A Multiuser Interference Cancellation Technique Utilizing Convolutional Codes and Orthogonal Multicarrier Modulation for Wireless Indoor Communications

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Abstract—This paper suggests that multicarrier modulation reduces the complexity and the delay caused by the multiuser interference cancellation process utilizing convolutional codes.

For spread spectrum multiple access, multiuser interference (interference due to signals from other users) limits the performance of the communication link. To remove this interference, a multiuser interference cancellation technique which utilizes orthogonal convolutional codes has been proposed. In CDMA systems, however, this technique requires the large interleavers and the huge memory, or the artificial multipath diversity and the RAKE system to achieve sufficient coding gain if it is applied for wireless indoor communications and fading is slow compared to the data rate.

To reduce the complexity of the canceller, multicarrier modulation is employed as it provides frequency diversity gain and coding gain without the interleavers or the RAKE system. This paper shows that the multicarrier modulation reduces the complexity of the canceller and still provides sufficient coding gain in order to cancel the multiuser interference. The canceller with decoding in the initial decision and multicarrier modulation improves capacity by a factor of 1.4 as compared with the canceller without decoding.

I. INTRODUCTION

RECENTLY, there has been much interest in application of code-division multiple-access (CDMA) in cellular and wireless personal communications. One of the reasons is its possibility of achieving a greater capacity per unit bandwidth than other multiple access schemes such as frequency-division multiple-access (FDMA) or time-division multiple-access (TDMA).

To increase CDMA capacity, especially for the uplink which limits the capacity of the cellular CDMA systems [1], a detection scheme which utilizes received signals of multiple users has been developed more than ten years ago [2], [3]. The optimum detector (minimum error probability detector) was proposed by Verdu [4], which consists of a matched filter front-end followed by a Viterbi algorithm (VA). Although notable performance gains are obtained, the detector requires the signature waveforms and the received signal amplitudes of all users and its complexity grows exponentially with the number of users.

Suboptimum multiuser detectors whose complexities are less than the optimum multiuser detector have also been considered. In [5], sequential decoding is applied instead of the VA as used in the optimal detector. In [6]–[9], the decorrelating detector, which multiplies the inverse cross-correlation matrix with the matched filter outputs, has been investigated. In [10] and [11], the decorrelator is combined with a decision-feedback detector. Although these multiuser detectors show performance close to the optimum detector with reasonable computation, these detectors must calculate the inverse cross-correlation matrix. In [12], the minimum mean-square error (MMSE) detector was proposed, which calculates the inverse matrix adaptively. This detector outperforms the decorrelating detector when the background noise dominates the performance. Adaptive algorithms are also utilized for multiuser detection in [13] and [14].

In [15] and [16], tentative decision-based multiuser detectors have been investigated. In a multistage structure, the first stage consists of a bank of conventional detectors. The second and third stages assume that the previous decisions are correct, calculate the co-channel interference caused by undesired users' signals, and remove them from the correlator output of the desired user's signal.

The multiuser detectors mentioned above require the receiver to have, a priori or through training, the knowledge of the sequences' cross-correlation. Unfortunately, these detectors are not useful if long pseudo noise (PN) sequences are used as signature sequences to separate the users' channels since the cross-correlation varies with different phases of the signature sequence. If the receiver has to calculate the cross-correlation for each symbol, the computational complexity grows exponentially with the number of users.

Multiuser detection techniques which do not use the knowledge of the sequences' cross-correlation and whose complexity grows only linearly with the number of users, have also been proposed [17]–[19]. In [19], users are detected at once while in [17] and [18] they are detected successively. In these techniques, the receiver reconstructs the other users' transmitted signals by using initial decisions about the other users' signals. These receivers then use the estimates (recon-
structured signals) to remove the co-channel interference from the composite received signal. These methods do not require the knowledge of the cross-correlation between the spreading sequences. However, there is residual interference due to symbol errors in the initial decision [20]. The performance of the canceller depends on the performance of the initial decision. Therefore, it is desirable to improve the accuracy of the initial decision.

In [21], [22], a new co-channel interference cancellation technique which utilizes orthogonal convolutional codes to improve the accuracy of the initial decision has been proposed. In this technique, received signals are both demodulated and decoded. Then the resulting bit streams are re-encoded and respread to be subtracted from the composite received signals. On account of the error correcting capability of orthogonal convolutional codes, the residual interference is reduced as the accuracy of the initial decision is increased. The proposed cancellation technique achieves capacity improvement by a factor of 1.5 to 3 as compared with the conventional canceller.

