Optimal Bandwidth Allocation for Scalable H.264 Video Transmission Over MIMO Systems

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4 authors, including:

Mohammad Jubran
Birzeit University

Lisimachos Paul Kondi
University of Ioannina

INTERFERENCE RESISTANT SCALABLE VIDEO TRANSMISSION OVER DS-CDMA CHANNELS

Manu Bansal, Elizabeth S. Pynadath, Lisimachos P. Kondi
Department of Electrical Engineering,
State University of New York at Buffalo,
Buffalo, NY 14260.
Email: {mbansal, pynadath, lkondi}@eng.buffalo.edu
and
John D. Matyjas, Michael J. Medley, Stephen P. Reichhart, Kurt A. Turck
Air Force Research Laboratory/IFGC,
525 Brook Rd.,
Rome, NY 13441.
Email: {john.matyjas, medleym, reichharts, kurt.turck}@rl.af.mil

ABSTRACT

In this work, we demonstrate the interference mitigation capabilities of the auxiliary vector (AV) receiver for scalable video transmission over direct-sequence code division multiple access (DS-CDMA) systems using a hardware testbed. The proposed receiver design is also compared to the conventional RAKE matched-filter (RAKE-MF) and minimum variance distortionless response (MVDR) receivers. The DS-CDMA video data stream is transmitted over an RF channel under "real world" Rayleigh-faded multipath channel conditions, emulating open and/or urban battlefield environments. The state-of-the-art Agilent E4438C Vector Signal Generator and Baseband Studio Fader is used to provide a configurable "real time" RF channel. In this work, the "foreman" video sequence is source encoded using an MPEG-4 compatible video codec and channel-coded using rate-compatible punctured convolutional (RCPC) codes. After spreading and modulating, the resultant bitstream is transmitted over a user-defined Agilent wireless channel emulation. Upon chip-matched filtering and sampling at the chip-rate on a hardware testbed, the received data are despread/demodulated using the AV, RAKE-MF and MVDR receivers and, subsequently, channel and source decoded. The resultant video clips exemplify that the AV receiver outperforms the MVDR and the RAKE-MF receiver counterparts under a wide range of rates and channel conditions.

I. INTRODUCTION

In the last few years, there has been a significant amount of research work done in the area of multimedia communication over direct sequence-code division multiple access (DS-CDMA) channels. In this paper we consider the scalable video transmission over "real world" DS-CDMA multipath fading channels in a multiuser environment and demonstrate the interference suppression capabilities of the auxiliary-vector (AV) receiver [1] for such systems. The choice of the AV receiver was dictated by realistic channel fading rates that limit the data record available for receiver adaptation and redesign. Under small sample support adaptation, the AV filter short-data-record estimators have been shown to exhibit superior bit-error-rate performance in comparison with least-mean-squares (LMS), recursive-least-squares (RLS), sample-matrix-inversion (SMI), diagonally loaded SMI, or multistage nested Wiener filter implementations [1], [2], [3].

In [4], video transmission from one transmitter to one receiver using binary phase-shift-keying (BPSK) modulation was analyzed. The channel behavior was modeled as non-frequency-selective Rayleigh fading. In [5], video transmission over a DS-CDMA link was considered. A frequency-selective (multipath) Rayleigh fading channel model was used. At the receiver, an adaptive antenna array auxiliary-vector (AV) linear filter that provides space-time RAKE-type processing (thus, taking advantage of the multipath characteristics of the channel) and multiple-access interference suppression was employed [1]. The tradeoffs of source coding, channel coding and spreading for image transmission in CDMA systems were considered in [6]. In [7] and [8], video transmission via a single-rate CDMA channel was compared against transmission via a combination of multi-code multirate CDMA and variable sequence length multirate CDMA under frequency selective Rayleigh fading.

