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Ultra-wideband MB-OFDM channel estimation with complementary codes

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Abstract

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Keywords

UWB, ultra-wideband, MB-OFDM, equalization, complementary, estimation

Disciplines

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Ultra-Wideband MB-OFDM Channel Estimation with Complementary Codes

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Abstract—In this paper, we design complementary codesets that significantly improve the quality of channel estimation in orthogonal frequency division multiplexing (OFDM) communication systems, with a focus on the recent MB-OFDM ultra-wideband (UWB) standard. The proposed time-domain OFDM channel estimation technique incurs only a nominal increase in computational complexity and is able to be readily retrofitted into the existing MB-OFDM standard. The underlying complementary codesets, found via an evolutionary algorithm, combine with the existing preamble synchronization sequences to yield asymptotically ideal auto-correlation functions (ACFs). We show how improvements exceeding 1 dB can be achieved in end-to-end packet error rate relative to conventional zero-forcing OFDM equalization.

I. INTRODUCTION

Coherent signalling over frequency-selective channels requires that the receiver estimate and equalize the channel prior to demodulation. To that end, one of the most significant advantages of orthogonal frequency division multiplexing (OFDM) systems [1], typified by the block diagram of Fig. 1, is that the highlighted channel estimation and equalization stage is conceptually and computationally simple. Since this allows an OFDM transceiver to be both low-power and low-cost, OFDM systems are becoming increasingly popular in applications ranging from wireless personal area networks (WPANs) to digital television.

The simplest OFDM channel estimation is a zero-forcing (ZF) approximation of N complex coefficients to rotate and scale each of the N subcarriers in the received OFDM symbol. Such channel estimation is most commonly performed using an explicit training sequence known to both sender and receiver. Although blind estimation [2] techniques exist, longer convergence times make them more appropriate for channel tracking than for initial channel estimation.

Most OFDM channel estimation sequences possess a flat power-spectral density so that each subcarrier can be estimated to the same accuracy. For example, in multi-band OFDM (MB-OFDM) [3], the channel estimation sequence is obtained by taking the inverse fast fourier transform (IFFT) of N equal-magnitude random quadrature phase shift keying (QPSK) symbol constellations. This allows the receiver to perform a ZF frequency-domain equalization (FDE) at the cost of a single fast fourier transform (FFT) and N complex multiplications. Unlike many iterative time-domain equalization (TDE)

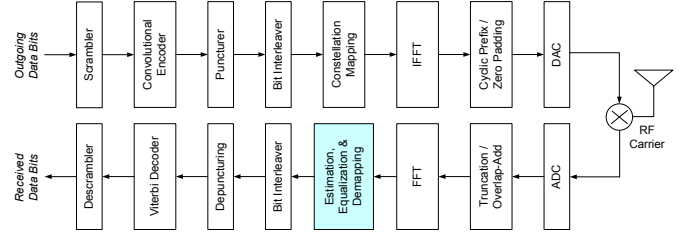


Fig. 1. Block diagram of a generic OFDM transmitter.

techniques, ZF FDE is guaranteed to converge and is low in computational complexity.

In this paper, we present an alternative estimation and equalization scheme that reuses the correlation of the synchronization sequence. In other words, we derive a channel estimation sequence that, while possessing a non-ideal auto-correlation function (ACF) in-and-of-itself, can be combined with an existing preamble synchronization sequence to create a complementary codeset with an asymptotically ideal ACF. As the receiver hardware must already implement a correlator for packet detection, this approach to channel estimation requires only nominal additional logic. Once the estimate of the channel impulse response (CIR) has been obtained, the equalization can be performed using an existing ZF equalizer by taking the FFT of the CIR estimate.

This approach to OFDM channel estimation mandates the use a complementary codeset since there are no individual finite-length sequences that have an impulse ACF with an acceptable peak-to-average power ratio (PAPR). In previous work, evolutionary algorithms (EAs), also referred to as genetic algorithms (GAs), have proven effective in optimizing arbitrary criteria [4]. For example, EAs have been used to find OFDM synchronization sequences with minimal ACF sidelobes [5]. In this paper, we use an EA to find a channel estimation sequence that is complementary to an existing preamble synchronization sequence.

Although this work is relevant to many OFDM systems, including IEEE 802.11a and 802.11g, we focus on MB-OFDM, the first ultra-wideband (UWB) technology to obtain international standardization [6]. MB-OFDM is representative of contemporary OFDM wireless systems in that it supports several data rates. This flexibility allows communication over

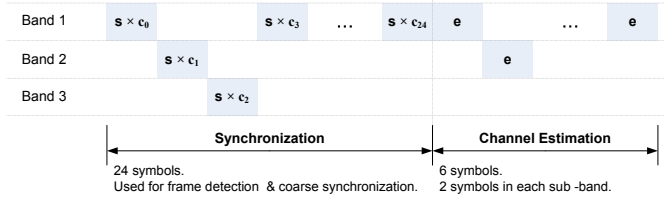


Fig. 2. Preamble sequence in MB-OFDM.

separations of less than 10 cm to well over 10 m and leads to significant variation in channel conditions. With longer UWB channels susceptible to significant frequency-selective fading, it is necessary that the channel estimation is robust and accurate. At the same time, given that an MB-OFDM receiver operates at at least 528MSPS, the total computational complexity must be kept to an absolute minimum in order for cost-effective digital hardware to be realizable.

