APPROXIMATE BEST LINEAR UNBIASED CHANNEL ESTIMATION WITH CFAR DETECTION FOR FREQUENCY SELECTIVE SPARSE MULTIPATH CHANNELS WITH LONG DELAY SPREADS

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ABSTRACT

We provide a non-iterative channel impulse response (CIR) estimation algorithm for communication systems which utilize a periodically transmitted training sequence within a continuous stream of information symbols. The non-iterative channel estimate is an approximation to the Best Linear Unbiased Estimate (BLUE) of the CIR, achieving almost similar performance, with much lower complexity. We first provide a formulation of the received data and correlation processing with the adjacent symbol correlation taken into account, and we then present the connections of the correlation based CIR estimation scheme to the ordinary least squares CIR estimation, and the approximate BLUE CIR estimation. Simulation results are provided to demonstrate the performance of the novel algorithms for 8-VSB ATSC Digital TV system.

1. INTRODUCTION

For the communications systems utilizing a periodically transmitted training sequence, *least-squares* (LS) based channel estimation or the *correlation* based channel estimation algorithms have been the most widely used two alternatives [1]. Both methods use a stored copy of the known transmitted training sequence at the receiver. However the accuracy of most channel estimation schemes is degraded due to the *baseline noise* term which occurs due to the correlation of the stored copy of the training sequence with the unknown symbols adjacent to transmitted training sequence, as well as the additive channel noise [1, 10]. In the sequel, we provide (semi-blind) approximate *Best Linear Unbiased Estimate* (a-BLUE) channel estimator for communication systems using a periodically transmitted training sequence [3, 8]. Although the examples following the derivations of the a-BLUE channel estimator are drawn from the ATSC digital TV 8-VSB system [2], to the best of our knowledge it could be applied with minor modifications to any digital communication system with linear modulation which employs a periodically transmitted training sequence. The novel algorithm presented in the sequel is targeted for the systems that are desired to work with channels having long delay spreads L_d; in particular we consider the case where $(NT+1)/2 < L_d < NT$, where NT is the duration of the available training sequence. For instance the 8-VSB digital TV system has 728 training symbols, whereas the delays spreads of the terrestrial channels have been observed to be at least 400-500 symbols long [6, 7, 8]. In addition many channels exhibit a sparse characteristic. In order to exploit the sparseness of the channel we formulate a Constant False Alarm Rate (CFAR) type detector to detect the channel taps which are non zero.

2. OVERVIEW OF DATA TRANSMISSION MODEL

The baseband symbol rate sampled receiver pulse-matched filter output is given by

$$\begin{aligned} y[n] &\equiv y(t)|_{t=nT} = \ _k I_k h[n-k] + \ [n] \\ &= \ _k I_k h[n-k] + \ _k [k] q^*[-n+k], \end{aligned} \quad (1)$$

where $I_k = \begin{cases} a_k, & 0 \le k \le N-1 \\ d_k, & N \le n \le N'-1, \end{cases} \in \mathscr{A} \equiv \{\ _1, \cdots, \ _M\} \subset \mathbb{C}^1$ is the M-ary complex valued transmitted sequence, and $\{a_k\}$ denote the first N known training symbols within a *frame* of N' symbols; $[n] = \ [n] * q^*[-n]$ denotes the complex (colored) noise process after the (pulse) matched filter, with $[n]$ being a zero-mean white Gaussian noise process with variance ^{2} per real and imaginary part; $h(t) = q(t) * c(t) * q^*(-t) = \ _{k=-K}^{L} c_k p(t - \ _k)$ is the complex valued impulse response of the composite channel, and $p(t) = q(t) * q^*(-t)$ is the convolution of the transmit and receive filters where $q(t)$ has a finite support of $[-T_q/2, T_q/2]$, and the span of the transmit and receive filters, T_q, is an even multiple of the symbol period T;$\{c_k\} \subset \mathbb{C}^1$ denote complex valued physical channel gains, and $\{\ _k\}$ denote the Time-Of-Arrivals (TOA). $c(t)$ is assumed to be a static inter-symbol interference (ISI) channel, at least throughout the training period. The symbol rate sampled composite CIR $h[n]$, can be written as $h = [h[-N_a], \cdots, h[-1], h[0], h[1], \cdots, h[N_c]]^T$ where N_a and N_c denote the number of anti-causal and causal taps of the channel, respectively, and $L_d = (N_a + N_c + 1)T$ is the delay spread of the channel (including the pulse tails). The matched filter output which includes *all* the contributions from the known training symbols (including the adjacent random data) is