Nevertheless, the complexity of the proposed canceller depends on channel characteristics. From some measurements, the indoor wireless channel shows very slow Rayleigh fading with short delay spread and multipath is hardly available if the system bandwidth is about 10 MHz [23], [24]. If fading is slow compared to the data rate, large interleaving is required to provide sufficient coding gain [25]. This is not preferable since the proposed canceller requires additional interleavers and de-interleavers and the delay owing to the cancellation process increases. If delay spread is short and only correlated paths are available, combination of artificial multipath and a RAKE system may be required to achieve antenna diversity [26]. This system also increases the complexity of the canceller since the RAKE system has to follow the delays of those paths by employing a delay line and the canceller requires the estimations of the attenuation, the gain, and the delay of each additional path.

Multicarrier (MC) modulation is attractive in terms of the cancellation as it eliminates multipath and achieves frequency diversity without complex RAKE systems or interleavers. Therefore, it is possible to construct a simple and effective canceller which utilizes convolutional codes and requires short processing delay. In this paper, we investigate a multiuser interference (interference due to the other users on the same carrier and from the different carriers) cancellation technique utilizing convolutional codes for a MC CDMA system for high speed indoor communications.

The MC CDMA system considered in this paper is based on the system described in [27]. Even though the MC CDMA system investigated in [28] also utilizes convolutional codes and shows good performance results, it is difficult to employ the cancellers since the power of the transmitted signals on each carrier is too small to be estimated and reconstructed. The proposed cancellation techniques are evaluated under the same total processing gain with different number of carriers in order to compare their cancellation effects. No interleaving is considered in this paper to reduce processing delay as some applications such as low resolution video are delay sensitive even if the interleaving further improves the performance of the system. Although a multiuser interference cancellation technique for MC CDMA has also been proposed by Sourour and Nakagawa [29], it requires very high speed signal processing for the first stage of the cancellation and does not utilize any coding scheme. Thus, it is suitable for voice, but not for high speed data communications.

This paper is organized as follows. In Section II, a MC CDMA system is presented. In Section III, multiuser interference cancellation techniques without decoding and with decoding in the initial decision are described and their performance is derived. Performance results are shown in Section IV. Section V presents our conclusions.

II. System Model

A. Radio Channel Model

Concerned high speed indoor communication systems are supposed to accommodate not only high quality voice services, but also data, facsimile, and video services for personal use [30]. Thus, the data rate is up to 128 kb/s while the maximum Doppler shift is assumed as 2 Hz in this paper [26], [31]. To evaluate the performance of the system, the exponential multipath profile model is assumed as shown in Fig. 1(a) [24], [32]. Correlation between the envelopes of the signals at two different frequencies is given by

$$\rho = \frac{1}{1 + (\Delta \omega \cdot \sigma)^2}$$  \hspace{1cm} (1)

where $\Delta \omega$ is the frequency separation and $\sigma$ is the delay spread [31]. Fig. 1(b) shows the relationship between the number of carriers and the chip duration on each carrier. As the number of carriers increases, the chip duration exceeds the multipath delay as compared in Fig. 1(a) and (b). Channel measurement results show that they have more severe fading than the best fit Rician distribution [23]. Therefore, the channel on each carrier can be modeled as a very slow Rayleigh-fading channel. The second and the following resolvable paths are neglected as the most of the signal energy is concentrated on the first resolvable path [23]. In addition, it is difficult to track those weak paths at low signal-to-noise ratio (SNR) and utilize them as multipath diversity with multiuser interference [33].

B. Orthogonal Multicarrier CDMA System

A system model is based on the system proposed in [27] with convolutional codes. A structure of a transmitter is shown in Fig. 2(a). Suppose there are $K$ users in a single cell. The bit streams with bit duration, $T_b$, are first encoded by a rate $1/R$ convolutional code. Thus, the symbol duration is $T_s = T_b / R$. The same coded symbol is fed to $S$ different parallel branches in the transmitter and then spread by a part of a long PN sequence (a $m$-sequence in this paper) assigned for each user. Different parts of the PN sequence are used for the different branches to accommodate more users compared to the system which employs the same PN sequence on every branch. The period of the PN sequence is much longer than the total processing gain and the part used for spreading changes symbol by symbol on each branch for the security purpose.
The length of the spreading sequence per symbol on each branch is adjusted in order to maintain the same total processing gain per each symbol while the number of branches influences the total processing gain in the system described in [27]. The reason that the assumed system employs the same total processing gain with the different number of carriers is to compare the effectiveness of the cancellation techniques.

