SLAC-PUB-5691
November 1991
(A/I)

HIGH RESOLUTION BPM FOR FFTB*

H. Hayano
National Laboratory for High Energy Physics◇
Oho 1-1, Tsukuba, Ibaraki 305 Japan
and
Stanford Linear Accelerator Center▷
Stanford University, Stanford, CA 94309 USA

J.-L. Pellegrin, S. Smith and S. Williams
Stanford Linear Accelerator Center
Stanford University, Stanford, CA 94309 USA

Abstract

The beam position monitor (BPM) for the Final Focus Test Beam (FFTB) is newly designed to have 1 μm precision and repeatability within 500 μm from center. The system design and the performance test of the prototype electronics are described.

Submitted to Nuclear Instruments and Methods A

*Work supported by Department of Energy contract DE-AC03-76SF00515.
◇Permanent address.
▷Present address.

1. Introduction

The Final Focus Test Beam (FFTB) proposed at SLAC is an optical test beam line to realize submicron spot size for future linear colliders. Optical tunings will be made by measuring beam position and then correcting magnet positions and field strengths to make final spot size very small. The beam position monitors (BPM) are required to have 1 μm precision and repeatability within 500 μm from center during optics tuning operation at the 1×10^{10} single bunch beam intensity. Another requirement is 20 μm detection precision for 2 mm bump of beam. The new BPM system is designed to have less noise and better accuracy than the SLC BPMs [1,2], of which it is an extension. In the following we describe the design and performance of the FFTB BPM system.

2. Strip Line Monitor

Strip lines were selected as a pickup for their enhanced low frequency response. Since the low frequency components in the signal are almost proportional to the strip line length, a longer strip line provides a better signal to noise ratio. In this application the strip line has been recessed into the space between the pole tips of the quadrupole magnet to maintain clearance to the beam. The strip line length was determined by the available space inside the bores of the quadrupole magnets. As a result, 457.2 mm (18 inches) was chosen for its length. One end of the strips is shorted to the chamber wall. Figure 1 shows a cross section of the monitor. The vacuum chamber is made by aluminum extrusion. It has four hyperbolic surfaces to fit and mount to the quadrupole pole tips. The extrusion aperture was picked to accommodate a beam stay clearance of $10\,\sigma$, and the electrode transverse dimensions were determined with the code "POISSON" to present a 50 Ω characteristic impedance.

3. Signal Processing Scheme

The strip line signal consists of two pulses of opposite polarities. Each pulse has the shape of the beam bunch. The pulses are 3 ns apart which corresponds to twice the length of the strip line. Since the bunch length is less than a millimeter, the induced pulse width on the strip line will be a few picoseconds. We can treat them as impulses since the electronics works in the low frequency region, and since the high frequency components are eliminated by long coaxial cables and low-pass filters in the electronics. The expected signal waveform is calculated by an approximation formula [3] which estimates an impulse response for long dispersive cables. The impulse areas are estimated using the charge in the bunch and the beam coupling coefficient to the pickup electrode. A geometrical coupling coefficient 0.042 is used and has good agreement with POISSON's result within 5% error for a centered beam. Figure 2 shows an example of the expected signal at the front end of the processing electronics. The waveform of the test pulser (manufactured by AVTECH CO.) together with an extra 30 m (100 ft) RG–223/u cable is shown in the same figure. It shows good agreement within the oscilloscope bandwidth of 400 MHz. The AVTECH pulser can be used as a signal source which simulates the beam signal after 100 ft of cable.

Four RG–223/u coaxial cables are used in one BPM for the signal transmission from the strip line pickups to the electronics front end. The four cables have identical lengths matched to within 100 ps. Unit lengths vary from 40 m (130 ft) to 85 m (280 ft) depending on the distances from the electronics hut. Since the electronics works at about 50 MHz, signal reduction by the coaxial cables is expected to be −4.3 dB to −9.4 dB at that frequency.

