ON THE USE OF BLIND CHANNEL EQUALIZATION IN THE HF COMMUNICATION

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ABSTRACT

Blind equalization algorithms have been so far addressed in stationary environments. In reality, communication channels are non-stationary and the performance of blind algorithms on these channels are yet to be discovered. In this study, it has been shown experimentally that close performance to the Minimum Mean Square Estimation (MMSE) based equalization can be achieved by using blind algorithms. Hence, psuedo-MMSE equalization can be achieved without training in a class of time-varying channels. Representative high frequency (HF) channels which are defined in related NATO STANAG 4285 have been used for simulation purposes. The robustness tests of algorithms to frequency and time spread has been conducted. It is seen that the performance of blind algorithms in terms of the Symbol Error Rate (SER) are slightly worse than the MMSE algorithm. Since there is no supervised training for blind algorithms, bandwidth utilization efficiency is very large which compensates for the slight decrease in SER performance.

1. INTRODUCTION

Blind equalization has been the focus of extensive research effort because the need to transmit a training sequence is alleviated, so that the equalizer can remove the effect of InterSymbol Interference (ISI) solely by using its own output. Therefore, the bandwidth utilization efficiency is increased and the adaptation of the equalizer is performed continuously, which is in sharp contrast with conventional equalizers. In this context, blind equalizers are useful tools either as standalone or support units to conventional systems. Hence, a thorough investigation and evaluation of blind algorithms are required. However, almost null effort has been spent concerning this fact, mainly due to proliferating technical problems of blind equalizers, such as existence of undesired local solutions which do not remove sufficient ISI. Recently, improved blind algorithms have been proposed using a deterministic cost criterion [1] which are improvements over Constant Modulus Algorithm (CMA)-like algorithms in stationary environments. However, the performance of both type of algorithms in non-stationary environments are widely unknown. Soft Constraint Satisfaction (SCS) algorithms have many desirable features including the removal of some undesired local solutions [1] which has sparked the interest to evaluate blind algorithms in time-varying (tracking) applications. In this context, HF channel models defined in NATO STANAG 4285 [2] have been adopted and the tracking performance of the algorithms are evaluated.

HF channels are nonstationary due to the time-varying conditions in the ionosphere. There are various methods to predict the HF channel parameters, especially the noise level [3]. Unfortunately however, even the most recent predictions often fail to yield sufficiently accurate results. Therefore, the statistical properties of the HF noise are often assumed to be rather unpredictable and are still under investigation [4]. Typical HF communication takes place between 60-3600 bps range in severe fading conditions. The prediction of the reflection-based fading are possible only using problem specific methods including, the beam propagation and the ray tracing [3]. The distant reflections may take place after several symbol periods reducing the effectiveness of communication. 2400 bauds with PSK modulation is the most common method of communication in NATO compliant HF modems. 4800 bauds transmission can be considered as the next goal to be achieved in the design of modems for HF links. In this study, the realizability of 4800 bauds communication is addressed and the possible incorporation of blind algorithms in the equalizer design is discussed.

2. FADING SCENARIO AND EVALUATION METHODOLOGY

Consider

$$W_{k+1} = W_k + \frac{1}{\|X_k\|^2}\mu(2 - |y_k|)y_k X_k \qquad (1)$$

$$W_{k+1} = W_k + \mu(2 - |y_k|)y_k X_k \qquad (2)$$

where (1) and (2) define the SCS algorithm [1] and its CMA-like "unnormalized" counterpart, respectively. W_k, X_k are the equalizer tap and input vectors respectively; y_k is the equalizer output and μ is the step-size. Baseband equalization is addresses in a baseband communication system model as shown in Figure 1. Two ray models for frequency selective Rayleigh fading HF communication channels have been used for simulation purposes as defined in NATO STANAG 4285 [2]. Eleven tests are defined in the STANAG, two of which characterize the effect of the Doppler spread (test 6 in [2]) and multipath spread (test 7 in [2]). Both experiments are based on a two ray model for several values of delay spread and Doppler frequency. The taps of both models are independent Gaussian distributed complex random variables so that the envelope of the taps are Rayleigh distributed.