In recognition of the variety of UWB channel conditions, the IEEE 802.15.3a working group defined 4 UWB channel models [7], denoted as CM1 through CM4, to model line-of-sight (LOS) and non-line-of-sight (NLOS) environments possessing average RMS delay spreads of 5 ns to 25 ns. These channel models all use a parameterized Saleh-Venezuela (S-V) model. In this paper, we use these channel models to show how the proposed equalization is effective in both short and long LOS and NLOS channels.

We begin by describing how we will use complementary codesets for channel estimation in Section II. Then, in Section III, we present an EA that, based on a predefined preamble synchronization sequence, searches for a complementary channel estimation sequence with a given PAPR. The results of Section IV then quantify end-to-end packet error rate (PER) and give an example of several high-performing channel estimation sequences appropriate for MB-OFDM. Finally, we present our conclusions in Section V.

II. COMPLEMENTARY CODESETS

Consider the 30 symbol MB-OFDM preamble shown in Fig. 2. The first 24 symbols are based on a 128 chip synchronization sequence $s(n)$. The receiver uses the auto-correlation of this sequence for frame detection and symbol synchronization as well as potentially frequency and sampling offset estimation. Immediately following this, the next 6 symbols are based on a 128 chip channel estimation sequence $e(n)$. Also shown in a 24 chip cover sequence c_k . This cover sequence has no impact on the ACF and is used only to delineate the synchronization and channel estimation stages of the preamble. Note that, depending on the selected time-frequency code (TFC), there may be frequency hopping. This is of little impact on the channel estimation presented in this paper since it is assumed that each sub-band is estimated independently.

We denote the ACF of the original synchronization sequence, obtained from the MB-OFDM standard [6], as $\phi_{ss}(n)$ and the ACF of our new ideal complementary channel estimation sequence as $\phi_{ee}(n)$. Our objective is therefore to

obtain a channel estimation sequence such that

$$\phi_{ss}(n) + \phi_{ee}(n) = 2\delta(n) \quad (1)$$

where $\delta(n)$ denotes an impulse function where $\delta(0) = 1$ and $\delta(n) = 0$ for $n \neq 0$. We can rearrange (1) to conclude that ideal channel estimation sequence must have an ACF $\phi_{ee}(n) = -\phi_{ss}(n)$ for $n > 0$.

MB-OFDM is different from many other OFDM systems in that it uses a 37 sample zero-pad (ZP) rather than the more conventional cyclic prefix (CP). In addition, the frequency hopping used by some TFCs can potentially completely eliminate inter-block interference (IBI). For both these reasons, it is necessary to base all ACFs on linear convolution instead of circular convolution. Although we do not explicitly consider systems other than MB-OFDM, we note that using a CP will require modification to (2).

Using linear convolution, we denote the ACF of the sought-after channel estimation sequence as

$$\phi_{ee}(n) = \sum_{m=n}^{N-1} e(m)e(m-n) \quad (2)$$

Since the n^{th} tap of this ACF is the sum of $N-n$ coefficients, it is necessary to limit evaluation of the fitness of the ACF to a region-of-interest (ROI) of $0 \leq n < L$, where L denotes channel maximum delay spread. It is important that $L \ll N$, such as $L = 64$ in a $N = 128$ MB-OFDM system, so that there are sufficient degrees of freedom in the code-space.

To balance computational complexity against performance, we restrict the candidate channel estimation sequences to being real-valued at the transmitter and binary at the receiver. In other words, we can denote the ACF of practical sequences as

$$\phi_{\hat{e}\hat{e}}(n) = \sum_{m=n}^{N-1} \hat{e}(m)\text{sgn}[\hat{e}(m-n)] \quad (3)$$

Although this simplification has negligible impact on performance, it significantly reduces complexity since the N multiplications used in the receiver correlation can be simplified to N additions/subtractions.

Fig. 3 shows the relevant normalized ACFs for TFC 1 of the MB-OFDM standard. Note that the ACF at $n = 0$ is not shown since the normalization forces it to 1. In this figure, the maximum channel delay was set to $L = 64$. We can see that the practical ACF, obtained using the EA of Section III, is almost ideal when $n < L$ and totally useless when $n \geq L$. Since an OFDM system assumes that the majority of channel energy arrives within the ZP or CP, lest IBI cripple performance, this causes little or no loss so long as L exceeds the ZP or CP. We conclude from Fig. 3 that practical complementary codesets can be used to obtain an almost ideal ACF.

