Novel Adaptive B-Spline Filter for Hybrid Echo Cancellation

Ratnakar Dash[1], Rutuparna Panda[2], Bibhudendra Acharya[3] and Banshidhar Majhi[4]
[1,4]Department of Computer Science & Engineering, NIT Rourkela, India
[2]Department of Electronics & Telecommunication Engineering, OIT, Burla, India
[3]Department of Electronics & Communication Engineering, NIT Rourkela, India
ratnakar.dash@ gmail.com[1], r_pp2002@yahoo.com[2], bibhudendra@gmail.com[3], bmajhi@nitrkl.ac.in[4]

***Abstract*—In this paper direct design of Generalized B-spline filters has been presented. It has been shown how to choose filter parameters to meet the desired specifications. These filters exhibit maximally flat magnitude response and low side lobe energy as compared to Kaiser Window filter. Also a new type of adaptive B-spline filter has been developed in this paper. The application of the proposed filters for hybrid echo cancellation on digital cellular has been extensively studied. Results reveal the suitability of the proposed filters for such applications. This filter minimizes the mean square error using proportionate adaptive LMS algorithm. The proposed adaptive filter makes a perfect estimate of the hybrid echo.**

Keywords-Adaptive B-spline filter, Hybrid Echo, Kaiser Window, LMS algorithm

I. INTRODUCTION

With the advent of digital cellular technology, there is a strong need to achieve superior signal quality. Cellular service providers are searching for new technologies to improve the speech quality. The hybrid echo produced in the transmission path degrades the voice quality. Multipath propagation of the transmitted signal also produces distortion in the signal. This motivates to suggest some new techniques to minimize the effects of hybrid echo. Echo is the repetition of a waveform due to reflection from points in the medium through which the wave propagates. Echo is useful in various applications like sonar and radar for detection and exploration purposes. In telecommunication, echo degrades the quality of service, i.e. the speech quality. So echo cancellation is an important aspect of communication systems. The development of echo cancellation began in the late 1950s, and continues today as new integrated landline and wireless cellular and mobile networks put additional constraint on the performance of echo cancellers [1].

There are two types of echoes that can exist in communication systems: acoustic echo and telephone line hybrid echo. Acoustic echo results from a feedback path set up between the speaker and the microphone in a mobile phone, hands-free phone, teleconference or hearing aid system.

Telephone line echoes result from an impedance mismatch at telephone exchange hybrids where the subscriber's 2-wire line is connected to a 4-wire line [2]. The perceptual effects of an echo depend on the time delay between the incident and reflected waves, the strength of the reflected waves, and the number of paths through which the waves are reflected. Telephone line echoes, and acoustic feedback echoes in teleconference and hearing aid systems, are undesirable and annoying and can be disruptive.

In this paper, it is shown how to develop the generalized B-spline [3] FIR filter and its use in the adaptive filter design for effective cancellation of echo. Also the idea of proportionate adaptation [4] and its particular solution to the network echo cancellation problem is presented.

The organization of the paper is as follows. Following the introduction, brief note on B-spline smoothing is outlined in section II. Section III discusses about the design of generalized B-spline filter. In section IV, Adaptive Echo Cancellation is presented. Results are discussed in section V. Finally, section VI describes the concluding remarks.

II. B-SPLINE SMOOTHING

In 1999, M.rorak and M.A.Escabi developed trapezoidal FIR filters by frequency convolution of the ideal prototype with a smoothing rectangular function [3]. In case of direct design of FIR filters B-splines have been used to replace the sharp transition edges of the magnitude response. In B-spline design of maximally flat filters, the convolving function $\psi(\omega)$ is obtained by repeated convolution of the rectangular function. Here we have used a smoothing exponential function rather than a rectangular as starting point. Then the convolving function $\psi(\omega)$ can be generated by ρ^{th} order convolution of exponential function with the same function [5]. The smoothing is better served by the order of the magnitude response description. Smoothing of the transition region may be continued indefinitely by the convolution process. The filter response is obtained by convoluting the ideal response with the convolving function obtained by repeated convolution.