$$\begin{aligned} y_{[-N_a:N+N_c-1]} &= Ah + Dh + Q \ _{[-N_a-L_q:N+N_c-1+L_q]}, & (2) \\ &= Ah + Hd + Q \ _{[-N_a-L_q:N+N_c-1+L_q]}, & (3) \end{aligned}$$

where $D = \mathscr{T}\{[\underbrace{0,\cdots,0}_{N} d_N, \cdots, d_{N_c+N_a+N-1}]^T, [0, d_{-1}, \cdots, d_{-N_c-N_a}]\}$, $A = \mathscr{T}\{[a_0, \cdots, a_{N-1}, \underbrace{0, \cdots, 0}_{N_a+N_c}]^T, [a_0, \underbrace{0, \cdots, 0}_{N_a+N_c}]\}$, where A is a Toeplitz matrix of dimension $(N + N_a + N_c) \times (N_a + N_c + 1)$ with first column $[a_0, a_1, \cdots, a_{N-1}, 0, \cdots, 0]^T$, and first row $[a_0, 0, \cdots, 0]$, and D is a Toeplitz matrix which includes the adjacent unknown symbols, prior to and after the training sequence. $q = [q[+L_q], \cdots, q[0], \cdots, q[-L_q]]^T$ is the receiver pulse matched filter, and $Q = \begin{bmatrix} q^T & 0 & \cdots & 0 \\ 0 & q^T & \cdots & 0 \\ \vdots & \vdots & \ddots & \vdots \\ 0 & 0 & \cdots & q^T \end{bmatrix}$ and

$$\begin{aligned} H &= \mathscr{H} S^T, & (4) \\ \bar{h} &= [h[N_c], \cdots, h[0], \cdots, h[-N_a]]^T = Jh, & (5) \end{aligned}$$

$$J = \begin{bmatrix} 0 & \cdots & 0 & 1 \\ 0 & \cdots & 1 & 0 \\ \vdots & \vdots & & \vdots \\ 1 & 0 & \cdots & 0 \end{bmatrix}_{(N_a+N_c+1)\times(N_a+N_c+1)} \quad (6)$$

$$\mathscr{H} = \begin{bmatrix} \bar{h}^T & 0 & \cdots & 0 \\ 0 & \bar{h}^T & \cdots & 0 \\ \vdots & \vdots & \ddots & \vdots \\ 0 & 0 & \cdots & \bar{h}^T \end{bmatrix}_{(N+N_c+N_a)\times(N+2(N_a+N_c))} \tag{7}$$

and $d = S\tilde{d}$, or equivalently $\tilde{d} = S^T d$, where

$$\tilde{d} = [d_{-N_c-N_a}, \cdots, d_{-1}, 0_{1\times N}, d_N, \cdots, d_{N+N_c+N_a-1}]^T \tag{8}$$

$$d = [d_{-N-N_a}, \cdots, d_{-1}, d_N, \cdots, d_{N+N_c+N_a-1}]^T \tag{9}$$

$$S = \begin{bmatrix} I_{N_a+N_c} & 0_{(N_a+N_c)\times N} & 0_{(N_a+N_c)} \\ 0_{(N_a+N_c)} & 0_{(N_a+N_c)\times N} & I_{N_a+N_c} \end{bmatrix} \tag{10}$$

where S is $(2(N_c+N_a)) \times (N+2(N_a+N_c))$ dimensional *selection* matrix which retains the random data, eliminates the N zeros in the middle of the vector $\tilde{d}$. where $\bar{h}$ is the time reversed version of h (re-ordering is accomplished by the permutation matrix J), and H is of dimension $(N+N_a+N_c)\times(2(N_c+N_a))$ with a "hole" inside which is created by the matrix S.