In the video transmission system proposed here, the scalable video bit stream is first channel encoded using a particular channel coding rate. The channel coded information is then spread using a spreading code and carrier modulated for transmission over the wireless channel. The transmission over an RF channel with Rayleigh-faded multipath channel conditions, emulating open and/or urban battlefield environments, is done by means of the Agilent E4438C Vector Signal Generator and Baseband Studio Fader. At the receiver the information is demodulated and despread. These despread data are then decoded using a channel decoder and the video sequence is finally reconstructed using the source decoder.

The rest of this paper is organized as follows. In section II, we describe the elements of the proposed video transmission system, i.e., scalable video coding (section II-A), channel encoding (section II-B), channel modeling (section II-C), the received signal (section II-D), and auxiliary-vector (AV) filtering (section II-E). Experimental results are presented in section III and conclusions are drawn in section IV.

II. VIDEO TRANSMISSION SYSTEM

A. Scalable Video Coding

A scalable video encoder produces a bit stream that consists of embedded subsets. Each embedded subset can be decoded and produce a video sequence of a certain quality. Thus, a single compression operation can produce bit streams with different rates and reconstructed quality. A subset of the original bit stream can be initially transmitted to provide a base layer quality with extra layers subsequently transmitted as enhancement layers.

In this work, an MPEG-4 compatible video source codec is used for scalable video coding. Scalability is obtained in terms of SNR where enhancement in quality translates in an increase in the SNR of the reconstructed video sequence [9]. Also, error-resilience tools of MPEG-4 such as resynchronization markers, data partitioning and reversible variable length codes (RVLC) were enabled.

B. Channel Coding

Rate-Compatible Punctured Convolutional (RCPC) codes [10] are used to provide the unequal error protection (UEP) to the bitstream. The rate of a convolutional code is defined as k/n where k is the number of input bits and n is the number of output bits. For variable rate coding, a higher rate code can be obtained by puncturing the output of a “mother” code of rate $1/n$. For rate compatibility, higher rate codes are chosen to be subsets of lower rate codes. Decoding of the convolutional codes is carried out by the Viterbi algorithm which is a maximum-likelihood sequence estimation (MLSE) procedure.

C. Channel Modeling

The Agilent E4438C RF Arbitrary Waveform Generator and the N115A Baseband Studio fader hardware/software package were used to create a “real time” RF Rayleigh fading channel with multipath and Doppler frequencies. The E4438C RF Waveform Generator converts baseband I/Q data to RF. The N115A Baseband Studio obtains non-faded I/Q data from the E4438C arbitrary waveform generator and based on operator selected fading parameters, it computes and inserts, in “real time”, faded I/Q data back into the waveform generator. This setup is used to emulate open and/or urban battlefield environments with up to 48 paths and various Doppler frequencies.

D. Received Signal

We model the baseband received signal as the aggregate of the received multipath spread-spectrum (SS) signal of interest with signature code $\mathbf{S}_o$ of length L (if T is the symbol period and T_c is the chip period then $L = T/T_c$), $K-1$ received DS-SS interferers with unknown signatures $\mathbf{S}_k$, $k = 1, ..., K-1$, and white Gaussian noise. For notational simplicity and without the loss of generality, we choose a chip-synchronous signal set-up. We assume that the multipath spread is of the order of a few chip intervals, P, and since the signal is bandlimited to $B = 1/2T_c$ the lowpass channel can be represented as a tapped delay line with $P+1$ taps spaced at chip intervals T_c. After conventional chip-matched filtering and sampling at the chip rate over a multipath-extended symbol interval of $L+P$ chips, the $L+P$ data samples from the antenna element are organized in the form of a vector $\mathbf{r}$ given by $\mathbf{r} = \sum_{k=0}^{K-1} \sum_{p=0}^{P} c_{k,p} \sqrt{E_k} (b_k \mathbf{S}_{k,p} + b_k^- \mathbf{S}_{k,p}^- + b_k^+ \mathbf{S}_{k,p}^+) + \mathbf{n}$, where, with respect to the kth SS signal, E_k is the transmitted energy, b_k, b_k^-, b_k^+ are the present, the previous and the following transmitted bits, respectively, and $\{c_{k,p}\}$ are the coefficients of the Rayleigh fading multipath channel emulated using the Agilent E4438C Vector Signal Generator and Baseband Studio Fader. $\mathbf{S}_{k,p}$ represents the zero-padded by P, p-cyclic-shifted version of the signature of the kth SS signal $\mathbf{S}_k$, $\mathbf{S}_{k,p}^-$ is the 0-filled $(L-p)$-left-shifted version of $\mathbf{S}_{k,0}$ and $\mathbf{S}_{k,p}^+$ is the 0-filled $(L-p)$-right-shifted version of $\mathbf{S}_{k,0}$. Finally, $\mathbf{n}$ is the additive complex Gaussian noise.