III. DESIGN OF GENERALIZED B-SPLINE FILTER

Direct or practical design technique exhibits a transition band in the response i.e. the smoothing of the sharp transition. The trapezoidal filters were developed by direct design approach for FIR filters. In the direct approach, a B-spline has been developed by convolving rectangular pulse of width

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$2\alpha\omega_c / \rho$ and height $\pi / \alpha\omega_c$ [5]. Further smoothing in the transition band was obtained by convolving two rectangular functions of half the width and twice the height. The ρ^{th} order polynomial can be generated by repeating the convolution process to replace the sharp transition edges.

$$H(\omega) = H_I(\omega) * \phi(\omega) * \phi(\omega) * ....\phi(\omega) \quad (1)$$

$$= H_I(\omega) * \psi(\omega)$$

$H_I(\omega)$ represents the ideal response and $\psi(\omega)$ is the convolving function.

Here a generalized exponential B-spline has been developed by taking an exponential function rather than a rectangular function as shown in Fig. 1.

In this case, a set of generalized B-spline functions have been developed from the basic principle that the generating function for such generalized B-spline of n^{th} degree has $(n-1)$ continuous derivatives and satisfies an $(n-1)$ th order linear differential equation. We can define a generalized B-spline of degree n and type r as follows.

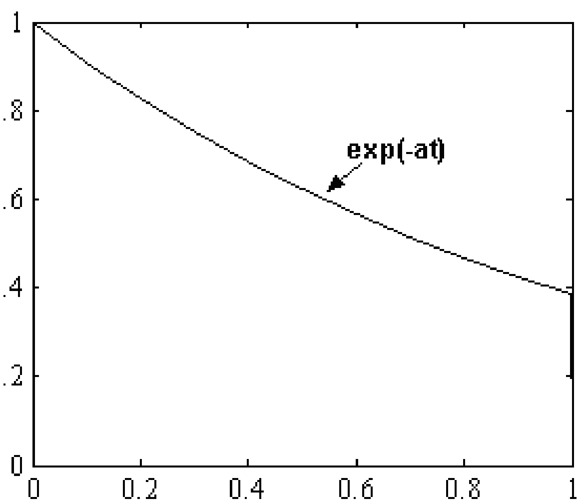

Figure 1. Generalized B-spline with an exponential function

$$B^{nr}(x) = L\sum_{j=0}^{n+1} W_j g^{nr}(x - x_j)U(x - x_j) \quad (2)$$

Where L is the constant for normalized B-spline, $U(x)$ is the unit step function, g^{nr} is the generating function for the n^{th} order spline and w_j is the constant of multiplication in the j^{th} segment of the B-spline. r denotes the type. There can be number of generalized B-spline functions of any degree. Therefore, the r different types of generalized B-splines for a particular degree are presented in terms of type $'r'$. The polynomial B-spline $B^{n1}(x)$ is one of these generalized splines. In (2), following relations can be obtained

$$L = n + 1 \quad (3)$$

$$W_k = \frac{1}{\prod_{\substack{j=0 \\ j\neq k}}^{n+1}(x - x_j)} \quad (4)$$

$$g(x) = x^n \quad (5)$$

To satisfy the continuity of the $(n-1)^{th}$ derivatives $g(x)$ must satisfy the following conditions:

$$g(x) = |g(-x)| \quad (6)$$

For (6) to satisfy, $G(s)$ must also be symmetrical about the imaginary axis. Thus, for generalized B-spline the poles of $G(s)$ have four quadrant symmetry.

$$B^{nr}(x) = B^{pr}(x) * B^{qr}(x) \quad (7)$$

$n = p + q + 1$ $\quad p, q \geq 0$ and * denotes the convolution.

