# “Adaptive Noise Cancellation in a Wireline Telemetry Receiver.”

A Literature Survey

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March 10, 1997

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# 1. Introduction

One of the many challenges that we face in wireline telemetry is how to operate high-speed data transmissions over non-ideal, poorly controlled media. The key to any telemetry system design depends on the system's ability to adapt to changing environment. While adaptive equalization can account for frequency-dependent cable attenuation by inverting the channel distortion effect, there still exists the need to reduce other sources of noise. When data and power cable are intertwined, for example, crosstalk noise may exist on the data cable. This literature survey will explore the basis of adaptive noise cancellation in reducing the crosstalk interference. This paper will also outline the next steps in the specification of a cancellation algorithm using a homogeneous synchronous dataflow (HSDF) graph and implementation on an embedded DSP processor.

# 2. Periodic In-Band Noise from Crosstalk

The source of noise is crosstalk interference. The noise can be described as a collection of high frequency pulses superimposed on top of a slow varying 60 Hz sine wave.

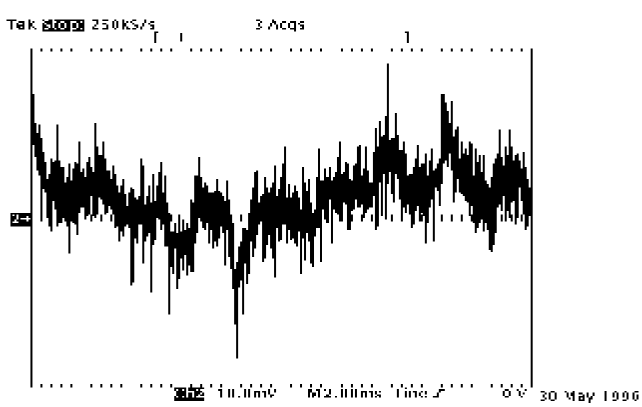
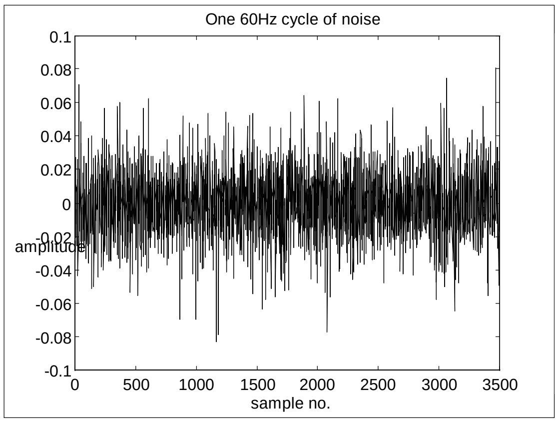



**Figure 1. Oscilloscope capture of crosstalk interference.**

However, the 60 Hz component is removed by an analog high pass filter in the receiver, thus, the noise, that we are truly interested in, is the periodic repetition of the high frequency pulses at a period of 1/60 seconds.

**Figure 2. One period of cyclic noise.**



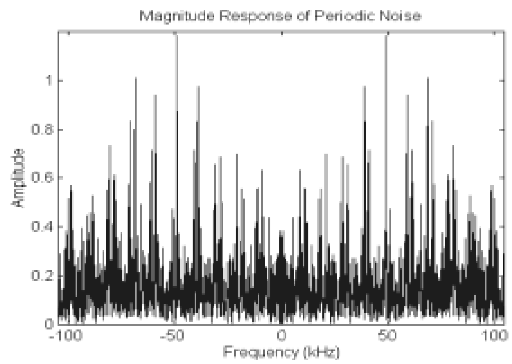




**Figure 3. FFT of the periodic noise, $f_s$=210 kHz.**

Since the noise consists of impulses, it maps to a wide-band interference in the frequency domain. Therefore, the noise affects not only the initial operation of receiver synchronization, but also the steady-state data transmission which uses QAM with a symbol rate of 70 kHz.