III. EVOLUTIONARY SEARCH

Finding a finite-length sequence possessing a specific ACF can require the searching of a very large code space. To make this problem computationally feasible, we employ the

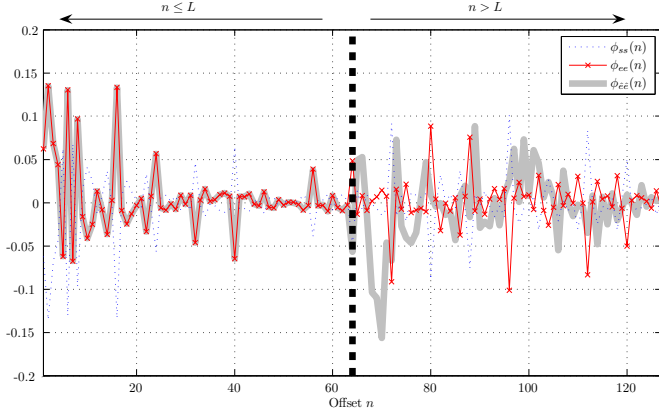


Fig. 3. Comparison of auto-correlations $\phi_{ss}(n)$, $\phi_{ee}(n)$ and $\phi_{\hat{e}\hat{e}}(n)$.

Alg. 1 Evolutionary Algorithm.

- 1) Generate initial population of P sequences of length N .
- 2) Calculate the scaled fitness F_p of each sequence.
- 3) With probability of selection proportional to relative fitness, select P sequences for the next generation.
- 4) For all chips in all selected sequences, apply a probability of mutation of α .
- 5) Iterate through Step 2 until termination condition.

evolutionary algorithm of Alg. 1. Each chromosome \hat{e}_p in the population represents a candidate channel estimation sequence. The algorithm works by biasing future generations towards high performing sequences from the current generation. Like [8], we found that deriving new sequences by crossing over old sequences did not improve the speed of the search. We therefore use mutation as the sole means for genetic diversity.

We define the raw fitness F'_p of sequence p as the inverse of the normalized mean square error (MSE). In other words,

$$F'_p = \sum_{n=0}^{L-1} \frac{\phi_{ss}(n)}{\phi_{ss}(0)} - \frac{\phi_{\hat{e}_p \hat{e}_p}(n)}{\phi_{\hat{e}_p \hat{e}_p}(0)} \quad (4)$$

To help the population retain sufficient diversity to avoid local minima, we employ fitness sharing [9]. Fitness sharing reduces the raw fitness of a given chromosome proportional to its commonality. We therefore denote the scaled fitness as

$$F_p = \frac{F'_p}{\sum_{k=0}^{P-1} \Gamma_{k,p}} \quad (5)$$

where

$$\Gamma_{k,p} = \begin{cases} 1 - \left[\frac{\|\hat{e}_k - \hat{e}_p\|}{\sigma} \right]^\kappa, & \text{when } \|\hat{e}_k - \hat{e}_p\| < \sigma \\ 0, & \text{otherwise} \end{cases} \quad (6)$$

where $\|\hat{e}_k - \hat{e}_p\|$ denotes the Euclidean distance between sequences \hat{e}_k and \hat{e}_p . σ and κ are arbitrary real constants.

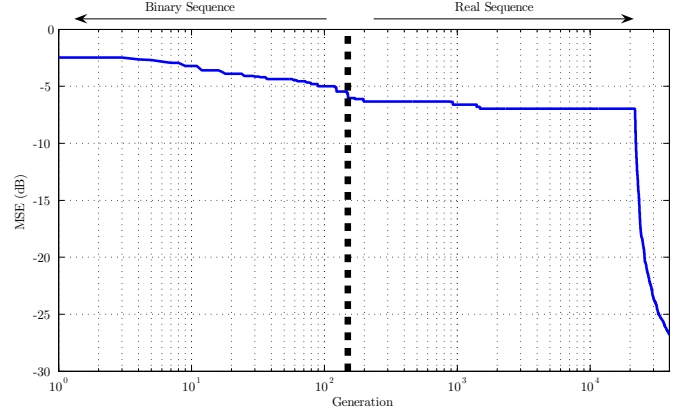


Fig. 4. MSE of population's best chromosome vs. generation count.

Even through we are seeking real-valued sequences, the PAPR

$$\text{PAPR}_{\hat{e}} = \frac{\max_{0 \leq n < N} \hat{e}^2(n)}{\text{mean}_{0 \leq n < N} \hat{e}^2(n)} \quad (7)$$

of the estimation sequence must not exceed the capability of the transceiver. The EA considers this limit by repeating the mutation until the PAPRs of all sequences are within the limit $\text{PAPR}_{\text{MAX}}=10$. We contend that this value is reasonable for two reasons. First is because the OFDM data symbols have a higher worst-case PAPR of $N = 128$. Second is due to ease with which transmitter non-linearities can be mitigated through pre-distortion of the channel estimation sequence.

To help our search for real-valued sequences avoid local-minima, we define two distinct phases of mutation. The first phase limits the sequences $\hat{e}(m)$ to $\{+1, -1\}$ and defines mutation as the inversion of the relevant chips. This phase ends when the best chromosome in the population has been stable for several generations. In the second phase, the binary restriction on the transmitted sequence is relaxed in accordance with (3). During this phase, each mutated chip is selected from a Gaussian distribution with unity variance. The mutation for each sequence is repeated until the PAPR of the sequence is less than PAPR_{MAX} .