The generating function of a generalized B-spline of any degree can be computed from zero degree generalized B-splines. Based on the properties provided by [5], different types of generalized B-splines can be generated. The noble feature of such generalized B-spline is that they can provide compact support. The number of generalized B-splines for higher degree more than 3 goes on increasing rapidly with n.

The generating function for zero degree B-spline as stated in the previous paragraph has been taken as i.e. $g^{01}(x) = \exp(-a * x)$, 'a' is the scaling parameter. Thus the concept of generalized B-spline differs from the concept of the B-spline design used by the authors previously [3]. In their case, the generating function is a rectangular function. On the other hand, here we have introduced another parameter 'a' called scaling parameter. This parameter will help us to approximate filter bandwidth efficiently.

The generalized B-spline of zero degree can be written as

$$B^{01}(x) = \frac{a}{(1 - e^{-a})}\left[e^{-ax}U(x) - e^{-a}e^{-a(x-1)}U(x-1)\right] \quad (8)$$

$a = 0$, this is same as zero order B-spline i.e. a centered rectangular pulse.

The first degree generalized B-spline can be generated by recognizing the fact that the poles of G(s) for first degree have four quadrant symmetry. They are

$$s^2 = 0$$
$$s = \pm a,$$
$$s = \pm ja$$

Similarly number of splines can be generated. Any, one of them can be taken as the window function for designing the generalized B-spline filter of any order. As the degree of the generalized B-spline functions increases, the frequency response closes to the desired response which will be shown. In this approach of filter design we have used the third degree generalized B-spline of type three.

The convolving function for the generalized B-spline can be expressed as

$$B(\omega) = L\rho \frac{\pi}{\alpha\omega_c} \sum_{k=0}^{\rho} w_k g\left(\frac{\rho}{2}\left(\frac{\omega}{\alpha\omega_c}+1\right)-k\right) \tag{9}$$

$B(\omega)$ is a function defined over the interval $-\alpha\omega_c$ to $\alpha\omega_c$ and for other values it is zero. The convolving function for the $H(\omega)$ in (1) obtained by repeated convolution process described is shown in Fig. 2. Note that $L = \dfrac{a}{(1-e^{-a})}$.

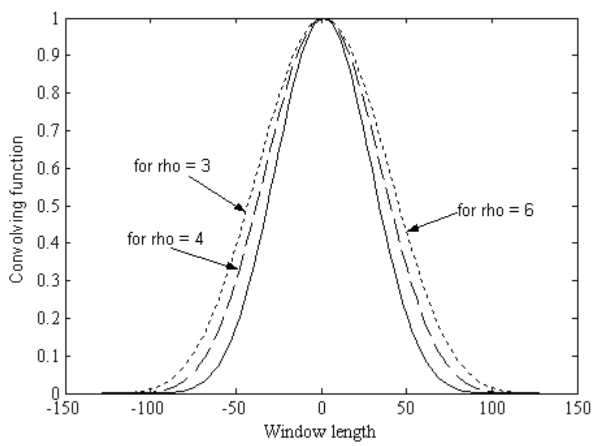

Figure 2. Convolving function for the generalized B-spline filter

The lowpass filter response obtained by applying the operation $H(\omega)*B(\omega)$ can be expressed as

$$H_I(\omega) = 1 - \frac{1}{\rho!}\sum_{k=0}^{\rho}(-1)^k \binom{\rho}{k}\left[\frac{\rho}{2}\left(\frac{|\omega|-\omega_c}{\alpha\omega_c}+1\right)-k\right] \tag{10}$$

The discrete time impulse response can be obtained by taking inverse fourier transform of (10) $H(\omega)$. Therefore impulse response for the low pass filter obtained as

$$h[n] = \frac{\omega_c}{\pi}\frac{\sin(n\omega_c)}{n\omega_c}\left(\frac{a^2}{1-\cos a}\right)\frac{(\cos(n/M+1)-\cos a)}{a^2-(n/M+1)}(\mathrm{sinc})^{\rho} \tag{11}$$

where $\sin c = \dfrac{\sin(pi*x)}{(pi*x)}$

These filters remain unity throughout the passband and zero throughout the stopband. So they are maximally flat. The ideal filter applied before to $B(\omega)$ may be any filter which is required to avoid sharp transition. Here a lowpass filter was developed. The impulse response for the ideal highpass, bandpass, and bandstop filter can be developed by multiplying the ideal response with the window function.

