have knowledge of the cross-correlation between user sequences. This method leaves residual interference after the cancellation caused by errors in the initial decisions. To reduce the residual interference and improve the initial decisions, the proposed scheme utilizes the error-correcting capability of orthogonal convolutional codes. This paper evaluates the performance of this scheme. Our results show that the proposed CCI canceller offers an improvement in capacity by a factor of 1.5 ~ 3 as compared with a conventional canceller on a multipath Rayleigh fading channel. The proposed canceller works in the presence of residual interference due to imperfect cancellation. The proposed canceller also has a capacity improvement with the use of soft handoff in a multihand configuration.

I. INTRODUCTION

RECENTLY, spread spectrum techniques have received a large amount of attention in wireless and cellular communication applications. This is in part because spread spectrum techniques have multiaccess capability, antijamming capability, and antijamming capability [1].

There are three basic multiple access schemes: 1) frequency division multiple access (FDMA); 2) time division multiple access (TDMA); and 3) code division multiple access (CDMA) [2]. Unlike FDMA and TDMA capacities, which are primarily bandwidth limited, CDMA capacity is interference limited [3]. This is one of the reasons that CDMA may be potentially spectrally efficient for cellular communications for which the additional spectrum will unlikely be allocated [2].

Multituser detection has been developed to increase the CDMA system capacity for more than ten years. In [4], the other users’ interference can be minimized through the use of maximum-likelihood sequence detection. Although this scheme is optimum, the complexity of the detection is high and grows exponentially with the number of users. In [5]–[7], the interference is removed by utilizing matrix operations. These methods do not require complex receiver structures, and provide a cross-correlation free signal at the output. They are not useful, however, if the signature sequences are used to separate the users’ channels, as these methods require the receiver to have knowledge of the sequences’ cross-correlation, which varies with different relative phase of the signature sequences. If these multituser detectors are employed in the reverse link of the CDMA system specified by the IS-95 (North American Digital Cellular Standard), a receiver must calculate the cross-correlation and the computational complexity increases exponentially with the number of users [8].

CCI cancellation methods, whose complexity grows only linearly with the number of users, have also been proposed [9]–[11]. In [10], users are detected jointly and the CCI is removed. In this method, the receiver reconstructs the other users’ transmitted signals by using the initial decisions about the other users’ signals. This receiver then uses the estimates (reconstructed signals) to remove the CCI from the composite received signal. This method does not require knowledge of the cross-correlation between the spreading sequences. This method suffers from a significant amount of residual interference, which is caused by symbol errors in the initial decisions [12], [13]. Thus, the performance of the canceller depends on the error performance of the initial decision. Therefore, it is desirable to improve the accuracy of the initial decisions.

To improve the accuracy of the initial decisions, we have proposed a CCI cancellation technique that utilizes orthogonal convolutional codes [14]. In this method, received signals are both demodulated and decoded by a soft decision Viterbi decoder. The resulting bit streams are re-encoded and re-spread, and this reconstructed signal is then removed from the composite received signal. Due to the error-correcting capability of orthogonal convolutional codes, the residual interference is reduced since the decisions on the other users’ signals are improved compared to a noncoded canceller. Thus, the BER performance of the receiver is improved. Since the decoding of the orthogonal convolutional codes increases the processing delay, this system is most suitable for packet data communications. In [14], this CCI cancellation technique was investigated on the AWGN channel. A similar technique pre-
sented at the same time as [14] utilizes cascaded combination of cancelling CCI and hard decision decoding of a Hamming code [15]. In this paper, we extend our results for our cancellation technique to a multipath Rayleigh fading channel.

This paper is organized as follows. In Section II, the system model is presented. In Section III, the conventional CCI canceller is described. In Section IV, the new CCI canceller is described and its performance is derived. In Section V, the effects of imperfect cancellation are considered. In Section VI, a multicell configuration is assumed and intercell interference and soft handoff are considered. Simulation results and theoretical performance results are shown in Section VII. Section VIII presents our conclusions.

II. SYSTEM MODEL AND DEFINITIONS

The system model is of the same type as the CDMA system specified by IS-95. The receiver structure is based on that in [16] followed by a CCI canceller and a decoder for an orthogonal convolutional code. There are \( K \) active users in a cell. Each user encodes its data using an orthogonal convolutional code and the constraint length as \( v \). Depending on the output of the convolutional code, one of \( M \) orthogonal sequences is chosen. Then, the resulting orthogonal sequence is combined with long pseudo-noise (PN) sequences, which separate the users’ channels, and is modulated using Offset-QPSK. The transmitted OQPSK signal of the \( i \)th user during one symbol interval \( T_s \) is

\[
\tilde{s}_i(t) = \left\{ \sqrt{P_W} w^r(t) c_i(t) p_I(t) + j\sqrt{P_W} w^r(t-T_d) c_i(t-T_d) p_Q(t-T_d) \right\} \times \exp(-j\omega_c t) \quad 0 \leq t \leq T_s
\]  

