Deconvolution Approach to Carrier and Code Multipath Error Elimination in High Precision GPS

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ABSTRACT

This paper proposes a novel technique based on the deconvolution approach for the simultaneous estimation and compensation of the multipath estimation errors in both the carrier phase lock and the code delay lock loops in GPS receivers. Simulation results are presented showing that the proposed GPS receiver algorithm achieves high precision in the range and differential range estimation in various GPS precision applications such as spacecraft attitude control, C\ 'C'J) in a relatively severe multipath environment. The proposed architecture and algorithm inherently involve tradeoffs among the hardware/software implementation complexity, the extent of the multipath expected in the specific application, and the degree of multipath cancellation

INTRODUCTION

The superiority of GPS receiver technology over other competing technologies, in terms of the cost, reliability, mass, size, and power considerations, as proven in its various applications such as navigation, spacecraft orbit determination, and surveying, has constantly prompted the extension of GPS to other important areas that require increasingly more exacting performance from the GPS system One such area, for example, is the application of GPS to the attitude determination of aircraft and earth orbiting satellites.

For the GPS receivers to provide the requisite precision, the existing error sources in the GPS system must be eliminated or greatly reduced. At the current state of GPS technology, the most significant error source is the signal multipath propagation. For example, in spacecraft operational environments, the GPS signal is reflected from various structural components of the spacecraft and these reflected components are received by the GPS along with the desired direct line-of-sight (LOS) path. The reflected signals differ from the desired LOS path signal in terms of their delays, amplitudes and phases. The carrier phase tracking loop provides no inherent discrimination against the multipath signals and thus tracks the phase of the composite signal corrupted by multipath components. The resulting differential carrier phase estimation error can be orders of magnitude higher compared to the case of no multipath propagation in many GPS applications. For example, the measurements obtained by the RADCAL satellite GPS-ADS (Attitude Determination System) experiments have shown that the differential range error in such environment is of the order of 1 cm corresponding to an attitude determination error of about 0.5 degree. Thus for the GPS receivers to provide precision pointing knowledge (order of 1 arcmin or better with 1 meter antenna baseline) or a differential range accuracy of about 0.3 mm or better, the multipath effects must be suppressed by orders of magnitudes Similar accuracy may also be desirable in other GPS precision applications in the presence of multipath signals such as GPS based geophysical measurements.

Among the past approaches to deal with the multipath problem, one approach involves reducing the early-late delay spacing among the correlators in the GPS receiver code lock loop. However while this approach reduces the code range errors to some extent, it does not aid in the carrier phase measurements accuracy that is the basis of most GPS precision applications, Moreover even the reduction in the code range error is limited and if the early-late spacing is smaller than the initial delay error due to multipath (easily the case with many multipath situations), then the loop error can be very high and the loop may not even track, although the probability of such an event may be small.

In a recent paper [1] based on the maximum likelihood (ML) estimation theory and the theory developed earlier in [2,3], a set of implicit equations are derived for the ML estimates of the parameters of interest, i.e. the amplitudes, phases and delays of the multipath signals. The paper proposes to solve these highly nonlinear implicit equations in a recursive manner. Reference [1] also presents some simulation results showing significant reduction in the multipath errors in the code phase measurements as compared to the delay lock loop. However, reference [1] does not present any results on the carrier phase measurements, the subject of most interest in the present paper.

This paper presents techniques for dealing with the multipath problem in a comprehensive manner. These techniques are based on the application of the optimal deconvolution approach in a somewhat unconventional manner, as compared to its application in other fields such as seismology and telecommunications. The proposed method consists of first estimating the impulse response of the effective multipath channel by a least squares algorithm. This step is followed by obtaining an inverse filter which equalizes the multipath channel response to the desired ideal multipath free response to the maximum extent possible within the specified constraints of the implementation complexity. Note that there is a possible tradeoff between the hardware/software complexity, the dynamics tracked and the extent of the multipath elimination. From the equalized response one then estimates the true carrier phase and code delay. The simulation results demonstrate that the proposed method is capable of completely eliminating the multipath distortion that is more severe than may possibly take place in any realistic precision GPS applications environment.

MULTIPATH ELIMINATION BY DECONVOLUTION APPROACH

In the more conventional telecommunication applications of the deconvolution approach or the equalizer theory [4], the multipath propagation channel is modeled as,

$$y_j = \sum_{k=-q_1}^{q_2} h_k u_{j-k} + v_j \qquad (1)$$

where $\{u_j\}$ represents the transmitted symbol sequence, $\{y_j\}$ is the channel output sequence and $\{h_{-q_1}, h_{-q_1+1}, \ldots, h_0, h_1, \ldots h_{q_2}\}$ represents the discrete channel impulse response. The additive noise sequence $\{v_j\}$ is usually assumed to be a zero-mean white Gaussian. The basic process of deconvolution involves the estimation of the transmitted input sequence $\{u_k\}$ on the basis of noisy observations $\{y_k\}$ assuming that the discrete channel impulse response $\{h_k\}$ is known to the receiver. In the case of unknown channel response, adaptive equalization techniques [5,6] are used wherein first an approximate estimate of $\{h_k\}$ is obtained on the basis of a training sequence known in advance to the receiver and the channel output $\{y_k\}$ and subsequently the estimate of $\{h_k\}$ is refined adaptively with u_k replaced by its estimated/detected version in the adaptive algorithm. From the real time estimate of $\{h_k\}$, a time-varying inverse filter is derived which then filters $\{y_k\}$ to obtain the estimate of $\{u_k\}$. In practice the two steps of estimating $\{h_k\}$ and then finding the corresponding inverse filter are combined into a single step of finding directly the adaptive equalizer coefficients. The problem wherein the adaptive equalization is achieved without any training sequence is relatively more difficult and has also received considerable attention in the literature.

There are a number of other important applications of the adaptive deconvolution approach in various other fields such as seismology [7] and antenna signal reconstruction

[8]. All these situations with their respective terminologies are modeled by equation (1) or its higher dimensional versions Adaptive algorithms are then derived following the above deconvolution approach.

In the following, the deconvolution approach is extended to the problem of multipath elimination in the GPS receiver code tracking and carrier phase lock loops. The following derivation of the signal model shows both the similarities and the differences between the GPS application of this paper and the other applications of the deconvolution approach.

In the absence of the multipath, the input signal to the GPS receiver is given by:

$$s(t) = A_c \cos[\omega_c t + a(t)\frac{\pi}{2}] \tag{2}$$

where the receiver noise is not considered in the first instance, A_C is the received signal amplitude and $a(t)=\pm 1$ is the pseudo-random code waveform that phase modulates the carrier, and it is assumed that the data modulation is removed in a decision-directed manner. in the presence of N multipaths in addition to the direct line-of-sight path, the input signal may be characterized as

$$s_m(t) = s(t) + \alpha_1 s(t-\tau_1) + \ldots + \alpha_N s(t-\tau_N) \tag{3}$$

where α_i and τ_i denote respectively the amplitude and delay of the i th multipath for i=1,2,...,N. After substituting (2) in (3), the composite received signal may be expressed as follows.

$$s_m(t) = A_c \cos(\omega_c t + a(t)\frac{\pi}{2}) + \alpha_1 A_c \cos(\omega_c t + \frac{\pi}{2} a(t-\tau_1) + \theta_{m_1})$$
$$+ \ldots + \alpha_N A_c \cos(\omega_c t + \frac{\pi}{2} a(t-\tau_N) + \theta_{m_N}); \theta_{m_i} = -\omega_c \tau_i \tag{4}$$

As in the conventional delay-lock discriminator the signal in (4) is correlated with the reference signals

$$s_L(t) = 2\cos(\omega_0 t + a(t-\tau-\tau_d)\frac{\pi}{2})$$
$$s_E(t) = 2\cos(\omega_0 t + a(t-\tau+\tau_d)\frac{\pi}{2}) \tag{5}$$

where τ is the delay tracking error, τ_d is the offset delay referred to as early and late correlator delay and ω_0 is the reference frequency. The resulting correlation functions are given by

$$\overline{s_m(t)}.\ s_I(t) = A_c a(t)\, a(t\ 7\quad \tau_d)\cos(\omega_1 t) + \ldots$$

$$+ A_c \alpha_N a(t\ \tau_N).\, a(t\quad \tau - \tau_d).\cos(\omega_1 t + \theta_{m_N}) \qquad (6)$$

where $\omega_1 = \omega_c - \omega_0$, some intermediate frequency. The signal in equation (6) is then demodulated by the reference signal $c_i(t)$ provided by the carrier phase lock loop

$$c_i(t) = 2\cos(\omega\ (\ \text{-}1\ \theta_c) \qquad (7)$$

The resulting demodulated signal denoted by $R_{1i}(t)$ may be written in the form below.

