JOINT BLIND SIGNAL DETECTION AND CARRIER RECOVERY OVER FADING CHANNELS

Lang Tong
Department of Electrical and Systems Engineering
University of Connecticut, Storrs, CT 06269-3157
Tel: (203) 486-2192, EM: ltong@eng2.uconn.edu

Abstract
A scheme of joint blind signal detection and carrier recovery is developed for intersymbol interference channels. Using a blind sequence estimation scheme, the proposed approach does not require explicit channel identification. It shows robustness against timing error and carrier phase uncertainty. The efficacy of the proposed approach is demonstrated in a simulation involving frequency selective fast fading channel with substantial doppler shifts.

1. INTRODUCTION

Reliable and efficient communications over bandlimited wireless fading channels presents enormous technical challenges. One of the most difficult tasks is the identification and equalization of time varying channels without sacrificing the transmission efficiency. The so-called blind equalization, first proposed by Sato [10], equalizes a channel without requiring the transmission of training signals. Recently, there have been increasing interests in "blind" techniques for the equalization of time-varying channels due to the potential gain of transmission efficiency. Many new ideas [15, 4, 17, 6, 7, 11, 9]. have been proposed recently since the publication of a closed-form solution in [14].

While blind equalization techniques have the potential to to track time-varying channels and, in the event of unexpected communication outage, bootstrap the system back to normal operation, several important issues have not been addressed adequately. One of the difficult issues is the problem relating to frequency offset and its effects on blind equalization schemes. In practical wireless transmission, for example, frequency-offset, phase jitter, and Doppler shifts are unavoidable. Traditionally, decision-feedback/decision-directed types of approaches are used (see, *e.g.*,[1]). The application of such techniques, however, is non-trivial when *blind* equalization methods are involved. Some of the existing techniques used in blind equalization (eg. [2, 8, 3]) follow closely to the traditional approaches that assume the equalizer has achieved convergence and detected symbols are relatively error-free. Such assumptions may not be satisfied for a relatively rapid changing channel. As a consequence, catastrophic error propagation may occur. Although it can be shown that frequency offsets do not affect the eigenstructure-based blind channel identification algorithm proposed recently (*e.g.*,[14, 7, 5]), the blind equalizer constructed from the identified channel often produces rapidly rotating equalized symbols due to frequency offset and Doppler shifts. Such rotation not only increases detection error, but also complicates the possibilities of combining blind techniques with regular decision feedback equalizers (DFE). In summary, there is a cogent need to develop robust schemes for blind equalization under unknown and perhaps time varying frequency offsets.

While most blind equalization methods focus on the identification or equalization of the channel, very few approaches the problem of blind signal detection directly without channel identification or equalization. To our knowledge, [13, 12] is perhaps the first such attempt. However, the issue of carrier recovery and frequency offset was ignored. In this paper, we present an approach of joint blind sequence estimation and carrier recovery. The efficacy of the proposed approach is demonstrated in a simulation involving frequency selective fast fading channel with significant doppler shifts.

2. THE CHANNEL MODEL

We consider an array of M receivers. The passband model of the ith frequency selective fading channel is illustrated in Figure 1. Sampled at a frequency F_s that is an integral multiples of the symbol rate $\frac{1}{T}$, the (discrete-time) received signal $x_i(t)$ can be modeled by

$$x_i(t) = \sum_k s_k g_i(t-kT) e^{j(\omega t+\phi_i(t))} + n_i(t) \qquad (1)$$

where $\{s_k\}$ is the sequence of information symbols, T is the symbol period, $\omega = w_c - w_d$ is the frequency offset, in some cases, the Doppler shift. $\phi_i(t)$ represents the phase jitter and perhaps the Doppler spread. $g_i(t)$ and $n_i(t)$ are the (equivalent) impulse response of the baseband channel between the source and the i^{th} receiver, and the additive

This work was supported in part by the National Science Foundation under Contract NCR-9321813

0-7803-2431-5/95 $4.00 © 1995 IEEE

noise, respectively. Some of the major assumptions made in this paper are now given. We assume that $\{s_k\}$ is a zero mean and uncorrelated sequence, i.e., $E(s_i s_j^*) = \delta(i-j)$. The noise $n_i(t)$ is zero-mean and uncorrelated with the information sequence $\{s_k\}$. We also assume that the channel impulse response $g_i(t)$ has a finite duration no longer than dT, and it satisfies the blind identifiability condition [15], namely, the M channels $\{g_i(t)\}$ do not share a common set of uniformly $\frac{2\pi}{T}$-spaced zeros.