Fig. 4 shows how the MSE, defined as $\min_p \frac{1}{F'_p}$, improves with each generation. For all results in this paper, the population size is $P = 100$ and the chip mutation probability is $\alpha = 0.01$. The fitness sharing is parameterized with $\sigma = 20$ and $\kappa = 4$. The termination condition is that the MSE is less than -45 dB. This condition was selected so that the error resultant from the non-ideal ACF is less than the receiver's thermal noise.

IV. RESULTS

The MB-OFDM standard defines seven synchronization sequences such that each TFC has a unique preamble. In this section, we provide seven channel estimation sequences that are complementary to their respective synchronization sequences. We also quantify the potential PER improvements

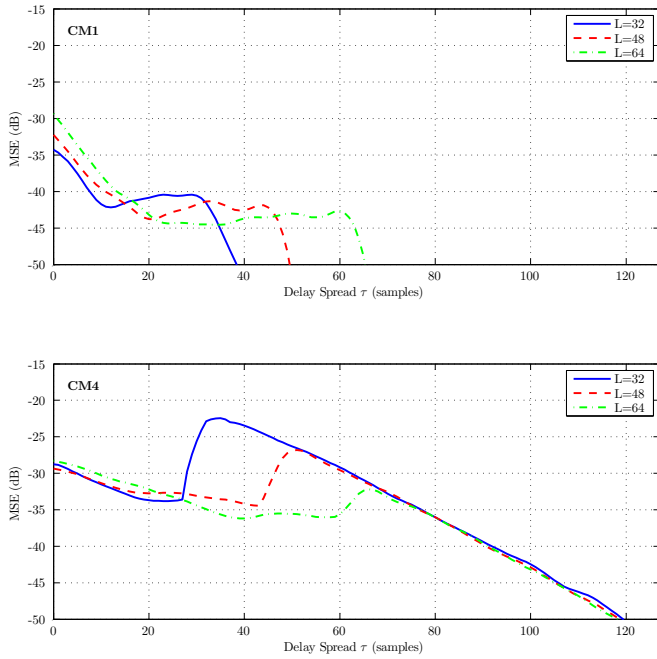


Fig. 5. MSE in channel estimation with respect to delay spread τ .

that can be achieved by employing this approach to channel estimation.

A. ACF Region-of-Interest

It is important that the ACF ROI L is appropriate for the expected channel delay spread. If L is too small, the ACF of (1) will have significant sidelobes at noticeable delays and will introduce error into the channel estimation. Conversely, if L is too large, the accuracy of dominant lower taps of the ACF will be degraded in favor of reducing fringe sidelobes that will never be encountered. We must therefore find an L for which the error due to the non-ideal channel estimation sequence is balanced with the energy lost due to truncation of the CIR.

To determine L , Fig. 5 shows the MSE of the estimated CIR as a function of channel delay for ACF ROIs of 32, 48 and 64 samples. Note that each channel estimation sequence was obtained via Alg. 1 and that both LOS and NLOS channels are considered via CM1 and CM4 respectively. We observe that CM1 channels lose little energy even under truncations as aggressive as $L = 32$. This is not the case in CM4 channels where the ROI must be larger to avoid spikes in the MSE. We conclude that an $L = 64$ ROI is an effective compromise. Although this does lead to a slightly higher MSE for taps $n > 30$ in CM1, the discarded energy is so small that it is of little consequence.

B. Channel Estimation Error

Fig. 6 compares the MSE of FDE and TDE in both CM1 and CM4 at a range of SNRs. The ZF FDE channel estimation is performed by calculating the impact of the channel on the magnitude and phase of symbol's N subcarriers. Although this estimation is computationally simple, it does not consider any

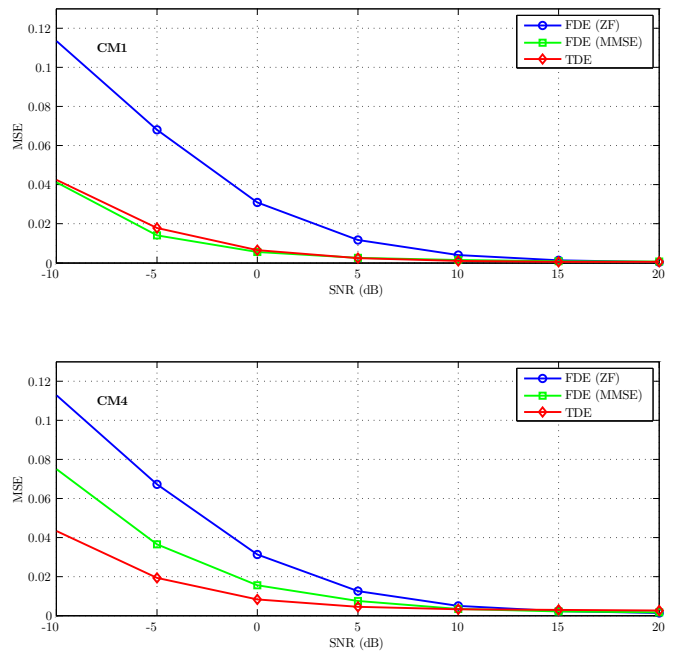


Fig. 6. MSE in channel estimation.

correlation that may exist between subcarriers. This makes it perform poorly in short CM1 channels.