$$R_{1i}(\tau) = h_{0i} R_c(\tau\ \text{-}1\ \tau_d) + h_{1i} R_c(\tau + \tau_d - \tau_1) + \ldots + h_{Ni} R_c(\tau + \tau_d - \tau_N)$$

$$h_{ki} = \alpha_k A_c \cos(\theta_{m_k} - \theta_c);\ k = 0, 1, \ldots, N \qquad (8)$$

In equation (8) $R_c(\tau)$ represents the code autocorrelation funct ion, $\alpha_0 = 1$ and $\theta_{m_0} = 0$. Similarly the corresponding signal obtained with s_I replaced by s_i in (6) and denoted by $R_{2i}(t)$ may be written in the following form

$$R_{2i}(\tau) = h_{0i} R_c(\tau - \tau_d) + h_{1i} R_c(\tau\ \ \text{i}\ ,\ 7,\ ) + h_{Ni} R_c(\tau\, \tau_d\, \tau_N) \qquad (9)$$

The difference between R_{1i} and R_{2i}, the so called discriminator funct ion $D_i(\tau)$ is now given by

$$D_i(\tau) = h_{0i} g_c(\tau) + h_{1i} g_c(\tau - \tau_1) + \ldots + h_{Ni} g_c(\tau - 7\text{A}) \qquad (1\text{o})$$

where

$$g_c(\tau) = R_c(\tau + \tau_d)\quad R_c(\tau - \tau_d)$$

is the discriminator function in the ideal case. Clearly if $\alpha_k = 0$ for all $k \neq 0$ then D,(7) $= g_c(\tau)$ corresponding to the case of no multipath propagation. The standard delay lock loop converges to the solution τ_m of the equation $D_i(\tau) = \text{O}$ instead of converging to O which is the solution of $g_c(\tau) = 0$. The multipath error τ_m can be excessive depending upon the actual multipath environment encountered as will be illustrated by severs] simulation examples. In the approach of this paper the discriminator function is measured first and then via equation (10) is used to estimate all the unknown variables including the multipath delays, amplitudes, and phases and the estimates of the multipath errors both in the phase lock and code lock loops. The multipath errors are then compensated for in arriving at the final differential phase and code range estimates.

To accomplish the above objectives, one also obtains the quadrature phase version of equation (1 O) by replacing $c_i(t)$ in (7) by its quadrature phase version:

$$c_q(t) = -2\sin(\omega_1 t + \theta_c)$$

Repeating the steps as in equations (6) to (10) one obtains the expression for the phase quadrature version $D_q(\tau)$ of the discriminator function as

$$D_q(\tau) = h_{0q} g_c(\tau) + h_{1q} g_c(\tau - \tau_1) + \ldots + h_{Nq} g_c(\tau - \tau_N)$$

$$h_{kq} = \alpha_k A_c \sin(\theta_{m_k} - \theta_c); k = 0,1,\ldots,N \qquad (11)$$

For the purpose of estimating and filtering, the measurement equations (10) and (11) may be combined into the following equation

$$D(\tau) = h_0 g_c(\tau) + h_1 g_c(\tau - \tau_1) + \ldots + h_N g_c(\tau - \tau_N) + n(\tau) \qquad (12)$$

$$D(\tau) = D_i(\tau) + jD_q(\tau); h_k = h_{ki} + jh_{kq}; k = 1,\ldots, N \qquad (13)$$

In equation (12) above $j = \sqrt{-1}$ and $n(\tau)$ represents the noise at the correlator output corresponding to the noise at the receiver input. In order to apply the adaptive digital signal processing techniques, the multipath delays are approximated by integer multiples of Δ, where Δ may be selected to be sufficiently small to provide the required multipath resolution, leading to the following desired discrete form:

$$y_j = \sum_{k=-q_1}^{q_2} h_k \bar{g}_{j-k} + n_j; j = -M_1,\ldots,0,\ldots M_2 \qquad (14)$$

where $y_j = D(j\Delta); \bar{g}_j = g_c(j\Delta); n_j = n(j\Delta)$ and (M_1,M_2) represents the interval over which the measured discriminator function is significant. Note that equation (14) is of somewhat more general form than (12) in that this form may also include negative values of the multipath delays, Standard estimation techniques are now applied to estimate the channel impulse response coefficients $\{h_k\}$ and the corresponding equalizer filter coefficients. Letting

$$\mathbf{h}^T = [h_{-q_1}, \ldots, h_{-1}, h_0, h_1, \ldots, h_{q_2}]$$

$$\mathbf{g}_j^H = [\bar{g}_{j+q_1}, \ldots, \bar{g}_{j+1}, \bar{g}_j, \bar{g}_{j-1}, \ldots, \bar{g}_{j-q_2}]; j = -M_1,\ldots,0,\ldots,M_2$$

the least squares estimate of the unknown channel impulse response vector is given by

$$\hat{\mathbf{h}} = \Big(\sum_{j=-M_1}^{M_2} \mathbf{g}_j \mathbf{g}_j^H + \sigma_n^2 \mathbf{I}\Big)^{-1} \sum_{j=-M_1}^{M_2} \mathbf{g}_j y_j \tag{15}$$

where $\sigma_n^2 = E[\bar{n}_j^2]$.

Now denoting by χ_0 the truncated (or zero padded depending upon whether $q_i > K_i$ or vice versa ; $i = 1,2$) version of $\hat{\mathbf{h}}$, i.e., with

$$\chi_0 = [\hat{h}_{K_1}, \ldots, \hat{h}_0, \ldots, \hat{h}_{-K_2}]$$

and by χ_j the j times shifted version of χ_0,

$$\chi_j = [\hat{h}_{K_1+j}, \ldots, \hat{h}_j, \ldots, \hat{h}_{-K_2+j}]$$

the optimum parameter vector $\hat{\mathbf{f}}$ is given by

$$\hat{\mathbf{f}} = \Big(\sum_{j=-(K_1+q_1)}^{(K_2+q_2)} \chi_j \chi_j^H + \frac{\sigma_n^2}{\kappa} \mathbf{I}\Big)^{-1} \chi_0 \tag{16}$$

Note that in the definition of χ_j the entries for $\hat{h}_j$ are set equal to zero if j falls outside the interval $[-q_1, q_2]$. Also the constant κ is given by

$$\kappa = \frac{1}{L} \sum_{j=-L_1}^{L_2} \bar{g}_j^2 ; \; L = L_1 + L_2 + 1$$

The equalized channel response $\mathbf{z}$ is obtained by convolving the two sequences $\{h_{-q_1}, \ldots, h_0, \ldots, h_{q_2}\}$ and $\{\hat{f}_{-K_1}, \ldots, \hat{f}_0, \ldots, \hat{f}_{K_2}\}$. The elements of $\mathbf{z}$ are denoted by $\{z_{-(q_1+K_1)}, \ldots, z_0, \ldots, z_{(q_2+K_2)}\}$. Therefore the equalized discriminator response is given by

$$D_{eq}(\tau) = \sum_{i=-(q_1+K_1)}^{(q_2+K_2)} z_i g_c(\tau - i\Delta) \tag{17}$$

For the case of perfect equalization $z_0 = 1$, $z_i = 0$ for $i \neq 0$ and $D_{eq}(\tau) = g_c(\tau)$. In practice $\mathrm{Re}\{D_{eq}(\tau)\} \approx g_c(\tau)$ and the solution of the equation

$$\mathrm{Re}\{D_{eq}(\tau)\} = 0 \tag{18}$$

is the estimate of the multipath error in delay estimation which can be compensated for in the range estimate.