A. Windows of MF filters

A window function for the filter developed here can be recognized from (11) as follows.

$$W(n) = \left(\frac{a^2}{1-\cos a}\right)\frac{(\cos(n/M+1)-\cos a)}{a^2-(n/M+1)}(sinc)^{\rho} \tag{12}$$

The (12) contains the parameters a, ρ and M. These parameters were used to control the different specifications of the filter. The Kaiser window and the generalized B-spline window are shown in Figs. 3, 4 and 5 below for the same window length and order.

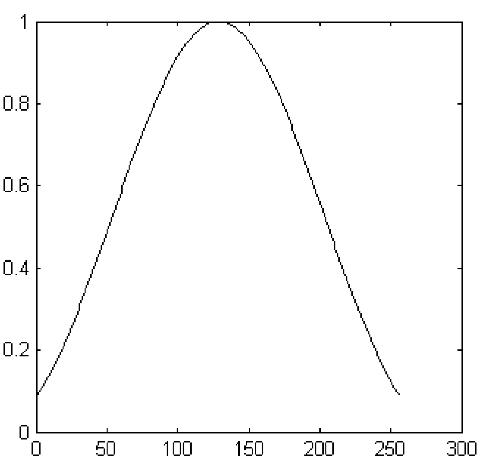

Figure 3. Kaiser window with β = 4

The Generalized B-spline window shows the properties of sinc function .The window function is flat at the order ρ-1 which was previously known. [5]. The window function was truncated smoothly in order to limit the energy of the convolving function $B(\omega)$ that in turn minimizes the passband and stopband error. It was heuristically argued by the previous authors for optimal value of $\alpha = 1$. It was proposed that for principally flat filter the relation between the window parameters is

$$\alpha = \frac{\rho}{(\omega_c / \pi)(M+1)} \tag{13}$$

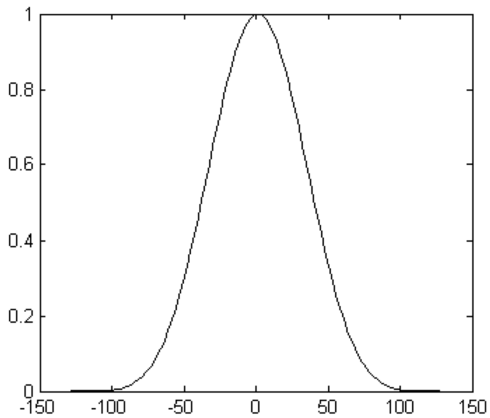

Figure 4. Generalized B-spline window with ρ = 4

The window function for the generalized filter discussed here little bit differs from the B-spline principally flat window function. There is an extra parameter 'a' is introduced with a constant function.

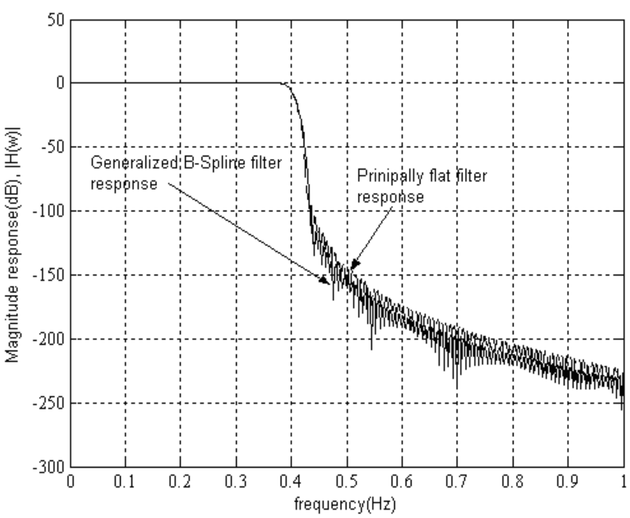

Figure 5. Comparison of filter responses

B. Design parameters

Error in the pass band is given by

$$E_p^2(\omega) = \frac{1}{\omega_1} \int_0^{\omega_1} \left[1 - \hat{H}(\omega)\right]^2 W_A(\omega) d\omega \tag{14}$$

