IMPROVED DOPPLER TRACKING AND CORRECTION FOR UNDERWATER ACOUSTIC COMMUNICATIONS

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ABSTRACT

The performance of coherent acoustic communication systems involving moving platforms (e.g., underwater vehicles and ships) is adversely effected by Doppler shift resulting from relative motion of the transmitter and receiver. This paper presents a series of innovations which, together, dramatically improve the response to Doppler shift of a widely-used adaptive receiver algorithm. The innovations include a frequency-shift estimator, time-scale interpolator and robust phase-locked loop (PLL). These techniques reduce the computational load of the coherent equalizer and provide accurate Doppler tracking. Results from at-sea testing are presented to illustrate the performance of the combined algorithm.

1. INTRODUCTION

The feasibility of bandwidth-efficient underwater acoustic communications has been recently established [1]. There is now a growing interest in the development of more sophisticated signal processing algorithms which will enable communication in adverse conditions of multipath and rapid phase variation. These development efforts aim to improve existing algorithms in a manner which will make them suitable for emerging applications. One such application is communication with an autonomous underwater vehicle (AUV), a task seriously complicated by large and variable Doppler shifts.

The adaptive decision feedback equalizer (DFE), in the kernel of many coherent communications receivers [2], is capable of decoding signals with moderate levels of Doppler shift. However, this requires the equalizer to rapidly adjust feedforward parameter phase, which is both computationally intensive and introduces adaptation noise [3]. An improved DFE algorithm containing an embedded phase-locked-loop (PLL) to remove carrier shift due to Doppler has been in use [1,4], but its performance has been unsatisfactory under realistic field conditions.

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To alleviate this problem, we investigate in this paper an alternative method for phase synchronization and equalization in the presence of large Doppler shifts. The proposed receiver algorithm performs signal detection in two steps. In the first step, the motion-induced Doppler shift is coarsely estimated and removed from the received signal. In the second step, residual Doppler shift is eliminated in the equalizer by the PLL which is modified from the original presented in [1]. We show that by performing carrier phase synchronization in a theoretically suboptimal way, the system is more robust to the choice of tracking parameters and has improved stability properties. Such features are desirable when the Doppler shift is subject to rapid variations as is the case with AUV communications.

The paper is organized as follows. In Section 2 the Doppler estimator and pre-processor is presented. In Section 3 details and performance trade-offs of the PLL within the equalizer are discussed. Finally, in section 4 the performance of the algorithm with experimental data is examined.

2. DOPPLER ESTIMATION AND COMPENSATION

Figure 1 shows the Doppler pre-processor portion of the receiver. First the frequency shift is estimated from the training (i.e., *a priori* known) data at the start of each received packet using an ambiguity function method. To minimize computation, the ambiguity function is computed across the range of Doppler shifts $\omega_{min} < \omega < \omega_{max}$ expected in a given application and at a resolution somewhat finer than the Doppler spread. The estimate is found by computing

$$\hat{\omega}_d = \max_{\omega_{min} < \omega < \omega_{max}} \sum_{k=0}^{n-1} e^{j\omega k} d_k x_k \qquad (1)$$

where x is the received data sequence and d is the length n training sequence. The performance of the estimator is proportional to both n and the signal to noise ratio. The Cramer-Rao lower bound on the variance of the estimator is given by

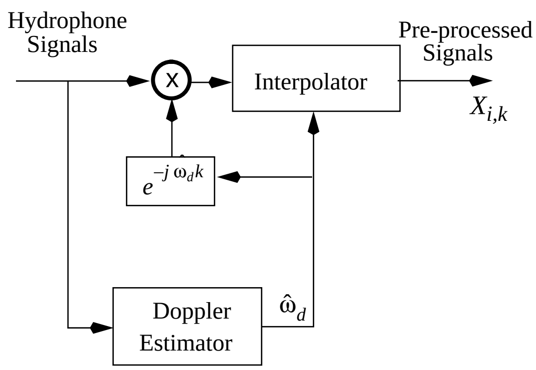

Figure 1: Doppler pre-processor which estimates frequency shift of the incoming data packet, removes that offset, then interpolates to shift the time scale of the data.

$$E\left[\left|\hat{\omega}_d - \omega_d\right|^2\right] \geq \frac{1}{\eta\tau^2} \qquad (2)$$

where τ is the observation time and η is the output SNR after matched-filtering [7]. The sequence length n is large (several hundred symbols) and data rates are modest (1000-5000 symbols per second), thus yielding very accurate Doppler estimates. Fig. 3 shows results of using the estimator during at-sea trials and demonstrates the variability of Doppler under realistic field conditions.