Delay Estimation :

One may note in the development above that the solution to the estimation problem is not yet complete, as the measurement equation (12) has not taken into account the initial delay uncertainty τ_p arising as a result of the unknown propagation delay. To take into account this uncertainty, the equation (12) is now modified as

$$D(\tau) = h_0 g_c(\tau - \tau_p) + h_1 g_c(\tau - \tau_1 - \tau_p) + \ldots + h_N g_c(\tau - \tau_N - \tau_p) + n(\tau) \quad (19)$$

In order to simultaneously estimate the channel response vector h and the delay uncertainty τ_p, let

$$\tau_p = k_0 \Delta + \tau_\varepsilon ; |\tau_\varepsilon| < \Delta$$

for some signed integer k_0 and with Δ selected above. The equation (19) may be rewritten as

$$\begin{aligned} D(\tau) &= h_0 g_{k_0}(\tau) + h_1 g_{k_0}(\tau - \tau_1) + \ldots + h_N g_{k_0}(\tau - \tau_N) + n(\tau) \\ g_{k_0}(\tau) &= g(\tau - k_0\Delta), \; g(\tau) = g_c(\tau - \tau_\varepsilon) \end{aligned} \quad (20)$$

Assuming that $|k_0| < N$ for some integer N and that $g(\tau) \approx g_c(\tau)$ (true for Δ small), then k_0 is the integer k that minimizes the following index

$$\min_{-N \le k \le N} \left\| D(\tau) - (\hat{\mathbf{h}}^k)^T \mathbf{g}^k(\tau) \right\| \quad (21)$$

where $\hat{\mathbf{h}}^k$ is the channel response obtained on the basis of replacing k_0 by k in (20),

$$\mathbf{g}^k(\tau) = [g_k(\tau), g_k(\tau - \tau_1), \ldots, g_k(\tau - \tau_N)]^T$$

and $g_k(\tau)$ is approximated by $g_c(\tau - k\Delta)$. In the estimation of $\hat{\mathbf{h}}^k$ for any integer k, the discretization and estimation procedure of (14)-(16) is followed. If $\hat{k}_0$ is the solution of the minimization in (21), then the overall channel estimate is given by $\hat{\mathbf{h}} = \hat{\mathbf{h}}^{\hat{k}_0}$. Substitution of $\hat{\mathbf{h}}$ in (16) provides the coefficients of the inverse filter and the solution of

(18) yields $\hat{\tau}_p$ Note that if it is known that the multipath error τ_p is smaller than A in magnitude, then this additional procedure is not necessary, However for $k_0 > 0$, considerable error can otherwise ensue in the estimate of h.

Multipath Phase Estimation :
With $\hat{\tau}_p$ denoting the estimate of the multipath delay error, one now solves equation for **h** but with $\bar{g}_j$ equal to $g_c(j\Delta\hat{\tau},)$. Denoting the resulting least squares solution by $\hat{\mathbf{h}}^f$, then the multipath phase error estimate is given by the argument of the zeroth component of the estimated impulse response, i e , $\hat{\theta}_p = -\arg(\hat{h}_0^f)$. This error is compensated from the carrier phase-lock loop phase estimate in the following estimation cycle.

Figure 1 depicts the block diagram of the proposed GPS receiver implementation. Note that the number of correlators $Q = Q_1 + Q_2 + 1$, depends upon the value of A, the spread oft he expected multipath, and the deconvolution filter order selected, The linear combiner simply takes the appropriate differences of the correlators' outputs according to

$$D(\tau + i\Delta) = R(\tau + i\Delta + \tau_d) - R(\tau + i\Delta - \tau_d); i = -Q_1, \ldots, 0, \ldots, Q_2$$

where τ_d is selected to be integer multiple of A, The channel es[irI-l:{tc)r/equalizer block computes the multipath phase and delay errors according to equations (1 5) - (21). These error estimates then appropriately compensate the code and carrier generator phase.

SIMULATION RESULTS

In this section some simulation examples are presented depicting the performance of the proposed multipath cancellation algorithm First some additional notations are introduced to present these results. Let T and T_c denote the sampling period for the signal processing system and the code chip period respectively Also let τ d denote the normalized delay τ_d/T_c and y be the sample signal-to-noise power ratio given by $\gamma = (P_c T / N_0)$ where $P_c = A_c^2 / 2$ is the signal power received by the direct line-of-si{r,l]t path and N_0 is the one-sided power spectral density of the receiver thermal noise. Further denote by a and θ_m the vectors consisting of the amplitudes and phases of the discrete channel response h in equation (14). The motivation] behind the research presented in this paper has been the GPS application to precision attitude determination of LEO (low earth orbit) spacecrafts. Therefore the achievable carrier phase estimation errors are interpreted in terms of the equivalent attitude pointing errors It may be easily seen that for 1 meter antenna baseline the conversion factor between the carrier phase error in radians and the corresponding pointing error in arcmin is approximately equal to 109.4. Several simulation examples are presented below in terms of the notations introduced in this Section Recall that $(q_1 + q_2)$ is equal to the number of discrete multipaths, (K_1+K_2+1) is the number of filter taps and y is the sample SNR.

Example 1. $q_1=0$; $q_2=8$; $K_1=12$; $K_2=1$; $\gamma=10^3$; $\tau_d=0.5$

α=[1 .2 .5.8.9 .7 1 2.9]

θ_m=[0 -.5 .71.:' ,8.6 1.3.2-1.1]

Figure 2 plots the ideal discriminator response $g_c(\tau)$, distorted response $D(\tau)$ of (12) and the discrete version of the equalized response $D_{eq}(\tau)$ of (17). Clearly while the zero-crossing of $D(\tau)$ is about .4 T_c the zero-crossings of both $g_c(\tau)$ and $D_{eq}(\tau)$ are equal to zero. Reducing the tap spacing to τ_d=0.1 does not make any significant difference in the zero-crossing of the function $D(\tau)$ as shown in Figure 3 which plots the various discriminator functions for the case of τ_d = (~.]. Figure 4 shows the convergence of the delay error as a function of the number of samples processed with an initial normalized delay of -.498 chips. Notice that the steady state error without multipath correction will remain approximately equal to .4 T_c while with the correction algorithm the error is approximately equal to ,001 T_c. Note also that with a 50 dB-Hz CNR at the receiver input this period of 100 iterations corresponds to only 1 sec of real time as per the definition of y. Figure 5 *shows* the carrier phase error expressed in arcmin for the GPS attitude determination application. Note that the initial error of more than 30 arcmin is reduced to about .1 arcmin.

Example 2. $q_1=0$; $q_2=25$; $K_1=35$; $K_2=4$; $\gamma=10^4$; $\tau_d=0.1$

α = *[1 ,8 ,5.3.90.2.5.3 .7 .1.050 .(S .1 ,03 ,02.1 07 05 .04,1 0.05 .08 .01]*

θ_m =[0 .5 .7 -.71 .3 .5-.61.5 1 .82 1 01 - .5 - ,3 ,6.1 .s -.8 -,5,91.2 -.61]

Figures 6 and 7 depict respectively the convergence *of* carrier phase and the code delay errors versus the number of samples In this example the sample SNR is equal to 40dB and thus corresponds to an update period of 0.1 sec for a 50dB-Hz CNR at the GPS receiver input. Note that the multipaths with delay greater than 9Δ are weak in that their relative amplitudes are less than 0.1 of the direct path which makes the identification of these multipaths more difficult. As may be inferred from these figures a residual pointing error of less than 1 arcmin and a normalized delay error of .005 chips is obtained in less than 50 iterations. Since the amplitudes of multipaths for delays exceeding 9Δ are relatively small, one may consider ignoring these and thus select a filter order that is much smaller than q. Figure 8 shows the resulting error performance when the filter order K_1 is selected to be 10. As is apparent from the figure, the residual error while much smaller compared to the initial error of about 30 arcmin is much higher than that achieved with K_1=35 filter case.