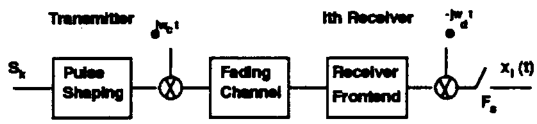

Figure 1: A Passband Model.

The equivalent baseband model can be obtained [1] by letting

$$h_i(t) = g_i(t)e^{j\omega t}, \qquad z_k = s_k e^{j\omega kT}. \tag{2}$$

From (1)

$$x_i(t) = e^{j\phi_i(t)} \sum_k z_k h_i(t-kT) + n_i(t). \tag{3}$$

Under the assumption that $h_i(t,\tau)$ lasts no more than d symbol intervals, a vector representation of the received signal at the ith receiver is obtained by letting

$$\begin{aligned}
\mathbf{x}_i(t) &= [x_i(tT), \cdots, x_i((t-1)T+1)]^t, && (4)\\
\mathbf{n}_i(t) &= [n_i(tT), \cdots, n_i((t-1)T+1)]^t, && (5)\\
\mathbf{h}_i(t) &= [h_i(tT), \cdots, h_i((t-1)T+1)]^t, && (6)\\
\mathbf{H}_i &= [\mathbf{h}_i(0), \cdots, \mathbf{h}_i(d-1)], && (7)\\
\mathbf{D}_i(t) &= diag(e^{j\phi_i(tT)}, \cdots, e^{j\phi_i((t-1)T+1)}), && (8)\\
\mathbf{z}(t) &= [z_t, z_{t-1}, \cdots, z_{t-d+1}]^t, && (9)\\
\mathbf{E}_t &= diag(e^{j\omega Tt}, \cdots, e^{j\omega T(t-d)}), && (10)\\
\mathbf{s}(t) &= [s_t, s_{t-1}, \cdots s_{t-d+1}]^t. && (11)
\end{aligned}$$

We now have, similar to that derived in [14, 16],

$$\begin{aligned}
\mathbf{x}_i(t) &= \mathbf{D}_i(t)\mathbf{H}_i\mathbf{z}(t) + \mathbf{n}_i(t) && (12)\\
&= \mathbf{D}_i(t)\mathbf{H}_i\mathbf{E}_t\mathbf{s}(t) + \mathbf{n}_i(t). && (13)
\end{aligned}$$

Putting all the data from M receivers in a single matrix form, let

$$\begin{aligned}
\mathbf{x}(t) &= [\mathbf{x}_1^t(t), \cdots, \mathbf{x}_M^t(t)]^t, && (14)\\
\mathbf{n}(t) &= [\mathbf{n}_1^t(t), \cdots, \mathbf{n}_M^t(t)]^t, && (15)\\
\mathbf{D}(t) &= diag(\mathbf{D}_1(t), \cdots, \mathbf{D}_M(t)), && (16)\\
\mathbf{H}(t) &= [\mathbf{H}_1^t, \cdots, \mathbf{H}_M^t]^t. && (17)
\end{aligned}$$

The vector representation is given by

$$\mathbf{x}(t) = \mathbf{D}(t)\mathbf{H}\mathbf{E}_t\mathbf{s}(t) + \mathbf{n}(t). \tag{18}$$

The above equation indicates different effects of fading of a time varying channel. $\mathbf{D}(t)$ is primarily caused phase jitters and flat fading and $\mathbf{H}$ by frequency selective fading. $\mathbf{E}_t$ represents the effect of frequency offset. Our objective is to estimate the information sequence $\mathbf{s}(t)$, and at the same time, compensate $\mathbf{D}(t)$ and $\mathbf{E}_t$ adaptively ($\mathbf{H}$ is time invariant). The technique presented below takes advantage the statistical property of $\mathbf{s}(t)$ and special structures $\mathbf{D}(t)$ and $\mathbf{E}_t$ which are both unitary.

3. BLIND SEQUENCE ESTIMATION AND CARRIER RECOVERY

The key idea of blind sequence estimation rests on the construction of the so-called property preserving transform $\mathcal{T}_t$ such that $\mathcal{T}_t(\mathbf{x}(t))$ preserves properties of $\{s_t\}$ from which $\{s_t\}$ can be recovered. The distinct feature of this type of approach is that the identification of the channel is not required.