Using the coarse Doppler estimate, the centre frequency of each received signal is shifted and the time-scale is then adjusted using an interpolator. As the change in sampling-rate is typically very small (<1%), a polyphase interpolator provides an efficient implementation. In this method, the received signal is processed by a sequence of filters with increasing (or decreasing, depending on the sign of the Doppler shift) time delay, selected from a pre-computed filter bank.

The computational load of the interpolator is proportional to the filter length and may be considerable if low spectral distortion is required over the full input frequency range. In coherent acoustic communications, a fractional-spaced equalizer is used to provide accurate timing alignment and so the input sampling rate to the equalizer is usually twice that indicated by the Nyquist criterion [2]. This implies that the interpolator need only perform well in the half band from 0 to 1/4 of the input sampling rate. The interpolator filter bank can then be generated using a least-squares technique to minimize spectral distortion in the band of interest. The resulting interpolator offers satisfactory performance with extremely short filters. For example, with length-3 filters, the signal-to-distortion ratio is >20dB over the half band (a level of performance sufficient for QPSK signals), and with length-5 filters, 40dB is attained.

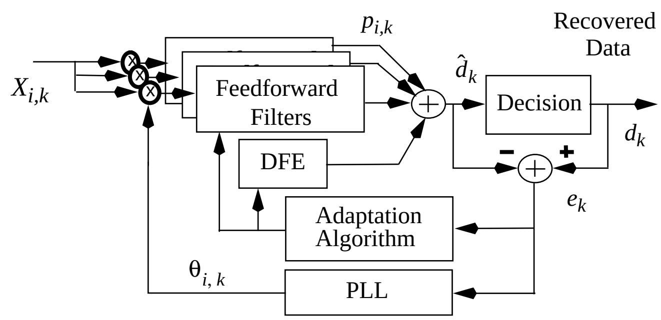

Figure 2: DFE equalizer with fine-scale phase shift correction in advance of the individual channel feedforward filters. Feedforward and feedback outputs are combined and the error term is generated by a decision device. The RLS algorithm is used for adaptation.

3. PHASE-LOCKED LOOP

In phase-coherent underwater communications, an adaptive equalizer is essential to overcome signal distortion due to multipath propagation and fading. In the equalizer structure of Fig. 2, the adaptive feedforward sections combine the Doppler-corrected input signals so as to maximize the signal-to-noise ratio of the direct arriving sound and remove inter-symbol-interference (ISI) due to dispersion. The DFE section is tasked with removing ISI due to multipath propagation. The feedforward and DFE sections are jointly adapted so as to minimize the mean-square-error (MSE) in the output symbol sequence. As noted above, the equalizer is itself able to correct for moderate levels of Doppler shift by continuously varying the phase of the feedforward filters. However this is a computationally expensive approach as the equalizer is forced to update at a much higher rate than would otherwise be necessary to

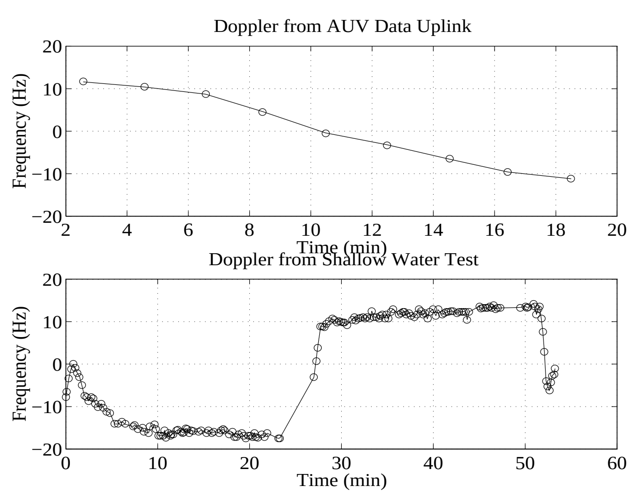

Figure 3: Doppler estimates obtained using the method of equation (1) with actual field data. Top: Doppler shift experienced when a AUV passes a stationary vessel. Bottom: Doppler shift over time for a shallow water experiment with two moving vessels.

track changes in the acoustic environment. A technique overcoming this limitation, proposed in [1], is to embed a multi-channel PLL in the equalizer designed to correct phase shifts in the output of the feedforward channels due to Doppler shift. The PLL is adjusted jointly with the equalizer using a gradient estimate of the form

$$g_{i,k} = \mathrm{Im}\{p_{i,k}(e_k + p_{i,k})^*\} \tag{3}$$

where i and k are the channel and time index, respectively, $p_{i,k}$ is the output of the ith feedforward section and e_k is the symbol error. The gradient estimate is combined with a second order filter to form a servo control loop.