The processing electronics is derived from the SLC Arc electronics which has two identical channels for two pickups and detects signals by Track&Hold

(T&H) circuits. The electronics for two pickups consists of a pulse stretcher amplifier (called Head Amp), T&H and digitization (NiTNH), and a pulse generator for electronics calibration (TPG). The signals through the long cables are fed into the Head Amp, then amplified and stretched by Gaussian low-pass filters to get wider pulses, with a good signal to noise ratio (S/N), that are acceptable for the T&H circuit. The output of the Head Amps are connected to the input of the NiTNH by matched RG–223/u cables. Inside the NiTNH, the trigger signal is generated from its input by creating a difference between itself and its 3 ns delay. Its circuits work like a coupler, though the derived signal is like a differentiated one and its zero cross point does not move with signal amplitude. This trigger generation is derived from the stretched pulse, and presents the advantage of being less sensitive to the Head Amp input waveform. Because of this stretching effect, we do not need to pay much attention to the calibration pulser waveform. The main signals come out of the coupler are delayed to compensate for the delay of the trigger generation circuit. The signals are tracked and held at their first extreme by the self-generated trigger. After holding, they are digitized by 16-bit ADCs and latched until the read operation is completed. To remove the electronics offset, pedestal levels are measured in advance during calibration and subtracted from the readings.

A beam position will be calculated by

$$X' = k \times \frac{V2 - Gx \times V4}{V2 + Gx \times V4}, \qquad Y' = k \times \frac{V1 - Gy \times V3}{V1 + Gy \times V3},$$

where k is geometrical coefficient and Gx and Gy are the gain ratios $V2c/V4c$ and $V1c/V3c$, respectively between two channels, which are obtained by calibration in advance. Since the electronics is adjusted to have a very flat gain ratio with the amplitude, the gain ratios are taken as a constant, or a linear function of the amplitude summation ($V2+V4$ or $V1+V3$). These calculated positions X' and Y' are the first order approximation only. To obtain 1 μm precision, we should go to the next higher order correction. That is, using the other geometrical coefficients k' and k'',

$$X = X' + k' \times X' \times (X'^2 - Y'^2) - k'' \times \{4 \times X'^3 - 3 \times (X'^2 + Y'^2) \times X'\} ,$$

$$Y = Y' + k' \times Y' \times (Y'^2 - X'^2) - k'' \times \{4 \times Y'^3 - 3 \times (X'^2 + Y'^2) \times Y'\} .$$

In the case of a simple cylindrical BPM, geometrical coefficients k, k' and k'' can be written analytically by:

$$k = \frac{r/2}{\sin \Delta\phi / \Delta\phi} ,$$

$$k' = \frac{\sin 2\Delta\phi / \Delta\phi}{r^2} ,$$

$$k'' = \frac{\sin 3\Delta\phi / 3\Delta\phi}{(\sin \Delta\phi / \Delta\phi) \times r2} ,$$

where r is the location of the pickup electrode from the center and $\Delta\phi$ is the half opening angle of the pickup electrode from the center. This formula gives a good approximation with less than 0.01 μm error for a 500 μm off-centered beam, and less than 3 μm error for a 2 mm off-centered beam.

4. Design of the Electronics

The design of the electronics for high resolution and high precision was made for both Head Amp and NiTNH. For high resolution, the thermal noise in the signal had to be reduced. For high precision, the electronics was required to have good linearity within a dynamic range for off-centered beam signals. Figure 3 shows the block diagram of the electronics after several trials and errors. The design effort was made on the following items.

4.1 Noise reduction

The thermal noise voltage e_n is expressed by

$$e_n = \sqrt{4kTBR} \, ,$$

produced by the resistor R. Normally we are working with 50 Ω impedance at room temperature, and the bandwidth B is the only parameter we can control. Since the amplitude of the signal pulse is proportional to the square of bandwidth in the case of a low-pass filter, the signal-to-noise ratio (S/N) is proportional to $B\sqrt{B}$. Though wider bandwidth is better for high S/N, the bandwidth is limited by the T&H slew rate and the acquisition speed, in order to get enough accuracy for holding the largest pulses. The 25 MHz bandwidth that stretches the 3 ns input pulse to a 10 ns pulse is used because of the T&H speed.

