SPACE-TIME EQUALIZATION AND INTERFERENCE CANCELLATION WITH CONVOLUTIONAL CODING FOR MIMO OFDM

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ABSTRACT

Future broadband cellular wireless communication systems will be characterized by high delay spread and the need to maximize the system capacity. Aggressive frequency reuse, one technique of maximizing capacity, subjects the mobile to high levels of interference when the mobile is near the cell edge. Traditional interference suppression techniques designed for synchronized OFDM systems do not perform optimally when the desired and interfering signals are asynchronous at the receiver. Methods of time aligning asynchronous signals (i.e., signals whose channel plus delay is longer than the cyclic prefix length) were recently presented in [1]. This paper presents results of the synchronization algorithms for coded systems. The results show that the algorithms give significantly different results when coding is applied.

1. INTRODUCTION

Future broadband cellular wireless communication systems will be characterized by high delay spread and the need to maximize the system capacity especially in light of the high cost of radio frequency spectrum. One of the techniques to maximize the system capacity is aggressive frequency reuse where adjacent cells are on the same time-frequency channels. Thus mobiles near the edge of the cells will be subjected to high levels of interference. For the case of no co-channel interference but intersymbol interference due to frequency selectivity of the channel, cyclic-prefix based communication systems, such as orthogonal frequency division multiplexing (OFDM), are employed in practice to compensate for intersymbol interference. The cyclic prefix enables the use of low complexity receiver adaptive antenna combining algorithms assuming that the signal is synchronized and the channel is not longer than the cyclic prefix length. When an interfering signal is present, this signal's cyclic prefix may not align in time with that of the desired signal especially because of the differing propagation delays between the desired signal and interfering signal. Traditional antenna combining techniques, such as linear MMSE combining, cannot perfectly suppress the interferer when its channel plus delay is greater than the cyclic prefix length [2] (i.e., an asynchronous interferer).

Methods of aligning asynchronous OFDM signals, based on channel shortening ideas, were recently presented in [1]. The techniques model the unsynchronized OFDM signal as being synchronized (i.e., its cyclic prefix is time-aligned with the cyclic prefix of the desired signal) at the time of transmission but having transversed a channel with some initial delay en route to the receiver. Processing is performed at the receiver which effectively aligns the channels so that the cyclic prefixes of the desired and interfering signals are aligned at the receiver. The techniques also attempt to keep the length of the channels (meaning the number of taps from the first non-zero tap to the last non-zero tap) at or below the maximum length that is guarded against by the cyclic prefix so that inter-carrier interference is not present in the received signal. Then, frequency-domain diversity combining and interference suppression techniques for synchronous signals may be applied.

Because of frequency selectivity OFDM subcarriers may experience different bit error rates. In discrete multitone (DMT) systems, of which OFDM may be viewed to be a subset, the overall bit rate may be optimized by tailoring the number of bits sent on each of the subcarriers, requiring feedback. In high data rate systems that experience fast fading where feedback may not be fast enough, such as future cellular systems, error correction coding is often employed to improve the overall bit rate. This paper presents simulations which show the effect of adding soft decision convolutional coding to the techniques in [1].

Recent work in space-time processing for OFDM systems includes [3], in which minimum mean squared error (MMSE) diversity combiner parameters are estimated which are then used to demonstrate suppression of both synchronous and asynchronous interference in OFDM systems. A two-stage adaptive array architecture that uses combined spatial diversity and beamforming to combat co-channel interference in the presence of fading is presented in [4]. Frequency domain techniques for channel identification and multiuser demodulation are presented in [5].

2. THE OFDM MIMO DATA MODEL

In general, OFDM systems can be described as follows. The system processes a block of K symbols at a time, and the block number, or OFDM baud, is denoted n_b. Once the K data symbols $X(0, n_b), \ldots, X(K-1, n_b)$ are available, an N-point inverse DFT ($N \geq K$) is used to create the transmitted values $x[0, n_b], \ldots, x[N-1, n_b]$. The last L_{cp} values are copied to the beginning of the block to create an extended block containing a *cyclic prefix*. Subsequent blocks are concatenated and transmitted serially through a channel. At the receiver, blocks are again formed. Then the first L_{cp} values are discarded and a DFT operation is performed on each block to yield the received symbols $Y(0, n_b), \ldots, Y(K-1, n_b)$. In OFDM systems the data sym-

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bols $X(0,n_b),\ldots,X(K-1,n_b)$ are referred to as being in the frequency domain and the transmitted values $x[0,n_b],\ldots,x[N-1,n_b]$ are said to be in the time domain.