Although shown to work in some situations, the method of [1] suffers from slow oscillatory convergence and, in certain cases, instability. This is a result of positive error feedback in the equalizer adaptation whenever phase correction term has a negative real part (e.g., between $\pi/2$ and π). Positive feedback is a well known cause of divergence in RLS adaptations [5]. The problem is eliminated by reversing the order of the phase correction and feedforward terms as shown in Fig. 2. Such re-ordering involves a slow-convergence assumption on the equalizer and PLL which appears to be fulfilled in all practical situations.

Unfortunately, a new problem is introduced by the reordering: the loop gain of the PLL is now dependant on the instantaneous value of the feedforward coefficients. This results in unpredictable PLL convergence rate and difficulty in setting the PLL tracking gain parameter. In order to overcome this problem, a modified phase gradient is proposed which is sensitive only to the phase of the product in (3)

$$\mathrm{g}_{i,k} = \mathrm{Im}\{\log(p_{i,k}(e_k + p_{i,k})^*)\}\,. \tag{4}$$

Use of this gradient, although theoretically sub-optimal, has been found to give very similar converged MSE performance to (3). However, the transient response of the PLL using (4) is more predictable and can be set independently of the equalizer coefficients.

The resulting equalizer equations for m input channels are given below. Fractional spacing is accommodated by assuming that r new observations are made on each channel at each time step k.

Step 1: update regressor vector with new observations $X_{i,k}$ using PLL output $\theta_{i,k}$

$$\Psi_{i,k} = \begin{bmatrix} e^{-j\theta_{i,k}} X_{i,k} \\ \Psi_{i,k-1}(1 \dots n_i - r) \end{bmatrix} \quad \text{for} \quad i = 1 \dots m \tag{5}$$

where n_i is the length of the ith feedforward section.
Step 2: compute feedforward section output using coefficient vector $A_{i,k}$

$$p_{i,k} = A_{i,k}^H \; \Psi_{i,k} \tag{6}$$

Step 3: compute feedback section output using the coefficient vector B_k applied over the vector of previous outputs D_{k-1}

$$t_k = B_k^H \; D_{k-1} \tag{7}$$

Step 4: compute symbol prediction $\hat{d}_k$ and error e_k

$$\hat{d}_k = t_k + \sum_{i=1}^{m} p_{i,k} \tag{8}$$

$$e_k = d_k - \hat{d}_k \tag{9}$$

where d_k is the nearest valid symbol to $\hat{d}_k$.
Step 5: update phase-locked loop

$$\theta_{i,k} = G\left(q^{-1}\right)\left(\mathrm{Im}\{\log(p_{i,k}(e_k + p_{i,k})^*)\}\right) \tag{10}$$

using a second order filter with gain g

$$G\left(q^{-1}\right) = \frac{g\left(1.1 - q^{-1}\right)}{1 - 2q^{-1} + q^{-2}}\,. \tag{11}$$

Step 6: update equalizer coefficient matrix $\mathbf{R}$

$$\mathbf{R}_{k+1} = \mathbf{R}_k + \lambda \begin{bmatrix} \Psi_{1,k} \\ \dots \\ \Psi_{m,k} \\ D_{k-1} \end{bmatrix} \begin{bmatrix} \Psi_{1,k}^H & \dots & D_{k-1}^H \end{bmatrix} \tag{12}$$

$$\begin{bmatrix} A_{1,k+1} \\ \dots \\ A_{m,k+1} \\ B_{k+1} \end{bmatrix} = \begin{bmatrix} A_{1,k} \\ \dots \\ A_{m,k} \\ B_k \end{bmatrix} + \mathbf{R}_{k+1}^{-1} \begin{bmatrix} \Psi_{1,k} \\ \dots \\ \Psi_{m,k} \\ D_{k-1} \end{bmatrix} \tag{13}$$

where $\lambda < 1$ is the forgetting factor. Note that (12) and (13) are performed implicitly using a square-root RLS algorithm [2].