Example 3. $q_1=0$; $q_2=35$; $K_1=35$; $K_2=4$; $\gamma = 10^3$; $\bar{\tau}_d = 0.1$

α= [1 ,8 .5.3.90.2,5.3.7,1 .5.6.2 01 .7.5.2.3,5.71.4.8.3.71.7.8.6 1 .5 .30 .7]

θ_m=[0 .5.7-.7 1 .3. S-,61,5 1.82 .1 01 -,:1- .3. 6. 1. 5. ,8-. 5.91.2-.6 1 .7-1 -.7.8.6.5-.81.4-.91

Figures 9 and 10 plot the residual carrier phase and code delay errors respectively, for [γ]= 30 dB. As may be inferred from these figures, for this case of very severe multipath propagation, the pointing error is reduced from about .32 arcmin to only 4.5 arcmin while the residual delay error is about .01 chips.

CONCLUSIONS

This paper has presented novel techniques based on the deconvolution approach for the simultaneous estimation and compensation of tile multipath estimation errors in both the carrier phase and code delay based precision GPS applications. Of particular interest is the application of the algorithm to GPS based attitude determination of LEO (low earth orbiting) satellites. From the simulations of the algorithm it turned out that for a one meter antenna baseline, an attitude error of ..5 deg to 1 deg may result due to multipath errors. This result is consistent with the earlier RADCAL satellite experiment. In the code pseudo range measurements errors of the order of 34 chips were observed in the simulations. From the simulations it also became apparent that while reducing the early late correlators delay spacing reduces the effects of thermal noise, it does little to mitigate the effects of multipath propagation.

When the proposed algorithm is applied to these simulation examples, the attitude determination errors are reduced to order of an arcmin in most cases. in some simulation examples a residual error of only .1 arcmin can be achieved while for very severe multipath situation involving 35 multipaths a residual error of about 4 arcmin is achieved. In terms of code pseudo range errors the error is reduced to order of ,01 chips. As expected, generally the higher the deconvolution filter order the smaller are the residual errors. However beyond a certain limit the increase in filter order or the SNR does not reduce the errors any further ,i.e. , the residual errors remain of the order of one arcmin as described above in various simulation examples. With the order of errors thus achieved the proposed algorithm should make GPS based attitude determination a reality.

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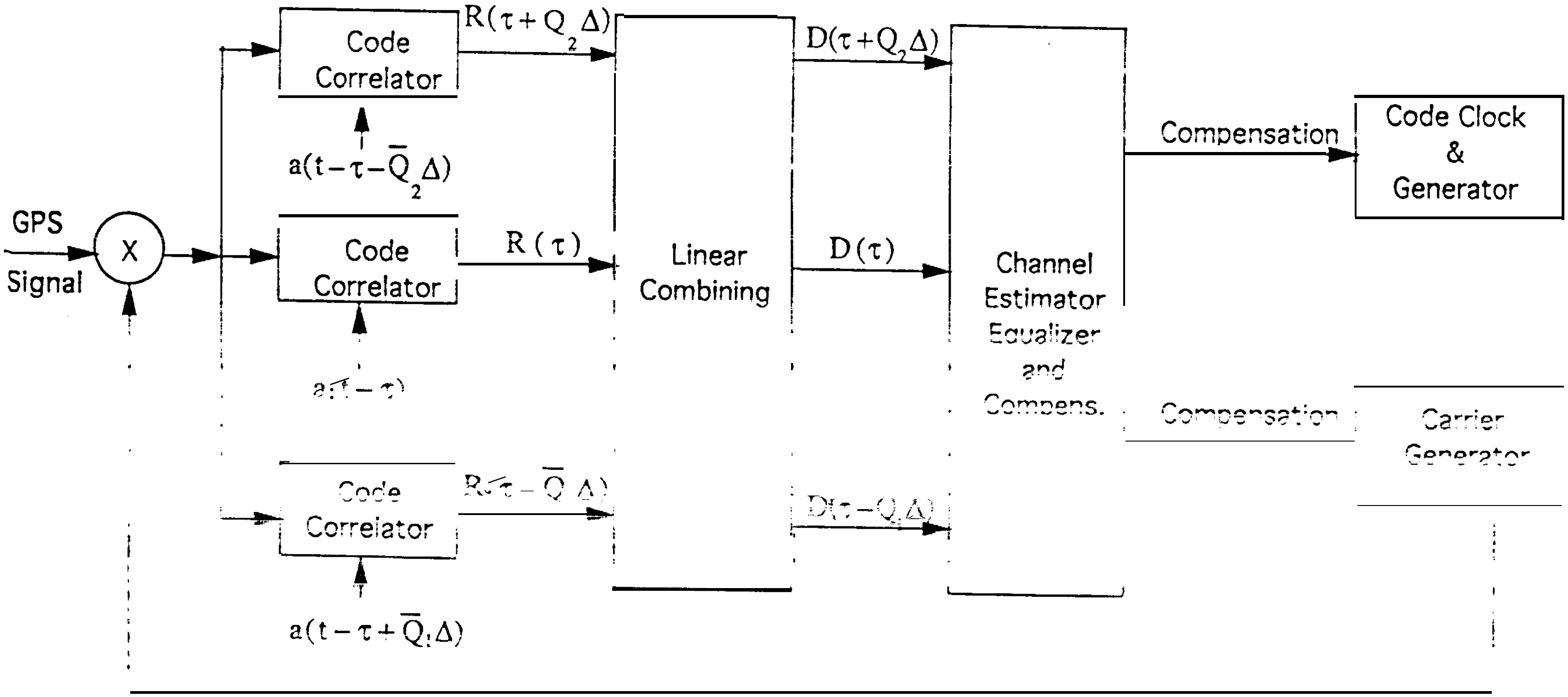
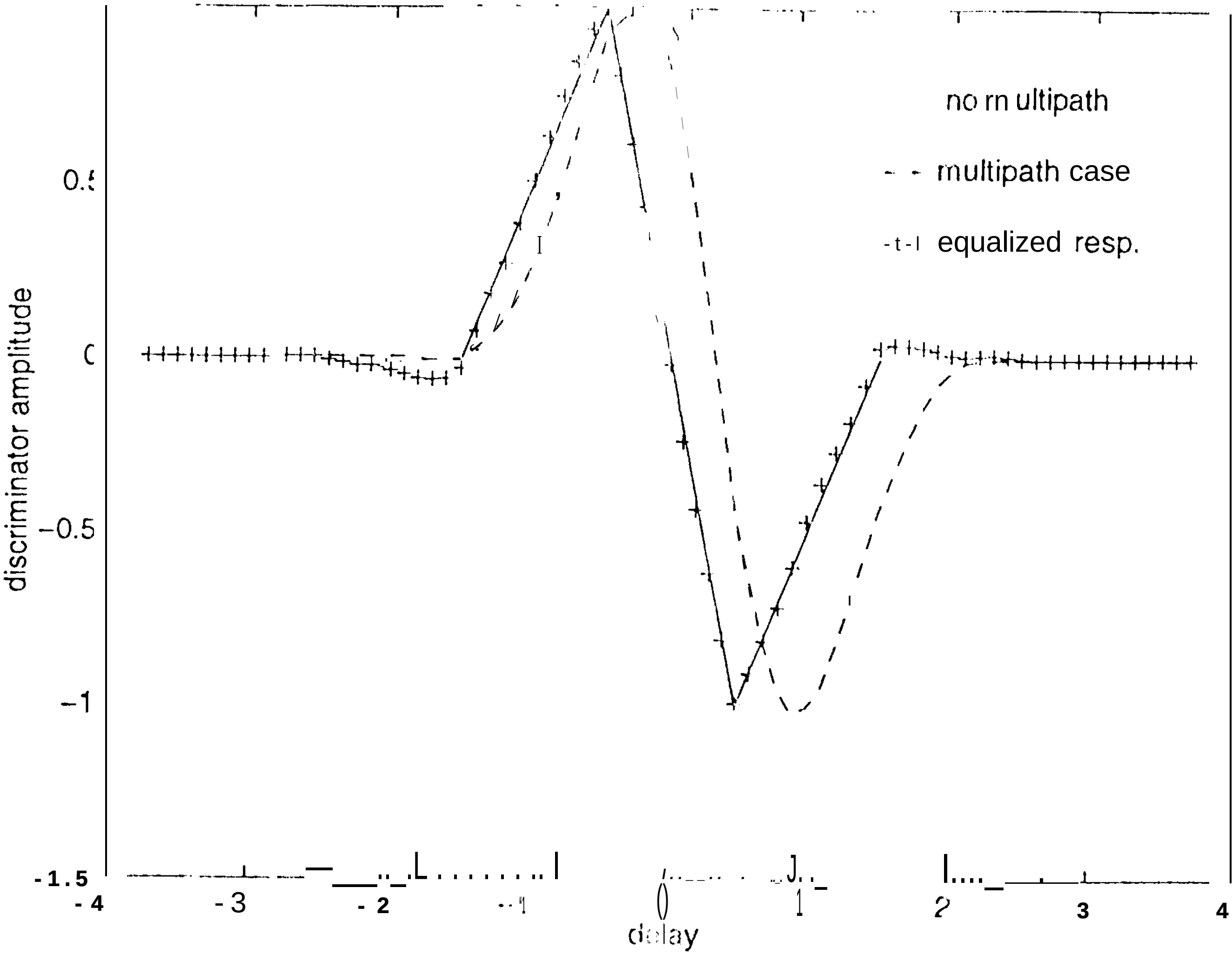