3. APPROXIMATE BLUE CIR ESTIMATION

For comparison purposes we first provide the well known correlation and ordinary least squares based estimators, where correlations based estimation is denoted $\hat{h}_u$ (the subscript $_u$ stands for the *uncleaned* CIR estimate) and is given by

$$\hat{h}_u = \frac{1}{r_a[0]} A^H y_{[-N_a:N+N_c-1]}, \tag{11}$$

with $r_a[0] = \sum_{k=0}^{N-1} \|a_k\|^2$, and the ordinary least squares CIR estimate is denoted by $\hat{h}_c$ (the subscript $_c$ stands for the *cleaned* CIR estimate) and is given by

$$\hat{h}_c = (A^H A)^{-1} A^H y_{[-N_a:N+N_c-1]}, \tag{12}$$

where "cleaning" is accomplished by removing the known sidelobes of the aperiodic correlation operation which is accomplished in (11).

We denote the two terms on the right side of Equation (3) by $v = Hd + Q \quad_{[-N_a-L_q:N+N_c-1+L_q]}$. Hence we rewrite (3) as

$$y_{[-N_a:N+N_c-1]} = Ah + v. \tag{13}$$

By noting the statistical independence of the random vectors d and , and also noting that both vectors are zero mean, the covariance matrix, K_v of v is given by

$$\mathrm{Cov}\{v\} = K_v \equiv \frac{1}{2}E\{vv^H\} = \frac{\mathscr{E}_d}{2}HH^H + \;^2 QQ^H, \tag{14}$$

where $\mathscr{E}_d$ is the energy of the transmitted information symbols, and equals to 21 if the symbols $\{d_k\}$ are chosen from the set $\{\pm1, \pm3, \pm5, \pm7\}$. For the model of (13) the generalized least squares objective function to be minimized is $J_{GLS}(h) = \left(y_{[-N_a:N+N_c-1]} - Ah\right)^H K_v^{-1}\left(y_{[-N_a:N+N_c-1]} - Ah\right)$ Then the generalized least-squares solution to the model of (13) which minimizes the objective function of $J_{GLS}(y)$ is given by

$$\hat{h}_K = (A^H K_v^{-1} A)^{-1} A^H K_v^{-1} y_{[-N_a:N+N_c-1]}. \tag{15}$$

The estimator of (15) is called the *best linear unbiased estimate* (BLUE) [9] among all *linear* unbiased estimators if the noise covariance matrix is *known* to be Cov{ } = K . The problem with Equation (15) is that the channel estimate $\hat{h}_K$ is based on the covariance matrix K_v, which is a function of the true channel impulse response vector h as well as the channel noise variance ^{2}. In actual applications the BLUE channel estimate of Equation (15) can not be exactly obtained. Hence we need an *iterative* technique to calculate generalized least squares estimate of (15) where every iteration produces an updated estimate of the covariance matrix as well as the noise variance. Without going into the details, a simplified version of the iterations, which yield a closer approximation to the exact BLUE CIR estimate after each step, is provided in [8].

Alternatively one may achieve nearly the same quality as the results produced by the algorithm described in [8] while at the same time requiring much less computational complexity (i.e., requiring about the same number of multiplications necessary to implement Equation (12)) and having storage requirements similar to that of Equation (12). The initial least squares estimation error can be reduced by seeking an approximation in which it is assumed that the baseband representation of the physical channel $c(t)$ is distortion-free (no multipath); that is $c(t) = \;(t)$ which implies

$$h(t) = p(t) * c(t) = p(t). \tag{16}$$

Thus we can assume that our finite length channel impulse response vector can be approximated by