For conceptual and notational simplicity we may rewrite the received data equation as follows: $\mathbf{r} = \sqrt{E_0} b_0 \mathbf{w}_{R-MF} + I + \mathbf{n}$, where $\mathbf{w}_{R-MF} = E_{b_0}\{\mathbf{r} b_0\} = \sum_{p=0}^{P} c_{0,p} \mathbf{S}_{0,p}$ is the effective channel-processed signature (RAKE Matched-Filter) of the SS signal of interest (signal-0) and $\mathbf{I}$ identifies comprehensively both the Inter-Symbol and the SS interference present in $\mathbf{r}$. In this work, as the fading coefficients are not known, $E_{b_0}\{.\}$, the statistical expectation with respect to b_0 is used to estimate $\mathbf{w}_{R-MF}$. This is done by using pilot bits (of the SS signal of interest) which are assumed to be available at the receiver error-free.

E. Auxiliary-Vector Filtering

After carrier demodulation, chip-matched filtering, and chip-rate sampling, auxiliary-vector (AV) filtering [1] provides multiple-access-interference (MAI) suppressing despreading. The AV receiver was chosen based on the realistic channel fading rates that limit the data record available for receiver adaptation. Under small sample support adaptation, AV filter short-data-record estimators have been shown to exhibit superior bit error rate (BER) performance in comparison to least mean squares (LMS), recursive least squares (RLS), sample matrix inversion (SMI), diagonally-loaded SMI, or multistage nested Wiener filter implementations. The AV algorithm generates a sequence of AV filters making use of two basic principles: (i) The maximum magnitude cross-correlation criterion for the evaluation of the auxiliary vectors (ii) The conditional mean-squared optimization criterion for the evaluation of the scalar AV weights. This algorithm is more clearly explained below.

The AV algorithm generates an infinite sequence of filters $\{\mathbf{w}_k\}_{k=0}^{\infty}$. The sequence is initialized at the RAKE filter

$$\mathbf{w}_0 = \frac{\mathbf{w}_{\text{R-MF}}}{\|\mathbf{w}_{\text{R-MF}}\|^2}, \tag{1}$$

which is normalized to satisfy $\mathbf{w}_0^H \mathbf{w}_{\text{R-MF}} = 1$. At each step $k+1$ of the algorithm, $k = 0, 1, 2, \ldots$, we incorporate in $\mathbf{w}_k$ an "auxiliary" vector component $\mathbf{g}_{k+1}$ that is orthogonal to $\mathbf{w}_{\text{R-MF}}$ and weighted by a scalar μ_{k+1} and we form the next filter in the sequence,

$$\mathbf{w}_{k+1} = \mathbf{w}_k - \mu_{k+1} \mathbf{g}_{k+1}. \tag{2}$$

The auxiliary vector $\mathbf{g}_{k+1}$ is chosen to maximize, under fixed norm, the magnitude of the cross-correlation between its output, $\mathbf{g}_{k+1}^H \mathbf{r}$, and the previous filter output, $\mathbf{w}_k^H \mathbf{r}$, and is given by

