Fermilab-TM-2208

# Calibration of the Resistive Wall Monitor for the SBD and FBI Systems

Jim Crisp and Brian Fellenz
June 1, 2003

## INTRODUCTION

Calibration of the Tevatron (Tev) Resistive Wall Monitor (RWM) used by the Fast Bunch Integrator (FBI) and the Sampled Bunch display (SBD) is discussed. A 1% error would result from a 0.09db difference in gain. Care must be taken to insure the Network Analyzer used to measure components of the system is calibrated and that test cables are properly terminated and taken into account. Calibration of the Main Injector (MI) RWM and FBI is also presented. Appendix B provides a comparison of the Tev and MI RWM's.

Frequency response of the system must be understood from dc to several hundred MHz to insure the measurement is independent of bunch shape. The expected bunch lengths are between 1.3 and 2.6nsec sigma t or 120 and 60 MHz sigma f respectively. On average, 95% of the beam will fall within ±1.96 sigma.

Calibration with an arbitrary wave form generator agrees to within 1% of the calibration empirically found by comparing the FBI to the DCCT with beam.

## RECENT IMPROVEMENTS

The integrator amplifier was replaced with an AD8055. The limited bandwidth of the original part caused the measured intensity to change by as much as 15% depending on the timing.

Nearly half of the cable attenuation and thus dispersion was being generated in RG58 cable used to run the RWM signal upstairs. The RWM heliax cables were extended by about 4 feet to a bracket on the front of the FBI relay rack. A short good quality cable is used from there to the FBI input. Cable dispersion still causes a 3% gain change with gate width. A correction for this should be implemented in software once the scale factor is verified.

## RESISTIVE WALL MONITORS

Two RWM's are installed a few feet apart in the Tevatron. They have small differences in response even though they were constructed identically. The upstream RWM is used for general purposes. The downstream RWM is dedicated to FBI and SBD

measurements and is the one discussed here. These RWM's have a calibrate or test port designed into them. Figure 1 shows the internal connections and resistor values.