## 3. Quadrature Amplitude Modulation

Quadrature Amplitude Modulation avoids the spectral inefficiency of Double Sideband Amplitude Modulation by mapping a stream of bits onto a constellation and modulating the coordinates of the constellation with two orthogonal carriers $90^o$ apart in phase [Samueli 94]. Thus, the transmitted signal is

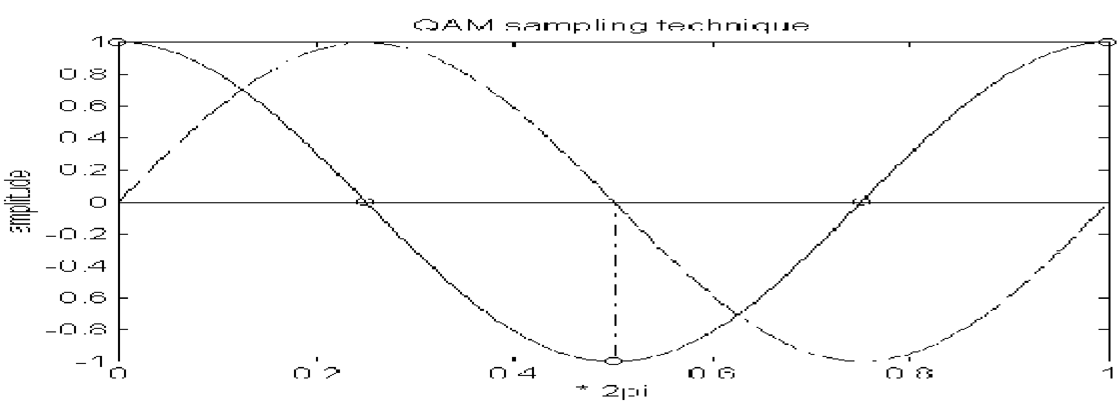
$$s(t) = x_p(t)\cos(\omega_c t) - x_q(t)\sin(\omega_c t)$$

At the receiver end, s(t) is multiplied by $\cos(\omega_c t)$ and $-\sin(\omega_c t)$ to recover the original data, the products are

$$y_p(t) = x_p(t)\cos^2\omega_c t - x_q(t)\sin\omega_c t\cos\omega_c t = \frac{x_p(t)(1+\cos 2\omega_c t) - x_q(t)\sin 2\omega_c t}{2}$$

$$y_q(t) = -x_p(t)\cos\omega_c t\sin\omega_c t + x_q(t)\sin^2\omega_c t = \frac{x_p(t)\sin 2\omega_c t + x_q(t)(1-\cos 2\omega_c t)}{2}$$

The sidebands of the second harmonics of the carriers are then removed by low-pass filtering, and the receiver baseband signals $y_p(t)$ and $y_q(t)$ are then within a factor 2 to the originals [Bingham 88]. However, the process described above is too computationally intensive to be used in the current telemetry system; there is simply not enough processing power to convolve a long $\frac{\sin x}{x}$ filter while operating at the 210 kHz sampling rate. Instead, Schlumberger developed and patented a technique to eliminate the need for signal reconstruction [Montgomery 93]. The two QAM carriers are $90^0$ out of phase; when one carrier is at zero, the other is at its peak. Schlumberger's technique separates the carriers by sampling the signal 4 times a period, at the zero crossings of either carrier. Since the symbol rate is at 70 kHz, every 3 samples will represent one symbol with either 1 or 2 samples representing the X-coordinate and either 1 or 2 samples representing the Y-coordinate.




Intensive convolution (lowpass filtering) is, therefore, avoided using this technique.

# 4. Existing Noise Cancellation Techniques

One possible interference cancellation technique can be found in [Solo 95]. A primary signal $s_k$ is contaminated with noise $n_k$ and recorded in the presence of another noise $m_{ok}$ as $y_k$. A measurement of $n_k$ is available through a noisy channel of transfer function $H(q^{-1})$ as $x_k$. A cleaned signal $e_k$ is obtained by subtracting $\hat{s}_k$ from $y_k$ as seen in