$$\mathbf{g}_{k+1} = \mathbf{R}\mathbf{w}_k - \frac{\mathbf{w}_{\text{R-MF}}^H \mathbf{R} \mathbf{w}_k}{\|\mathbf{w}_{\text{R-MF}}\|^2} \mathbf{w}_{\text{R-MF}} \tag{3}$$

where $\mathbf{R}$ is the input autocorrelation matrix, $\mathbf{R} = E\{\mathbf{r}\mathbf{r}^H\}$. The scalar μ_{k+1} is selected such that it minimizes the output variance of the filter $\mathbf{w}_{k+1}$ or equivalently minimizes the mean-square (MS) error between $\mathbf{w}_k^H \mathbf{r}$ and $\mu_{k+1}^* \mathbf{g}_{k+1}^H \mathbf{r}$. The MS-optimum μ_{k+1} is

$$\mu_{k+1} = \frac{\mathbf{g}_{k+1}^H \mathbf{R} \mathbf{w}_k}{\mathbf{g}_{k+1}^H \mathbf{R} \mathbf{g}_{k+1}}. \tag{4}$$

The AV filter recursion is completely defined by (1)-(4). Theoretical analysis of the AV algorithm was pursued in [1]. The results are summarized below in the form of a theorem.

Theorem 1: Let $\mathbf{R}$ be a Hermitian positive definite matrix. Consider the iterative algorithm of eqs. (1)-(4).

(i) Successive auxiliary vectors generated through (2)-(4) are orthogonal: $\mathbf{g}_k^H \mathbf{g}_{k+1} = 0$, $k = 1, 2, 3, \ldots$ (however, in general $\mathbf{g}_k^H \mathbf{g}_j \neq 0$ for $|k-j| \neq 1$).

(ii) The generated sequence of auxiliary-vector weights $\{\mu_k\}$, $k = 1, 2, \ldots$, is real-valued, positive, and bounded: $0 < \frac{1}{\lambda_{\text{max}}} \leq \mu_k \leq \frac{1}{\lambda_{\text{min}}}$, $k = 1, 2, \ldots$, where λ_{max} and λ_{min} are the maximum and minimum, correspondingly, eigenvalues of $\mathbf{R}$.

(iii) The sequence of auxiliary vectors $\{\mathbf{g}_k\}$, $k = 1, 2, \ldots$, converges to the $\mathbf{0}$ vector: $\lim_{n\to\infty} \mathbf{g}_n = \mathbf{0}$.

(iv) The sequence of auxiliary-vector filters $\{\mathbf{w}_k\}$, $k = 1, 2, \ldots$, converges to the minimum-variance-distortionless-response (MVDR) filter: $\lim_{k\to\infty} \mathbf{w}_k = \frac{\mathbf{R}^{-1}\mathbf{w}_{\text{R-MF}}}{\mathbf{w}_{\text{R-MF}}^H \mathbf{R}^{-1} \mathbf{w}_{\text{R-MF}}}$. $\square$

If $\mathbf{R}$ is unknown (as in practice) and is sample-average estimated from a packet data record of D points, $\hat{\mathbf{R}}(D) = \frac{1}{D}\sum_{d=1}^{D} \mathbf{r}_d \mathbf{r}_d^H$, then Theorem 1 shows that

$$\hat{\mathbf{w}}_k(D) \underset{k\to\infty}{\longrightarrow} \hat{\mathbf{w}}_\infty(D) = \frac{\left[\hat{\mathbf{R}}(D)\right]^{-1} \mathbf{w}_{\text{R-MF}}}{\mathbf{w}_{\text{R-MF}}^H \left[\hat{\mathbf{R}}(D)\right]^{-1} \mathbf{w}_{\text{R-MF}}} \tag{5}$$