If the length of the possibly time-varying (but stationary over an OFDM block) channel with impulse response $h[n,n_b]$ is no greater than the length of the cyclic prefix L_{cp} plus one then then the received symbols $Y(0,n_b),\ldots,Y(K-1,n_b)$ are simply the transmitted symbols times a scalar (the channel's frequency response at the respective frequency) plus noise. When U synchronous OFDM signals, possibly transmitted by different transmit antennas, impinge upon a set of M receive antennas, the received signals can be sampled and stacked into an $M \times 1$ vector and modeled as

$$\mathbf{Y}(k,n_b) = \sum_{u=1}^{U} \mathbf{H}^{(u)}(k,n_b)X^{(u)}(k,n_b) + \mathbf{V}(k,n_b), \qquad (1)$$

where $\mathbf{H}^{(u)}(k,n_b) = \left[H_1^{(u)}(k,n_b),\ldots,H_M^{(u)}(k,n_b)\right]^T$ is the channel vector for signal u at subcarrier k during baud interval n_b, $H_i^{(u)}(k,n_b)$ is the channel parameter for the path from transmit antenna u to receive antenna i, $X^{(u)}(k,n_b)$ is signal u's data symbol, and $\mathbf{V}(k,n_b)$ is a vector of additive noise. Again, this model assumes that the channels have length $L_{cp}+1$ or less.

3. SPACE-TIME PROCESSING TO ALIGN ASYNCHRONOUS SIGNALS

Consider a receiver with two receive antennas, spaced for diversity, which receives the desired m-ary QAM OFDM signal and simultaneously receives a second OFDM signal from another basestation operating on the same time-frequency channel. When the signals are not synchronous, the expression in (1) is not valid and the performance of traditional antenna combining techniques is degraded. This degradation may be significantly reduced by designing pairs of time-domain equalizers that effectively align the cyclic prefixes of the desired and interfering signals and that constrain the lengths of the effective channels from the transmit antennas to the output of the partial equalizers to be less than or equal to the length of the cyclic prefix plus one.

We denote the delay of the interferer's cyclic prefix relative to the beginning of the desired signal's cyclic prefix as T and model it as T zeros at the beginning of the impulse responses for the channels from the interfering basestation. The channels from basestation 1 are length L or less (meaning that the number of taps from the first non-zero tap to the last is L or less) and the channels from basestation 2 (the interferer) have length $L+D$ or less, where D ($D \geq 0$) is the excess length of the channels from the interfering basestation. The channels are assumed to be known, and the cyclic prefix length at both basestations is L_{cp}, where $L_{cp} \geq L-1$. If either the cyclic prefix delay T or the excess length D causes any of the channels to be non-zero outside some window of $L_{cp}+1$ samples then partial equalization is needed, and we proceed by forming the sets of equalizers.

In the time domain, the antenna output of the i-th antenna is denoted $y_i[n]$. For the $U=M=2$ scenario we have

$$y_1[n] = h_1^{(1)}[n] * x^{(1)}[n] + h_1^{(2)}[n] * x^{(2)}[n] + \nu_1[n] \qquad (2)$$

$$y_2[n] = h_2^{(1)}[n] * x^{(1)}[n] + h_2^{(2)}[n] * x^{(2)}[n] + \nu_2[n] \qquad (3)$$

where $x^{(u)}[n]$ is the time-domain transmitted signal for transmit antenna u, and where $h_i^{(u)}[n]$ is the impulse response of the channel from basestation u to receive antenna i. Note that $h_1^{(2)}[n]$ and $h_2^{(2)}[n]$ will be zero for $n < T$ to account for the cyclic prefix delay relative to the desired basestation's signal.

A set of equalizers $g_1^{(a)}[n]$ and $g_2^{(a)}[n]$, with respective lengths $N_{g1}^{(a)}$ and $N_{g2}^{(a)}$, is applied to the antenna outputs as

$$y^{(a)}[n] = y_1[n] * g_1^{(a)}[n] + y_2[n] * g_2^{(a)}[n] \qquad (4)$$

Collecting terms convolved with $x_1[n]$ and $x_2[n]$ reveals the effective channels from each of the basestations to the output of equalizer pair (a). The effective channel impulse responses are $\tilde{h}_1^{(a)}[n]$ and $\tilde{h}_2^{(a)}[n]$, where

$$\tilde{h}_1^{(a)}[n] = h_1^{(1)}[n] * g_1^{(a)}[n] + h_2^{(1)}[n] * g_2^{(a)}[n] \qquad (5)$$

and

$$\tilde{h}_2^{(a)}[n] = h_1^{(2)}[n] * g_1^{(a)}[n] + h_2^{(2)}[n] * g_2^{(a)}[n]. \qquad (6)$$

By properly choosing the equalizers, $\tilde{h}_1^{(a)}[n]$ and $\tilde{h}_2^{(a)}[n]$ can be tailored to have desired properties, such as specific lengths.