Here ω_1 is the accumulation interval. Usually the accumulation interval is taken as the passband frequency ω_p. $W_A(\omega)$ is the weighting function taken as unity. Another measure for passband error is the mean rms error frequency ω_m given [5] by the (15)

$$\omega_m = \frac{\int_0^{\omega_1} \omega E_p(\omega) d\omega}{\int_0^{\omega_1} E_p(\omega) d\omega} \quad \text{for} \quad 0 < \omega_1 \le \omega_p \tag{15}$$

Where ω_p Gives the centroid frequency error in the passband frequencies $0 < \omega_1 \le \omega_p$.

Impulse response functions provided by (11) were used to calculate the frequency responses of the filter by varying the filter parameters over the ranges 0< ρ <15, 0.001< α <0.8, 4<= M <=250, and 0.05 π <= ω_c <=0.5 π . Stopband attenuation and transition width and pass band error were measured for each filter. It was seen that variations in the filter specifications from the actual value were negligible. The frequency responses for different values of filter parameter were plotted. The comparison between the frequency responses was also seen and it was clearly observed that the Generalized B-spline filter gives an improved version of principally flat filter developed by previous authors. The frequency response plot is shown in the Fig. 5.

IV. Adaptive Echo Cancellation

The echo canceller can be an infinite impulse response (IIR) or a finite impulse response (FIR) filter. The main advantage of an IIR filter is that a long-delay echo can be synthesized by a relatively small number of filter coefficients. In practice, echo cancellers are based on FIR filters. This is mainly due to the practical difficulties associated with the adaptation and stable operation of adaptive IIR filters [11]. The Fig. 6 shows one side of a typical hybrid echo canceller.

Assuming that the signal on the line from speaker B to speaker A , $y_B(n)$ is composed of the speech of speaker B, $x_B(k)$, plus the echo of speaker A we have

$$y_B(n) = x_B(n) + x_A^{echo}(n) \tag{16}$$

In practice, speech and echo signals are not simultaneously present on a phone line. This, as pointed out shortly, can be used to simplify the adaptation process.

Assuming that the echo synthesizer is an FIR filter, the filter output estimate of the echo signal can be expressed as

$$\hat{x}_A^{echo}(n) = \sum_{k=0}^{L-1} w_k(n) x_A(n-k) \tag{17}$$

Where $w_k(n)$ are times varying coefficients of an adaptive FIR filter and $x_A^{echo}(n)$ is an estimate of the echo of speaker A on the line from speaker B to speaker A. The residual echo signal, or the error signal, after echo subtraction is given by

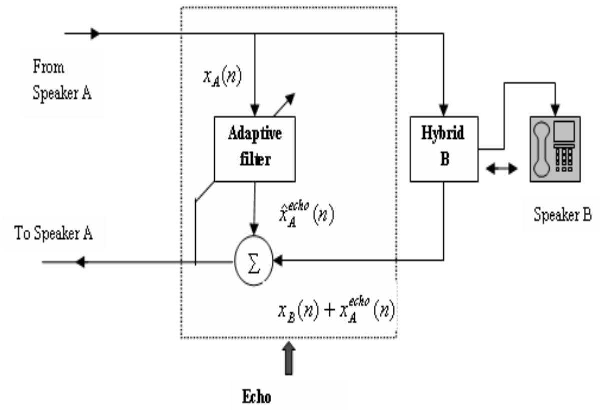

Figure 6. Block Diagram of Echo cancellation using Adaptive filter

$$\begin{aligned} e(n) &= y_B(n) - \hat{x}_A^{echo}(n) \\ &= x_B(n) + x_A^{echo}(n) - \sum_{k=0}^{L-1} w_k(n) x_A(n-k) \end{aligned} \tag{18}$$