where $\hat{\mathbf{w}}_\infty(D)$ is the widely used MVDR filter estimator known as the sample-matrix-inversion (SMI) filter [11]. The output sequence begins from $\hat{\mathbf{w}}_0(D) = \frac{\mathbf{w}_{\text{R-MF}}}{\|\mathbf{w}_{\text{R-MF}}\|^2}$, which is a *0*-variance, fixed-valued, estimator that may be severely biased ($\hat{\mathbf{w}}_0(D) = \frac{\mathbf{w}_{\text{R-MF}}}{\|\mathbf{w}_{\text{R-MF}}\|^2} \neq \mathbf{w}_{\text{MVDR}}$) unless $\mathbf{R} = \sigma^2 \mathbf{I}$ for some $\sigma > 0$. In the latter trivial case, $\hat{\mathbf{w}}_0(D)$ is already the perfect MVDR filter. Otherwise, the next filter estimator in the sequence, $\hat{\mathbf{w}}_1(D)$, has a significantly reduced bias due to the optimization procedure employed at the expense of non-zero estimator (co-)variance. As we move up in the sequence of filter estimators $\hat{\mathbf{w}}_k(D)$, $k = 0, 1, 2, \ldots$, the bias decreases rapidly to zero while the variance rises slowly to the SMI ($\hat{\mathbf{w}}_\infty(D)$) levels (cf. (5)).

An adaptive data-dependent procedure for the selection of the most appropriate member of the AV filter estimator sequence $\{\hat{\mathbf{w}}_k(D)\}$ for a given data record of size D is presented in [12]. The procedure selects the estimator $\hat{\mathbf{w}}_k$ from the generated sequence of AV filter estimators that exhibits maximum $\mathcal{J}$-divergence between the filter output conditional distributions given that $+1$ or -1 is transmitted. Under a Gaussian approximation on the conditional filter output distribution, it was shown in [12] that the $\mathcal{J}$-divergence of the filter estimator with k auxiliary vectors is

$$J(k) = \frac{4\,E^2\left\{b_0 \mathrm{Re}\left[\hat{\mathbf{w}}_k^H(D)\mathbf{r}\right]\right\}}{Var\left\{b_0 \mathrm{Re}\left[\hat{\mathbf{w}}_k^H(D)\mathbf{r}\right]\right\}}. \qquad (6)$$

To estimate the $\mathcal{J}$-divergence $J(k)$ from the data packet of size D, the transmitted information bits b_0 are required to be known. We can obtain a blind *approximate* version of $J(k)$ by substituting the information bit b_0 in (6) by the detected bit $\hat{b}_0 = \mathrm{sgn}\left[\mathrm{Re}\left\{\hat{\mathbf{w}}_k^H(D)\mathbf{r}\right\}\right]$ (output of the sign detector that follows the linear filter). In particular, using $\hat{b}_0$ in place of b_0 in (6) we obtain the following $\mathcal{J}$-divergence expression:

$$\begin{aligned} J_B(k) &= \frac{4\,E^2\left\{\hat{b}_0 \mathrm{Re}[\hat{\mathbf{w}}_k^H(D)\mathbf{r}]\right\}}{Var\left\{\hat{b}_0 \mathrm{Re}[\hat{\mathbf{w}}_k^H(D)\mathbf{r}]\right\}} \\ &= \frac{4\,E^2\left\{\left|\mathrm{Re}[\hat{\mathbf{w}}_k^H(D)\mathbf{r}]\right|\right\}}{Var\left\{\left|\mathrm{Re}[\hat{\mathbf{w}}_k^H(D)\mathbf{r}]\right|\right\}} \end{aligned} \qquad (7)$$

where the subscript "B" identifies the blind version of the $\mathcal{J}$-divergence function. To estimate $J_B(k)$ from the data packet of size D, we substitute the statistical expectations in (7) by sample averages. The following criterion summarizes the corresponding AV filter estimator selection rule.