A single pair of equalizers provides one effective channel from each basestation to the receiver. A second set of equalizers can be formed to provide a second channel from each basestation to the receiver and to assist in interference suppression. Figure 1 shows how these two pairs of equalizers $g_i^{(a)}[n], i = 1,2$, and $g_i^{(b)}[n], i = 1,2$, can be used to effect synchronization and provide the same number of effective antenna outputs to the input of the frequency domain equalizer as there are receive antennas.

4. EFFECTIVE CHANNEL CHOICES

Equalizers for several choices of effective channel impulse response pairs were discussed in [1] and are briefly outlined here. Each equalizer pair sought to align cyclic prefixes and simultaneously keep the effective channels to length $L_{cp}+1$ or less.

The first solution sought to nullify the interfering signal while maintaining the length of the channel from the desired basestation to length L or less. This was referred to as the zero forcing channel shortening approach (ZF-CS). This approach sought to avoid noise enhancement associated with pure zero forcing solutions by not completely equalizing the desired signal but by allowing its channel to have length greater than one. Another approach, the traditional zero forcing (ZF) approach, seeks to nullify the interferer while making the channel from the desired basestation an impulse. Both of these approaches allow single tap frequency domain equalization after the DFT at the receiver.

In the constrained channel shortening (C-CS) approach certain taps in the effective channels were constrained to chosen values while others were left unconstrained. The constraints were changed for the second equalizer pair in a way that attempted to insure that the effective channels for each of the solutions had different frequency responses, ideally with no common nulls. The outputs of the technique were followed by MMSE frequency domain equalization.

A final solution was based on energy considerations and sought to maximize the energy transferred by certain channel taps while minimizing the energy transferred by other taps. Best results were

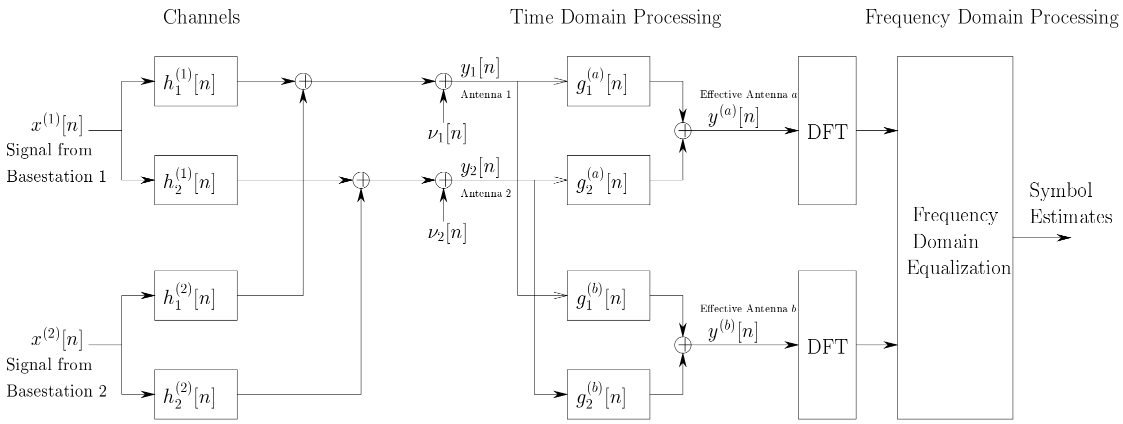

Fig. 1. Block diagram of the overall processing.

reported for cases where the first and second equalizer pairs maximized the energy transferred by a single tap (a different tap for each pair) in the effective channel from the desired basestation while minimizing the energy transferred by all the other taps. This solution was again followed by MMSE frequency domain equalization.

5. SIMULATION RESULTS

This section presents simulation results for the synchronization and interference suppression methods presented in [1] with and without coding for OFDM signals with symbols taken from QPSK, 16-ary QAM, and 64-ary QAM constellations. In the simulations, signals from two basestations, each received with equal power, were present, and the receiver had two antennas. The FFT length was 64 and the number of used subcarriers was 49. Each channel had four non-zero taps (the last three being placed randomly within the channel length) with complex, independent, and normally distributed amplitudes. Channel estimation was not performed by the algorithms, but, rather, the channels were assumed to be known perfectly. The simulations used a maximum channel length of $L = 25$, and the cyclic prefix length was $L_{cp} = 25$. The time-domain equalizers were given a length of 75 taps per antenna, roughly three times the length of the channels. Channels were static over a single block. The bit error rate (BER) is plotted versus the average signal to noise ratio (SNR) that would have existed if no interferer had been present.