If the total processing gain is increased so as to preserve the same transmission bandwidth, the multiuser interference is reduced without cancelling it. The chip duration $T_c$ is then equal to $T_s/N$ where $N = G_p/RS$ is the processing gain on each branch, $G_p$ is the total processing gain, and $S$ is the number of branches. The spread symbol is then modulated using BPSK and by the carrier signals. The frequencies of the carriers are assigned so that the frequency spectrum of the modulated signals overlaps each other and still achieves orthogonality as shown in Fig. 2(b). For the special case, when the number of carriers $S$ is equal to one, the described system turns to be the conventional CDMA system.

Let $b_k(t)$ indicate the coded symbol stream and $a_{k,m}(t)$ denote the spreading sequence waveform of the $m$th carrier of the $k$th user. The transmitted signal of the $k$th user is written as

$$s_k(t) = \sum_{m=1}^{S} \sqrt{2W} b_k(t) a_{k,m}(t) \cos(\omega_m t + \phi_{k,m})$$

(2)

where $W$ is the transmitted power of the signal per carrier, $\omega_m$ is the frequency of the $m$th carrier, and $\phi_{k,m}$ is the random phase of the $m$th carrier of the $k$th user. The chip pulse shape is assumed to be rectangular. To achieve orthogonality between the carriers, the frequencies of the carriers are assigned as

$$\omega_m = \omega_1 + (m - 1) \frac{2\pi}{T_c}.$$  

(3)

At the base station, the signals from the $K$ users arrive asynchronously. As the multipath delay spread is small on the indoor channel, the coherent bandwidth is larger than the bandwidth of the signal modulated by each carrier. Therefore, it is assumed that the channels on which the carriers are transmitted show Rayleigh fading and are correlated one another. The received signal is then given by,

$$r(t) = \eta(t) + \sqrt{2W} \sum_{k=1}^{K} \sum_{m=1}^{S} \beta_{k,m} b_k(t - \tau_k) \times a_{k,m}(t - \tau_k) \cos(\omega_m t + \phi_{k,m})$$

(4)

where $\eta(t)$ is the additive white Gaussian noise (AWGN) with zero mean and two sided power spectral density $N_0/2$, $\{\beta_{k,m}\}$ account for the Rayleigh-fading envelope of the $m$th carrier of the $k$th user and

$$\phi_{k,m} = (\phi_{k,m} + \gamma_{k,m} - \omega_m \tau_k) \mod 2\pi$$

(5)

where $\gamma_{k,m}$ and $\tau_k$ account for the phase shift and the delay due to asynchronous transmission of the $m$th carrier of the $k$th user. $\{\gamma_{k,m}\}$ and $\{\tau_k\}$ are uniform random variables in $[0, 2\pi]$ and $[0, T_s]$ respectively. It is assumed that $E[\beta_{k,m}^2] = 2$, where $E[x]$ is the expected value of $x$. $\{\beta_{k,m}\}$ are correlated among the different carriers by the correlation coefficient $\rho$ given in (1).

Without loss of generality, we assume that receiver 1 is the reference user and let $\tau_1 = 0$. A model of a MC CDMA receiver for the user 1 is shown in Fig. 3. After the demodulation, the output of the correlator for the $q$th carrier is

$$Z_{1,q} = \int_{0}^{T_s} r(t) a_{1,q}(t) \cos(\omega_{q1} t + \varphi_{1,q}) dt$$

$$= T_s \sqrt{\frac{W}{2}} (\eta_{1,q} + D_{1,q} + I_{1,q} + J_{1,q})$$

(6)
where $\eta_{q,t}$ is the Gaussian random variables (rv’s) with zero mean and variance $1/(2\gamma_t)$, where $\gamma_t = WT_s/N_0$, and $D_{1,q} = \beta_{1,q}b_1(t)$ is the desired output signal. $J_{1,q}$ and $J_{1,q}$ are the co-channel interference and the adjacent channel interference on the carrier $q$. The co-channel interference on the carrier $q$ is given by

$$I_{1,q} = \frac{1}{T_s} \sum_{k=2}^{K} \beta_{k,q} \cos(\varphi_{k,q} - \varphi_{1,q})$$

$$\times \int_{0}^{T_s} b_k(t - \tau_k)a_{k,q}(t - \tau_k)a_{1,q}(t)dt$$

$$= \frac{1}{T_s} \sum_{k=2}^{K} g_{k,q} [b_{k-1,1,q}(\tau_k) + b_{0,1,q}(\tau_k)]$$

(7)

where

$$g_{k,q} = \beta_{k,q} \cos(\varphi_{k,q} - \varphi_{1,q})$$

(8)