Since the T&H IC has a bandwidth of about 150 MHz in order to track fast signals, in itself it has much thermal noise. However, its contribution to the S/N is decreased by the gain of the Head Amp. To realize 1 μm resolution, the S/N should be over 4243. A wide-band amplifier and low-pass filter scheme is adopted for larger signal gain and less noise. In this scheme, the signal is amplified over the frequency range of the filter cutoff, and the noise is eliminated outside the bandwidth of the filter. Figure 4 shows the required gain and noise figure (NF) in the Head Amp, modeled by one amplifier followed by one low-pass filter. From this figure, we can see that the amplifier is required to have around 38 dB gain and less than 2 dB NF. We estimated the output power of the amplifier, using the "PSpice" program.

To avoid use of an expensive, noisy, and power-consuming amplifier for a single amplifier and filter scheme, we have divided the circuit into two stages of amplification and filtering, sharing gain and power output. The selected models are Anzac AM–147 and AM–146, respectively. The low-pass filters have Gaussian responses in order to get a flat crest for accurate holding.

4.2 Linearity

Even if pickup signals are not linear with the beam position, there is an advantage to pick up signals by linear electronics for further higher order corrections. The 1 μm detection precision of relative beam movement within 500 μm from center is required for FFTB optics tuning. From this requirement, we find by numerical estimation that the electronics needs less than 0.024%/dB slope of non-linearity over a 1.45 dB range. The other optics requirement is 20 μm detection precision for a 2 mm beam bump. It requires less than 0.088%/dB slope for a range of 5.83 dB. These ranges should be taken for the signal of a centered beam of 1×10^{10}. This corresponds to about half the amplitude of ADC saturation in the case of 100 ft cable transmission.

The main nonlinearity components come from saturation and distortion of the amplifiers in the Head Amp, and the T&H trigger slewing with input amplitude.

Since the RF amplifiers we found have 'lossless feedback,' based on a directional coupler feedback circuit that provides a lower noise figure and a higher linear output, they have relatively poor reverse isolation. To avoid distortion caused by output impedance miss-match, we adopted matching inductances just before the Gaussian low-pass filters. For the first stage amplifiers, the low-pass filters are placed in the input to eliminate unwanted high frequency components which will cause distortions. Since it was difficult to meet both low NF and high output power, we compromised to the low NF side for a compact and low power-consuming Head Amp. So there is saturation beginning in the region starting at slightly over 1×10^{10} that comes from the first stage AM–147 amplifier.

When the trigger slews with its amplitude, the T&H holds the signal away from the crest and loses linearity because the crest is not flat enough. To easily achieve the above requirements, we have to adjust the trigger timing to cross the signal crest just at 1×10^{10} for a centered beam signal. In that case the allowable slewing will be less than 74 ps/dB, and it is easily attainable.

4.3 Position offset by the electronics

The contribution of the electronics to beam position offset comes from an electronics offset and a gain difference between the four channels. The electronics offset is corrected by subtracting the pedestal level. The gain difference is also corrected, by using gain-ratio functions that are measured using the calibration pulser fed into the calibration port of the Head Amp. The gain ratio measurements are done by changing the amplitude of the signal for the entire ADC dynamic range, and the correction functions are obtained by a linear fit as a function of summation of two ADC readings. This correction function should be flat, except for nonsimilarity of channels caused from saturation and distortion. The nonflat gain ratio of the correction function expresses a different nonlinearity for each channel, and will produce an error for the correction in the case of an off-centered beam.

To flatten a gain ratio, it is important (1) to make the circuit identical for each channel, and (2) to adjust the signal propagation delay difference, which mainly comes from delay lines and filters, and results in the T&H trigger slewing and sweeping over the signal crest. For this purpose, the NiTNH has an independently adjustable trigger for each channel that has the same slewing. In order to retain exchangeability of the Head Amp, we should adjust the propagation delay difference for the Head Amp itself. The required adjustment accuracy depends on the magnitude of trigger slewing, and is around 5 ps for 1 μm offset.

5. Performance of the Electronics

Because the NiTNH is still under fabrication, the TwinTNH that is employed for the SLC final focus BPMs is used to evaluate the performance of Head Amp and T&H. Since the NiTNH has been upgraded (especially its trigger circuit), we could easily modify the TwinTNH into an NiTNH, except for the trigger circuit. All the results shown here were obtained from the newly designed Head Amp and modified TwinTNH, with only two channels. The position calculation is done by;

$$X = k\,\frac{V2 - V4}{V2 + V4} \quad ,$$

and no higher order correction. The electronics test was carried out by using an AVTECH test pulser and the input signal was split into both channels simulating a centered beam signal.