2/4/8-PSK modulation have been chosen for 1200/2400/3600 bps NATO HF modems [2] with a fixed baud rate of 2400 bauds. The 4-PSK modulation is also adopted herein. The use of a Decision Feedback Equalizer (DFE) is recommended with an oversampling factor of 2 in the 16-tap feedforward section. The recommended feedback section has 8 taps [2]. In this study, the equalizer is a 32 tap fractionally spaced blind equalizer with an oversampling factor of 2. Pulse shaping in the baseband is chosen as $\alpha = 0.25$ in compliance with the STANAG. The use of blind algorithms with DFEs are not feasible yet due to the convergence of the equalizer taps to undesired stationary points. The approach pursued herein is to evaluate only the feedforward section of the equalizer.

In blind equalization, delay of decisions must also be estimated. To find the correct delay of the decisions in the receiver with respect to the transmitter, the combined space of the channel and equalizer is computed. In the combined space, the optimum delay is found as the vector index of $\|S_k\|_\infty$, where S_k represents the combined channel and equalizer. In the ideal case, the combined space must represent an impulse function and the vector index of the unit sample is the delay of the equalization. In the non-ideal case the combined space has a maximum with non-zero elements around it. The level of these elements determine the residual ISI left after equalization. Based on this fact, the Open-Eye-Measure (OEM), an important measure to assess the performance of equalizers, is defined as

$$\mathrm{OEM}(k) \triangleq \frac{\|S_k\|_1 - \|S_k\|_\infty}{\|S_k\|_\infty} \qquad (3)$$

If $\mathrm{OEM}(k) > 0$ dB, the channel eye is said to be closed and decision errors may occur due to residual ISI. If $\mathrm{OEM}(k) < 0$ dB then the eye is open and ISI has no effect in the decision process, the decision errors are caused only by the additive noise.

Also the phase shift of the receiver output must be calculated as

$$\angle \|S_k\|_\infty \qquad (4)$$

where $\|S_k\|_\infty$ is the tap that has the maximum magnitude in the combined space. The phase of the output of the decision device is corrected according to (4) in the receiver. The output is then compared to the delayed complex transmitted symbol. If both symbols are identical the decision is correct, otherwise wrong. The ratio of the number of wrong decisions and total number of symbols yields the essential evaluation paradigm, Symbol Error Rate (SER).

Although the assumption that combined space is known to the receiver is not realistic at all, this setup is particularly useful insofar as a minimum attainable SER level is found that the algorithms can maintain. However, if the procedure to estimate the optimum delay in practice yields accurate results, the minumum SER can be attained.

3. SIMULATIONS

The simulation conditions are identical for all individual runs, i.e, the input symbols, the channel coefficients, the additive noise are identical sample-by-sample in each run. Test 1 is the characterization of the SER vs Doppler spread (0 to 8 Hz) performance. Two independent equal power fading paths are seperated 1 ms apart (multipath spread). Test 2 is the characterization of the SER vs multipath spread (0 to 8 ms) for two independent equal power fading paths. Each path has 0.5 Hz Doppler spread. Both experiments are performed with no additive noise with a carrier frequency of 20 MHz.

Cold-start initialization has been used, i.e., the center taps of the filters in subchannels of the equalizer are set to unity. As a benchmark the conventional NLMS algorithm has also been evaluated. The training of the NLMS algorithm has been performed continuously and the a priori output of the equalizer has been compared to the transmitted signal as outlined in the previous section to find the SER. The conventional NLMS algorithm can be regarded as the non-blind (active) version version of the SCS algorithm. If the nonlinearity $\psi(.)$ in (1) is replaced with the training sequence d, the NLMS algorithm is obtained.