We assume that $\{g_{k,q}\}$ are the Gaussian rv’s with zero mean and unit variance. This approximation is known to provide good results for a large number of users and long codes [29]. $b_{k}^{q}$ is the $k$th symbol of the $q$th user, and $R_{k,q}(\tau)$ and $\hat{R}_{k,q}(\tau)$ are the continuous time partial code cross-correlations of the spreading sequences between the $k$th and the $q$th user and the $k$th user on the $q$th carrier, which are defined in [35]. From [27] and [35], the variance of (7) is given by

$$E[(I_{1,q})^2] = \frac{1}{3N3} \sum_{k=2}^{K} r_{k,q,1,q}$$

(9)

where

$$r_{k,q,1,q} = 2\mu_{k,q,1,q}(0) + \mu_{k,q,1,q}(1),$$

(10)

$$\mu_{k,q,1,q}(n) = \sum_{l=-1-N}^{N-1} C_{k,q,1,q}(l)C_{k,q,1,q}(l+n)$$

(11)

and $C_{k,q,1,q}$ is the discrete aperiodic cross-correlation function between the spreading sequences of the $k$th user and the $q$th user on the $q$th carrier [35]. The adjacent channel interference on the carrier $q$ is given by

$$J_{1,q} = \frac{1}{T_s} \sum_{k=2}^{K} \sum_{m=1}^{S} \beta_{k,m} \int_{0}^{T_s} b_k(t - \tau_k)a_{k,m}(t - \tau_k)a_{1,q}(t)$$

$$\times \cos(\omega_m - \omega_q)t + \varphi_{k,m} - \varphi_{1,q})dt$$

$$= \frac{1}{N} \sum_{k=2}^{K} \sum_{m=1}^{S} g_{k,m,q}$$

$$\times \{[b_{k-1,1,q}(l_k + 1 - N) - C_{k,m,1,q}(l_k - N)]$$

$$+ b_{0,1,q}(l_k + 1) - C_{k,m,1,q}(l_k)]\}$$

$$\times x_k \text{Sinc}(\pi(m - q)x_k)$$

(12)

where

$$g_{k,m,q} = \beta_{k,m} \cos[\pi(m - q)x_k + \varphi_{k,m} - \varphi_{1,q}].$$

(13)

$\{g_{k,m,q}\}$ are approximated as the zero mean Gaussian rv’s with unit variance. $\{l_k = [\tau_k/T_s]\}$ are independent and identically distributed (i.i.d.) discrete rv’s uniform in $[0, N-1]$, $\{\tau_k = [\tau_k/T_c]\}$ are i.i.d. rv’s uniform in $[0,1)$, and $C_{k,m,1,q}$ is the discrete aperiodic cross-correlation function between the spreading sequences of the $m$th carrier of the $k$th user and of the $q$th carrier of the $1$st user [35]. The variance of $J_{1,q}$ is given by

$$E[(J_{1,q})^2] = \frac{1}{\pi^2 N^3} \sum_{m=1}^{S} \sum_{n=0}^{S} \sum_{m\neq n} \frac{1}{(m - q)^2}$$

$$\times \sum_{k=2}^{K} \mu_{k,m,\gamma,1,q}(0) - \mu_{k,m,\gamma,1,q}(1)$$

(14)

where

$$\mu_{k,m,\gamma,1,q}(n) = \sum_{l=-1-N}^{N-1} C_{k,m,1,q}(l)C_{k,m,1,q}(l+n).$$

(15)

Though estimations of the delay, $\{\tau_k\}$, the attenuation, $\{\beta_{k,m}\}$, and the phase shift, $\{\varphi_{k,m}\}$, should be available for the cancellation, and make maximal ratio combining possible, equal gain combining is employed throughout this paper as following [29]. Due to the equal gain combining, the symbol decision variable is given by

$$Z_q = \sum_{q=1}^{S} Z_{1,q}.$$  

(16)

The demodulated symbols are then decoded by the soft decision Viterbi decoder. The upper bound on the bit error probability can be obtained as [36, p. 459]

$$P_b < \sum_{d=d_{\text{min}}}^{\infty} B_d P_d(d)$$

(17)

where $B_d$ is the total number of nonzero bits on all weight $d$ paths on the trellis diagram [37, p. 327]. $d_{\text{min}}$ is the minimum distance. Using a Gaussian approximation, the probability of selecting the incorrect path on the trellis diagram by a soft decision Viterbi decoder, $P_d(d)$, is

$$P_d(d) = \frac{1}{\sqrt{2\pi}} \int_{0}^{\infty} \frac{1}{\sqrt{2\pi}} \text{erfc} \left( \sqrt{\frac{d}{2V(E)}} \right) f(E) dE$$

(18)

where

$$E = \sum_{m=1}^{S} \beta_{1,m},$$

(19)

$$V = \frac{S}{2\gamma_s} + \sum_{q=1}^{S}$$

$$\times \left\{ \frac{1}{3N^3} \sum_{k=2}^{K} r_{k,q,1,q} + \frac{1}{\pi^2 N^3} \sum_{m=1}^{S} \sum_{n=0}^{S} \frac{1}{(m - q)^2} \right\}$$