In order to estimate the resolution, we made 50 position measurements during each signal amplitude and took the standard deviation as a resolution. Since the AVTECH pulser can simulate a beam signal after 100 ft of cable, we added another cable to estimate the resolution for longer cables. Figure 5 shows the result of the resolution measurements for 100-, 248- and 396-ft long cables. In the FFTB, most of the BPMs will have about 130 to 280-ft long cables, and we can obtain about 1 μm resolution for 1×10^{10} beam in all cases.

The linearity measurement is performed by placing an attenuator before the Head-Amp input and taking the ratio of the output with and without attenuator. This ratio gives the difference from a linear function, and we can obtain the nonlinearity by integrating it. The measured nonlinearity of the Head Amp and modified TwinTNH is shown in Fig. 6(a). To compare the results with the requirements, the slope of the nonlinearity as a function of amplitude is shown in Fig. 6(b). As a result, we could almost meet the requirement of $< 0.024\%/\mathrm{dB}$ for the 1.45 dB range and $< 0.088\%/\mathrm{dB}$ for the 5.83 dB range by the careful adjustment of the trigger position.

Figure 7 shows the gain ratio for the two channels. Even after careful adjustment of the propagation delays by means of coaxial phase trimmers, there is an nonflat part in the high amplitude region. It shows that there is a nonsimilarity between the two channels because of different saturation characteristics of the amplifiers. The achieved flatness of the gain ratio is 0.25%. It corresponds to about 7.5 μm offset slewing with beam intensity.

The repeatability is measured by overnight running with measuring the temperature in the crate. We can expect a position drift of 1 μm/°C. Therefore, we need good temperature control of the electronics to obtain a repeatability of 1 μm during optics tuning.

Acknowledgements

The authors would like to thank J.-P. Coulon of ORSAY for his early contribution to the Head Amp design, and David Burke for his leadership of the FFTB project at SLAC.

Reference

[1] J.-C. Denard, G. Oxoby, J.-L. Pellegrin and S. Williams, SLAC–PUB–3058, (1983).

[2] J.-L. Pellegrin, M. C. Ross, B. D. Scott and D. S. Wilson, SLAC–PUB–4209, (1987).

[3] R. L. Wigington and N. S. Nahan, "Transient Analysis of Coaxial Cables Considering Skin Effect," Proc. IRE 45, pp. 166–174, 1957.

Figure Captions

1. Transverse cross section of the FFTB BPM pickup. The inner wall has 10 mm radius. The inner conductors which are placed 12 mm from the center are supported by three ceramic discs.

2. Expected waveform at the front of electronics. The circles show the measured waveform of the AVTECH pulser with 100 ft cable.

3. Block diagram of the electronics.

4. Signal-to-noise ratio estimation for the amplifier and filter scheme varying gain and NF.

5. Measured resolution of the electronics for three cases; 100 ft RG–223/u, 248 ft and 396 ft.

6. (a) Nonlinearity of the electronics; (b) slope of nonlinearity.