$$\tilde{h} = [\underbrace{0, \cdots, 0}_{N_a-N_q}, \underbrace{p[-N_q], \cdots, p[0], \cdots, p[N_q]}_{\text{raised cosine pulse}}, \underbrace{0, \cdots, 0}_{N_c-N_q}]^T \tag{17}$$

with the assumptions of $N_a \geq N_q$ and $N_c \geq N_q$, that is the tail span of the composite pulse shape is well confined to within the assumed delay spread of $[-N_aT, N_cT]$. Then the approximation of (17) can be substituted into Equations (4-10) to yield an initial (approximate) channel convolution matrix $\tilde{H}$ and is given by $\tilde{H} = \tilde{\mathscr{H}} S^T$ where $\tilde{\mathscr{H}}$ is formed as in Equation (7) with $\tilde{\bar{h}} = J\tilde{h}$. We can also assume a reasonable received Signal-to-Noise (SNR) ratio measured at the input to the matched filter which is given by SNR $= \frac{\mathscr{E}_d\|(c(t)*q(t))|_{t=nT}\|^2}{2} = \frac{\mathscr{E}_d\|q\|^2}{2}$.For instance we can assume an approximate SNR of 20dB yielding an initial noise variance of $\tilde{\;}^2 = \frac{\mathscr{E}_d\|q\|^2}{100}$. Then combining $\tilde{H}$ and $\tilde{\;}^2$ we can pre-calculate the initial approximate covariance matrix where the covariance matrix of the approximate channel is given by

$$\tilde{K}_v(\tilde{H}) = \frac{1}{2}\mathscr{E}_d \tilde{H}\tilde{H}^H + \tilde{\;}^2 QQ^H, \tag{18}$$

which further leads to the initial channel estimate of

$$\hat{h}_{\tilde{K}} = \underbrace{\left(A^H[\tilde{K}_v(\tilde{H})]^{-1}A\right)^{-1} A^H[\tilde{K}_v(\tilde{H})]^{-1}}_{\text{pre-computed and stored}} y_{[-N_a:N+N_c-1]}. \tag{19}$$

Equation (19) is the resulting a-BLUE CIR estimate. The key advantage of the a-BLUE method is that the matrix $\left(A^H[\tilde{K}_v(\tilde{H})]^{-1}A\right)^{-1} A^H[\tilde{K}_v(\tilde{H})]^{-1}$ is constructed based on the initial assumptions that the receiver is expected to operate, and can be *pre-computed* and *stored* in the receiver. By using the pre-stored matrix $\left(A^H[\tilde{K}_v(\tilde{H})]^{-1}A\right)^{-1} A^H[\tilde{K}_v(\tilde{H})]^{-1}$ as in Equation (19) we obtain a CIR estimate with much lower computational complexity than the BLUE algorithm.

3.1 Analysis of Baseline Noise and CFAR Thresholding

The channel estimates $\hat{h}_c$ or $\hat{h}_{\tilde{K}}$ have contributions due to unknown symbols prior to and after the training sequence, which are elements of the vector d, as well as the additive channel noise. These contributions due to unknown symbols and channel noise is called *baseline noise*, and we can give an expression which summarizes the baseline noise for two different estimators of Equations (12), and (19). The general channel estimate can be written in the form

$$\hat{h} = h + \quad = h + B\left(Hd + Q \quad_{[-N_a-L_q:N+N_c-1+L_q]}\right) \tag{20}$$

where the baseline noise vector is defined by

$$= \quad B\left(Hd + Q \quad _{[-N_a-L_q:N+N_c-1+L_q]}\right) \tag{21}$$

and the matrix B takes one of the two following different forms depending on the estimator used:

$$B = \begin{cases} \left(A^H A\right)^{-1} A^H, \text{ for } \hat{h}_c \\ \left(A^H[\tilde{K}_v(\tilde{H})]^{-1}A\right)^{-1} A^H [\tilde{K}_v(\tilde{H})]^{-1}, \text{ for } \hat{h}_{\tilde{K}} \end{cases} \tag{22}$$