Figure 1. GPS Receiver Multipath Compensator

Figure 2. Discriminator Function of the Delay Lock Loop



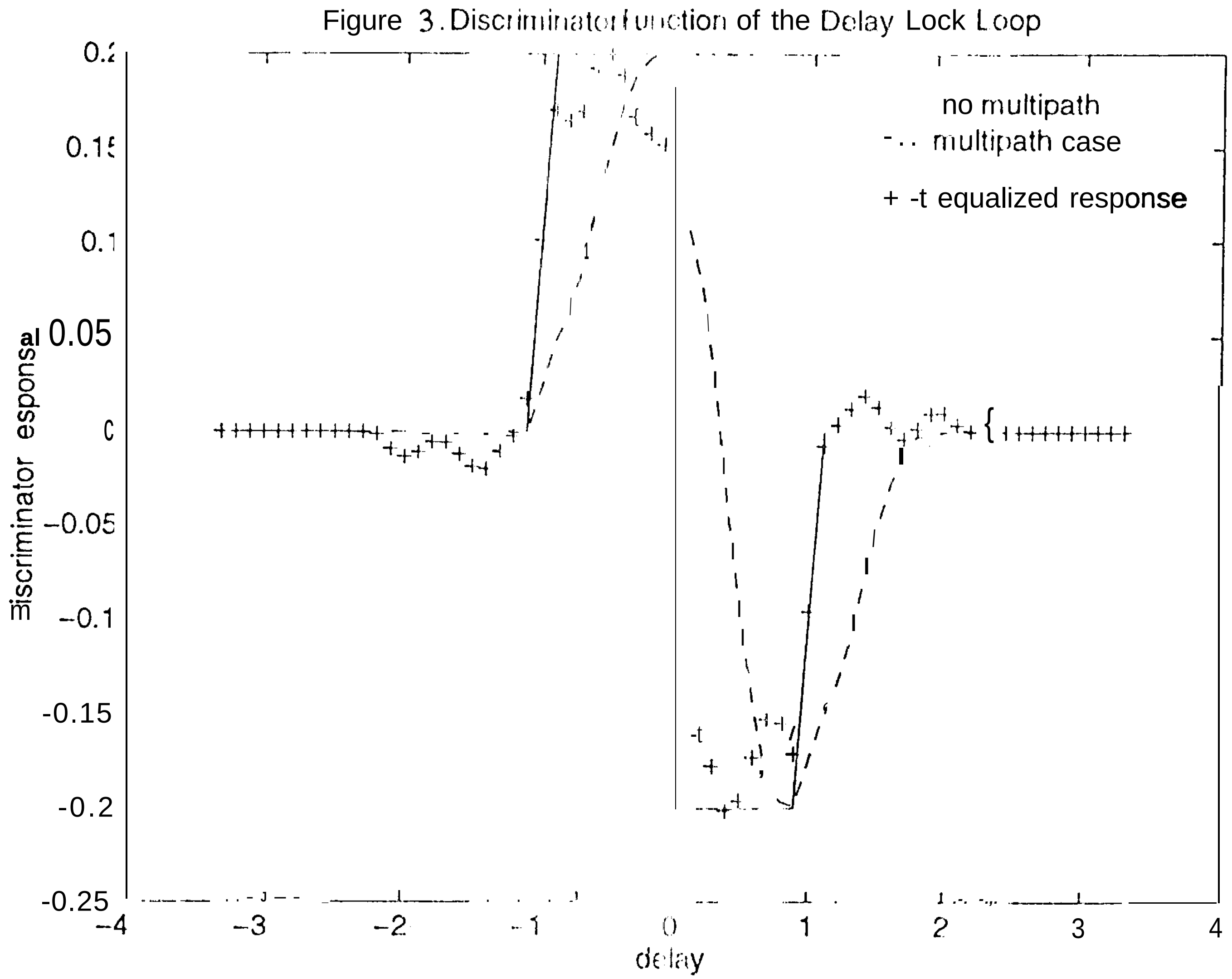


Figure 3. Discriminator function of the Delay Lock Loop
discriminator response
delay
no multipath
multipath case
equalized response
0.2
0.15
0.1
0.05
0
-0.05
-0.1
-0.15
-0.2
-0.25
-4
-3
-2
-1
0
1
2
3
4