Minimum mean square error (MMSE) FDE [10] uses a pre-computed estimate of the degree of subcarrier correlation to improve the initial ZF estimate. Although this does significantly reduce the MSE, it is extremely computationally intensive as it relies on implementing a long finite-impulse response (FIR) filter with complex coefficients.

The TDE techniques proposed in this paper offer performance that is at least as good as MMSE estimation but at a fraction of the computational complexity. In other words, since the correlator hardware is reused from the synchronization stage, the complexity of the TDE is almost equivalent to that of ZF FDE. In addition, since MMSE FDE requires off-line statistical modeling to calculate the expected channel auto-covariance matrices, TDE actually outperforms MMSE FDE in long channels that are difficult to characterize.

C. Packet Error Rate

Fig. 7 and Fig. 8 show the PER of the proposed equalization scheme for low and high rates respectively in CM1. Fig. 9 and Fig. 10 show the same data rates in CM4. It can be seen that TDE can outperform ZF FDE by over 1 dB. Although MMSE FDE can achieve PER performance similar to TDE, its tremendous computational complexity makes it infeasible in practical receivers.

D. Optimal Channel Estimation Sequences

Effective channel estimation sequences for all seven TFCs in MB-OFDM are provided in Table I. These sequences were obtained via the evolutionary algorithm of Section III and the TFC 1 sequence was used in the PER simulations.