Although we can derive the exact probability distribution of the baseline noise term, we can alternatively make the assumption of *normality* (having Gaussian distribution) of the baseline noise. This assumption can be asserted by invoking the central limit theorem[4]. The baseline noise vector has covariance matrix $K \quad = \mathrm{Cov}\{ \quad \}$

$$K \quad = B(\frac{\mathscr{E}_d}{2} HH^H + \quad ^2 QQ^H)B^H = BK_v B^H \tag{23}$$

where K_v is given in (14), and we make the approximation

$$\sim \mathscr{N}(0, B(\frac{\mathscr{E}_d}{2} HH^H + \quad ^2 QQ^H)B^H) = \mathscr{N}(0, BK_v B^H) \tag{24}$$

by invoking the central limit theorem, where B takes one of the appropriate forms as displayed in Equation (22).

We also provide the probability distribution of $\| \quad _k\|^2$ where subscript k denotes the kth element of the baseline noise vector $= [\quad _1, \ldots, \quad _{N_a+N_c+1}]^T$. Based on (24) we can show that $_k$ has a Gaussian marginal distribution with zero mean and variance[4]

$$^2_k \equiv \frac{1}{2} E\{ \quad _k \quad ^*_k\} = 1_k^T BK_v B^H 1_k \tag{25}$$

that is $_k = 1_k^T B(Hd + Q \quad _{[-N_a-L_q:N+N_c-1+L_q]})$, and

$$_k \sim \mathscr{N}(0, \underbrace{1_k^T BK_v B^H 1_k}_{\ ^2_k}), \tag{26}$$

where B takes one of the appropriate forms as displayed in Equation (22), and $1_k = [\underbrace{0, \ldots, 0}_{k-1}, 1, 0, \ldots, 0]^T$ is the vector of zeros of appropriate dimension with a 1 at the kth position.

Now we state an important fact about the probability distribution of the square-norm of the complex Gaussian random variables [11]. Let $= \quad _r + j \quad _q$ be a complex valued random variable, with statistically independent real and imaginary parts $_r$ and $_q$. Given that is Gaussian with 0 mean and variance $^2 = \ ^2_r = \ ^2_q = \frac{1}{2}E\{ \quad ^*\}$, the random variable defined by $Z = \| \quad \|^2 = \ ^2_r + \ ^2_q$ is exponentially distributed, and its density is given by

$$p_Z(z) \quad = \quad \begin{cases} \frac{1}{2 \ ^2} e^{-\frac{z}{2 \ ^2}}, & r \geq 0 \\ 0, & r < 0. \end{cases} \tag{27}$$

Although it is apparent that the real and imaginary parts of the baseline noise $_k$ are not statistically independent, for the sake of obtaining a simple thresholding rule and for the special case of Digital TV system the correlation can be shown to be small, we will proceed as if the real and the imaginary parts of $_k$ are uncorrelated. With this simplified assumption $\| \quad _k\|^2$ is an exponentially distributed random variable with parameter $2 \ ^2_k$, and the density function is

$$p_{\| \ _k\|^2}(r) \quad = \quad \begin{cases} \frac{1}{2 \ ^2_k} e^{-\frac{r}{2 \ ^2_k}}, & r \geq 0 \\ 0, & r < 0. \end{cases} \tag{28}$$

where ^2_k is defined by Equation (25).

Right after obtaining a channel estimate, prior to using that channel estimate for noise variance, ^{2}, calculation and prior to building the channel convolution matrix H, the baseline noise has to be cleaned from the channel estimate. This cleaning can be achieved via thresholding. Previously we have used a fixed thresholding algorithm [6] to get rid of the baseline noise. We have observed that there can be significant performance loss if a fixed thresholding is applied at every iteration. This performance loss is inevitable due to getting rid of significant amount of pulse tails embedded in the channel impulse response while getting rid of the baseline noise. To overcome this problem we propose constant false alarmrate (CFAR) based thresholding, and the threshold calculation is based on the approximate statistical distribution of the baseline noise which is already provided in (26).