where \( w^r(t) \) is the \( r \)th \( M \)-ary orthogonal symbol of equal energy \( (r = 1, \ldots, M) \), \( c_i(t) \) is the long PN sequence for the \( i \)th user, \( p_I(t), p_Q(t) \), are the PN sequences for the \( I \) and \( Q \) channels, \( T_d \) is the time offset, which is equal to \( T_c/2 \), and \( T_c \) is the chip interval. The processing gain \( G_p = T_s/T_c \) where \( T_s \) is the symbol duration. \( P \) is the power signal, thus the energy per symbol is \( E_s = P \cdot T_s \).

If the signal written in (1) is transmitted over a multipath channel, the resulting signal \( y(t) \) at the receiver is the sum of delayed, phase-shifted, and attenuated versions of the input signal. The received signal for the \( i \)th user can be written as

\[
y_i(t) = \text{Re}\left\{ \sum_{j=1}^{L_i} g_{ij}(t-\tau_{ij}) s_i(t-\tau_{ij}) \right\}
\]  

where \( L_i \) is the number of paths for the \( i \)th user’s channel, \( \tau_{ij} \) and \( g_{ij}(t) \) are the delay and complex gain coefficients, respectively, for the \( j \)th path for the \( i \)th user, and the asterisk denotes complex conjugation. The complex gain coefficients are defined as

\[
g_{ij}(t) = \gamma_{ij}(t) \exp(j\phi_{ij}) \]  

where \( \gamma_{ij}(t) \) and \( \phi_{ij} \) account for the attenuation and phase shift, respectively. \( \gamma_{ij}(t) \) is Rayleigh distributed and \( \phi_{ij} \) is an uniform random variable in \( [0, 2\pi] \).

In a single-cell CDMA cellular system with \( K \) users, the signal arriving at the base station receiver is

\[
r(t) = \text{Re}\left\{ \sum_{i=1}^{K} \sum_{j=1}^{L_i} g_{ij}(t-\tau_{ij}) s_i(t-\tau_{ij}) \right\} + n(t)
\]  

where \( n(t) \) is additive white Gaussian noise with zero mean and two-side spectral density \( N_0/2 \). At the output of the receiver bandpass filter with a bandwidth of \( B(\approx 1/T_c) \), \( n(t) \) becomes narrowband noise that can be represented by

\[
n(t) = n_c(t) \cos \omega_c t + n_s(t) \sin \omega_c t
\]  

where \( n_c(t) \) and \( n_s(t) \) are independent lowpass Gaussian processes with zero mean and variance \( N_0B \).

Fig. 2 shows the receiver structure. At the output of the low-pass filter (LPF) of the upper path (\( I \)-channel)

\[
d_I(t) = \text{LPF}\{r(t) \cos \omega_c t\}
\]  

where \( \theta_{ij} = \phi_{ij} - \omega_c \tau_{ij} \). Also for the lower path (\( Q \)-channel)

\[
d_Q(t) = \text{LPF}\{r(t) \sin \omega_c t\}
\]  

where

\[
\begin{align*}
G_p &= T_s/T_c \\
T_s &= \text{symbol duration} \quad P \quad \text{power signal} \quad E_s = P \cdot T_s \\
\gamma_{ij}(t) &\sim \text{Rayleigh} \quad \phi_{ij} \sim \text{Uniform}(0, 2\pi) \\
\end{align*}
\]
For the $I$-channel, the output of the $m$th correlator for the real ($I$) part of the $\lambda$th path of the $k$th user is

$$Z^{\lambda}_{I,m}(m) = \frac{1}{\sqrt{T_s}} \int_0^{T_s} d(t)c_k(t - \tau_{k\lambda})p_I(t - \tau_{k\lambda})W^m(t - \tau_{k\lambda}) dt$$

$$= \int_0^{T_s} \gamma_{k\lambda}(t - \tau_{k\lambda}) \sqrt{\frac{P}{T_s}} W^m(t - \tau_{k\lambda}) dt \times \cos(\theta_{k\lambda}) + I_{I,m}^{\lambda,k} + I_{I,m}^{\lambda,i} + N_{I,m}^{\lambda}$$

where $I_{I,m}^{\lambda,k}$, $I_{I,m}^{\lambda,i}$, and $N_{I,m}^{\lambda}$ which are given in Appendix A, represent the self path interference, interference due to the other paths, interference due to the other users, and thermal noise corresponding to the real ($I$) part of the path, respectively. Owing to the quadrature nature of the PN sequences, the self path interference is small compared to the other interference (Appendix A). In this case, $I$-channel correlator output becomes