To illustrate this idea, suppose that there is an orthogonalisation transform $\mathbf{T}(t)$ such that

$$\mathbf{T}(t)\mathbf{D}(t)\mathbf{H} = \mathbf{V}, \tag{19}$$

where $\mathbf{V}$ is some orthogonal matrix. The existance of such a matrix is quite obvious and the structure of such a matrix will be examined shortly. Define

$$\mathbf{y}(t) = \mathbf{T}(t)\mathbf{x}(t) = \mathbf{V}\mathbf{E}_t\mathbf{s}(t). \tag{20}$$

We now have, for the noiseless case,

$$\begin{aligned}
\mathbf{y}^*(t)\mathbf{y}(t-1) &= \mathbf{s}^*(t)\mathbf{E}_t^*\mathbf{E}_{t-1}\mathbf{s}(t-1) && (21)\\
&= e^{-j\omega T}\mathbf{s}^*(t)\mathbf{s}(t-1). && (22)
\end{aligned}$$

We make two key observations.

1. The time varying term $e^{j\omega t}$ in (1) becomes the time invariant term $e^{-j\omega T}$ in (21).
2. The correlation function of $\mathbf{s}(t)$ can be recovered (up to a constant phase) from the observation process without knowing the channel.

It is the inner product preserving property (21) that enables a direct application of the Viterbi algorithm to the transformed observation $\mathbf{y}(t)$. Denote

$$\begin{aligned}
r_y(t) &= \mathbf{y}^*(t)\mathbf{y}(t-1), && (23)\\
r_s(t) &= \mathbf{s}^*(t)\mathbf{s}(t-1) = \sum_{l=0}^{d-1} s_{t-l}^* s_{t-l-1}. && (24)
\end{aligned}$$

(25)

We have,

$$r_y(t) = e^{j\omega T} r_s(t) + w(t), \tag{26}$$

where $w(t)$ accounts for the contribution of noise. For a fixed ω, the "optimal" sequence detection can be defined in the following sense

$$\min_{\{s_n\}} \sum_t |r_y(t) - e^{j\omega T} r_s(t)|^2. \tag{27}$$

and this optimization can be easily achieved by applying the Viterbi algorithm.

We examined a simple approach where a two-step approach is proposed:

- *Initial Carrier Estimation:*
  Find the ML estimation of ω at a t_0:

$$\hat{\omega}_0 = \arg \min_{\omega, r_s(t_0)} |r_y(t_0) - e^{j\omega t_0} r_s(t_0)|^2 \tag{28}$$

- *Sequence Estimation:*
  At step $l+1$, applying the Viterbi algorithm

$$\{\hat{s}_t\}_l = \arg \min_{\{s_n\}} \sum_t |r_y(t) - \hat{e}^{j\hat{\omega}_l T} r_s(t)|^2. \tag{29}$$

$$\hat{r}_{s,l} = \hat{\mathbf{s}}_l^*(t)\hat{\mathbf{s}}_l^*(t-1). \tag{30}$$

- *Carrier Recovery:*

$$e_{l+1}^{j\hat{\omega}T} = e^{j\angle \sum_t \hat{r}_{s,l}^*(t) r_y(t)}. \tag{31}$$

The orthogonalization transform $\mathbf{T}(t)$ plays a central role in our approach. One of the important questions is how to construct such a transform in a efficient way. Consider the vector representation (18), and denote the autocorrelation function of the observation by

$$\mathbf{R}_x(t,\tau) = E(\mathbf{x}(t)\mathbf{x}^*(t-\tau)). \tag{32}$$

We then have

$$\mathbf{R}_x(t,0) = \mathbf{D}(t)\mathbf{H}\mathbf{H}^*\mathbf{D}^*(t) + \sigma^2\mathbf{I} \tag{33}$$

$$= \mathbf{D}(t)(\mathbf{H}\mathbf{H}^* + \sigma^2\mathbf{I})\mathbf{D}^*(t). \tag{34}$$

Under the condition that the the M channels do not share a common set of uniformly $\frac{2\pi}{T}$-spaced zeros, it was shown that $\mathbf{H}$ is of full column rank [15], *i.e.*, $rank(H) = d$. It is well known that the singular value decomposition (SVD) of $\mathbf{R} = (\mathbf{H}\mathbf{H}^* + \sigma^2\mathbf{I})$ must have the following form

$$\mathbf{R} = \mathbf{U} diag(\pi_1^2, \cdots, \pi_M^2)\mathbf{U}^* \tag{35}$$

$$= \mathbf{U_S} diag(\lambda_1^2, \cdots, \lambda_d^2)\mathbf{U_S}^* + \sigma^2\mathbf{I}, \tag{36}$$

where $\mathbf{U}_s$ is the submatrix consisting the first d columns of U, and $\pi_i^2 = \lambda_i^2 + \sigma^2$. The so-called Mahalanobis orthogonalization transform, defined by

$$\mathbf{T}_m = \mathbf{\Lambda}_s^{-1}\mathbf{U}_s^*, \tag{37}$$

$$\mathbf{\Lambda_S} = diag(\lambda_1, \cdots, \lambda_d), \tag{38}$$

has the following simple but important property.