Figure 4 . Residual Delay Estimation Error

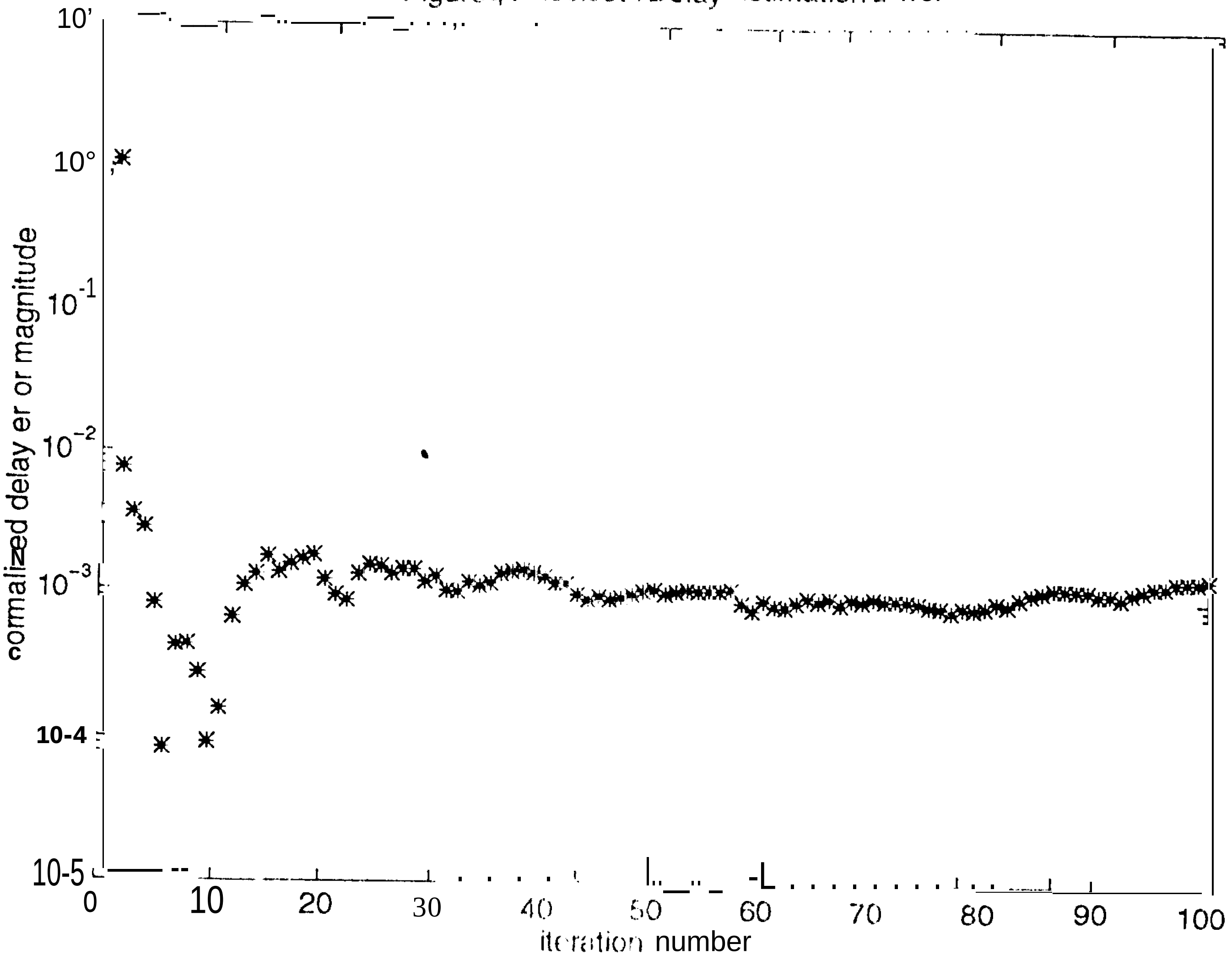
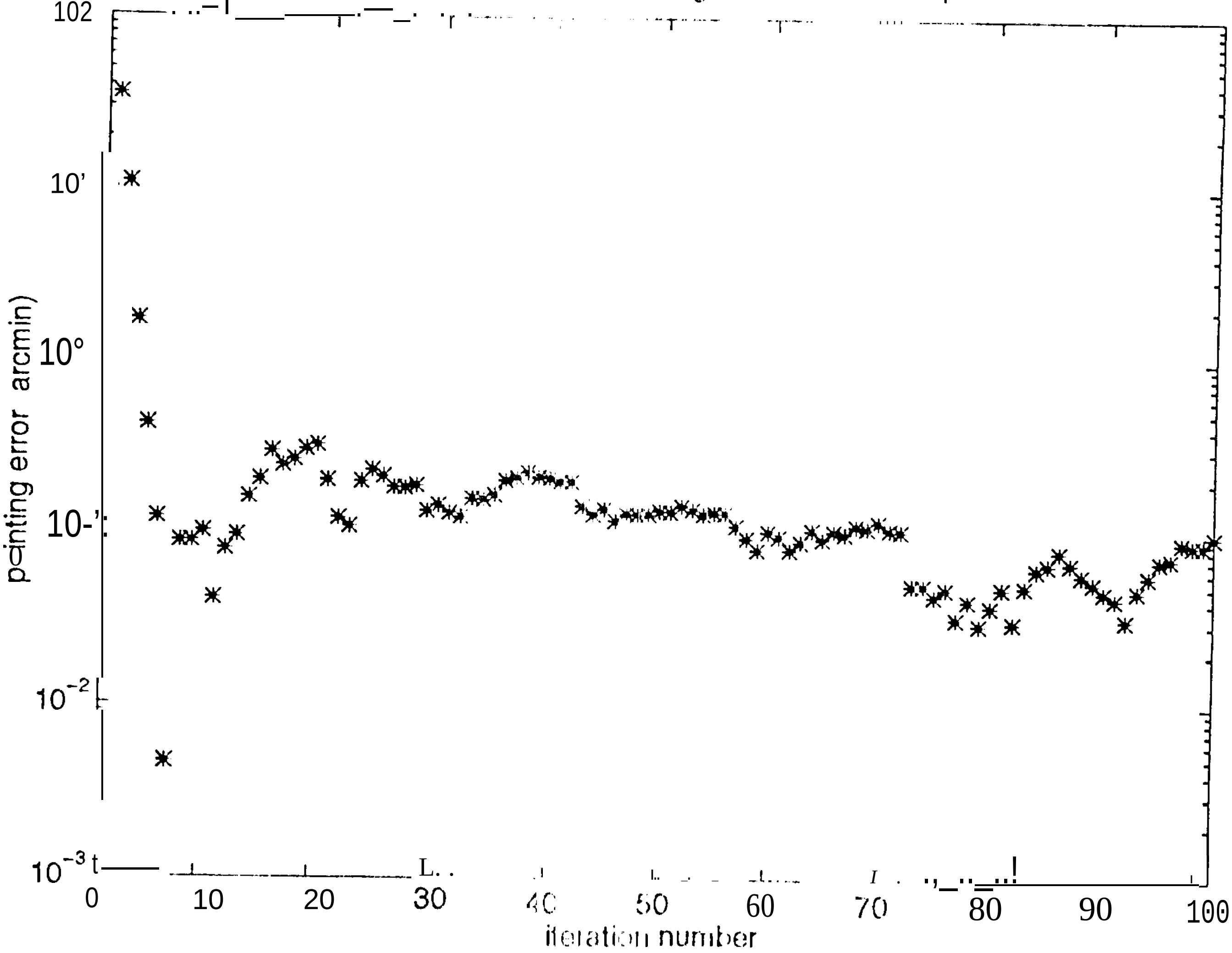


Figure 5 Residual Pointing Error due to Multipath



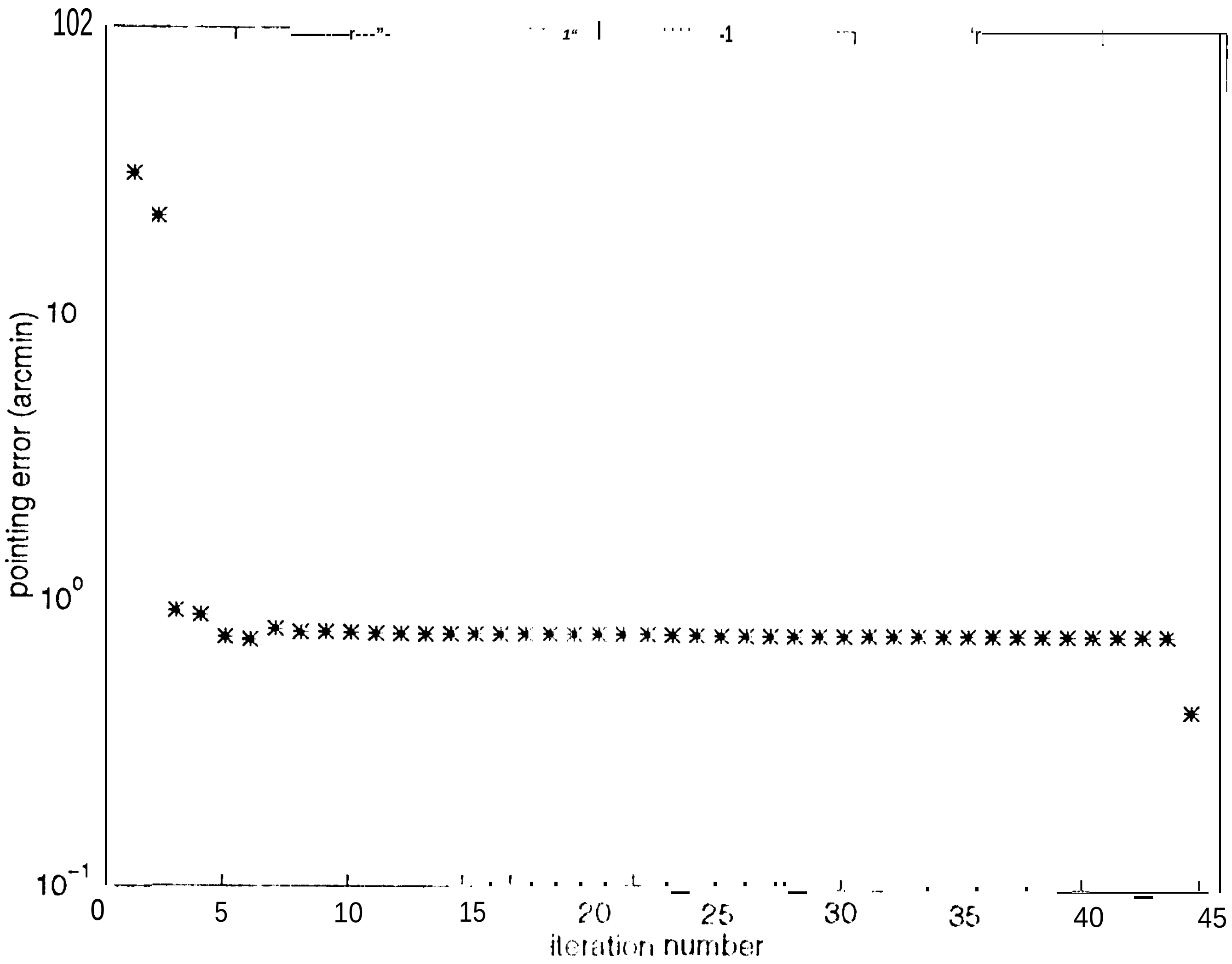


Figure 6. Residual Pointing Error due to Multipath (q=25); Amplitudes of Multipaths with Delay > .9 Δ Small

