Beam Position Monitor Electronics using DC Coupled Demodulating Logarithmic Amplifiers

G. Roberto Aiello and Mark R. Mills
Superconducting Super Collider Laboratory*
2550 Beckleymeade, MS-4005
Dallas, Texas 75237

Abstract

An electronics circuit operating up to 120 MHz suitable for Beam Position Monitor signal processing is described. Two different channels process signals from the electrodes. Each channel is realized with two cascaded DC coupled demodulating logarithmic amplifiers, providing an output voltage proportional to the logarithm of the input signal amplitude. The outputs from the two channels are processed by differential and summing amplifiers. The difference output produces a voltage proportional to the beam displacement between the electrodes, but both the difference and sum outputs are digitized in order to allow for a software correction of the gain and offset mismatches. The electronics show better characteristics than previous implementations utilizing log-amp circuits. The dynamic range has been increased, keeping the linearity error smaller than 1% over a 65 dB input signal range. The noise characteristics have been improved providing good resolution at low currents. The RF burst response has also been tested showing good characteristics for use on a Linac or Transfer Line. One prototype, working at 60 MHz, has been built and is planned for use on one or more machines at the SSC.

1. INTRODUCTION

The circuit described in this paper utilizes the AD640 integrated circuit from Analog Devices [1]. The chip is a logarithmic detector: it furnishes a DC output proportional to the logarithm of the input signal amplitude. The AD640 operates from DC to 120 MHz of input signal.

The main characteristics evaluated in this paper are:

- linearity
- dynamic range
- noise figure
- RF burst response
- frequency response
- stability
- channel matching

The parameters used in this evaluation are those of the SSC Low Energy Booster, but they are still quite general: some parameters are common for all the SSC machines.

Table 1.
Parameters used in the evaluation

sensitivity	S_x	=	0.72 dB/mm
beam current	I_b	=	1.5 - 480 mA
input frequency	f	=	47.5 - 60 MHz
bandwidth	BW	=	2 MHz

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2. DETECTOR SCHEME

The principle scheme of detection is shown in figure 1. The band pass filter (BPF) selects only the main RF component; the logamp circuit produces a DC output proportional to the logarithm of the input signal amplitude; the low pass filter (LPF) optimizes the signal to noise ratio; the differential and summing circuits are used to calculate the beam position in the beam pipe. The position is calculated in the following way:

$$x = \frac{1}{S_x} \times 20\log\left(\frac{A}{B}\right) = \frac{1}{S_x} \times \Delta V \qquad \text{(EQ 1)}$$

where A and B are the electrode potentials, S_x [dB/mm] is the sensitivity, and ΔV [dB] is the ratio of the two inputs.

Taking care of the electronics:

$$x = \frac{1}{S_x} \times \frac{1}{g_{LA}} \times \frac{1}{g_{\Delta}} \times V_{out} \qquad \text{(EQ 2)}$$

where g_{LA} [V/dB] is the logamp gain, defined as the ratio between the output voltage and the input power, g_D is the differential amplifier gain, and V_{out} [V] is the output from the differential amplifier.

In our case $g_{LA} = 0.046$, and $g_D = 9.552$, so:

$$x\,[mm] = 3.161 \times V_{out} \qquad \text{(EQ 3)}$$

The signal from the electrodes is simulated by a signal generator operating at 60 MHz. The low pass filters cut off frequency is 2 MHz. A 12 bit ADC board is used as a digitizer.

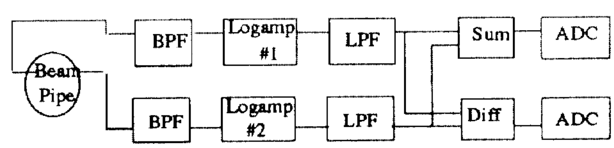

Figure 1. Principle scheme of detection.

3. LINEARITY CHARACTERISTICS

The circuit must be linear in order to get the same position measurement at different beam currents, and to get a signal proportional to the beam position when the beam is off center. The linearity error is here defined as the error, in percent, of the channel response (V_{out} vs. V_{in}) with respect to the ideal straight line.