Properties 1 *The Mahalanobis transform matrix* $\mathbf{T}_m$ *orthogonalizes the channel parameter matrix* $\mathbf{H}$*, i.e., there is an orthogonal matrix* $\mathbf{V}$ *such that*

$$\mathbf{T}_m\mathbf{H} = \mathbf{V}. \tag{39}$$

Recalling now that $\mathbf{D}(t)$ is itself unitary, we have immediately a time-varying orthogonal transformation

$$\mathbf{T}(t) = \mathbf{T}_m\mathbf{D}^*(t) \tag{40}$$

so that there exists a *time invariant* $\mathbf{V}$ such that

$$\mathbf{T}(t)\mathbf{D}(t)\mathbf{H} = \mathbf{V}. \tag{41}$$

The implication of (40) is that, although the orthogonalisation transform is time varying, only the diagonal phase $\mathbf{D}(t)$ changes with time. Therefore, the tracking of the orthogonalisation matrix $\mathbf{T}(t)$ is equivalent to the tracking of $\mathbf{D}(t)$, which may significantly reduce the complexity.

4. A SIMULATION EXAMPLE

We consider, in this simulation example, an antenna array with three receivers, each sampled at a rate four times of the symbol rate. The composite channels were given by

$$g_i(t) = \alpha_{i1}p(t) + \alpha_{i2}p(t-\tau_1) + \alpha_{i3}p(t-\tau_2), \quad i = 1,2,3. \tag{42}$$

where $p(\cdot)$ was a raised-cosine pulse with 90% roll-off. The delays were $\tau_1 = 0.9332T$ and $\tau_2 = 0.9811T$. The gain coefficients $\{\alpha_{ij}\}$ at the ith receiver with respect to the jth multipath were

$$\begin{pmatrix} 0.1883 + 0.9692j & 0.1735 - 0.5120j & 1.4965 - 0.0687j \\ -0.6797 + 0.7145j & -1.2401 + 0.2360j & -0.9172 - 0.6860j \\ 0.5612 + 0.5772i & -0.2201 - 0.1872j & 0.0145 - 0.5511j \end{pmatrix}, \tag{43}$$

which were generated from a zero mean unit variance Gaussian distribution. The frequency offset was such that $\omega T = \frac{4\pi}{3}$. To simulate the fading effects, the received signal is multiplied by a complex Gaussian process obtained from the output of a third-order Butterworth filter with the normalised fading bandwidth of 0.1. Specifically, the received signal $x_i(t)$ at the ith receiver satisfies

$$x_i(t) = e^{j\omega t} f_i(t) \sum_k s_k g_i(t-kT) + n_i(t), \tag{44}$$

$$f_i(t) = 1 + q(t) * \nu_i(t), \tag{45}$$

where $q(t)$ is the impulse response of the Butterworth filter and $\nu_i(t)$ are independently generated zero-mean Gaussian processes with variance $\sigma_{\nu_i}^2 = 0.16$. The source was an uncorrelated BPSK signal.

For every 100 bits of data, we estimated the covariance matrix $\mathbf{R}_x(0)$. We did not assume the dimension of the signal space was known. Starting with signal subspace dimension $d = 1$, the transformation matrix $\mathbf{T}_o$ was formed from the SVD of the estimated covariance matrix $\hat{\mathbf{R}}_x(0)$. The correlation function $r_y(t)$ of the transformed data was computed. The initial estimate of ω was obtained by taking the phase of $r_y(t_0)$ where $|r_y(t_0)|$ is maximum. The Viterbi algorithm was then used to estimate $\{s_k\}$. If there were errors in the symbol estimation, the signal dimension d was increased to obtain the sequence estimate with minimum error. Such a scheme may be implemented in practice if the source was coded with error detection We used this scheme for the purpose of finding the best performance that can be offered by this algorithm. Figure 2 shows the estimated bit-error-rate vs. SNR. Total 10^4 bits were tested for SNR below $7dB$, 10^5 bits for $SNR = 9 - 15dB$. and 10^7 bits for $SNR = 9dB$. For $SNR \geq 13dB$, there was no error found in 10^5 testing bits. It appears that using 100 symbols to estimate the necessary statistics was adequate.