7. Measured gain ratio.

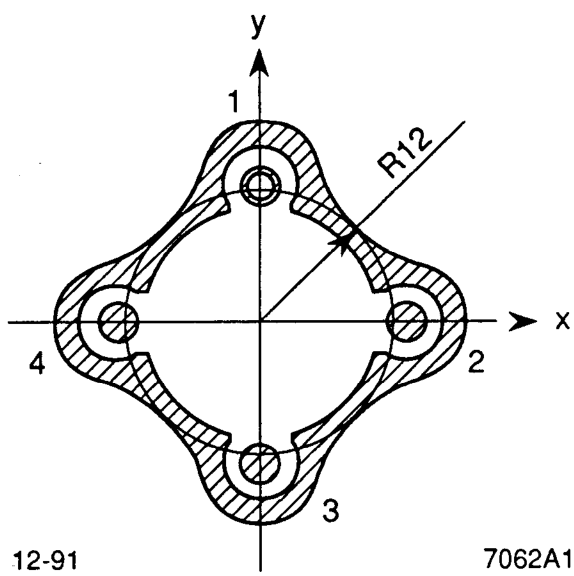


Fig. 1

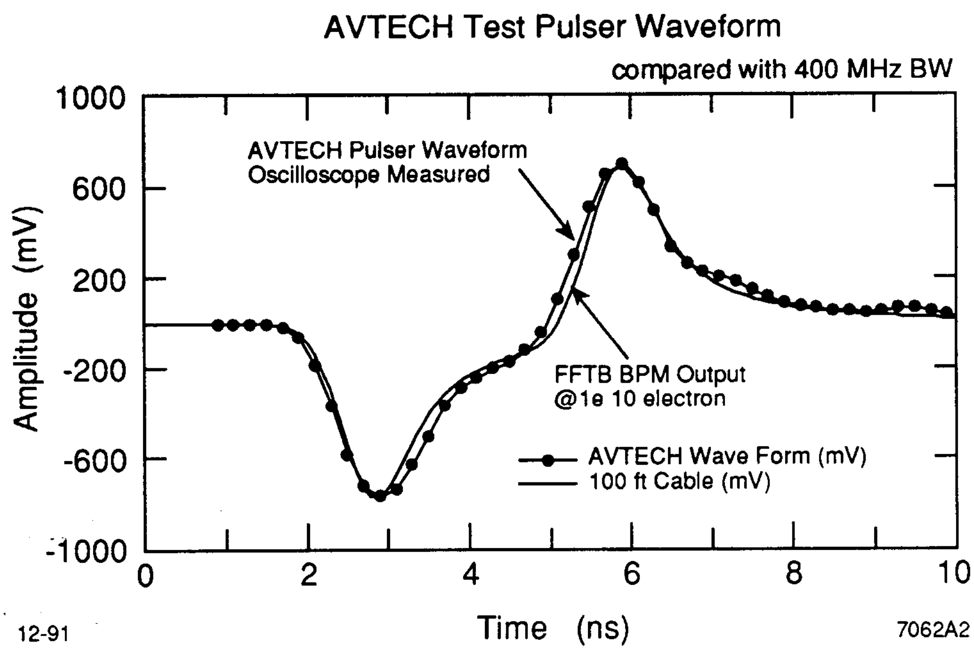

AVTECH Test Pulser Waveform
compared with 400 MHz BW
1000
600
200
-200
-600
-1000
Amplitude (mV)
0
2
4
6
8
10
Time (ns)
AVTECH Pulser Waveform
Oscilloscope Measured
FFTB BPM Output
@1e 10 electron
AVTECH Wave Form (mV)
100 ft Cable (mV)
12-91
7062A2