For those time instants when speaker A is talking, and speaker B is listening and silent, and only echo is present from line B to A, we have

$$\begin{aligned} e(n) &= x_A^{echo}(n) - \hat{x}_A^{echo}(n) \\ &= x_A^{echo}(n) - \sum_{k=0}^{L-1} w_k(n) x_A(n-k) \end{aligned} \tag{19}$$

where $e(n)$ is the error signal or residual echo.

A.. PNLMS Algorithm

The coefficient updation of the PNLMS [6, 7] (Proportionate Normalized LMS) algorithm can be formulated by

$$w_t = w_{t-1} + \frac{\mu G_t}{X_t^T G_t X_t + \alpha} X_t e_t \equiv w_{t-1} + \mu_t X_t e_t \tag{20}$$

Where $e_t = y_t - X_t^T h_{t-1}$, w and X are N dimensional filter weight and input vectors respectively. y is the desired output scalar $G_t = diag\{g_{t,1}\sqrt{g_t}......, g_{t,N}\sqrt{g_t}\}$ is a diagonal matrix with the elements $g_{t,n}$ defined as

$$g_{t,n} = \max\{\rho \max\{\delta, |w_{t-1,1}|,....|w_{t-1,N}|\}, |w_{t-1,n}|\}, \tag{21}$$

and $\overline{g_t} = \frac{1}{N}\sum_{n=1}^{N} g_{t,n.}$. The parameter μ is a step size or learning rate which is a scalar quantity. ρ, δ, and α are small positive parameters known as regularization factor. When G is an identity matrix i.e. $G = I$, (proportionate LMS reduces to normalized LMS algorithm).

V. Results and Discussions

Fig.7 shows the error convergence using the Kaiser window used to design the FIR filter which in turn uses the adaptive algorithm to make it an adaptive filter. Error convergence curves for different fractions of the signal are shown. The error performance curves of and adaptive B-spline filter is shown in Fig. 8. All the graphs were shown below using proportionate normalized LMS algorithm. This has been implemented in MATLAB 6.5.

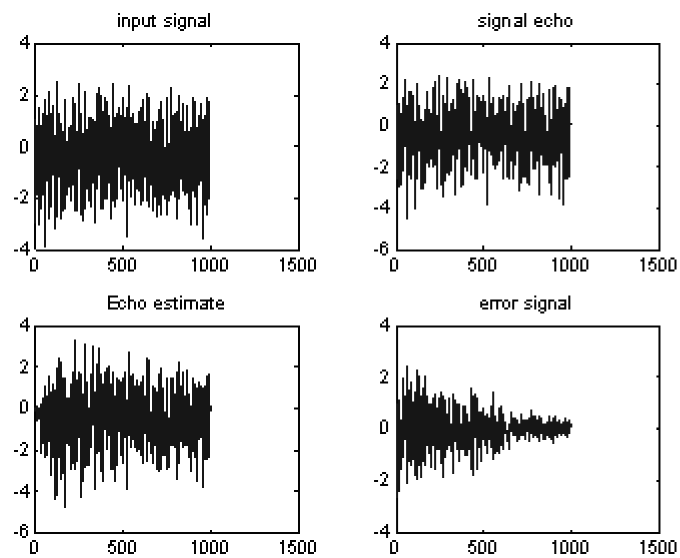

Figure 7. Echo estimate and error convergence curve using Kaiser Window adaptive filter: 30% of the signal returned back to the transmitter.

VI. Conclusion

The proposed filter has shown a significant improvement over the filters using Kaiser Window. There are two more advanced adaptation techniques that employ stochastic principles to minimize the probability of error named maximum a posteriori probability and maximum likelihood sequence estimation. Several adaptation schemes and alternate filter structures offer better performance than those methods discussed before in the earlier sections. Usually this performance improvement comes at the cost of increased complexity.