$$Z^{\lambda}_{I,m}(m) = \left\{ \begin{array}{ll}
\tilde{\gamma}_{k\lambda} \sqrt{\frac{E_s}{2}} + I_{I,m}^{\lambda,k} + I_{I,m}^{\lambda,i} + N_{I,m}^{\lambda} & m = r \\
N_{I,m}^{\lambda} + I_{I,m}^{\lambda,k} + I_{I,m}^{\lambda,i} & m \neq r
\end{array} \right.$$  (9)

where $\tilde{\gamma}_{k\lambda}$ is the average attenuation of $\gamma_{k\lambda}(t)$ over the symbol period. Similarly, the output of the $m$th correlator for the imaginary ($Q$) part of the $\lambda$th path of the $k$th user is

$$Z^{\lambda}_{Q,m}(m) = \left\{ \begin{array}{ll}
\tilde{\gamma}_{k\lambda} \sqrt{\frac{E_s}{2}} \sin(\theta_{k\lambda}) + I_{Q,m}^{\lambda,k} + I_{Q,m}^{\lambda,i} + N_{Q,m}^{\lambda} & m = r \\
N_{Q,m}^{\lambda} + I_{Q,m}^{\lambda,k} + I_{Q,m}^{\lambda,i} & m \neq r
\end{array} \right.$$  (10)

where $I_{Q,m}^{\lambda,k}$, $I_{Q,m}^{\lambda,i}$, and $N_{Q,m}^{\lambda}$ are the interference due to the other paths, interference due to the other users, and thermal noise corresponding to the imaginary ($Q$) part of the path, respectively. Due to the square-law combining, the decision variable for the $k$th user will be

$$S^k(m) = \sum_{j=1}^{L_m} \left\{ \left[ Z_{I,j}^{k}(m) + Z_{Q,j}^{k}(m) \right]^2 + \left[ Z_{I,j}^{k}(m) - Z_{Q,j}^{k}(m) \right]^2 \right\}.$$  (13)

From [16], [17], and Appendix A, the average SNR for the $\lambda$th path of the $k$th user after the correlation is

$$\text{SNR}_{k\lambda} = \frac{\tilde{\gamma}_{k\lambda} E_s}{2(S_{\lambda} + \sum_{k=1}^{K} \sum_{i=1}^{L_{\lambda}} \tilde{\gamma}_{k\lambda} E_s - \tau_{k\lambda} E_s) + N_0}.$$  (14)
For simplicity, $L_i$ is set to $L$ for all $i$. We assume all fading paths have equal mean strength, i.e., $\tau_{ij}$ is one for all $i$ and $j$. With these assumptions, the SNR per path becomes

$$\text{SNR} = \frac{E_s}{2(K-1)L_p E_s} + N_0. \quad (15)$$

III. CONVENTIONAL CCI CANCELLER

A. System Model

Fig. 3 shows a receiver model using a conventional CCI canceller with user 1 as a reference. Perfect chip synchronization for every user is assumed. In this method, as shown in Figs. 2 and 3, all of the user’s received signals are first demodulated and decorrelated with a bank of correlators. The output of the correlators for each path is squared and combined. Then, the most probable orthogonal sequence is selected for each user. Utilizing these initial decisions, the selected sequences for each user are then re-spread and re-modulated with delay $\tau_{ij}$, attenuation $\gamma_{ij}$, and phase shift $\theta_{ij}$ (error in estimating these parameters is considered in Section V). The re-modulated signals are removed from the composite received signal. Following this process, user 1’s signal is demodulated, decorrelated, deinterleaved, and decoded based on the orthogonal convolutional code employed.

B. Performance of the Conventional CCI C canceller

To emphasize the performance difference of the conventional and the proposed canceller, we consider the simple case. To analyze the performance of the conventional CCI canceller, we assume the interference from the other users is Gaussian distributed, the chip pulse shape is rectangular, the chip synchronization is perfect, and the canceller can estimate the phase shift, the attenuation, and the delay perfectly (if the parameters are not accurately estimated, performance of all cancellers will deteriorate [11], as will be shown in Section V).

The error probability of the initial decision on the orthogonal sequence is given by [18, p. 745]

$$P_{e_{cl}} = 1 - \int_{0}^{\infty} \frac{U^{L-1}e^{-\frac{U}{2(NR_{c1})}}}{(1 + SNR_{c1})L(L-1)} e^{-U} \left(1 - e^{-U} \sum_{h=0}^{L-1} \frac{U^h}{h!} \right) dU. \quad (16)$$

where $SNR_{c1}$ is the signal-to-noise ratio after the first correlation given in (15) and is given by

$$SNR_{c1} = \frac{E_s}{2(K-1)L_p E_s} + N_0. \quad (17)$$

When errors in the initial decisions arise, the CCI canceller doubles the interference power [12], [13]. The signal-to-noise ratio after the CCI cancellation is (Appendix B)