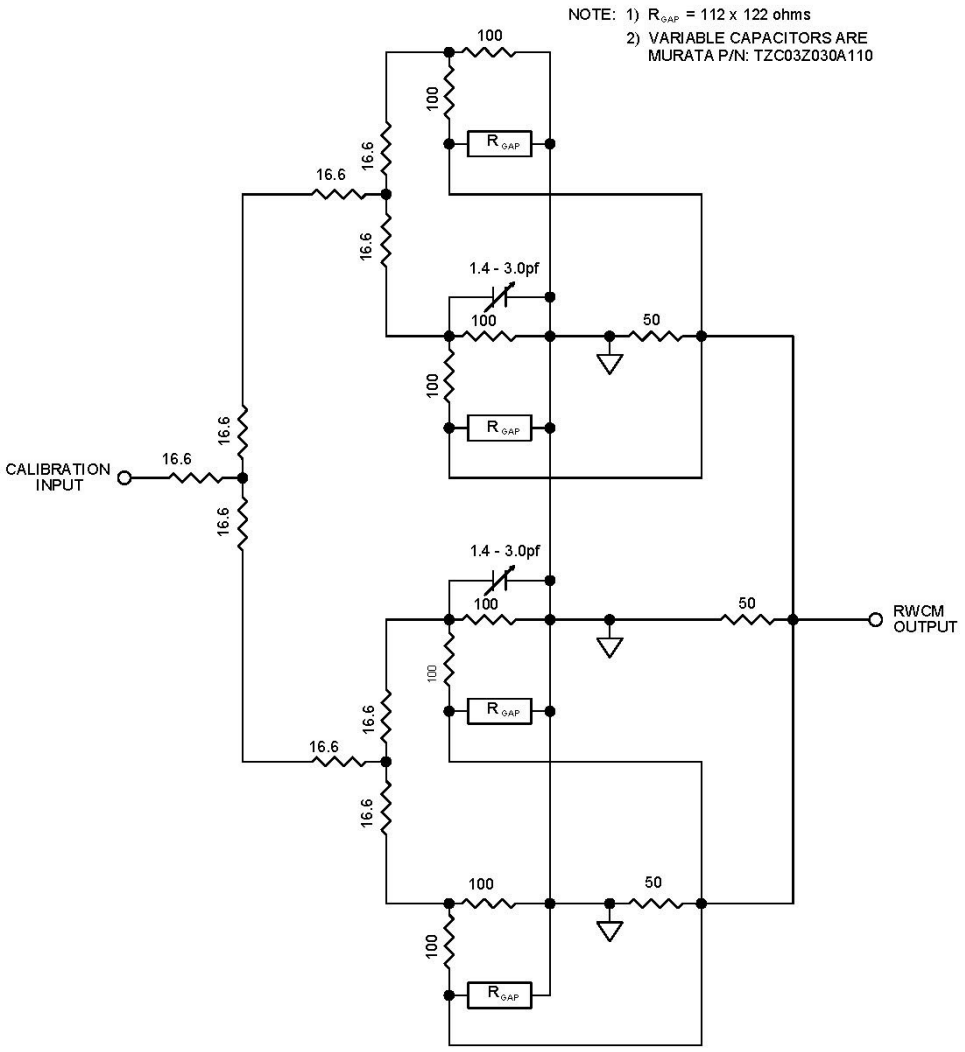



**Figure 1**. Internal connections and resistor values for the Tevatron RWM's.

The calibrate signal is divided with resistive splitters and connected to the gap through 100Ω at four equally spaced points. The 100Ω resistors were required to properly terminate the incoming signal. The output comes from four points equally spaced

between the calibrate connections combined through resistive combiners. The 1.288Ω gap impedance is determined by the 80x120Ω resistors, 4x133.3Ω calibration port connections, and 4x50Ω output port connections in parallel. The output is just the gap resistance times the beam current.

$$TevR_{gap} = \frac{1}{\frac{80}{120} + \frac{4}{133.3} + \frac{4}{50}} = 1.288\Omega$$

The MI RWM uses 112x122Ω resistors.

$$MIR_{gap} = \frac{1}{\frac{112}{122} + \frac{4}{133.3} + \frac{4}{50}} = 0.9727\Omega$$

The calibrate signal is attenuated by the voltage divider formed by the gap impedance and the 100Ω resistors at the connection points. The calibration port can be used to monitor the beam signal, however, it will be attenuated by 16.89db compared to the normal output for both Tev and MI.

$$Tev\frac{Vout}{Vcal} = \frac{1}{50}\left(\frac{\frac{100}{4}}{\frac{100}{4} + \left(\frac{100}{4} + \frac{1}{\frac{4}{50} + \frac{80}{120}}\right)}\right)\left(\frac{1}{\frac{4}{50} + \frac{80}{120}}\right) = .01304 \qquad \text{-37.69db}$$

$$MI\frac{Vout}{Vcal} = .009823 \qquad \text{-40.16db}$$

$$\frac{Vcal}{IbeamRgap} = \frac{\frac{50x100}{50+100}}{100 + \frac{50x100}{50+100}} \frac{50}{50 + 16.6 + \frac{2x16.6}{2} + \frac{16.6}{4}} = 0.1431 \qquad \text{-16.89db}$$

**MEASUREMENTS**

The response of the calibrate cable (1/2" heliax) and the output cable (7/8" heliax) connected through a type N bullet in the tunnel is compared to an ideal cable in figure 2. An overall delay of 638.78 nsec makes the measured and calculated phase agree below 1GHz.

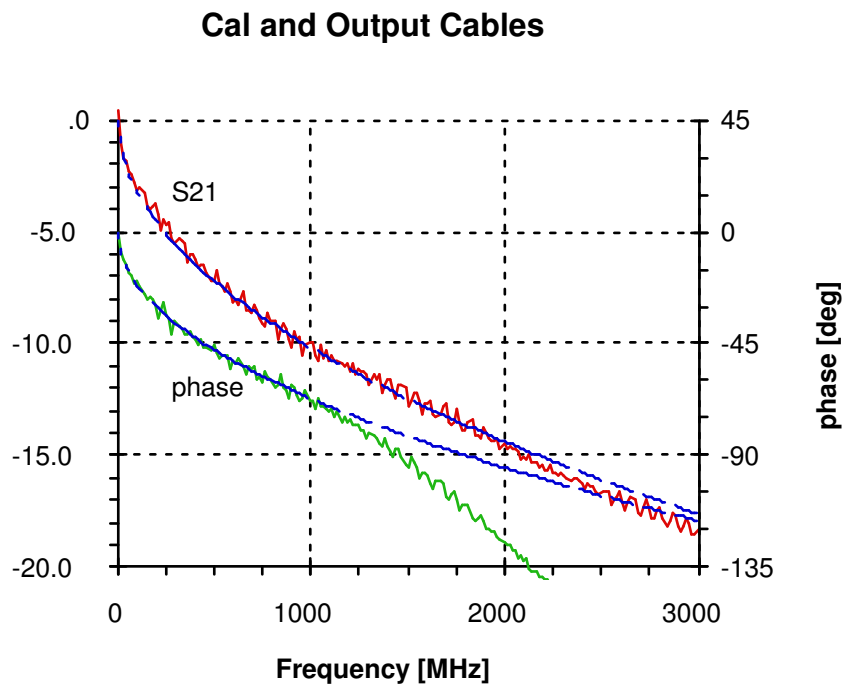



**Figure 2**. Calibrate and output cables in series for the Tev downstream RWM. The cables were connected through a type N bullet in the tunnel. The dashed traces are calculated using 638.78 nsec, v/c=.885, and 1.83db/100ft@1GHz