(20)
and $f(E)$ is the probability density function of the sum of the correlated Rayleigh rv's. If the weight spectra $B_d$ is known from $d_{cee}$ to infinity in (17), the upper bound can be calculated. However, the bound becomes looser as the bit-error rate (BER) becomes worse. The first seven terms are used to get a performance approximation through this paper [21], [22]. As for the term $f(E)$, though it is possible to calculate it, the calculation includes summations with infinite number of terms and the probability distribution is strongly influenced by the number of terms [29], [38]. Therefore, as following [29], $f(E)$ is generated by a computer up to 1 000 000 times, and substituted in the erfc($\cdot$).

III. INTERFERENCE CANCELLATION

A. Multisuser Interference Cancellation Without Decoding in the Initial Decision

The concept of a multisuser interference cancellation technique which is explained here is the same as [19]. It estimates the transmitted signals of the interfering users and reconstructs them. A model of a receiver using the multisuser interference canceller without decoding in the initial decision is shown in Fig. 4. User 1 is the reference user. As shown in Fig. 4, every user's received signals on each carrier are first despread and combined, and the transmitted symbols are detected. Utilizing these initial decisions, the coherent interference and the adjacent channel interference are reconstructed with the estimations of the delay, $\tau_k$, and the attenuation on the concerned carrier, $\tilde{g}_{k,m}$, and on the other carriers, $\tilde{g}_{k,m,q}$. The reconstructed interference is removed from the composite received signal on each carrier. After that, user 1's signal is despread, and decoded.

When an error in the initial decision arises, the multisuser interference cancellation process actually increases the interference power by four times instead of cancelling one [20]. Thus, the performance of the canceller depends on the performance of the initial decision.

After the cancellation, residual interference can be written as

$$I_{k,q}^{\text{res}} = I_{k,q} - \frac{1}{T_{\text{sc}}} \sum_{k=2}^{K} \tilde{g}_{k,q} X_k^{\text{res}}$$

and

$$J_{k,q}^{\text{res}} = J_{k,q} - \frac{1}{N} \sum_{k=2}^{K} \sum_{m,q} \tilde{g}_{k,m,q} X_k^{\text{res}}$$

where $\{X_k^{\text{res}}\}$ are i.i.d. rv's that take value $-1$ with probability $P_{\text{e}}^{\text{res}}$ and 1 with probability $1 - P_{\text{e}}^{\text{res}}$. $P_{\text{e}}^{\text{res}}$ is the symbol error probability of the initial decision which is given by

$$P_{\text{e}}^{\text{res}} = \int_0^\infty \frac{1}{2} \text{erfc} \left( \frac{E}{\sqrt{2V}} \right) f(E) dE.$$

As for the estimations of the attenuation, following [29], it is assumed that $\tilde{g}_{k,q} = g_{k,q} + \Gamma_{k,q}$ and $\tilde{g}_{k,m,q} = g_{k,m,q} + \Gamma_{k,m,q}$ where all $\{\Gamma\}$ are i.i.d., independent of all $\{g\}$, the Gaussian rv's, caused by the noise and the multisuser interference, with zero mean and variance $\varepsilon$. As for the estimations of the delay, it is assumed that $\tilde{\tau}_k = (l_k + \tilde{\tau}_k) T_{\text{sc}}$, where $\tilde{\tau}_k$ is a uniform in $[-b,b)$ due to the asynchronous transmission, where $b \leq 0.5$. Substituting (7) and (12) in (21) and (22), and taking the second moment of (21) and (22), we get

$$E \left[ (I_{k,q}^{\text{res}})^2 \right] = \frac{1}{3N^3} \sum_{k=2}^{K} \left( 2 + e \right) r_{kq,1q}$$

$$- 6H(b) (G_{kq,1q} + G_{kq,1q})$$

$$+ 12H(b) P_{\text{e}}^{\text{res}} (G_{kq,1q} + G_{kq,1q})$$

where $H(b)$ is the entropy of the channel, $G_{kq,1q}$ is the gain of the channel, and $P_{\text{e}}^{\text{res}}$ is the symbol error probability of the initial decision.
and

\[ E\left[ (J_{1,q}^{wo})^2 \right] = \frac{1}{2\pi^2 N^3} \times \sum_{k=2}^{K} \sum_{m=1}^{S} \frac{2 + \varepsilon - 4U(m - q, b) [1 - 2P_e] + \varepsilon}{(m - q)^2} \times (\hat{B}_{km,1q} + B_{km,1q}) \]  