Recall that the kth tap of the channel estimate vector can be expressed in the form

$$\hat{h}_k = h_k + \underbrace{1_k^T B\left(Hd + Q \quad _{[-N_a-L_q:N+N_c-1+L_q]}\right)}_{k}, \tag{29}$$

and $_k$ has a Gaussian distribution with zero mean and variance $^2_k = 1_k^T BK_v B^H 1_k$ where B takes one of the appropriate forms as displayed in Equation (22), and the random variable $\| \quad _k\|^2$ is assumed to have exponential distribution with parameter $2 \ ^2_k$.

The problem of deciding whether the kth tap estimate $\hat{h}_k$ is a zero tap or not can be formulated as a simple hypothesis testing problem. That is we consider

$$H_0 \quad : \quad \hat{h}_k = \quad _k, \tag{30}$$

$$H_1 \quad : \quad \hat{h}_k = h_k + \quad _k; \tag{31}$$

where under H_0 the hypothesis is that the kth channel tap is actually zero and we are observing only baseline noise, and under H_1 the hypothesis is that the channel tap is non-zero, and we are observing (non-zero) channel tap plus the baseline noise. We have shown that the probability distributions of the kth channel tap under each hypothesis is given by $H_0 : \hat{h}_k \sim \mathscr{N}(0, \ ^2_k)$, $H_1 : \hat{h}_k \sim \mathscr{N}(h_k, \ ^2_k)$. We can come up with different decision rules on how to threshold the channel estimate $\hat{h}$, however we choose to pursue CFAR based thresholding. False alarm probability based decision rule is chosen so that the resulting threshold rule does not require any a priori knowledge of the distribution of the hypothesis H_1, it is solely based on H_0. False alarm rate is the probability of choosing H_1 when H_0 is true. Our decision rule will be in the form of

$$\text{set } \hat{h}_k^{(th)} = \begin{cases} 0, & \text{if } \|\hat{h}_k\|^2 < \quad _k \\ \hat{h}_k, & \text{otherwise.} \end{cases} \tag{32}$$

Based on the rule of (32) the false alarm rate, denoted by p_{FA} is given by

$$p_{FA} = \Pr\{\|\hat{h}_k\|^2 \geq \quad _k | H_0 \text{ is true }\} = \int_{\ _k} \frac{1}{2 \ ^2_k} e^{-\frac{r}{2 \ ^2_k}} dr = e^{-\frac{\ _k}{2 \ ^2_k}}. \tag{33}$$

For the given level of false alarm probability p_{FA} the threshold $_k$ is

$$_k \quad = \quad -2 \ ^2_k \ln(p_{FA}) \tag{34}$$

where ^2_k is given by (25).

Although we end up with a very simple expression for the threshold of Equation (34), which should be applied to the channel estimate as in (32), we still have the problem of not knowing the true covariance matrix $B(\frac{1}{2}\mathscr{E}_d HH^H + \quad ^2 QQ^H)B^H$ and the kth diagonal element which we have denoted by $_k^2$. We can only have an estimate $\widehat{\ }^2_k$ available to be used in Equation (34). Thus it is natural to see some performance loss due to using the estimate $\widehat{\ }^2_k$ in place of the true variance as will be shown in the simulations. Indeed the thresholding step is going to be incorporated into the iterations of the channel estimation with covariance matrix updated at every iteration. Once the covariance matrix is updated at every iteration we would have a new, and presumably better, threshold $_k$ since we will get a better estimate $\widehat{\ }^2_k$ at every iteration.