Criterion 1: For a given data record of size D, the unsupervised (blind) $\mathcal{J}$-divergence AV filter estimator selection rule chooses the estimator $\hat{\mathbf{w}}_k(D)$ with k auxiliary vectors where

$$k = \arg\max_k \left\{\hat{J}_B(k)\right\} = \\ \operatorname*{argmax}_k \left\{ \frac{4\left[\frac{1}{D}\sum_{d=1}^{D}\left|\mathrm{Re}[\hat{\mathbf{w}}_k^H(D)\mathbf{r}_d]\right|\right]^2}{\frac{1}{D}\sum_{d=1}^{D}\left|\mathrm{Re}[\hat{\mathbf{w}}_k^H(D)\mathbf{r}_d]\right|^2 - \left[\frac{1}{D}\sum_{d=1}^{D}\left|\mathrm{Re}[\hat{\mathbf{w}}_k^H(D)\mathbf{r}_d]\right|\right]^2} \right\} \qquad (8)$$

Criterion 1 completes the design of the AV filter estimator.

III. EXPERIMENTAL RESULTS

In this section, we present the experimental results for the setup described above. We assume $K = 7$ users that employ direct-sequence SS signaling (the user-of-interest and six interferers). The SNR for all the users is fixed at 8 dB (all SNR values reported in this paper refer to the SNR per chip). An MPEG-4 compatible source codec was used to encode a "foreman" video sequence with two different source rates of 40 and 120 kbps. Rate-compatible punctured codes (RCPC) were used for channel coding by using a "mother" code rate of 1/4 and puncturing it down to the code rate of 1/2. Walsh-Hadamard codes of length $L = 16$ were used as spreading codes. The transmission over a Rayleigh fading channel was simulated using the Agilent RF Waveform Generator and Baseband Studio Fader with all CDMA signals $k = 0, 1, \ldots, K-1$ experiencing $P = 3$ resolvable multipaths and various Doppler values equal to 0, 4, 40 and 200 Hz. Three different receivers were assumed: The RAKE matched-filter (RAKE-MF), the conventional sample-matrix-inversion minimum-variance-distortionless-response (SMI-MVDR) filter, and the auxiliary vector (AV) filter (based on criterion 1, the best AV was selected out of 11 AV's that were produced). The channel decoding was performed using Viterbi algorithm.

The tabulated results obtained by this study are shown next. Since the resulting video quality of all the decoded frames is comparable, the standard by which the output is deemed superior is the number of frames that can be decoded. If there are multiple sets that produce fully-decodable video clips (all the frames), the best set is determined by the lowest number of pilot bits used (given in terms of percentage of the packet size). The packet size in the tables below refer to the size of the packet after channel encoding. For all the cases, the bit error rate (BER) is calculated after channel decoding at the receiver. Table I gives the results for the cases where there is zero Doppler. Table I corresponds to the simulations run with the video sequence (150 frames) encoded at the rate of 120 kbps. Using 13%/12% of the packet size (410 bits) as pilot bits, AV filter recovered the total/partial video sequence, respectively. For the same packet size and number of pilot bits, the RAKE-MF and the MVDR filter can not even produce decodable video.

The results for only the user-of-interest experiencing Doppler shifts of 4 Hz while the interfering users experience zero Doppler are given in Table II. The packet sizes were adjusted according to the Doppler frequency

TABLE I

SIMULATION RESULTS FOR 150 FRAME VIDEO SEQUENCE ENCODED AT 120 KBPS WITH ZERO DOPPLER.

Filter used	Packet size (bits)	Pilot bits	BER	no. of dec. frames
AV	410	12%	1.4E-05	54
RAKE-MF	410	12%	6.33E-02	0
MVDR	410	12%	4.20E-03	0
AV	410	13%	3.1E-06	150 (ALL)
RAKE-MF	410	13%	5.9E-02	0
MVDR	410	13%	5.77E-04	0

TABLE II

SIMULATION RESULTS FOR 300 FRAME VIDEO SEQUENCE ENCODED AT 40 KBPS WITH 4 HZ DOPPLER.