$$\text{SNR}_{c2} = \frac{E_s}{2Ap_{e}(K-1)L_p E_s} + \frac{2L-1)E_s}{3O_p} + N_0 \quad (18)$$

where

$$Ap_{e}(K-1)L_p = 2P_{e_{cl}}(K-1)L_p. \quad (19)$$

Here, we assume that the receiver can estimate the accurate value of received signal power. From SNR$_{c2}$ the upper bound on the bit error rate using a soft decision Viterbi decoder can be obtained as [18, p. 760]

$$P_{b_{cl}} < \sum_{w=0}^{\infty} N_s(w)P_{d_{c}}(w, L) \quad (20)$$

where

$$P_{d_{c}}(w, L) \quad \text{probability of selecting the incorrect path (Appendix C)}$$

$$= (p_e)^w \sum_{h=0}^{L-1} \left(\frac{wL-1+h}{h!}\right)(1-p_e)^h; \quad (21)$$

$$p_e \quad \text{probability of error for square-law combining between orthogonal sequences}$$

$$= \frac{1}{2 + SNR_{c2}}; \quad (22)$$

$N_s(w)$ weight spectra of the code which can be obtained from [19]; $d_{free}$ minimum free distance.

If the weight spectra $N_s(w)$ is obtained from $d_{free}$ to $\infty$ in (20), the upper bound can be calculated. The bound becomes looser, however, as the BER increases. Quite often, the first few terms in (20) are used to get a performance approximation of a Viterbi decoder. We use the first four terms as an approximation for both the conventional and the new CCI cancellation techniques.

IV. NEW CCI CANCELLER

A. System Model

In this section, we present the new CCI canceller utilizing orthogonal convolutional codes. As mentioned before, the performance of the CCI canceller depends on the initial decisions. Therefore if the signal-to-noise ratio, SNR$_{c1}$, decreases, the probability that the initial decisions are incorrect increases and the canceller may increase the CCI seen by a user. To improve the error performance of the initial decision, the proposed system utilizes the error-correcting capability of orthogonal convolutional codes.

Fig. 4 is a model of the proposed system. After making a maximum likelihood decision on the received symbol sequence, the proposed canceller deinterleaves and decodes the orthogonal convolutional code using a Viterbi decoder. The decoded data is then re-encoded, re-interleaved, assigned to an orthogonal sequence, combined with the long PN sequences, and re-modulated to an OQPSK signal with delay $\tau_{ij}$, attenuation $\gamma_{ij}$, and phase shift $\theta_{ij}$. During this demodulation, decoding, re-encoding, re-interleaving, and re-modulation
processes, the received signal is stored in the memory. The memory size is dependent upon the interleaving size and the delay caused by the decoding process. After the re-modulation, the new canceller follows the same steps as the conventional canceller. That is, the canceller removes the 2~K\textsuperscript{th} users' signals from the composite received signal, which is extracted from the memory, and user 1's signal is demodulated, decorrelated, deinterleaved, and decoded.

It is noted that Viterbi decoders needed in the proposed scheme are already utilized in the conventional CCI cancellation scheme where each user has a Viterbi decoder. If the reuse of the same Viterbi decoder after CCI cancellation creates an unacceptable delay (if we are dealing with the transmissions of packets, they may tolerate a relatively long delay and the receiver may not be busy for receiving consecutive packets), one can solve the problem by employing two Viterbi decoders per user. Later, we will show that, for the same performance, the complexity of each decoder may be reduced by more than a factor of 2 so that the overall decoding complexity is still reduced.

**B. Performance of the Proposed CCI Canceller**

All the assumptions are the same as those for the conventional canceller, which were presented in Section III-B. The number of symbol errors caused by choosing a wrong path after re-encoding is the same as the distance of the coded sequence along the wrong path in the state diagram (for example, the number of symbol errors is four in Fig. 5) from the coded sequence along the correct path. Therefore, an approximation of the expected number of interfering symbols after the re-encoding is (Appendix B)

\[
\begin{align*}
\mathcal{A}_{\text{re}}(K,L) & \approx \sum_{w=0}^{d_{\text{re}}+3} N_c(w) \cdot w \cdot \left\{2P_{\text{d}}(w, L) \cdot (K-1)L\right\} \\
\end{align*}
\]

where

\[
\begin{align*}
P_{\text{d}}(w, L) = & \left(p_{\text{d}1}\right)^w L \sum_{h=0}^{L-1} \left(\frac{wL - 1 + h}{h}\right) (1 - p_{\text{d}1})^h; \\
p_{\text{d}1} = & \frac{1}{2 + \text{SNR}_{\text{d}1}}; \\
\text{SNR}_{\text{d}1} = & \frac{E_s}{2(KL-1)E_s + N_0}; \\
N_c(w): & \text{ coefficient of the weight enumerator function of the orthogonal convolutional code [19].}
\end{align*}
\]