Figures 2 through 7 present the results of simulations with and without error correction coding. In the figures, solid lines represent systems where the interferer's prefix delay was $T = 25$ while dashed lines indicate simulations with no cyclic prefix offset. The simulations with coding and interleaving used the half-rate, constraint length seven convolutional code with polynomial generators (133) and (171). Coding and interleaving was performed over each OFDM block individually, and soft decision decoding was used. Figures 2 and 3 present uncoded and coded results, respectively, for OFDM symbols from a QPSK symbol constellation. Figures 4 and 5 present similar results for a 16-ary QAM symbol constellation, and Figures 6 and 7 present results for 64-ary QAM.

The curves labeled MMSE present the results of minimum mean squared error diversity combining in the frequency domain. For these curves, no preprocessing was performed in the time domain. In each of the figures, MMSE is seen to perform well for $T = 0$ and poorly for $T = 25$. It is interesting to note that in the figures with coding some of the curves are not monotonically decreasing. This is most pronounced in the MMSE curve of Figure 3. In both the coded and uncoded simulations the MaxPwr method shows the best performance at low SNRs. However, whereas in the uncoded simulations the C-CS method performs best at high SNRs (for $T = 25$), in the coded simulations the ZF technique gave the best results for QPSK OFDM signals (Figure 3) while the ZF-CS method performs best for OFDM signals with symbols from the 16-ary and 64-ary QAM constellations (Figures 5 and 7, respectively).

6. CONCLUSIONS

This paper has presented simulation results for the techniques of [1] in the presence of coding. The techniques, with coding added, have shown achievable bit error rates of better than 10^{-6} even when an asynchronous interferer impinges at the receiver with strength equal to that of the desired signal. Of the techniques, the MaxPwr technique performs best at low SNRs for both coded and uncoded signals. At high SNRs, the ZF technique performed best for coded QPSK OFDM, and the ZF-CS technique gave the best results for higher order modulations.

7. REFERENCES

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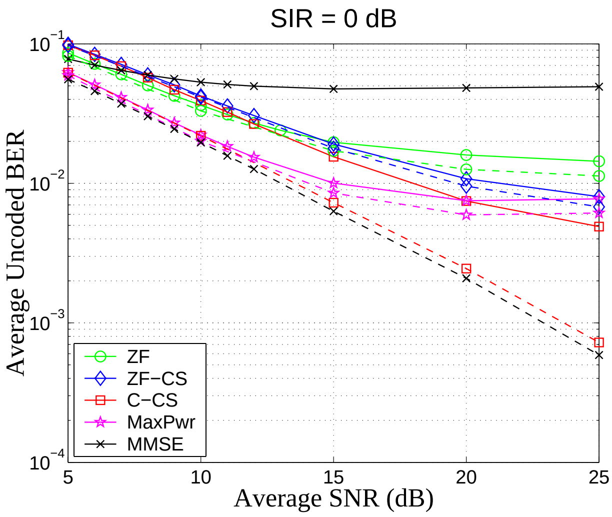

Fig. 2. Two bits per symbol. Dashed: $T = 0$. Solid: $T = 25$.

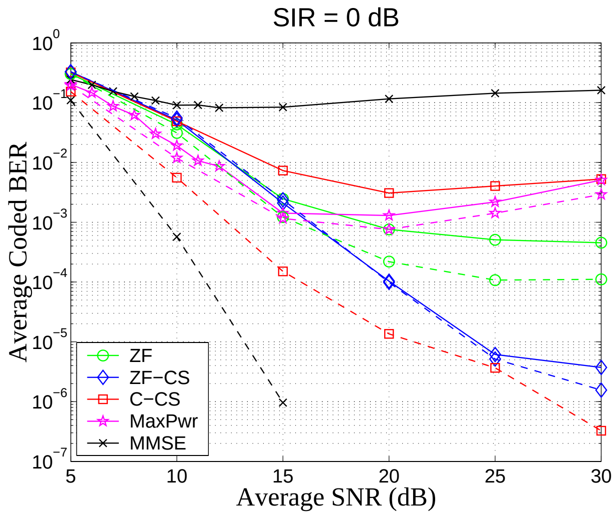

Fig. 5. Four bits per symbol. Dashed: $T = 0$. Solid: $T = 25$.