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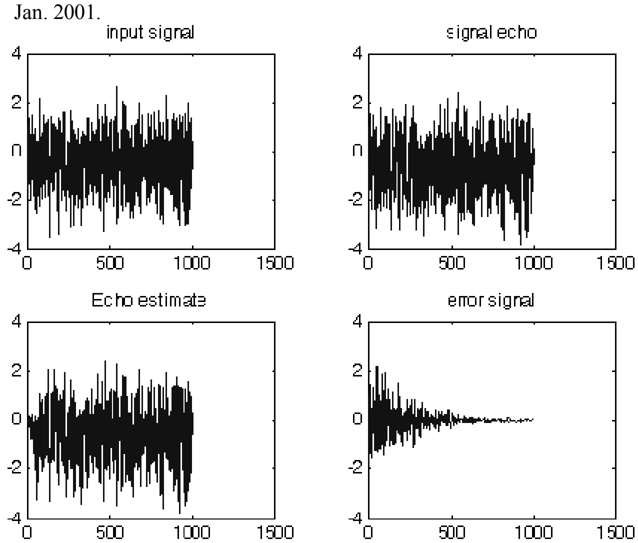

Figure 8. Echo estimate and error convergence curve using adaptive B-spline filter: 30% of the signal returned back to the transmitter.

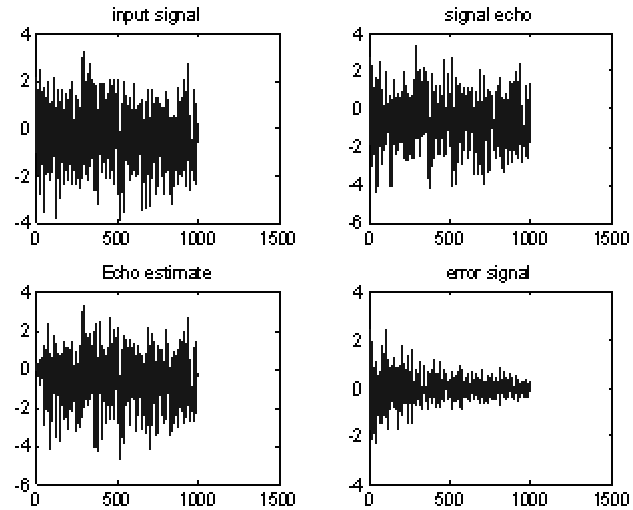

Figure 9. Echo estimate and error convergence curve using Kaiser window adaptive filter 40% of the signal returned back to the transmitter.

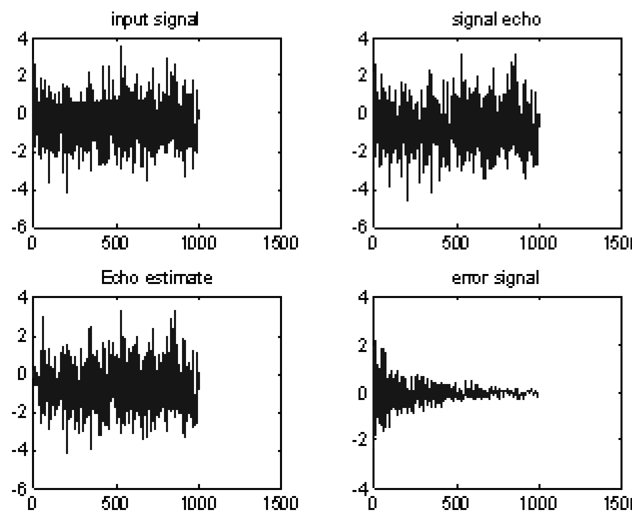

Figure 10. Echo estimate and error convergence curve using adaptive B-spline filter: 40% of the signal returned back to the transmitter