The two key user-selected parameters in the equalizer are the PLL gain, g, and the RLS forgetting factor, λ, both of which control the responsiveness of the equalizer to changing conditions. There is an essential redundancy in these parameters due to the fact that both the RLS-adaptation and the PLL can track phase changes. This redundancy gives rise to an ill-conditioned transient response in the equalizer whereby residual Doppler shift is compensated by a combination of adaptation and PLL action. As we are concerned only with the quality of the symbol estimate rather than the instantaneous values of the equalizer parameters, this ambiguity is not a problem except for the higher-than-necessary computation load it can cause. In practice, choosing the PLL tracking rate to be significantly higher than the equalizer tracking rate (parameterized by $1 - \lambda$, eliminates much of the competition between the

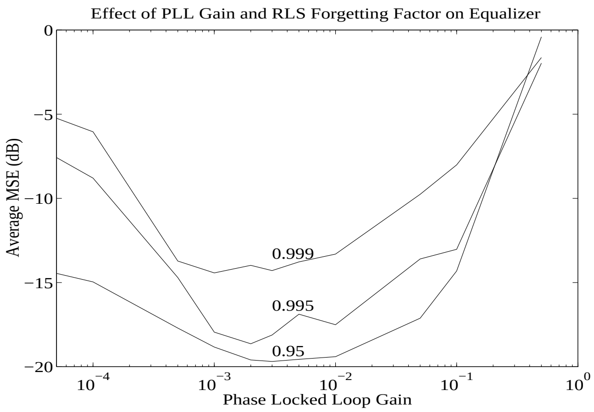

Figure 4: Performance of the equalizer with varying PLL gain and RLS forgetting factor, λ. The time constant of equalizer memory is $1/1-\lambda$ symbols. This in-water data set was Doppler compensated but has a time-varying Doppler shift.

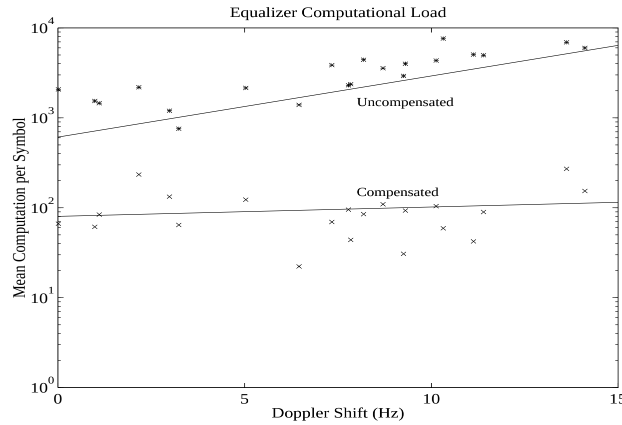

Figure 5: Comparison of computation burden while processing a variety of data packets over a range of Dopper shifts (the first 10 minutes from Fig. 3). The carrier is 12.5 kHz and the data rate is 1250 symbols/sec.

two systems. The use of MSE thresholding in the RLS update [6] also forces the PLL to dominate in tracking Doppler shift.

4. EXPERIMENTAL RESULTS

The performance of the new receiver in tracking and correcting Doppler shift has been evaluated with data from moving platforms at relative speeds of up to 6 knots. Evaluating the system with this data provides information on setting key parameters for the equalizer.

In Fig. 4 MSE as a function of the PLL and equalizer tracking parameters is shown. This is for a single QPSK packet at 1250 symbols/sec. Over a large range of RLS forgetting factors almost a decade of PLL gain adjustment is possible with little ill effect. However, as gain is decreased, the forgetting factor must be reduced to allow the equalizer to compensate phase not removed by the slowly-responding PLL. At high gain, overall equalizer performance suffers as phase noise introduced by the over-energetic PLL is added to each output symbol.

Fig. 5 shows the difference in the computation requirement per symbol needed to successfully decode a set of 21 consecutive packets spanning 0 to 15 Hz both with and without Doppler pre-compensation. The savings in computations comes about through two mechanisms. The first is that the number of feedforward parameters, which are required to act as interpolating filters, is reduced. The second savings arises because the equalizer is only updated when the MSE exceeds a certain threshold [6]. With preprocessing, fewer updates are needed to track the residual phase and time-scales changes in the signal. The computational requirement for an RLS update is $O(n^2)$, where n is the total number of parameters, and thus any reduction in parameter count or update rate is significant for real-time operation.

CONCLUSION

Doppler pre-processing is made possible by a sequence of known symbols in the data which allow accurate estimation of the carrier frequency shift. Interpolation via polyphase filters provides a computationally efficient method for adjusting the time scale of the data and this pre-processing removes the requirement for long feedforward sections which would otherwise be needed as interpolators. The total savings is very large, a factor of 50 in RLS update cost, for Doppler shifts greater than 0.1 percent. Additionally, the overall performance of the equalizer is expanded because parameterization and tracking concentrate on removing ISI, not frequency offsets and time shifts.

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