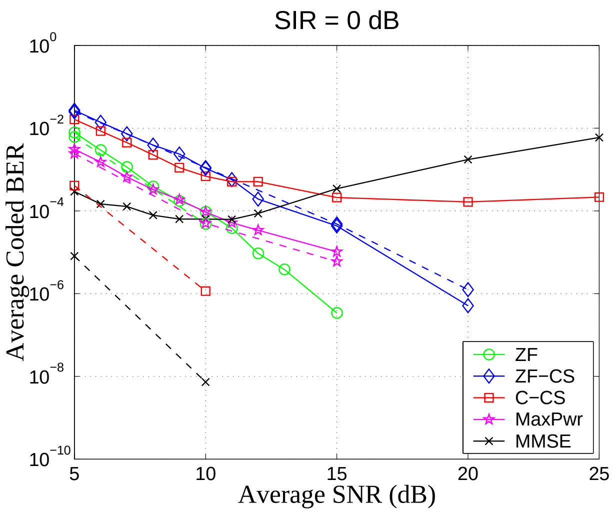

Fig. 3. Two bits per symbol. Dashed: $T = 0$. Solid: $T = 25$.

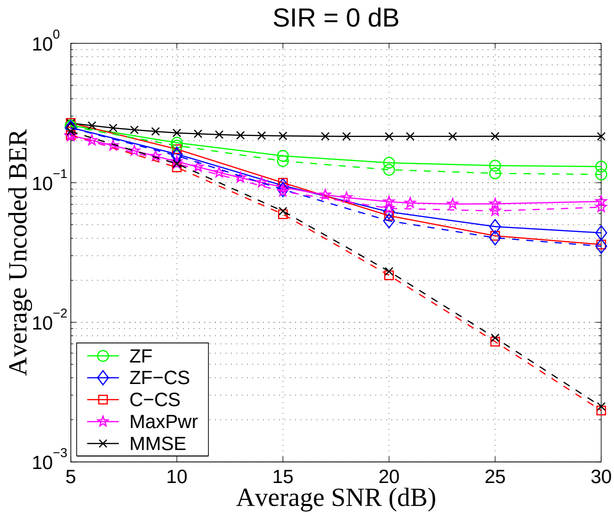

Fig. 6. Six bits per symbol. Dashed: $T = 0$. Solid: $T = 25$.

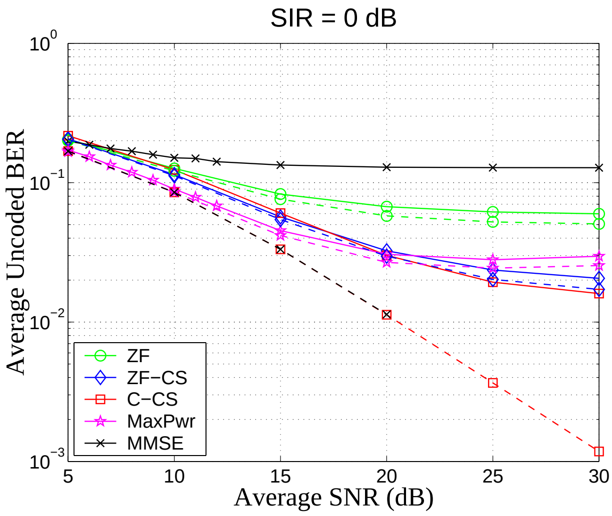

Fig. 4. Four bits per symbol. Dashed: $T = 0$. Solid: $T = 25$.

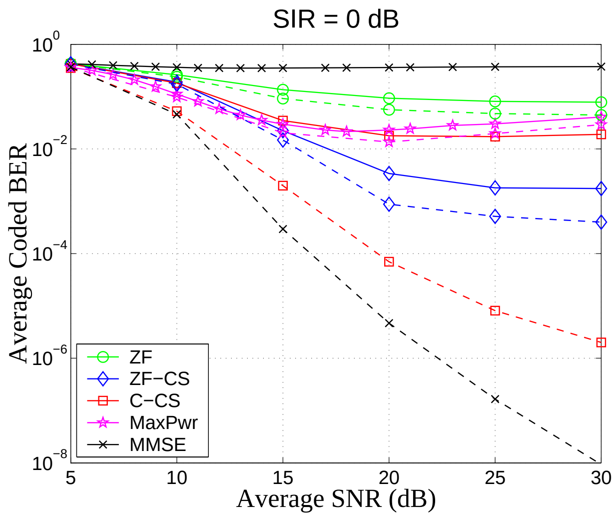

Fig. 7. Six bits per symbol. Dashed: $T = 0$. Solid: $T = 25$.