(a) Samples 1 . . . 64

	TFC 1	TFC 2	TFC 3	TFC 4	TFC 5	TFC 6	TFC 7
1	3.0967	-0.0113	1.6018	0.7410	0.0056	1.9453	-0.3981
2	-0.0076	-0.0076	1.0000	0.2811	0.0019	-1.4676	0.0232
3	0.0311	1.4329	-0.6233	0.5020	0.0022	1.6289	-1.0143
4	1.4040	-2.1373	2.0245	-2.4111	1.6342	-0.1946	0.5150
5	-0.0059	0.0734	0.0568	0.7938	-0.0010	0.4971	-1.0225
6	1.0723	-0.0003	-0.9004	-0.8126	1.7735	2.2154	-1.1657
7	0.1597	-1.0218	-0.0116	0.7967	0.7140	0.0227	-2.3474
8	-0.1714	-0.0335	0.8219	-0.1430	0.0034	-1.4017	-1.0281
9	-1.5204	0.8189	1.2646	2.5625	-1.9621	0.6027	-0.9118
10	-1.9859	-2.0216	0.6737	-1.5280	1.1517	-0.5410	-1.0617
11	1.1617	2.3473	-2.3353	-1.3864	0.9414	0.8444	-1.2883
12	-0.0011	0.6383	1.0708	-0.6797	-0.0047	-0.9182	-0.7165
13	1.7849	0.5407	-1.5883	0.0120	-0.0025	-1.1306	-2.2673
14	0.4617	0.9001	0.7420	-2.6034	-0.0017	-0.4401	-0.3438
15	0.6386	-1.1154	0.6738	0.4725	0.8390	1.6879	0.5741
16	-2.0351	-0.6311	1.9906	-0.3799	1.1587	1.4459	2.7439
17	1.1948	-0.3380	-0.4801	-2.0178	-1.4201	1.1615	-1.0220
18	-1.5833	1.2513	-1.5613	1.2552	0.4610	-1.0463	1.0169
19	1.4779	-0.8851	-1.0937	-1.0007	-0.0111	-0.3104	1.4586
20	1.5136	0.6919	-1.2467	0.9643	2.6271	-0.7841	-1.1982
21	1.1621	1.3744	1.5607	2.8268	2.2129	-1.1041	0.1301
22	1.4769	1.0162	-0.9285	-2.1326	1.9171	-1.0132	-1.2833
23	0.0034	1.9988	0.5692	0.8537	-1.5789	1.3988	-1.1773
24	-1.7864	1.0241	-1.2999	1.6839	-1.2704	-2.1574	2.4103
25	-1.2172	-0.7992	-1.0114	0.5800	-0.6384	-0.0000	-1.5466
26	-0.7650	0.3176	0.8367	-0.8879	-0.7602	0.8925	1.0590
27	1.8126	0.2414	-1.6040	0.9113	-1.1909	-1.5897	-0.8497
28	-1.5593	0.9585	-0.2797	-1.6924	0.5303	1.4649	1.0553
29	0.4103	0.9278	-0.5020	-0.6630	0.2454	0.5664	-0.8616
30	1.6778	-0.7228	1.2064	0.9468	1.3207	1.6169	-0.9315
31	1.3377	1.2293	1.4782	0.9456	-0.6480	1.0912	-0.6684
32	1.4785	1.4171	1.4924	-0.2491	1.0924	-1.8913	1.8057
33	-1.0434	-1.4698	-1.4157	-0.1044	1.2461	1.0533	-1.1490
34	-0.9313	0.7233	-0.8821	1.1603	-1.0636	-1.1090	-1.5624
35	0.6725	1.1684	-0.3149	1.3497	0.2857	0.8091	0.9787
36	0.1152	1.3487	0.3495	-0.9852	1.4859	-1.2121	-1.2162
37	1.6085	-0.4606	-0.2649	-1.0407	-0.4482	0.0073	-0.5072
38	2.1369	-0.7118	1.1599	-0.9219	0.8815	1.0422	0.8772
39	0.9745	-0.8256	-0.0052	1.0253	0.3178	-1.3392	-0.8684
40	0.2796	-1.0172	-0.8126	0.8328	1.3256	-1.1169	-1.0780
41	-0.7421	-1.0549	-0.9072	1.2253	1.1138	-0.4611	-0.7848
42	-0.3172	-0.1350	0.9960	-1.5046	1.6187	0.9210	0.0115
43	-0.5981	-0.0532	1.2655	-0.6487	-0.2972	-1.0844	-0.4779
44	-0.0141	0.1984	-0.7820	0.8832	0.2805	-1.6001	-0.5622
45	-1.0332	-1.5916	-1.0784	0.6158	-1.1082	0.1250	1.0522
46	-1.5122	-1.3496	0.5885	0.8076	1.0709	1.0864	-0.7957
47	-2.0664	-1.4318	0.9557	0.6395	-1.2515	-1.1813	-0.4366
48	-0.7741	0.8070	-1.1762	1.2472	0.5310	1.1007	-0.9666
49	-0.5790	1.0265	0.7644	-0.8077	0.3126	-0.0027	-0.8695
50	-0.0037	1.0359	0.9470	0.5916	-0.5851	0.7934	0.7351
51	1.2656	-0.9555	1.0197	-1.8758	0.0126	-0.9292	0.1952
52	0.8974	1.2953	0.8746	0.9067	1.0585	0.9616	0.9691
53	0.6514	-0.9777	1.0478	1.0713	-0.5819	0.7209	-1.2858
54	0.9544	-0.1112	-0.9992	1.4011	-0.0060	1.2180	-1.8018
55	1.4149	0.9402	0.8257	1.1650	-0.5180	0.9023	0.8531
56	1.0804	0.3469	1.0224	0.1600	-1.0509	0.7551	0.9848
57	1.4089	-1.0547	-1.0124	1.4590	0.8360	-0.3993	-1.1119
58	-1.0598	-0.0305	-0.0883	-1.3642	-1.2564	-1.0881	0.3743
59	1.6441	-1.0383	1.0014	1.1983	1.5468	0.6921	-0.4241
60	0.9501	-1.1384	-0.4994	0.0379	0.5591	0.6548	-1.0046
61	0.9754	0.2246	1.6335	0.0032	-1.1545	-0.4220	-0.9938
62	-0.2760	1.4309	-0.5987	1.2300	0.8527	-1.0995	-1.0023
63	-0.3367	0.8507	-1.2152	-0.2162	0.8350	1.1992	0.9859
64	1.1239	0.5341	-0.9050	-0.7915	-1.7449	1.2929	-0.0817