Fig. 2

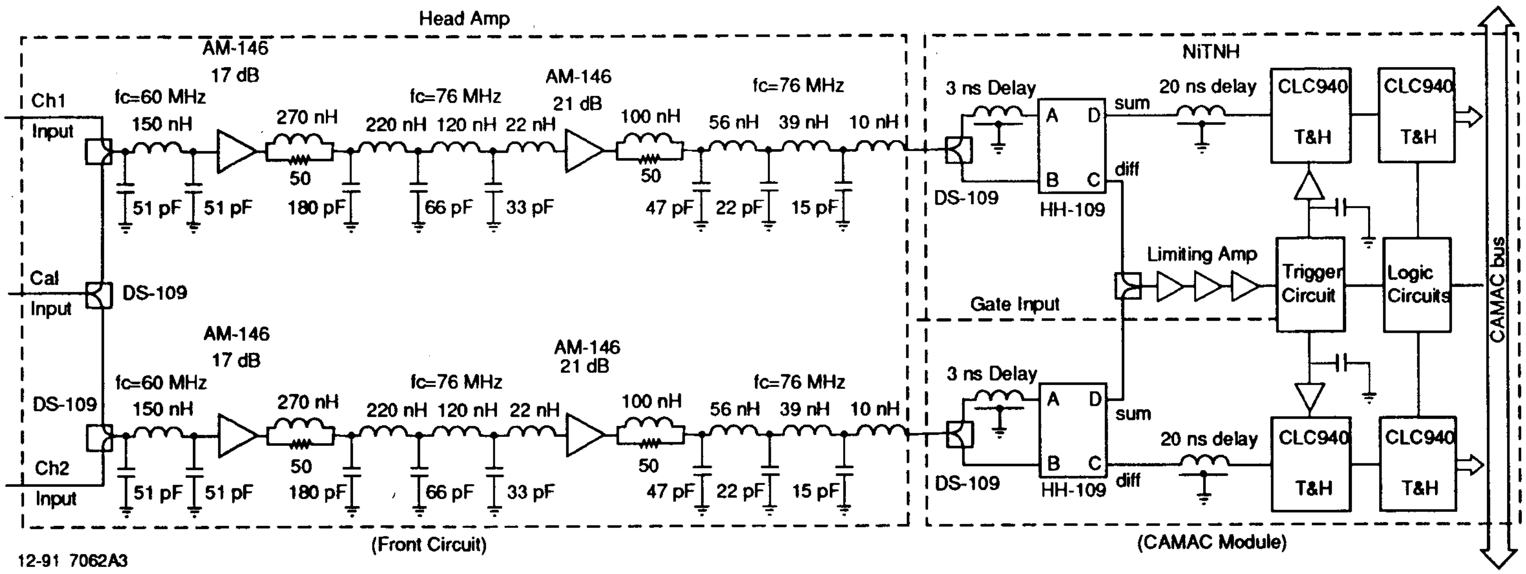
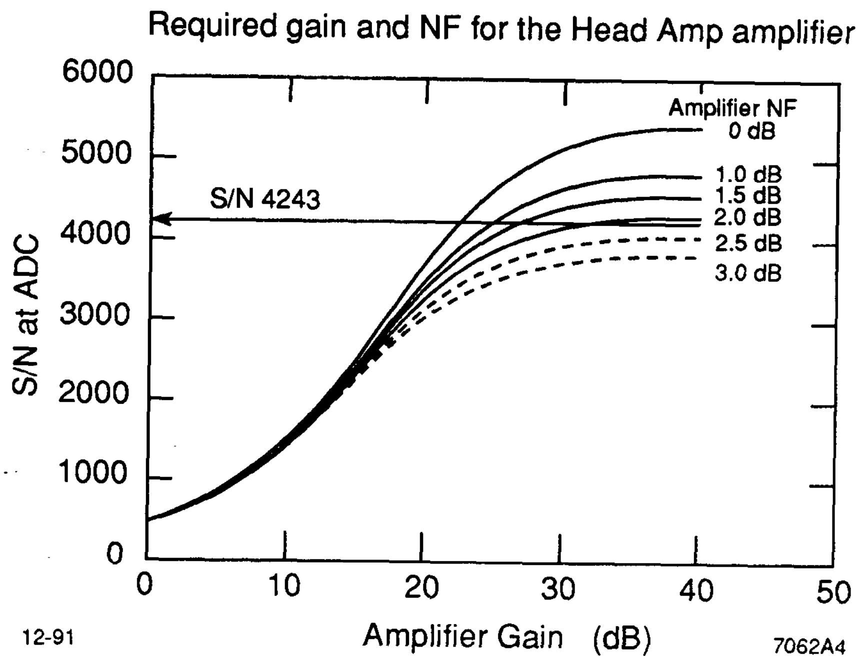
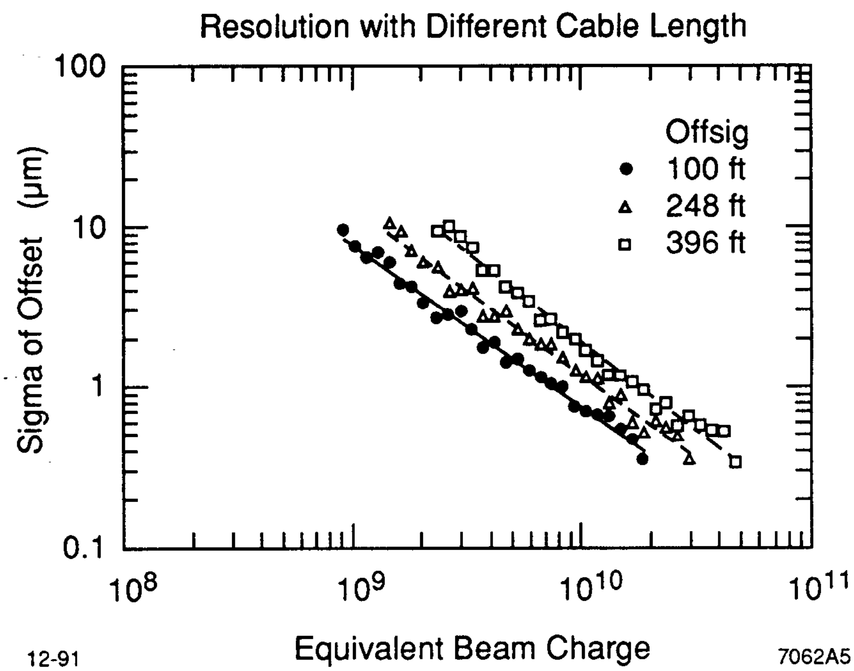