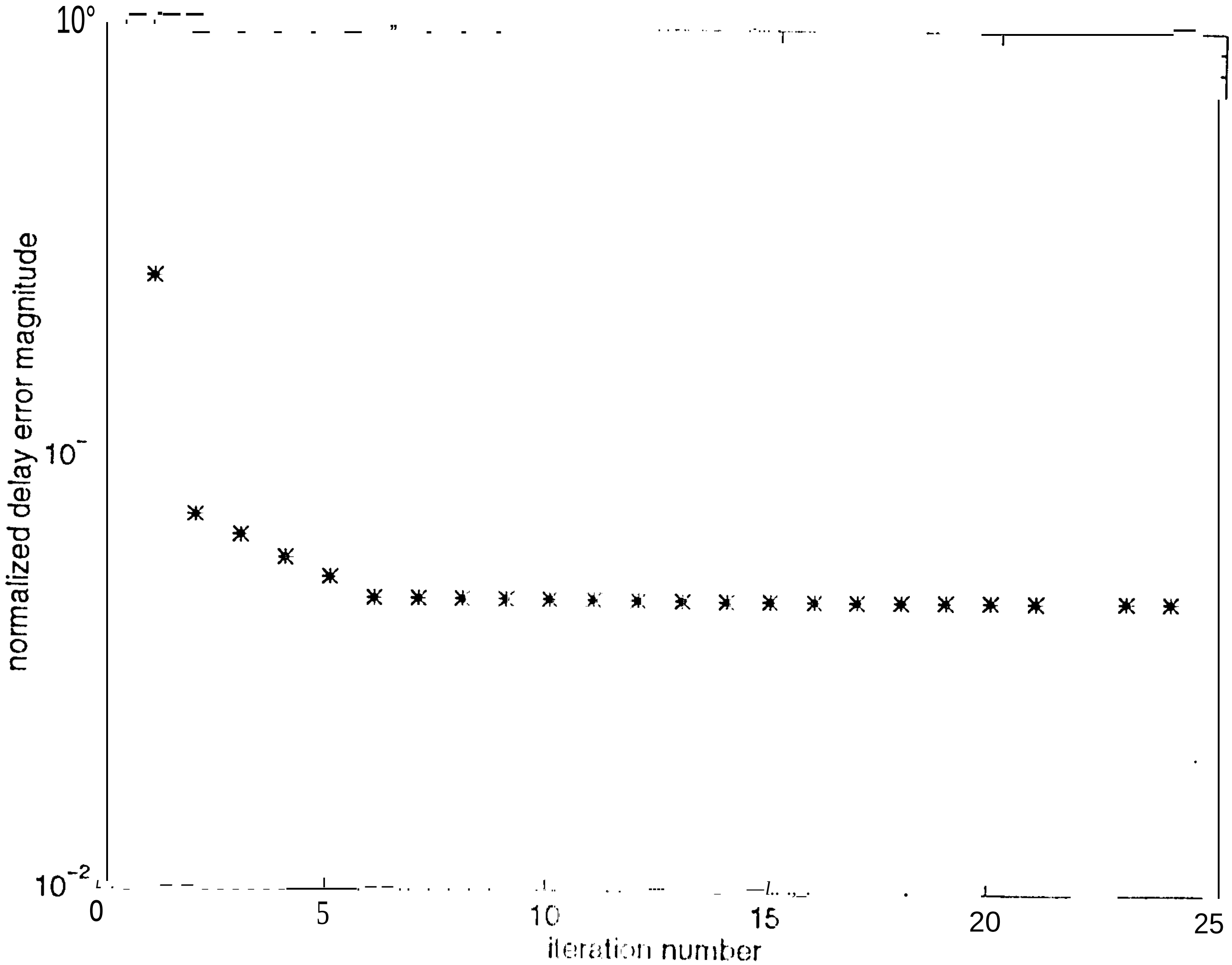


Figure 7. Residual Delay Estimation Error due to Multipath (q=25);
Amplitudes of Multipaths with Delay > .9 Δ Small

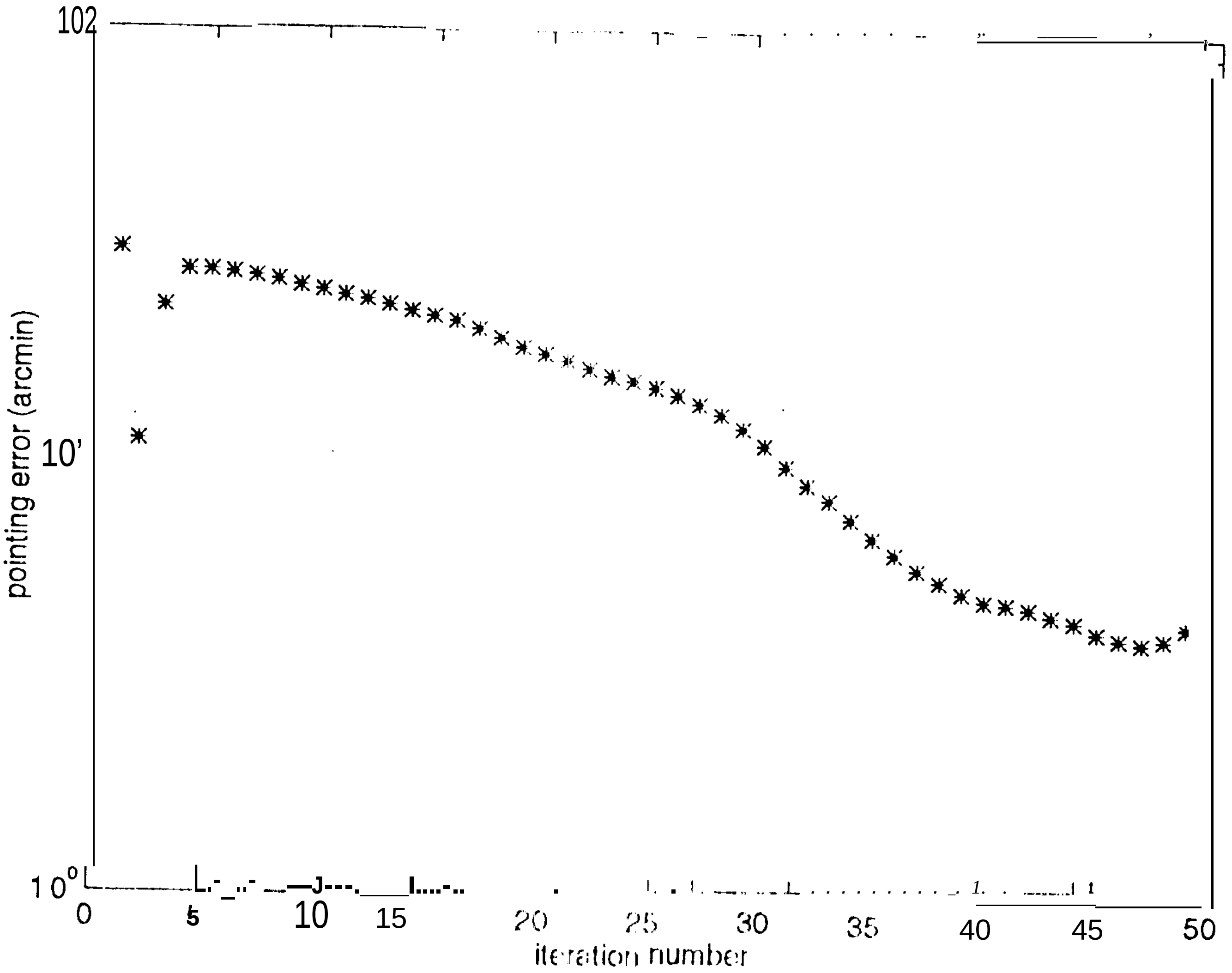


Figure 8. Residual Pointing Error due to Multipath (q=25); Filter Order Much Smaller than q.