(c) Samples 65 . . . 128

	TFC 1	TFC 2	TFC 3	TFC 4	TFC 5	TFC 6	TFC 7
65	1.4992	1.5937	-1.0075	0.5084	1.5503	0.7981	0.9814
66	0.4854	1.0062	0.1894	1.0154	1.0666	0.5292	0.8174
67	1.2874	-0.9281	-1.0289	1.3578	0.6652	0.1974	1.0127
68	-0.4520	1.0104	0.4700	0.2713	0.4706	-1.0897	-1.4155
69	-1.1701	-0.6125	0.9388	-1.0593	-0.9341	0.0135	1.6613
70	1.5227	-1.1338	-1.1251	0.5419	-1.2420	0.6515	0.8913
71	-0.3317	0.4576	1.0389	-0.9772	1.6277	-0.7518	-0.4097
72	-1.1454	-0.3547	1.0058	1.5766	0.6344	1.0552	-1.5819
73	0.9899	-1.7471	-0.8249	1.2299	-0.7574	1.2369	-0.0044
74	-1.0312	1.0562	0.4682	1.2882	0.0812	0.9845	1.3727
75	-2.1262	-1.5252	-0.8591	0.7677	0.4460	-1.3633	1.0698
76	1.1406	1.0751	-0.6382	1.3442	-0.8786	0.9948	1.0345
77	-1.0091	0.7283	-0.9838	0.9956	-0.0058	0.6011	1.8009
78	-0.7607	1.2378	-0.6055	0.1404	0.9070	1.2722	-1.1836
79	1.1905	-1.1167	-0.8925	0.4517	0.4027	0.9464	1.3130
80	-1.2160	0.0005	0.2153	-0.8706	-0.9291	1.3169	1.0968
81	0.9445	1.1348	-0.8190	-1.0266	-0.7664	0.4654	-0.1400
82	0.0024	0.8206	-0.8684	0.6744	0.3855	1.1951	-0.8169
83	-0.7934	0.8250	-0.8798	-1.1740	-1.2241	1.1067	1.0822
84	-0.8948	0.1262	-1.0895	0.0685	-1.1838	0.9938	0.6661
85	0.4475	-0.6099	0.9178	0.0869	0.4685	1.4398	0.6356
86	-1.5636	0.9098	0.9841	-0.5220	1.1926	0.8887	-1.0385
87	0.3876	0.4728	1.3279	0.7571	-1.4378	-1.2228	0.0630
88	-1.0356	1.8201	-1.5777	-0.9041	0.3874	0.7793	0.7943
89	-1.0469	-0.7564	-0.0954	1.0282	-1.1871	-0.7677	0.9187
90	-0.5597	1.0256	-1.3685	1.4686	-1.0848	0.3436	0.9083
91	-0.3670	-1.1906	-1.1945	-0.1371	0.0015	-0.5336	-0.4480
92	-1.0396	0.3965	-1.4513	-0.9481	-0.2429	0.0006	0.5887
93	0.8715	-0.2978	-1.0479	-1.0571	1.3195	1.0520	-0.9111
94	-1.9810	0.5952	-0.8051	-0.2364	1.6051	0.9904	-0.5643
95	-0.8104	-0.7957	0.9790	-0.9443	0.7856	-1.1090	0.7892
96	-1.0481	0.7412	-1.3610	-1.2709	-1.2594	-0.3923	-0.9977
97	-0.7492	-0.4459	0.8712	0.8444	-1.5789	-0.3564	-0.9166
98	0.8818	0.2438	-0.7655	-0.7990	-1.0629	-0.8142	-0.5766
99	-1.4309	1.3266	0.9829	-1.2177	-1.3412	-0.3976	1.3941
100	1.0485	-0.6089	-1.2158	-1.8743	0.5179	-1.4359	-0.9883
101	-2.0045	1.3896	-1.6772	0.9054	-0.8890	-1.0530	-1.6224
102	-0.0026	0.0390	-0.8467	-1.3045	-0.5164	0.9891	0.9890
103	-0.7028	1.0958	-0.2180	1.0311	0.3972	-1.3110	-1.8990
104	-0.7051	0.8693	-0.0901	-1.0074	1.1317	-1.1253	-1.3677
105	-0.8922	0.9774	0.9071	0.6766	-1.4413	0.0027	-0.2703
106	-1.0133	0.6492	-1.5684	1.1088	0.2985	-0.4740	-0.9527
107	-1.0113	0.7672	-0.5969	-0.8933	-1.4271	-0.3351	1.0078
108	0.8678	-0.6722	-0.8482	0.9879	0.0012	0.7481	-0.5318
109	0.9814	0.7062	-0.7216	-0.9640	-0.9384	-0.6180	-1.5952
110	-0.6137	0.7765	-0.9043	1.0768	-1.0870	-0.8800	0.2702
111	0.6292	1.0398	1.0679	-0.9491	0.9846	1.1281	0.3419
112	0.7830	0.9733	-1.0150	0.9571	-1.5938	1.0349	-0.7192
113	-0.7868	1.0191	1.1588	1.0451	1.4767	1.0109	0.9883
114	-0.4964	-1.3029	-1.0058	1.4849	-1.2787	1.1518	-0.0418
115	-1.1238	-0.9675	0.7411	-0.3374	0.3830	1.1300	0.8512
116	1.3608	-1.0485	-1.5607	0.8711	0.9930	-0.9575	-0.3651
117	-1.8799	1.0713	0.6698	-0.7731	0.8245	1.0257	1.0776
118	0.7429	-0.9991	-0.9241	1.0932	0.6973	0.9326	-0.9511
119	-0.5944	-1.1159	0.9405	-0.9855	-0.6812	-1.4521	0.8588
120	1.2252	-0.8415	0.1291	1.0658	0.7708	1.0707	0.3803
121	0.9940	1.2548	1.1897	0.9757	1.0660	-0.6667	-0.9768
122	1.1687	0.8708	0.6988	0.8954	0.8528	-0.9725	1.0348
123	-0.3805	0.6727	-1.0221	0.8433	1.0237	0.2983	1.2738
124	0.9386	0.8460	1.1414	0.2457	-1.0883	-0.7621	-0.9863
125	-0.7517	-0.8159	-1.0696	-1.2666	1.0869	1.2078	0.8017
126	-1.0210	-1.2087	1.2031	-0.5198	0.2239	0.6378	-1.5032
127	0.0039	0.5211	0.8950	-1.1812	0.1716	-1.5741	-1.5416
128	1.2494	-0.3989	-0.1580	-0.8098	0.7164	1.0775	0.7686

TABLE I
EFFECTIVE CHANNEL ESTIMATION SEQUENCES FOR MB-OFDM.