Head Amp
AM-146
17 dB
Ch1
Input
fc=60 MHz
150 nH
270 nH
50
51 pF
51 pF
180 pF
fc=76 MHz
220 nH 120 nH 22 nH
66 pF
33 pF
AM-146
21 dB
100 nH
50
fc=76 MHz
56 nH 39 nH 10 nH
47 pF
22 pF
15 pF
Cal
Input
DS-109
AM-146
17 dB
AM-146
21 dB
DS-109
Ch2
Input
(Front Circuit)
NiTNH
3 ns Delay
A D
B C
sum
diff
DS-109
HH-109
20 ns delay
CLC940
T&H
Limiting Amp
Trigger Circuit
Logic Circuits
Gate Input
CAMAC bus
(CAMAC Module)
12-91 7062A3

Fig. 3



Required gain and NF for the Head Amp amplifier
6000
5000
4000
3000
2000
1000
0
S/N at ADC
S/N 4243
Amplifier NF
0 dB
1.0 dB
1.5 dB
2.0 dB
2.5 dB
3.0 dB
0
10
20
30
40
50
Amplifier Gain (dB)
12-91
7062A4

Fig. 4



Resolution with Different Cable Length
Sigma of Offset (µm)
100
10
1
0.1
Offsig
100 ft
248 ft
396 ft
10^8
10^9
10^{10}
10^{11}
Equivalent Beam Charge
12-91
7062A5