To determine the optimal step-size approximately due to computational constraints, the following procedure has been applied: The equalizer performance in terms of SER has been observed over a dense grid of step-sizes for relatively short simulation lengths. A single optimal step-size is needed no a priori information about the enviroment is available. Thus, the one that leads to best SER in most of the cases has been chosen for the full simulation.

A safety constant, 0.05, is added to the denominator in (1) to prevent the steps from being too large, which may jeopardize the stability due to low powered input vectors.

In terms of the SER for several values of Doppler frequency, the NLMS algorithm has the best performance due to continuous training, which is apparently not practical at all. When the training is performed over discrete intervals, the SER performance would considerably fall, particularly for fast fading channels. Since the performance of the SCS algorithms are close to the NLMS algorithm it can be said that the training is not absolutely necessary, the equalizer can cope with the channel by using its own output. The "unnormalized" algorithm defined in (2) do not show satisfactory performance.

However, for several values of delay spread, SCS has better performance. This may be due to the fact that the NLMS algorithm could be sensitive to the equalizer order.

Trade-offs exist in this type of application, it is not possible to say one algorithm performs better than the other. Consequently, additional training sequence overhead in the NLMS algorithm can be considered as redundant.

It may be argued that when additive channel noise is at a significant level, the performance of blind algorithms would deteriorate considerably because it is known that the additive channel noise destroys the optimality of F-spaced equalizers [5]. There are multiple optimal solutions in the noise-free case when the length-and-zero condition is satisfied and all solutions remove the ISI completely. When the noise is added, some solutions tend to be affected more and the residual ISI for those solutions is higher than the least affected ones. In a time-varying channel the locations and the "types" (in the previously defined sense) of these solutions change continuously and the equalizer has to reconverge after singularities in the channel. Hence, without loss of generality we may assume that convergence of equalizer taps to any solution is equally likely.

4. CONCLUSIONS

The theme of this paper has been to show the viability of SCS type blind algorithms under realistic time-varying fading channels. It has been demonstrated that the use of a training signal is not absolutely necessary, almost redundant, for acceptable MMSE equalization, which is a rather important result in the sense that the available bandwidth is not wasted by transmitting known symbols. The profound advantage is the continuous operability which helps track changes in the channel. This is not possible in conventional approaches due to intermittent training and equalization periods. The future directions of research should focus on the SCS algorithms and their performance under realistic conditions and equalizer structures.

5. REFERENCES

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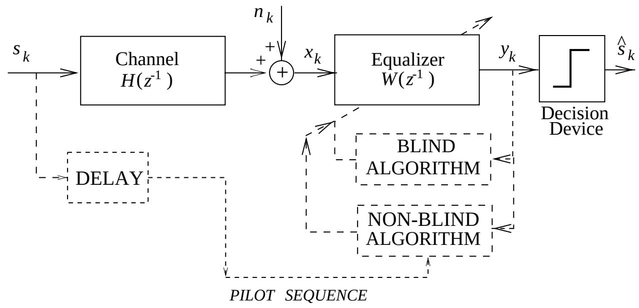

Figure 1: Baseband model of a communication system.

Doppler Fr. (Hz)	4800 bauds	
	NLMS	SCS
0.1	0.2e-2	0.52e-2
1	0.36e-2	0.92e-2
2	0.45e-2	0.12e-1
4	0.6e-2	0.12e-1
6	0.86e-2	0.14e-1

Table 1: Evaluation results for several values of Doppler frequency with two equal powered channel taps 1 ms apart.

Multipath Sp. (ms)	4800 bauds	
	NLMS	SCS
0	0	0
0.5	0.14e-2	0.72e-2
1	0.22e-1	0.86e-2
2	0.21e-1	0.94e-2
4	0.21e-1	0.94e-2
6	0.22e-1	0.96e-2

Table 2: Evaluation results for several values of delay spread with two equal powered channel taps of 0.5 Hz Doppler frequency.