(25)

where

\[ \hat{G}_{kq,1q} = \sum_{l=0}^{N-1} (C_{kq,1q}(l))^2 + (C_{kq,1q}(l + 1))^2 + \left( \frac{1}{H(b)} - 2 \right) C_{kq,1q}(l) C_{kq,1q}(l + 1) \]  

(26a)

\[ \hat{B}_{km,1q} = \sum_{l=0}^{N-1} (C_{km,1q}(l))^2 + (C_{km,1q}(l + 1))^2 - 2C_{km,1q}(l) C_{km,1q}(l + 1). \]  

(26b)

\[ G_{kq,1q} \] and \[ B_{km,1q} \] are found from (26a) and (26b), respectively, by replacing the letter "m" by "k - N" in all arguments inside the sum. Also, \[ H(b) = E[x_k, x_k] \] and \[ U(n, b) = E[\sin(n\pi x_k) \sin(n\pi x_k)] \] are given in [29] as

\[ H(b) = E[x_k, x_k] = 1/3 - b/4 + b^2/6, \]  

(26c)

\[ U(n, b) = E[\sin(n\pi x_k) \sin(n\pi x_k)] = \frac{1 - \varepsilon - b^2}{2n\pi^2} - \frac{\sin(n\pi b) \sin(2n\pi b)}{2(n\pi)^2b} + 4 \sin^2(n\pi b/2) \sin^2(n\pi/2) \]

\[ + \frac{\cos^2(n\pi/2) \sin(n\pi b - n\pi/2)}{n\pi}. \]

(26d)

The user I’s symbols are then decoded by the soft decision Viterbi decoder. The bit error probability can be approximated as

\[ P_e^\text{wo} \approx \sum_{d=d_{\text{min}}+6} B_d P_d^{\text{wo}}(d) \]  

(27)

where

\[ P_d^{\text{wo}}(d) = \int_0^\infty \frac{1}{2} \text{erfc} \left( \sqrt{\frac{d}{2V_{\text{wo}}}} \right) f(E) dE \]  

(28)

\[ V_{\text{wo}} = \frac{S}{2\gamma_s} \]

\[ + \sum_{q=1}^{S} \frac{1}{3N^3} \sum_{k=2}^{K} (2 + \varepsilon) \gamma_{kq,1q} \]

\[ - 6H(b)(\hat{G}_{kq,1q} + G_{kq,1q}) \]

\[ + 12H(b)P_e^{\text{wo}}(\hat{G}_{kq,1q} + G_{kq,1q}) + \frac{1}{2\pi^2 N^3} \]

\[ \times \sum_{k=2}^{K} \sum_{m=1}^{S} \frac{2 + \varepsilon - 4U(m - q, b) [1 - 2P_e] + \varepsilon}{(m - q)^2} \]

\[ \times (\hat{B}_{km,1q} + B_{km,1q}) \]  

(29)

B. Multiuser Interference Cancellation with Decoding in the Initial Decision

As mentioned in Section I, the problem of the multiuser interference cancellation technique which reconstructed the interfering signals was the residual interference due to the symbol errors in the initial decisions. Forward error correction was utilized to improve the performance of the initial decision [21], [22], [39]. However, in [22], the canceller required additional interleavers and Viterbi decoders. If fading is very slow, the size of the interleaver has to be long enough to provide sufficient coding gain. Thus, the MC modulation is attractive in terms of the cancellation as it provides frequency diversity gain and coding gain without large interleavers and complex RAKE systems.

A model of a receiver using a multiuser interference canceller with a convolutional code is shown in Fig. 5. After the despreading, the canceller decodes the received symbols by a soft decision Viterbi decoder. The decoded data are then re-encoded and re-spread on each carrier. During the process, on the other hand, the received signal is put in the memory.
The memory size is set to the several times of the constraint length. After the re-spreading, the signals of the other users and the interference from the other carriers are reconstructed and removed from the received signal on each carrier. After that, user 1's signal is despread and decoded.