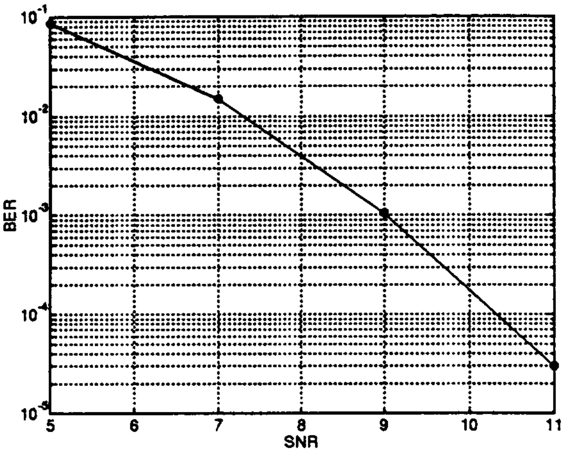

Figure 2: BER vs. SNR. Signal statistics were estimated every 100 symbols.

5. CONCLUSIONS

We presented in this paper a joint blind sequence estimation and carrier recovery method. We have shown that the identificaiton of the channel is not necessary when the objective is to recover the information symbols. The proposed approach is particularly useful in cases when the carrier is not known, or there are substantial phase jitter, timing errors and Doppler shifts.

REFERENCES

[1] D.D. Falconer. "Jointly adaptive equalization and carrier recovery in two-dimensional digital communication systems". *Bell Syst. Tech. J.*, 55(3), March 1976.

[2] D.N. Godard. "Self-recovering equalization and carrier tracking in two-dimensional data communication systems". *IEEE Trans. Commun.*, COM-28(11):1867–1875, November 1980.

[3] N. Jablon. "Joint Blind Equalization, Carrier Recovery, and Timing Recovery for High-Order QAM Signal Constellations". *IEEE Trans. Signal Processing*, June 1992.

[4] Y Li and Z. Ding. "Blind channel identification based on second-order cyclostationary statistics". In *Proc. IEEE Int. Conf. Acoust., Speech, and Signal Processing*, volume IV, pages 81–84, Minneapolis, MN, April 1993.

[5] H. Liu, G. Xu, and L. Tong. "A Deterministic Approach to Blind Identification of Multichannel FIR Systems". In *Proc. 27th Asilomar Conference on Signal, Syst., & Comp.*, Asilomar, CA, Oct. 1993.

[6] H. Liu, G. Xu, and L. Tong. "A Deterministic Approach to Blind Identification of Multichannel FIR Systems". In *Proc. ICASSP'94 Conf.*, Adelaid, Australia, April 1994.

[7] E. Moulines, P. Duhamel, J.F. Cardoso, and S. Mayrargue. "Subspace-Methods for the Blind Identification of Multichannel FIR Filters". In *Proc. ICASSP'94 Conf.*, Adelaid, Australia, April 1994.

[8] G. Picci and G.Prati. "Blind equalisation and carrier recovery using a stop-and-go decision directed algorithm". *IEEE Trans. Commun.*, COM-35(9):877–887, September 1987.

[9] D.J. Reader and J. B. Scholz. "Blind maximum likelihood sequence estimation for fading channels". In *Proc. ICASSP94 Conf.*, volume 4, pages 401–403, Adelaid, Australia, April 1994.

[10] Y. Sato. "A method of self-recovering equalization for multilevel amplitude-modulation". *IEEE Trans. Commun.*, COM-23(6):679–682, June 1975.

[11] D. Slock. "Blind Fractionally-Spaced Equalization, Perfect Reconstruction Filterbanks, and Multilinear Prediction". In *Proc. ICASSP'94 Conf.*, Adelaid, Australia, April 1994.

[12] L. Tong. "Blind sequence estimation". *IEEE Trans. Communications. to appear.*

[13] L. Tong. "Blind equalization of fading channels". In *Proc. 26th Asilomar Conference on Signal, Syst., & Comp.*, Asilomar, CA, Oct. 1992.

[14] L. Tong, G. Xu, and T. Kailath. "A new approach to blind identification and equalization of multipath channels". In *25th Asilomar Conf.*, Pacific Grove, CA, Nov. 1991.

[15] L. Tong, G. Xu, and T. Kailath. "Fast blind equalization of multipath channels via antenna arrays". In *Proc. ICASSP'93*, Minneapolis, MN., April. 1993.

[16] L. Tong, G. Xu, and T. Kailath. "Blind identification and equalization based on second-order statistics: A time domain approach". *IEEE Trans. Information Theory*, 40(2), March 1994.

[17] G. Xu, L. Tong, and H. Liu. "A New Algorithm for Fast Blind Equalization of Wireless Communication Channels". In *Proc. ICASSP'94*, Adelaid, Australia, April 1994.