The signal generator has been ramped from 0 dBm down to -70 dBm. The circuit output voltage is measured with a digital multimeter and, because of the signal generator non linearity, the circuit input power is measured with a vector voltmeter. The measured linearity error is less than 0.5% over 50 dB of input power and better than 1% over 65 dB of input power. The consequence of this error can be seen in figure 2.

No special attempt has been made to match the channels: all components used in the circuit are "off the shelf". An alternate method of hardware matching is described here. A software correction takes care of the different channel gains and offsets. This procedure requires the use of both the difference and sum information, and the measurements of all the circuit parameters: logamp gain and offset, differential and summing circuit gains and offsets.

The position calculated should be:

$$x = \frac{1}{S_x} \times 20\log\left(\frac{A}{B}\right) = \frac{1}{S_x} \times 20\left(\log(A) - \log(B)\right) \qquad \text{(EQ 4)}$$

with A and B being the input power from the electrodes. The two channels really give:

$$V_A = g_A \times \log(A) + \delta_A ; \ldots\ldots\ldots V_B = g_B \times \log(B) + \delta_B \qquad \text{(EQ 5)}$$

with g_A and g_B being the gain, δ_A and δ_B the offset. The calculated position is:

$$x = \frac{1}{S_x} \times \left(\frac{V_A - \delta_A}{g_A} - \frac{V_B - \delta_B}{g_B}\right) \qquad \text{(EQ 6)}$$

The difference and the sum from the two channels are:

$$\Delta = V_A - V_B ; \ldots\ldots\ldots \Sigma = V_A + V_B \qquad \text{(EQ 7)}$$

so we can express V_A and V_B in term of difference and sum:

$$V_A = \frac{\Sigma + \Delta}{2} ; \ldots\ldots\ldots V_B = \frac{\Sigma - \Delta}{2} \qquad \text{(EQ 8)}$$

The position can be expressed in terms of difference and sum in the following way:

$$x = \frac{1}{S_x} \times \left(\frac{\frac{\Sigma + \Delta}{2} - \delta_A}{g_A} - \frac{\frac{\Sigma - \Delta}{2} - \delta_B}{g_B}\right) \qquad \text{(EQ 9)}$$

The real difference and sum electronics also have a certain gain and offset:

$$\Delta_m = g_\Delta \Delta + \delta_\Delta ; \ldots\ldots\ldots\ldots \Sigma_m = g_\Sigma \Sigma + \delta_\Sigma \qquad \text{(EQ 10)}$$

with g_Δ and g_Σ being the gain, δ_Δ and δ_Σ offset;

$$\Delta = \frac{\Delta_m - \delta_\Delta}{g_\Delta} ; \ldots\ldots\ldots\ldots \Sigma = \frac{\Sigma_m - \delta_\Sigma}{g_\Sigma} \qquad \text{(EQ 11)}$$

Thus, the position calculated in terms of the digitized difference and sum, and of the electronics parameters is:

$$x = \frac{1}{S_x} \times \left(\frac{\frac{\frac{\Sigma_m - \delta_\Sigma}{g_\Sigma} + \frac{\Delta_m - \delta_\Delta}{g_\Delta}}{2} - \delta_A}{g_A} - \frac{\frac{\frac{\Sigma_m - \delta_\Sigma}{g_\Sigma} - \frac{\Delta_m - \delta_\Delta}{g_\Delta}}{2} - \delta_B}{g_B}\right) \qquad \text{(EQ 12)}$$

The result obtained applying this method is shown in figure 2. The signal generator power is ramped down from 0 dBm to -70 dBm, simulating different beam currents. A set of switched attenuators, giving different power ratios between the two channels, simulates different beam positions in the beam pipe. Many measurements are taken at 1 dB attenuation steps between the two channels. This correspond to 1.4 mm steps according to the sensitivity value shown in table 1.

The ideal plot would show many straight lines. In reality the lines are not straight because of the electronics linearity error, and the mismatch between the two channels. The largest fluctuation in amplitude is present at 10 dB power ratio between the two channels. This is due to the AD640s internal structure: each of the five internal stages has 10 dB of gain.