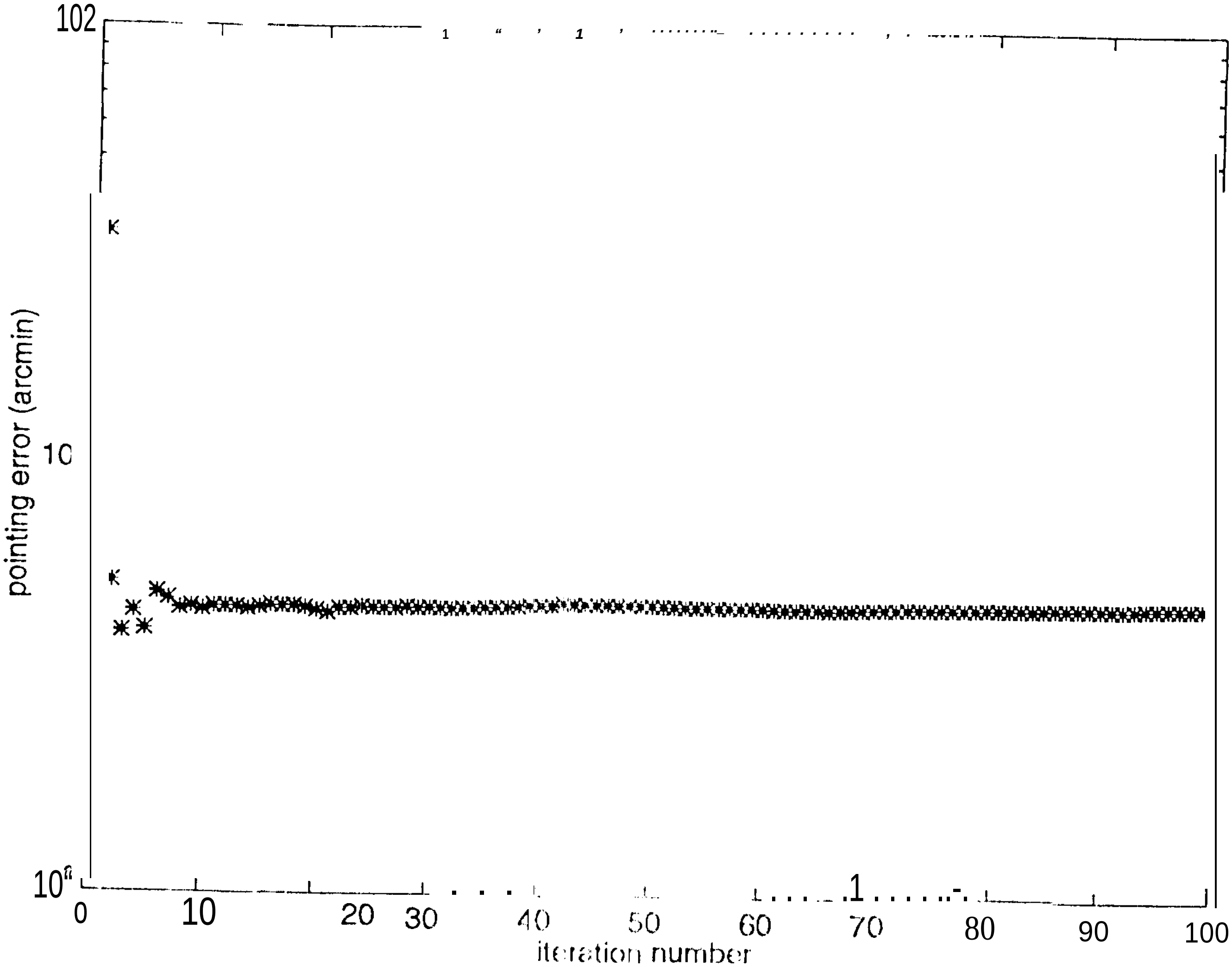


Figure 9. Residual Pointing Error due to 35 Strong Multipaths;
Filter Order $K_1 = 35$

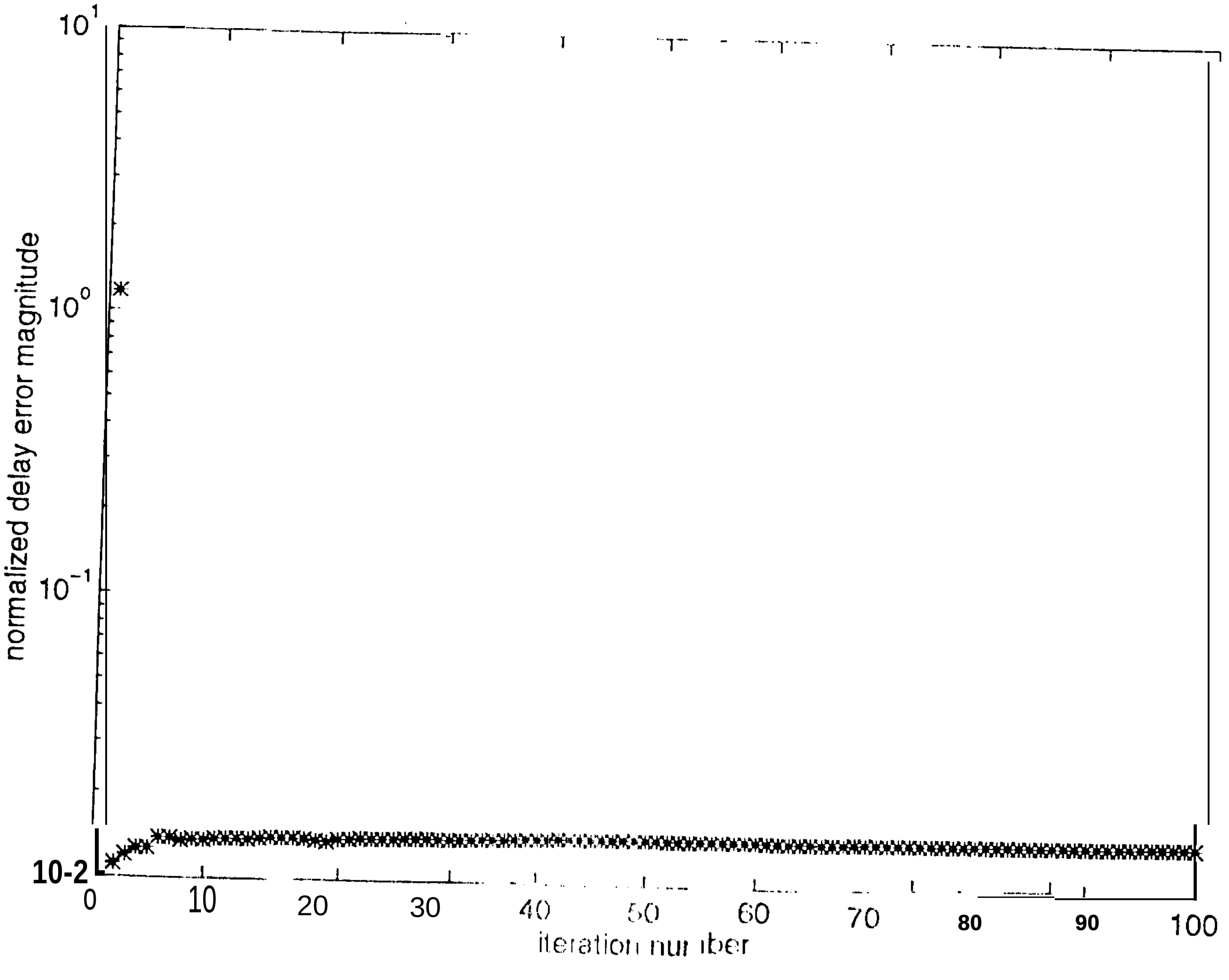


Figure 10. Residual Delay Estimation Error due to 35 Strong Multipaths; Filter Order K1=35