Then, the signal-to-noise ratio after the cancellation is

\[
\text{SNR}_{\text{2}} = \frac{E_s}{2\mathcal{A}_{\text{re}}(K,L)E_s + 2(L-1)E_s + N_0}. \\
\]

From SNR\textsubscript{2}, the approximate error rate after the cancellation can be obtained by the following equation, which is similar to (20):

\[
\begin{align*}
\mathcal{P}_{\text{d}2} & \approx \sum_{w=0}^{d_{\text{re}}+3} N_s(w) P_{\text{d}}(w, L) \\
\end{align*}
\]
where the $PD_{\nu 2}(w, L)$ is obtained by substituting $\text{SNR}_{w 2}$ into (21) and (22) instead of $\text{SNR}_{u 2}$

$$\begin{align*}
PD_{\nu 2}(w, L) &= (p_{\nu 2})^{uL} \sum_{h=0}^{uL-1} \binom{wL-1+h}{h} (1-p_{\nu 2})^h; \\
p_{\nu 2} &= \frac{1}{2 + \text{SNR}_{w 2}}.
\end{align*}$$

V. IMPERFECT CANCELLATION

In Sections III and IV, it is assumed that the canceller can estimate the phase shift, the attenuation, and the delay perfectly. There are always errors in estimation of these parameters, however, due to the noise and the co-channel interference. Therefore, even though the canceller reconstructs correct signals after the initial decisions, it might not remove the co-channel interference completely; there is residual interference caused by imperfect cancellation.

Suppose a fraction $\beta$ of cancelled signal power is left in the composite signal. For the conventional canceller, the expected number of cancelled symbols is $(\beta (K-1)L - Ap_c(K,L)/2)$.

so the SNR after the cancellation is

$$\text{SNR}_{2}(\beta) = \frac{E_s}{2Ap_c(K,L)E_s} + \frac{2\beta(K-1)L - Ap_c(K,L)/2}{2\beta(L-1)} + \text{No}.$$
For the proposed canceller, the expected number of cancelled symbols is given by \((K-1)L - Ap_s(K,L)/2\), so the SNR after the cancellation is

\[
\text{SNR}_{c2}(\beta) = \frac{E_s}{2A_p(K,L)E_s + 2B(K-1)L - Ap_s(K,L)/2} + \frac{2(B-1)E_s}{3C_p} + N_0.
\]

From \(\text{SNR}_{c2}(\beta)\) and \(\text{SNR}_{c2}(\beta)\), the approximate error rate after the cancellation can be obtained by equations similar to (20) and (28).

VI. INTERCELL INTERFERENCE AND SOFT HANDOFF

In previous sections, we assumed a single-cell configuration. In this section, we consider a multicell configuration, consisting of a cluster of seven cells as depicted in Fig. 6. Each cell is divided into three sectors. Due to the sectorization, the interference at the base station comes from only two adjacent cells. Assuming perfect power control, wave propagation with path loss proportional to the fourth power of distance, and uniformly distributed users in a sector, the intercell interference energy from one of the adjacent cells is

\[
I_{\text{int}}(d_mR) = 2E_{bs} \cdot 3K \left[ 2 \cdot d_m^{2N} \left( \frac{d_m^2}{d_m^2 - 1} \right) - 4 \cdot d_m^2 - 6 \cdot d_m^2 + 1 \right]
\]

where \(d_mR\) is the distance between the base stations (in this case \(d_m = 2\)), and \(E_{bs}\) is the total symbol energy defined as \(LE_s\). Therefore, in (15), the SNR per path becomes

\[
\text{SNR} = \frac{E_s}{2A_p(K,L)E_s + 2B(K-1)L - Ap_s(K,L)/2} + 2 \cdot \frac{2E_{bs}(2R)}{3C_p} + N_0.
\]

If soft handoff is employed, mobiles close to cell boundaries might transmit to two base stations. Suppose the handoff probability is \(P_{\text{off}}\), the period of the soft handoff is short, and the mobiles that are in the soft handoff process transmit their signals with energy \(E_{bs}\) to the base stations, the number of users using the soft handoff can be considered as

\[
K' = [(1 + P_{\text{off}})K]
\]

where \([X]\) is the nearest integer less than or equal to \(X\). Also, the intercell interference decreases to

\[
I_{\text{int}}'(2R) = I_{\text{int}}(2R) - E_{bs} \cdot P_{\text{off}} K.
\]

The performance of the CCI cancellers can be calculated by adding the intercell interference term and replacing \(K\) with \(K'\). Therefore, the path SNR after the cancellation for the conventional canceller is

\[
\text{SNR}_{c2} = \frac{E_s}{2A_p(K',L)E_s + 2B(K-1)L - Ap_s(K',L)/2} + \frac{2B(2R)}{3C_p} + N_0.
\]

and for the proposed canceller the path SNR after the cancellation is

\[
\text{SNR}_{c2} = \frac{E_s}{2A_p(K',L)E_s + 2B(K-1)L - Ap_s(K',L)/2} + \frac{2B(2R)}{3C_p} + N_0.
\]

VII. RESULTS

The following results assume full interleaving. Results for finite interleaving may also be readily obtained. Since they are application specific, we omit them in this paper. We also assume that the processing gain \(G_p\) is 128 and the coding rate \(1/3\) (8-ary orthogonal convolutional code) with optimum generator polynomials given in [19] for the constraint length of 3 to 7 and in [8] for the constraint length of 9.