An interesting idea to reduce this fluctuation is to use two pairs of cascaded AD640s per channel, each pair contributing to the same output. Both the pairs would receive the same input signal, with one input being attenuated by 5 dB. This would correct the internal stage 10 dB gain effect.

The fluctuation due to the two channels linearity error, visible in the plot, is less than 0.05 mm over 65 dB of beam current for the beam on center, and less than 0.3 mm over 55 dB for the beam 2.8 mm off center. No special test has been done in order to measure the circuit parameters stability, but the same slope and intercept have been measured over 5 months of testing.

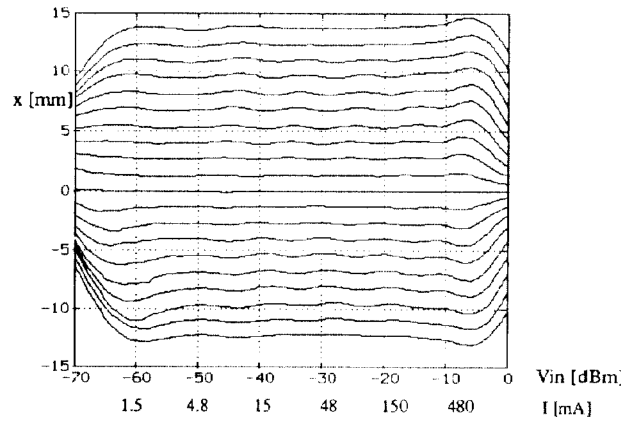

Figure 2. Simulated beam fluctuation.

4.0 NOISE MEASUREMENT

The signal to noise ratio, due to the thermal noise and the electronics, limits the measurement resolution [2]. The resolution due to the signal to noise ratio is:

$$\delta x = \frac{b}{2\sqrt{2}} \sqrt{\frac{P_N}{P_S}} \qquad \text{(EQ 13)}$$

with P_S and P_N in Watts, or:

$$\delta x = \frac{b}{2\sqrt{2}} 10^{\left(\frac{P_N - P_S}{20}\right)} \qquad \text{(EQ 14)}$$

with P_S and P_N in dBm. P_S is the input power, P_N the noise power, and b the beam pipe radius.

The thermal noise can be calculated with:

$$\begin{aligned} P_N[dBm] &= 10\log(4KTR) + 10\log(BW) + NF \qquad \text{(EQ 15)} \\ &= -174 + 10\log(BW) + NF \end{aligned}$$

R = 50 Ω, BW = bandwidth [Hz], and NF = noise figure [dB].

For this measurement the signal generator is providing the same input to both channels. Both an RF voltmeter and 12 bit ADCs are used to digitize the differential and sum circuits. The signal power is ramped, and the position calculated using (EQ 3).

The resolution can now be determined by calculating the standard deviation in either one of the following ways: digitize V_{out} with the ADC, calculate the position, and calculate the standard deviation of x, or measure the r.m.s. value of V_{out} with the RF voltmeter, and calculate the r.m.s. value of x (which is its standard deviation).

The results of the measurements are shown in figure 3. Both the experimental plot and the theoretical plot (considering NF = 0 dB) are shown.

The noise figure is represented by the distance between the theoretical and measured plots. The noise figure evaluated in this way is 17 dB.

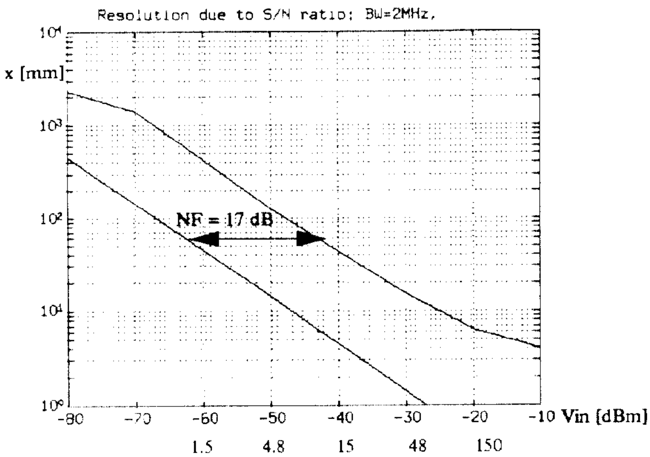

Figure 3. Noise measurement with BW = 2 MHz.