The probability with which the Viterbi decoder chooses a wrong path with the distance $d$ is given by the same equation as (18) and is

$$P_d(d) = \int_0^{\infty} \frac{1}{2} \text{erfc} \left( \sqrt{\frac{d}{2V_w}} E \right) f(E) dE. \tag{30}$$

Suppose there are $A_d$ paths which cause $d$ symbol errors after the re-encoding [37, p. 324]. Then the average symbol error probability caused by one user after the re-encoding is

$$P^{\text{w}}_e \approx \sum_{d=\text{dem}+1}^{d_{\text{dem}}+6} d A_d P_d(d). \tag{31}$$

Therefore, co-channel interference and the adjacent channel interference after the cancellation is given by

$$I^{\text{w}}_{1,q} = I_{1,q} - \frac{1}{N} \sum_{k=2}^{K} \sum_{m=1}^{S} g_{k,m,q} x_k^w$$

$$\times \left[ b_{k-1}^w R_{k,1,q} (\tilde{\tau}_k) + b_0^w \hat{R}_{k,1,q} (\tilde{\tau}_k) \right] \tag{32}$$

and

$$J^{\text{w}}_{1,q} = J_{1,q} - \frac{1}{N} \sum_{k=2}^{K} \sum_{m=1}^{S} g_{k,m,q} x_k^w$$

$$\times \left\{ b_{k-1} [C_{km,1,q} (l_k + 1 - N) - C_{km,1,q} (l_k - N)] + b_0 [C_{km,1,q} (l_k + 1) - C_{km,1,q} (l_k)] \right\} \times \tilde{x}_k \text{Sinc} (\pi (m - q) \tilde{x}_k) \tag{33}$$

where $\{x_k^w\}$ are the i.i.d. rv's that take value $-1$ with probability $P^{\text{w}}_e$ and 1 with probability $1 - P^{\text{w}}_e$. The second moments of (32) and (33) are

$$E \left[ (I^{\text{w}}_{1,q})^2 \right] = \frac{1}{3N^3} \sum_{k=2}^{K} (2 + \varepsilon)r_{kq,1q}$$

$$- 6H(b) (\hat{G}_{kq,1q} + G_{kq,1q})$$

$$+ 12H(b) P^{\text{w}}_e (\hat{G}_{kq,1q} + G_{kq,1q}) \tag{34}$$

and

$$E \left[ (J^{\text{w}}_{1,q})^2 \right] = \frac{1}{2\pi^2 N^3} \sum_{k=2}^{K} \sum_{m=1}^{S} \frac{2 + \varepsilon - 4U (m - q, b) [1 - 2P^{\text{w}}_e]}{(m - q)^2}$$

$$\times (\hat{B}_{km,1q} + B_{km,1q}). \tag{35}$$

The user 1's symbols are then decoded by the soft decision Viterbi decoder. The bit error probability is given by

$$P^{\text{w}}_b \approx \sum_{d=\text{dem}}^{d_{\text{dem}}+6} B_d P_d(d) \tag{36}$$

where

$$P_d(d) = \int_0^{\infty} \frac{1}{2} \text{erfc} \left( \sqrt{\frac{d}{2V_w}} E \right) f(E) dE \tag{37}$$

$$V_w = \frac{S}{2\gamma_s} + \sum_{q=1}^{S} \left\{ \frac{1}{3N^3} \sum_{k=2}^{K} (2 + \varepsilon)r_{kq,1q}$$

$$- 6H(b) (G_{kq,1q} + G_{kq,1q})$$

$$+ 12H(b) P^{\text{w}}_e (G_{kq,1q} + G_{kq,1q}) + \frac{1}{2\pi^2 N^3} \sum_{k=2}^{K} \sum_{m=1}^{S} \frac{2 + \varepsilon - 4U (m - q, b) [1 - 2P^{\text{w}}_e]}{(m - q)^2}$$

$$\times (\hat{B}_{km,1q} + B_{km,1q}) \right\}. \tag{38}$$

IV. NUMERICAL RESULTS

The following results assume that the data rate is 128 kb/s, the total process gain $G_p$ is set to 120 and the coding rate is 1/2 with the optimum generator polynomials given in [40]. The first seven terms of the transfer functions are used to get performance approximations and the accuracy of the approximations is ensured by computer simulations [22]. The delay spread is set to 100 ns so that the coherent bandwidth is about 1.6 MHz [24], [31]. The maximum Doppler shift is set to 2 Hz and no interleaving is utilized. To evaluate the performance, the probability density function of the sum of the correlated Rayleigh rv's, $f(E)$, is required. As mentioned in Section II-B, though it is possible to calculate $f(E)$, the calculation includes the summation with infinite number of terms and the probability distribution is strongly influenced by the number of terms [38]. Therefore, as following [29], rv is generated by a computer up to 10 000 000 times, and substituted in the erfc(·). The correlation between the envelopes of the signals at the two frequencies are replaced by the time correlation of the Rayleigh-fading channel created by the simulator given in [31] for simplicity though it can be obtained through the analysis shown in [29]. The rest of the evaluation is done by the theoretical analysis. The accuracy of the Gaussian approximation in (8) and (13) is also ensured in [29]. The performance is evaluated with the number of carriers $S$ equal to 1, 2, 3, 4, 6, and 12. The transmission bandwidth of each carrier is about 30.72 MHz, 15.36 MHz, 10.24 MHz, 7.68 MHz, 5.12 MHz, and 2.56 MHz, respectively. The total transmission bandwidth is 30.72 MHz, 23.04 MHz, 20.48 MHz, 19.2 MHz, 17.92 MHz, and 16.64 MHz, respectively. The correlation between the adjacent channels is about 0 (single carrier), 0.041, 0.088, 0.147, 0.279, and 0.607, respectively. As the transmission bandwidth between the adjacent channels overlaps each other, the total bandwidth is reduced as the number of carrier increases while
the system maintains the same total processing gain to evaluate the performance improvement only due to the cancellation techniques. It is also possible to keep the total bandwidth by increasing the processing gain. In this case, the capacity of the system should be better than the one which we will present in this section.