**Figure 4. Interference Cancellation.**

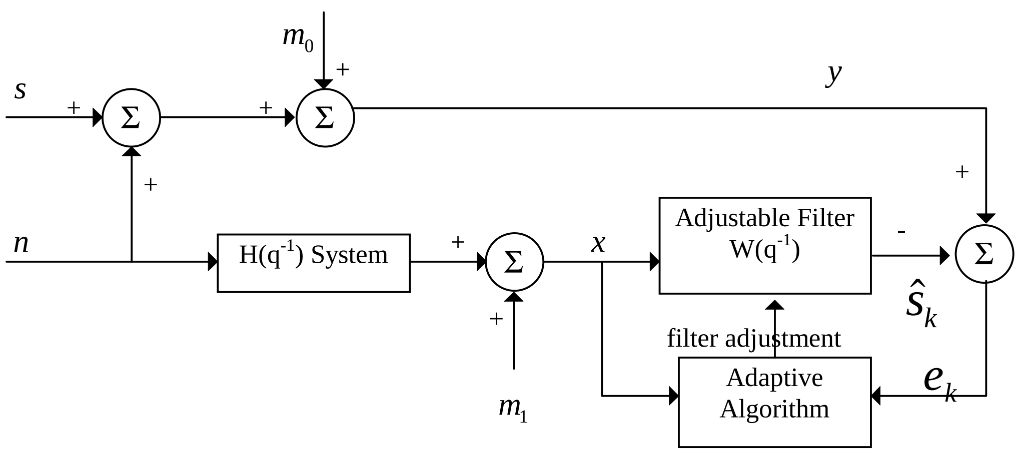



Figure 4. From this figure we know that

$$e_k = y_k - \bar{s}_k = y_k - W(q^{-1})x_k = s_k + n_k + m_{0k} - W(q^{-1})(m_{1k} + H(q^{-1})n_k)$$
$$e_k = s_k + n_k - \tilde{n}_k + m_{0k}$$

where $\tilde{n}_k = W(q^{-1})(H(q^{-1})n_k + m_{1k})$. Also $n_k - \tilde{n}_k$ is the noise cancelling error. From the above equations we can define the Mean Square Error (MSE) as

$$\varepsilon = E(e_k^2) = E(s_k^2) + E(n_k - \tilde{n}_k)^2 + E(m_{0k}^2).$$

Thus, minimizing MSE with respect to $W(q^{-1})$ is equivalent to minimizing the noise cancelling error power $E(n_k - \tilde{n}_k)^2$. To accomplish this task, we must come up with a method that will allow us to approximate our noise in an accurate manner. We, therefore, propose a noise cancellation technique that will acquire the noise characteristics or $\tilde{n}_k$ during the transmitter and receiver synchronization (startup) phase. The reason is that, for example, during startup in a telemetry system that performs well logging for seismic surveys, the downhole tool transmits only the carrier wave (a sine tone generated at 52.5 kHz) and a good noise approximation can be obtained by implementing a sharp cut-off band-pass filter centered at $f_c = 52.5kHz$ and subtracting the filter output from the received signal.

Other adaptive noise cancellation techniques also exist [Widrow 75], however, they focus more on the cancellation of a single tone rather than on a signal that contains multiple frequency components (which is the case that we encounter.)

## 5. Our cancellation technique

As described above, an accurate noise approximation can be obtained by implementing the lower portion of Figure 4 using a bandpass filter to extract the noise. Adaptation is then introduced to combat possible carrier drift. To accomplish this, an external zero-crossing reference is introduced from the power supply; the output of the filter is then stored in a buffer using the equation

$$buffer_i(n) = \frac{\alpha}{\beta} buffer_i(n-1) + (1 - \frac{\alpha}{\beta})\, filter\_output_i \,.$$

where $\frac{\alpha}{\beta}$ can be assigned a value of , for example, $\frac{31}{32}$ to reduce the effect of impulses.