Fig. 7 shows the BER performance of the convolutional orthogonal code on the multipath Rayleigh fading channel. In the figure, the points are the simulation results and the line is the theoretical performance calculated from the approximation with the first four terms in (20). For the results shown in Fig. 7, the number of paths is assumed to be three. As the simulated points are close to the theoretical values with \(w = d_{\text{ave}} = d_{\text{ave}} + 3\), we assume that the approximation is appropriate.

From Figs. 8–24, the performances without a canceller, with the conventional canceller, and with the proposed canceller are presented for different conditions. The performance without a canceller is calculated by (20)–(22) with \(\text{SNR}_{c2}\) replacing \(\text{SNR}_{c2}\). Figs. 8 and 9 show the system SNR versus equivalent Gaussian SNR with the number of paths equal to 1 and 3, respectively. The system SNR is the SNR just before the final maximum likelihood decision (after the cancellation), defined as \(L \cdot \text{SNR}_{c2}\) or \(L \cdot \text{SNR}_{c2}\). The equivalent Gaussian SNR/symbol is the SNR without the CCI, which is defined as \(E_{bs}/N_0 = LE_s/N_0\). Thus the difference between these two \(\text{SNR}\) is the CCI, which cannot be removed by the canceller, shown in Figs. 10 and 11. From Figs. 10 and 11, it is clear that the proposed canceller cancels more CCI and provides better SNR than the conventional canceller when the SNR/symbol is not too low. Especially when \(K = 20\) (the number of
users that transmit the signal at the same time), the proposed canceller removes most of the CCI and provides a nearly interference free signal for the final maximum likelihood decision. Also the proposed canceller works very well when there are multiple paths in the channel which cause more CCI. In Fig. 9, when $K = 20$, the system SNR of the conventional canceller does not improve in the high equivalent Gaussian SNR area since the CCI dominates the accuracy of the initial decision. On the other hand, the proposed canceller improves the system SNR as the equivalent Gaussian SNR becomes larger. If $K$ is 40, however, there are three paths, and the equivalent Gaussian SNR is less than 12 dB, then the re-encoding process produces too much interference. Thus, in this case, the proposed canceller is inferior to the conventional canceller.

Figs. 12 and 13 show the BER versus SNR with the number of paths equal to one and three, respectively. SNR/symbol is defined as $\frac{E_s}{N_0}$. When $K = 20$, the performance of the proposed canceller is very close to the performance with $K = 1$ and superior to the conventional canceller by 2 dB.
with one path and by 1 dB with three paths at $\text{BER} = 10^{-3}$. If $K = 40$, the difference between the BER of the two methods is much larger than the $K = 20$ case. For example, for constraint length 9, with one path, and at $\text{BER} = 10^{-3}$, the difference is more than 6 dB as seen in Fig. 12.

Figs. 14 and 15 show the BER versus the number of users for one and three paths, respectively. It is shown that the number of simultaneous users at $\text{BER} < 10^{-3}$ increases by a factor of 1.5~3 with the new canceller as compared with the conventional canceller. It is also clear that up to about 20 users the proposed canceller removes almost all CCI when SNR/symbol equals to 10 dB as the error floor due to thermal noise is observed.

Figs. 16 and 17 show the BER versus the constraint length of the convolutional code for one and three paths, respectively. From the figures, a significant performance improvement can be obtained by using a longer constraint length, especially when $K$ is large. As the constraint length becomes longer, the BER decreases linearly. Therefore the canceller should use
a convolutional orthogonal code that has a large constraint length.

The proposed cancellation scheme can also be viewed as an effective way of reducing the system implementation complexity as far as Viterbi decoding is concerned. For example, for $K = 20$ and required $BER < 10^{-3}$, from Fig. 16, the conventional canceller requires the code constraint length equal to 5, while the proposed canceller requires the constraint length equal to 4. Thus the number of states in the trellis diagram is reduced by a factor of four and the memory size for storing trellis paths is reduced by a factor of $6.67 (= 4 \times \frac{5}{3})$ as the decoding depth also reduces. This represents more than a factor of 4 reduction in Viterbi decoding complexity. Recall that the number of Viterbi decoders could be doubled in order to shorten the delay. Even with this taken into account, the overall decoding complexity is reduced by more than a factor of 2.