Filter used	Packet size	Pilot bits	BER	no. of dec. frames
AV	810	7.5%	0	300
RAKE-MF	810	7.5%	4.68E-02	0
MVDR	810	7.5%	1.09E-04	0
AV	810	10%	0	300
RAKE-MF	810	10%	2.54E-02	0
MVDR	810	10%	3.24E-05	300

to maintain the assumption of constant fading over each packet. As shown in Table II, the AV filter gives a fully decodable video with only 7.5% of the packet size used as pilot bits while the MVDR filter requires higher percentage for pilot bits. Using the RAKE-MF filter, no frames were recovered with a practical number of pilot bits.

Table III shows the results for the situation where all users experience the same Doppler frequencies of 4 Hz. There are three cases that produce fully-decodable video using AV filters. However, using 30% for a packet size of 266 or even 20% for a packet size of 1000 bits for pilot bits is not a practical condition. Hence using 12.5% for a packet size of 800 bits as pilot bits can be regarded as the best choice. Also, neither RAKE-MF filter nor MVDR filter recovered any part of the video sequence for any reasonable number of pilot bits.

Tables IV and V show the results for 40 and 200 Hz Doppler shifts. The values in these tables clearly show that the AV filter outperforms RAKE-MF and MVDR filters for all range of practical number of pilot bits used.

IV. CONCLUSIONS

We demonstrated the effectiveness of using an auxiliary vector (AV) receiver for scalable video transmission

TABLE III

SIMULATION RESULTS FOR 150 FRAME VIDEO SEQUENCE ENCODED AT 40 KBPS WITH 4 HZ DOPPLER.

Filter used	Packet size	Pilot bits	BER	no. of dec. frames
AV	800	10%	7.39E-05	70
RAKE-MF	800	10%	5.71E-02	0
MVDR	800	10%	1.20E-03	0
AV	800	12.5%	4.31E-05	150
RAKE-MF	800	12.5%	5E-02	0
MVDR	800	12.5%	3.90E-05	0
AV	1000	20%	7.18E-06	150
RAKE-MF	1000	20%	5.72E-02	0
MVDR	1000	20%	6.74E-05	70
AV	600	25%	5.41E-05	150
RAKE-MF	600	25%	4.48E-02	0
MVDR	600	25%	1.76E-04	0
AV	266	30%	2.15E-05	150
RAKE-MF	266	30%	4.97E-02	0
MVDR	266	30%	9.48E-04	0

TABLE IV

SIMULATION RESULTS FOR 300 FRAME VIDEO SEQUENCE ENCODED AT 40 KBPS WITH 40 HZ DOPPLER.

Filter used	Packet size	Pilot bits	BER	no. of dec. frames
AV	610	7.5%	6.01E-05	134
RAKE-MF	610	7.5%	5.8E-02	0
MVDR	610	7.5%	6.5E-03	0
AV	610	10%	0	300
RAKE-MF	610	10%	7.2E-02	0
MVDR	610	10%	1.5E-03	0

TABLE V

SIMULATION RESULTS FOR 300 FRAME VIDEO SEQUENCE ENCODED AT 40 KBPS WITH 200 HZ DOPPLER.

Filter used	Packet size	Pilot bits	BER	no. of dec. frames
AV	250	12.5%	4.62E-06	300
RAKE-MF	250	12.5%	5.7E-02	0
MVDR	250	12.5%	2.48E-04	0

over DS-CDMA systems with "real world" multipath channel conditions. The results clearly establish the interference mitigation capabilities of the AV receiver. The AV filter receiver was also shown to outperform the MVDR and the RAKE-MF receivers under a wide range of rates and channel conditions.

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