4. SIMULATIONS

We considered an 8-VSB [2] receiver with a single antenna. 8-VSB system has a complex raised cosine pulse shape [2]. The CIR we considered is given in Table 1. The phase angles of individual paths for all the channels are taken to be $\arg\{c_k\} = \exp(-j2\ \ f_c\ \ _k)$, $k = -1, \cdots, 6$ where $f_c = \frac{50}{T}$ and $T = 92.9$nsec. The simulations were run at 28dB Signal-to-Noise-Ratio (SNR) measured at the input to the receive pulse matched filter, and it is calculated by SNR $= \frac{\mathcal{E}_d \|\{c(t)*q(t)\}_{t=kT}\|^2}{\ ^2}$. Figure 1 shows the simulation results for the test channel provided in Table 1. Part (a) shows the actual CIR; part (b) shows the correlation based CIR estimate, of Equation (11) $\hat{h}_u$; part (c) shows the ordinary LS based CIR estimate of Equation (12) $\hat{h}_c$; part (d) shows the approximate BLUE CIR estimate of Equation (19) with an assumed SNR of 20dB; part (e) shows the BLUE based CIR estimate of Algorithm 1, after the first iteration, $\hat{h}_K[1]$, where we used CFAR based thresholding with $p_{FA} = 10^{-5}$; part (f) shows the ideal BLUE case for which the true covariance matrix K_v is known. Part (f) provides a bound for the rest of the BLUE algorithm. We note superior performance of the BLUE algorithm even after the first iteration, as compared to the correlation based and ordinary least squares based CIR estimation schemes. However iterative BLUE CIR estimation algorithm is computationally very demanding, thus in many applications the approximate BLUE, as shown in part (d), might be sufficiently acceptable as an initial estimate. The performance measure is the normalized least-squares error which is defined by $\mathcal{E}_{NLS} = \frac{\|h-\hat{h}\|^2}{N_a+N_c+1}$. Approximate BLUE significantly outperforms the ordinary least squares CIR estimation, but it has virtually identical computational complexity and storage requirement.

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Table 1: Simulated channel delays in symbol periods, relative gains. $L = -1, K = 6, L_d \approx (1+333+2N_q)T = 453T \approx 44\ \ $ sec, $N_q = 60$.

Channel taps	Delay $\{\ _k\}$	Gain $\{\|c_k\|\}$
$k = -1$	-0.957	0.7263
Main $k = 0$	0	1
$k = 1$	3.551	0.6457
$k = 2$	15.250	0.9848
$k = 3$	24.032	0.7456
$k = 4$	29.165	0.8616
$k = 5$	221.2345	0.6150
$k = 6$	332.9810	0.4900

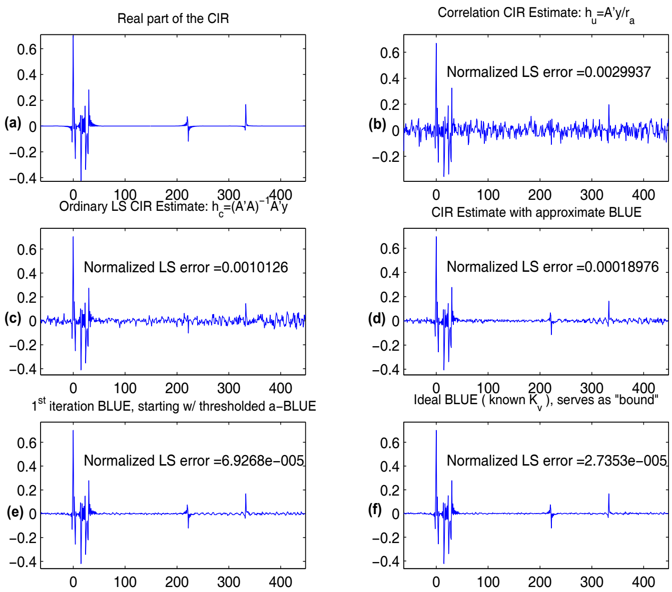

Figure 1: Part (a) shows the real part of the actual CIR; part (b) shows the correlation based CIR estimate of Equation (11) $\hat{h}_u$; part (c) shows the LS based CIR estimate of Equation (12) $\hat{h}_c$; part (d) shows the approximate BLUE CIR estimate of Equation (19) with an assumed SNR of 20dB; part (e) show the BLUE based CIR estimate of Algorithm 1, after the first iteration, $\hat{h}_K[1]$; part (f) shows the ideal BLUE case for which the true covariance matrix K_v is known, which provides a bound for the rest of the estimators.