Fig. 18 shows the BER versus number of paths that are available for the square-law combining. When $K$ is 20, the BER of both cancellers improves as the number of paths increases from one to two. If the number of paths further increases, however, the performance deteriorates and the proposed canceller is affected more than the conventional canceller. When $K$ is 40, as the number of paths increases, the BER becomes worse—as there is too much interference due to too many paths—and the low SNR produces more symbol errors in the decoding process.

Figs. 19 and 20 show the BER performance in the presence of imperfect cancellation for one and three paths. The estimation error $\beta$ is the fraction of cancelled signal power that is left in the composite signal. The estimation error has almost no affect on either the conventional or the proposed canceller when $\beta < 10^{-2}$, and both cancellers work well until $\beta = 10^{-1}$. From these results, it is concluded that CCI cancellers are useful even with an estimation error.

Figs. 21 and 22 show the BER versus the number of users with an estimation error for one and three paths, respectively. It is clear that the number of simultaneous users with an estimation error $\beta = 10^{-1}$ is almost the same as that without estimation error for both the conventional canceller and the proposed canceller. Therefore, the proposed canceller is effective and better than the conventional canceller even in the presence of the interference caused by imperfect cancellation.
Fig. 18. BER versus number of paths, SNR/symbol = 10 dB, code rate = 1/3, constraint length = 9, 8-ary orthogonal signalling.

Fig. 19. BER versus estimation error, number of users = 40, number of paths = 1, SNR/symbol = 10 dB, code rate = 1/3, constraint length = 9.

Figs. 23 and 24 show the BER performance for the multicell situation and the effects of the cancellation with soft handoff and one and three paths per channel. The number of users per sector is the number of users that transmit in one sector at the same time. Though the handoff probability $P_{\text{off}}$ depends on many parameters, $P_{\text{off}}$ is set to 0.087 following [21] as an example. It is shown that the combination of the soft handoff and the proposed canceller improves the capacity, while the conventional canceller does not make any significant difference in the capacity. The reason is that if the number of users per sector is small, very few users utilize the soft handoff.

Therefore, the impact of the canceller is not significant as it does not cancel signals from mobiles in an adjacent cell that are not in soft handoff to the reference cell. On the other hand, if there are a large number of users per sector, the conventional canceller does not work well. In this case, even though some mobiles outside the cell transmit to the base station during soft handoff, the conventional canceller might not cancel the interference from them. Since the proposed canceller still works with a large number of users in the low BER region, it shows the capacity improvement with the use of the soft handoff.
Fig. 22. BER versus number of users with estimation error, number of paths = 3, coding rate = 1/3, constraint length = 9, 8-ary orthogonal signalling.

Fig. 23. BER versus number of users/sector, number of paths = 3, SNR/symbol = 15 dB, code rate = 1/3, constraint length = 9, handoff probability = 0.087.

VIII. CONCLUSION

In this paper, we have proposed and evaluated a new co-channel interference cancellation technique on a multipath Rayleigh fading channel that improves the accuracy of the initial decision by utilizing the error correction capability of orthogonal convolutional codes. The performance of the proposed canceller has been computed on the multipath Rayleigh fading channel and compared with the conventional canceller. It has been shown that the proposed canceller offers 1.5~3 times higher user capacity than the conventional canceller.

We have also shown that the canceller significantly benefits from the use of a large constraint length of an orthogonal convolutional code. For the same performance, the proposed canceller can effectively reduce the decoding complexity. The proposed canceller works even in the presence of the interference due to imperfect cancellation. In the multicell configuration, the proposed canceller has been shown to improve the user capacity in combination with soft handoff.

APPENDIX

A. Interference

The self interference on the output of the I-channel’s mth correlator is

$$I_{I,d}^{m}(t) = \frac{P}{T_s} \int_0^{T_s} \sum_{j=1}^{L_c} \gamma_{Kd}(t - r_{Kd}) \gamma_{Kd}(t - r_{Kd}) W^m(t - r_{Kd}) W^m(t - r_{Kd}) dt + \frac{\cos \theta_{kj}}{2} + \frac{\sin \theta_{kj}}{2} dt$$  \hspace{1cm} (A1)

The interference due to the other paths of the same user is

$$I_{I,d}^{m}(t) = \frac{P}{T_s} \int_0^{T_s} \sum_{j=1}^{L_c} \gamma_{Kd}(t - r_{Kd})$$  \hspace{1cm} (A2)
The interference due to the other users is
\[
I_{ij}^{\text{other}} = \sqrt{\frac{P}{T_s} \sum_{t=1}^{L_i} \sum_{j=1}^{L_j} \int_0^T \gamma_{ij}(t) \, dt} \quad (A3)
\]
\[
\times \left[ W^T(t - \tau_{ij})W^m(t - \tau_{ij}) \cdot C_i(t - \tau_{ij})C_i(t - \tau_{ij})C_j(t - \tau_{ij})P_j(t - \tau_{ij})P_j(t - \tau_{ij}) \right.
\]
\[
\times \left. \frac{\cos \theta_i + W^T(t - T_d - \tau_{ij})W^m(t - \tau_{ij})}{2} \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \right. \r
C. Probability of Selecting the Incorrect Path