We conclude that even though TDE requires modification to the existing MB-OFDM standard, its significant performance improvements make it particularly attractive as a proprietary differentiator. We also note that since the synchronization sequence remains unchanged, the use of TDE would not impact legacy receivers in adjacent networks.

V. CONCLUSION

We have presented a novel approach to time-domain channel estimation for MB-OFDM systems that uses a complementary codeset to achieve a near-ideal ACF. We showed how an EA can use the preexisting preamble synchronization sequences to find complementary real-valued channel estimation sequences that facilitate efficient and accurate receiver equalization. We

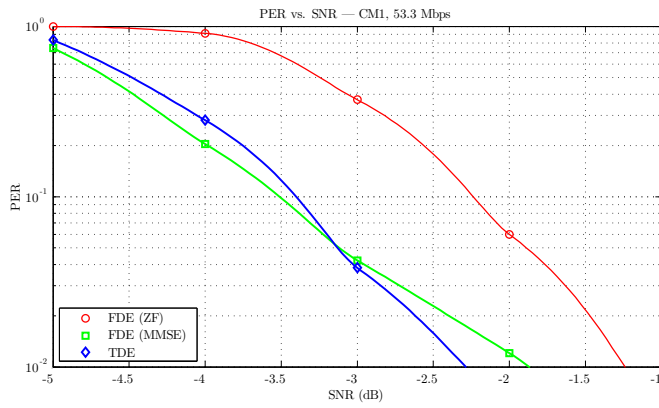


Fig. 7. PER at 53.3 Mbps for FDE and TDE for CM1.

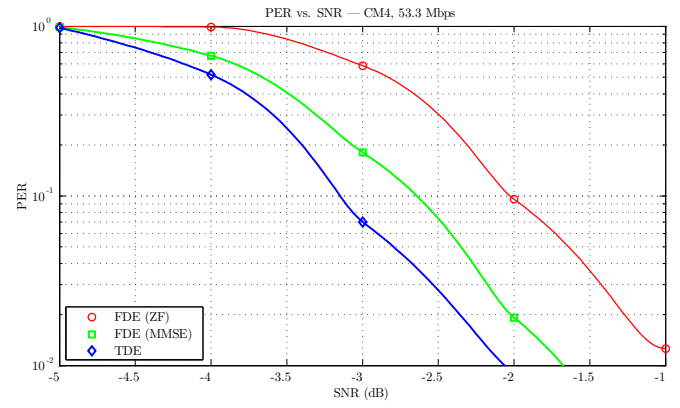


Fig. 9. PER at 53.3 Mbps for FDE and TDE for CM4.

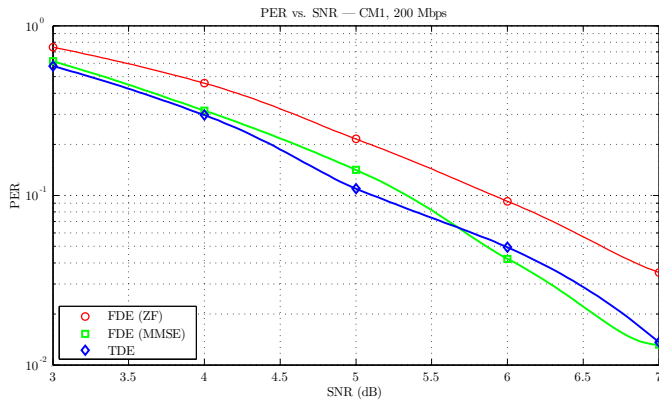


Fig. 8. PER at 200 Mbps for FDE and TDE for CM1.

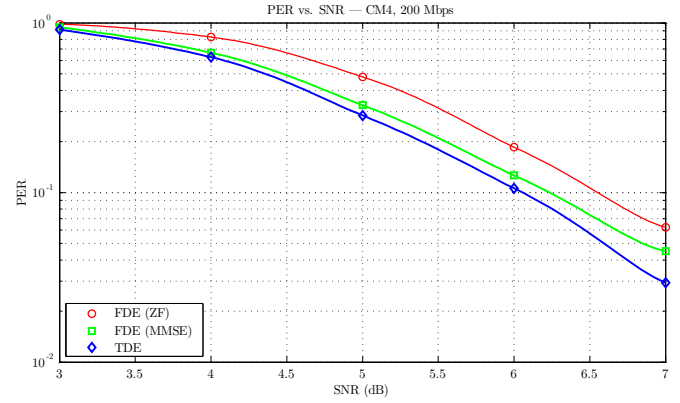


Fig. 10. PER at 200 Mbps for FDE and TDE for CM4.

provided example channel estimation sequences for all 7 TFCs and demonstrated PER improvements up to 1 dB with Monte Carlo simulations.

Future work will develop a multi-objective evolutionary algorithm that aims to derive entirely new synchronization and estimation sequences that offer an optimal balance between ACF, cross-correlation function, PAPR and quality of channel estimation.

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