5.0 RF BURST RESPONSE

It is useful to know the RF burst response if the circuit will be applied to the Linac or transfer lines, or if a gap is present in the circular machines.

Both the logamp input and the output signals are measured with an oscilloscope. The output signal from the logamp is measured without a low pass filter in order to get a response due only to the logamp circuit. The result is shown in figure 4. The circuit risetime corresponds to two cycles of input signal.

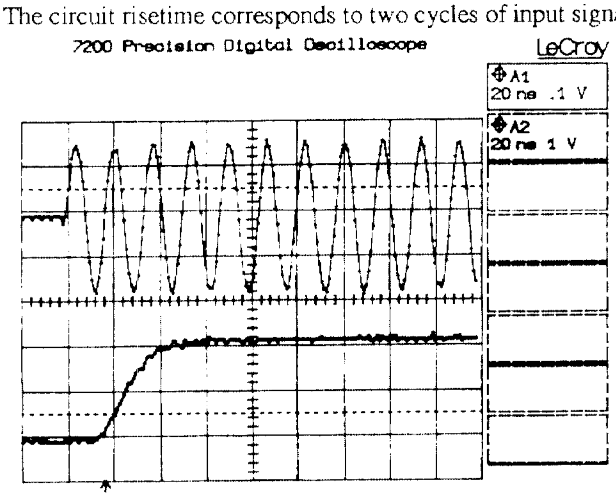

Figure 4. RF burst response.

6.0 FREQUENCY RESPONSE

The frequency response is particularly important in the LEB, where the RF frequency changes from 47.5 MHz to 60 MHz during the acceleration ramp.

The signal generator is set at a fixed power, and the frequency is swept from 47 MHz to 60 MHz. Many measurements are taken using different values of the variable attenuators. The position is calculated using (EQ 12). The results are shown in figure 5. The lines are not straight because the gain of the channels is not constant with frequency. This data indicate less than 0.1 mm of drift at all the calculated positions.

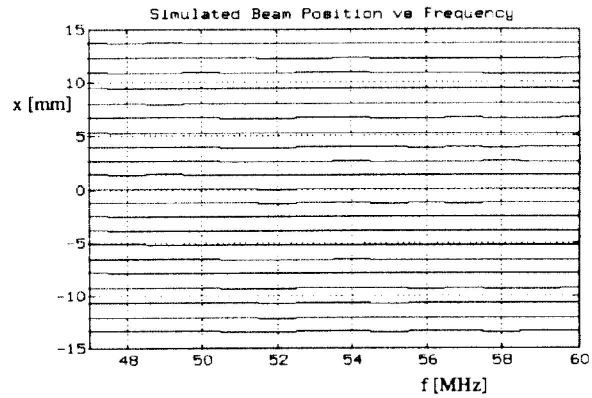

Figure 5. Frequency response.

7.0 PERFORMANCE

The summarized performance is as follows:

- *dynamic range at 1% error*: the value directly measured is 65 dB, and the plot in figure 2 confirms this value, showing a good working region for 55 dB of input current variation
- *noise*: two independent methods confirm the behavior predicted by theory, the noise figure being 17 dB
- *RF burst response*: the test at 60 MHz input frequency shows a risetime corresponding to two cycles of input signal
- *frequency response*: the range of interest, 47.5 MHz - 60 MHz, produces less than 0.1 mm error in the measured position
- *channel matching*: a software correction involving a calibration, measuring offset and gain, has been proven very useful

8.0 CONCLUSION

The results from this preliminary evaluation are encouraging (see also [3]). The performance is better than previous versions [4]. Dynamic range and fluctuation fit the requirements for all the SSC machines. The LEB frequency sweep produces less than 0.1 mm of error in the measured position; the stability, RF burst response and channel matching are all satisfying. A prototype for the SSC Linac will be built in a short time.

9.0 ACKNOWLEDGEMENTS

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10.0 REFERENCES

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