From (13), the decision variable for the \(n\)th user at the \(m\)th correlator is

\[
S_k(n) = \sum_{j=1}^{L_n} \left\{ [Z_{1,j}^k(n) + Z_{Q,j}^k(n)]^2 + [Z_{1,j}^k(n) - Z_{Q,j}^k(n)]^2 \right\},
\]

(C1)

Thus \(S_k(n)\) has a chi-square probability distribution with \(2L_n\) degrees of freedom. Assuming there are \(L_n = L\) paths and \(w\) is the distance of the coded sequence along the wrong path in the state diagram from the coded sequence along the correct path, the decision variable between paths has a chi-square probability distribution with \(2wL\) degrees of freedom

\[
\sum_{j=1}^{wL} \left\{ [Z_{1,j}^k(n) + Z_{Q,j}^k(n)]^2 + [Z_{1,j}^k(n) - Z_{Q,j}^k(n)]^2 \right\} = 2\sum_{l=1}^{wL} \{\tau_l\}^2.
\]

(C2)

Therefore, the probability of choosing a wrong path is

\[
Pd(w, L) = P \left[ \sum_{l=1}^{2wL} \{\tau_l\}^2 > \sum_{l=1}^{2wL} \{\tau_l\}^2 \right]
\]

(C3)

where \(\tau_l\) is the output of a matched filter corresponding to the correct path while \(\tau_l^*\) corresponds to the wrong path. Therefore \(\{\tau_l\}\) and \(\{\tau_l^*\}\) are statistically independent zero-mean Gaussian random variables with variance

\[
\begin{align*}
\langle\tau_l^2\rangle &= \frac{E[|E_s|] + N_0 + I_\infty}{2} \propto \sigma^2, \\
\langle\tau_l^*2\rangle &= \frac{N_0 + I_\infty}{2} \propto \sigma^2.
\end{align*}
\]

(C4)

(C5)

where \(N_0\) is the noise and \(I_\infty\) is the co-channel interference. The density function of a chi-square distribution with \(2wL\) degrees of freedom is [22, p. 542]

\[
p(\alpha) = \frac{1}{(wL - 1)! \sigma^2} \alpha^{(wL - 1)/2} \exp\left(-\frac{\alpha^2}{2\sigma^2}\right)
\]

(C6)

\[
p'(\beta) = \frac{1}{(wL - 1)! \sigma^2} \beta^{(wL - 1)/2} \exp\left(-\frac{\beta^2}{2\sigma^2}\right).
\]

(C7)

The probability of selecting the wrong path is

\[
Pd(w, L) = \int_{-\infty}^{\infty} p(\alpha)d\alpha \int_{-\infty}^{\infty} p'(\beta)d\beta.
\]

(C8)

Carrying out the integrations of (C8) by parts, we obtain first

\[
\int_{-\infty}^{\infty} p'(\beta)d\beta = \exp\left(-\frac{\alpha^2}{2\sigma^2}\right) \times \left[ 1 + \frac{\alpha^2 + \alpha^2/2\sigma^2 + \alpha^2/2\sigma^2/2}{1!} + \cdots \right.
\]

\[
\left. + \frac{\alpha^2/2\sigma^2}{(wL - 1)!} \right].
\]

(C9)

and then

\[
Pd(w, L) = \left( \frac{\sigma^2}{\sigma^2 + \sigma^2} \right)^wL 
\times \left[ 1 + \frac{wL}{1} \left( \frac{\sigma^2}{\sigma^2 + \sigma^2} \right)^2 + \cdots \right]
\]

(C10)

The error probability depends on the average received signal energy \(E_s = E[|E_s|]\) through

\[
p = \frac{\sigma^2}{\sigma^2 + \sigma^2} = \frac{1}{2 + E_s/(N_0 + I_\infty)}
\]

(C11)

and

\[
1 - p = \frac{\sigma^2}{\sigma^2 + \sigma^2}.
\]

(C12)

Therefore

\[
Pd(w, L) = p^{wL} \sum_{l=0}^{wL-1} \left( \frac{wL - 1 + h}{h} \right)(1-p)^h.
\]

(C13)

ACKNOWLEDGMENT

The authors would like to express their deep gratitude to the members of CITR Lab, especially to M. Wang and R. Kerr for their help in preparing this paper.

REFERENCES


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