Fig. 5

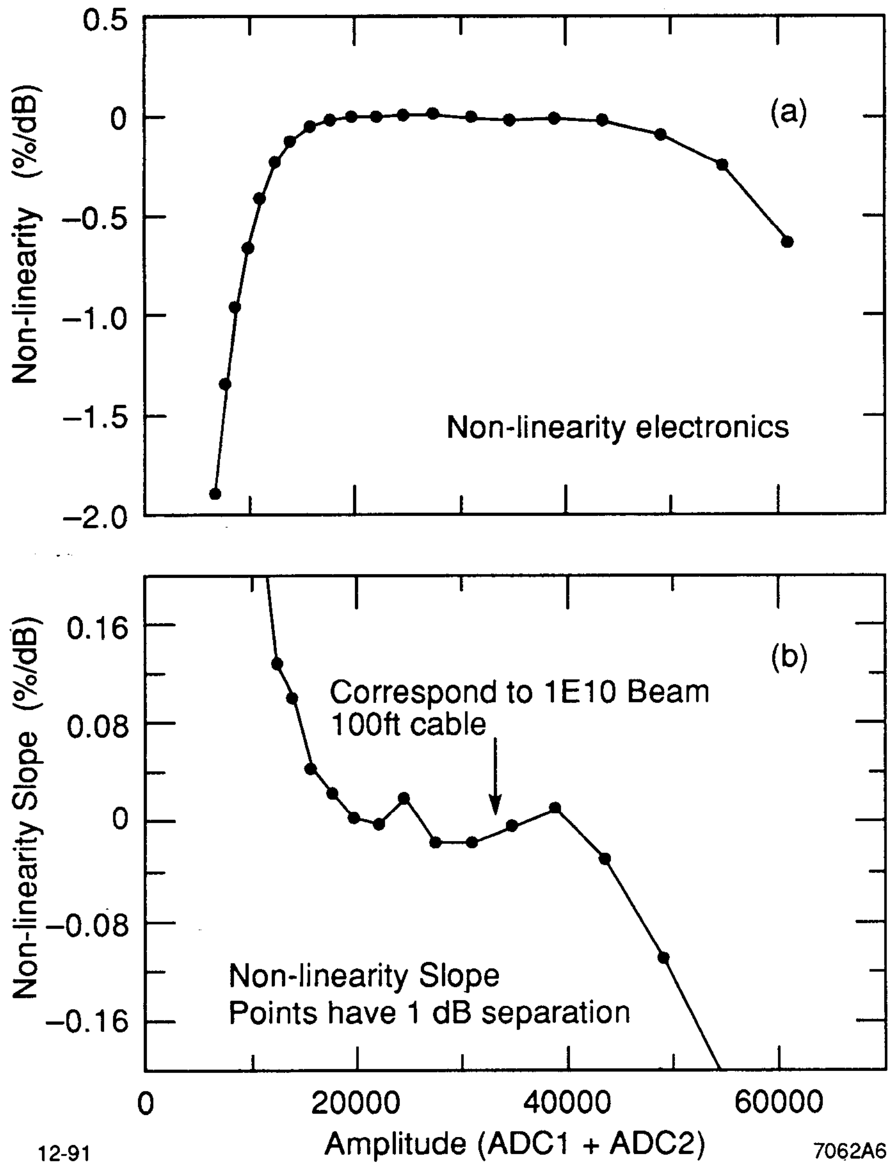

0.5
0
−0.5
−1.0
−1.5
−2.0
Non-linearity (%/dB)
(a)
Non-linearity electronics
0.16
0.08
0
−0.08
−0.16
Non-linearity Slope (%/dB)
(b)
Correspond to 1E10 Beam
100ft cable
Non-linearity Slope
Points have 1 dB separation
0
20000
40000
60000
Amplitude (ADC1 + ADC2)
12-91
7062A6

Fig. 6

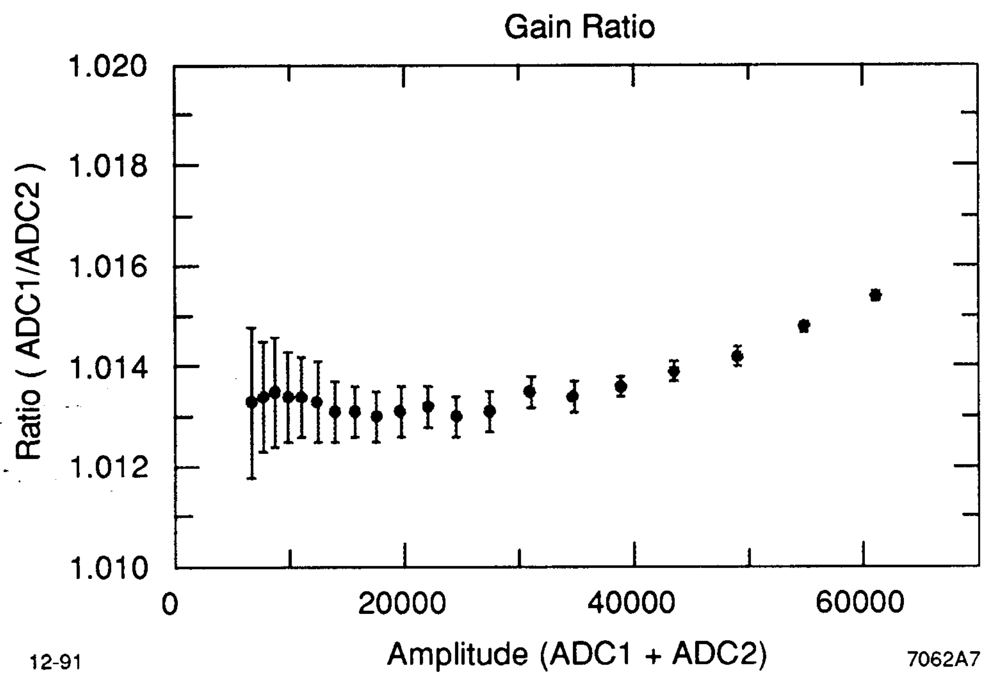

Gain Ratio
1.020
1.018
1.016
1.014
1.012
1.010
Ratio (ADC1/ADC2)
0
20000
40000
60000
Amplitude (ADC1 + ADC2)
12-91
7062A7

Fig. 7