Fig. 6 shows the BER versus $E_b/N_0$. $K$ is the number of users. As for a single carrier case, the multiuser interference cancellers do not work well. The cancellers both with and without decoding in the initial decision show almost the same performance and the BER has the floor about $5 \times 10^{-2}$. On the other hand, when $S = 6$, the performance of the canceller with decoding is close to the performance of the single-user case, i.e., $K = 1$. At BER $= 10^{-5}$, which is a required BER of low resolution video services, the performance with decoding is about 1.5 dB better than that of without decoding. From these results, it is clear that the MC modulation provides the frequency diversity gain and the coding gain, and without interleaving the multiuser interference canceller with decoding improves the BER performance.

Fig. 7 shows the BER versus the number of users. It is also shown that the cancellers do not work with the single carrier modulation. Regarding the MC modulation, at BER $= 10^{-5}$, the canceller with decoding improves the capacity by a factor of 1.4 as compared with the canceller without decoding. Again, it is clear that the MC modulation provides coding gain. At around BER $= 10^{-5}$ the BER increases gradually with both of the cancellers while it increases rapidly without canceller. Thus, the system with the multiuser interference canceller is more robust to variations in the number of users.

Fig. 8 shows the BER performance of the canceller with decoding versus the constraint length of the convolutional codes. The BER decreases as the number of carriers and the constraint length of the convolutional code increases. When the number of carriers is one, the BER performance improvement due to the increment of the constraint length is trivial. On the other hand, when the number of carriers is 12, the BER rapidly decreases as the constraint length increases. This figure also shows that the MC modulation provides not only diversity gain, but also coding gain without interleaving. It is desirable to use the larger number of carriers as long as the system allows.

The MC modulation also reduces the processing delay and the complexity of the decoding. If the single carrier
system has to accommodate the same number of users as the MC system; the transmitters and the receivers have to employ huge interleavers and de-interleavers. For example, the correlation of 0.279 (the same correlation value as the $S = 6$ case) is obtained between two symbols which are about 38800 symbols apart on the assumed Rayleigh-fading channel of the single carrier system. This is not preferable for the canceller with decoding as it utilizes additional de-interleavers and interleavers. On the other hand, for the MC modulation system, the additional processing delay due to the canceller with decoding is only several times of the constraint length of the convolutional code, which is caused by the Viterbi decoder. Thus, the MC modulation makes the cancellers possible to be implemented with the realistic amount of processing delay and complexity.

When the constraint length is eight, the BER is lower than that of the constraint length of nine. This is because the coefficient $A_d$ in (31) of the code with the constraint length of eight is smaller compared to the code with the constraint length of nine [40]. This result suggests the possibility of constructing "optimum" codes in terms of the multiuser interference cancellation. Figs. 9 and 10 show the BER versus the error in the gain estimation and the timing estimation. The performance of the cancellers both with and without decoding deteriorates rapidly as the estimation error increases. Therefore, it is important for the cancellers to estimate the channel accurately. On both figures, the canceller with decoding always shows better performance than the canceller without decoding.

V. CONCLUSION

This paper has investigated the performance of the multiuser interference cancellation technique utilizing convolutional codes and MC modulation. It has been shown that the MC modulation provides not only diversity gain, but also coding gain without interleaving. Therefore, the multiuser interference canceller with decoding in the initial decision could improve the capacity by a factor of 1.4 at the BER $= 10^{-5}$ as compared to the canceller without decoding. It has also been suggested the possibility of constructing optimum codes in terms of the multiuser interference cancellation. To achieve full capacity improvement, the significance of the estimations of the channel gain and delay has been shown. As the utilization of the MC modulation substitutes for large interleavers and shortens the delay due to the cancellation process, the multiuser interference canceller with decoding is possible to be implemented to high speed wireless indoor communications.

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Masao Nakagawa (M’81), for a photograph and biography, see this issue, p. 1486.