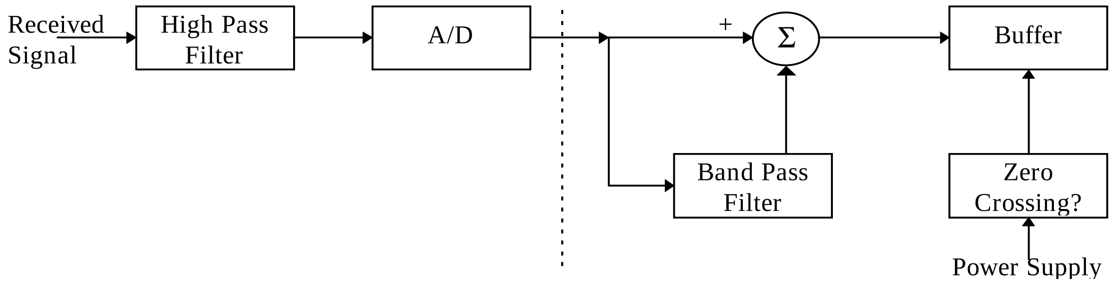



**Figure 5. Noise acquisition during startup.**

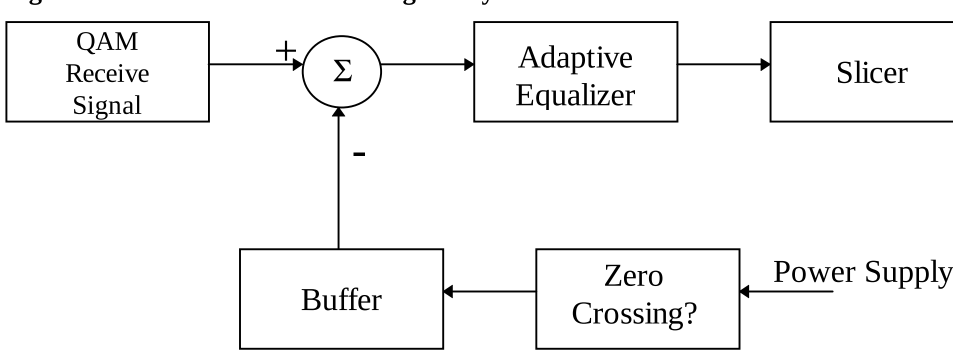
The variable $i$ will be reset to 0 at every alternate zero crossings. The band-pass filter is implemented by a second-order IIR biquad with the transfer function

$$\frac{Y(z)}{X(z)} = \frac{k_0(1-z^{-2})}{1-k_1 z^{-1} - k_2 z^{-2}}$$

where the coefficients $k_0, k_1, and\ k_2$ can be adjusted to obtain the desired cutoff and the desired filter build-up duration (currently $k_0 = 0.1$, $k_1 = 0$, and $k_2 = -0.8$.)

During steady state operation, QAM transmission begins, and we implement the upper portion of Figure 4. Sample-by-sample subtraction of the received signal and the buffered noise is performed. The buffer adapts to the carrier drift using the power

**Figure 6. Noise Cancellation during steady state.**




supply's zero crossing information. The buffer values are also re-updated during short training. The adaptive equalizer in Figure 6 is implemented using the LMS algorithm

[Campbell 96]. Each of the equalizer's filter coefficients is updated by the addition of a weighted error term

$$h_i(n) = h_{i-1}(n-1) + 2\mu\, y(n-1)e(n)$$

where $e(n)$ is the error term and $\mu$ is the step size. The adaptive equalizer is already implemented in the telemetry system and can be incorporated into the project with a few modifications. The slicer is a decision device that maps an X and Y coordinate into a valid symbol.

## 6. Homogeneous Synchronous Dataflow

Given an SDF graph G, if for each edge in G the number of tokens produced and consumed on the edge is one, then we say that G is a homogeneous SDF (HSDF) graph. HSDF fits nicely with our implementation of the noise cancellation algorithm. The procedure consists of two graphs, $G_1$, noise acquisition, and $G_2$, noise cancellation. In $G_1$, the firing sequence will be {input, band-pass, adder, buffer, zero-cross} according to Figure 5, with each actor consuming one token and producing one token. The buffer will be updated using post-modify addressing method. The index of the buffer is reset to zero with each zero-crossing trigger.

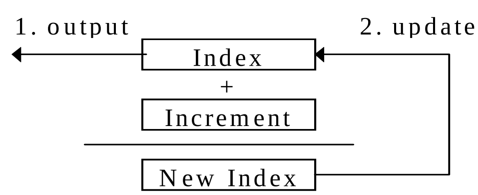



**Figure 7. Post-Modify with index update.**

In G2 the firing sequence will be {Receiver input, buffer, zero-cross, adder, adaptive equalizer, and slicer} according to Figure 6.

## 7. Future Directions

In this literature survey, we have defined our problem, explored how others have approached similar problems, mapped our proposed solution onto the HSDF domain, and laid the foundation for realization on a DSP chip. The next step involves writing SHARC code for each of the blocks defined in the above figures. The goal is to use the SHARC simulator to implement the HSDF noise cancellation algorithm using real noise data measured off of an oscilloscope. The simulation results should provide an answer to the effectiveness of the proposed cancellation algorithm and its practicality.

## 8. References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