**Table 1**. Various types of cable and their attenuation.

| | | db/100ft @1GHz | v/c | Ω/1000ft center/outer |
|---|---|---|---|---|
| 7/8" heliax | LDF5-50A | 1.31 | .89 | .35/.36 |
| 1/2" heliax | LDF4-50A | 2.34 | .88 | .50/.60 |
| 3/8" heliax | LDF2-50 | 3.50 | .88 | 1.17/.87 |
| 1/2" superflex | FSJ4-50B | 3.58 | .81 | .82/1.00 |
| 1/4" superflex | FSJ1-50 | 6.59 | .78 | 2.5/2.0 |
| RG213 | | 9.00 | .67 | 1.7/2.9 |
| RG58 | | 20.0 | .67 | 10.8/4.1 |
| RG174 | | 34.0 | .67 | 97/10.3 |

Assuming equal lengths of cable, the calibrate and output cables are 319 nsec long. The attenuation at 1GHz through the output and calibrate cables is 3.65 and 6.52db respectively. The excellent agreement between the calculated and measured attenuation using the equation provided in Appendix A suggests beam measurements can be accurately corrected for dispersion.

Figure 3 compares the RWM response as measured directly in the tunnel to the response measured from upstairs through but corrected for the calibrate and output cables. The agreement demonstrates the accuracy with which the RWM response can be verified remotely.

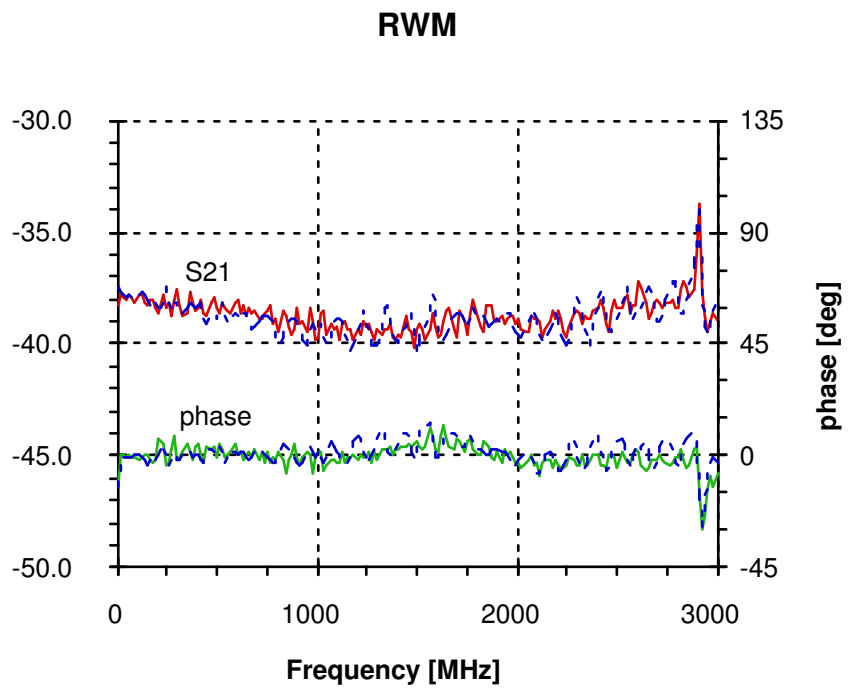



**Figure 3**. RWM response measured from upstairs through the cables compared to the response measured in the tunnel (dotted traces).

**FBI SCALE FACTORS**

The downstream RWM signal is brought directly to a high quality 3 way resistive splitter at the top front of the relay rack. Short solid jacketed cables bring the signal directly to the SBD, FBI, and MDAT FBI systems. The signals to the FBI systems are again divided into separate p and pbar integrator modules with two way resistive splitters. Currently, the pbar signals are amplified by a factor of ten before they enter the fast integrator module. The pbar amplifiers have 10% compression at 24e9/bunch.

The response of the front end of the fast integrator is shown in Figure 4. The signal monitor port is attenuated with a 499Ω series resistor (-20.81db) inside the fast integrator module. Using the average response from dc to 100MHz, the proton integrator response is -14.71+20.81 = +6.10db and the pbar integrator is +6.21db.

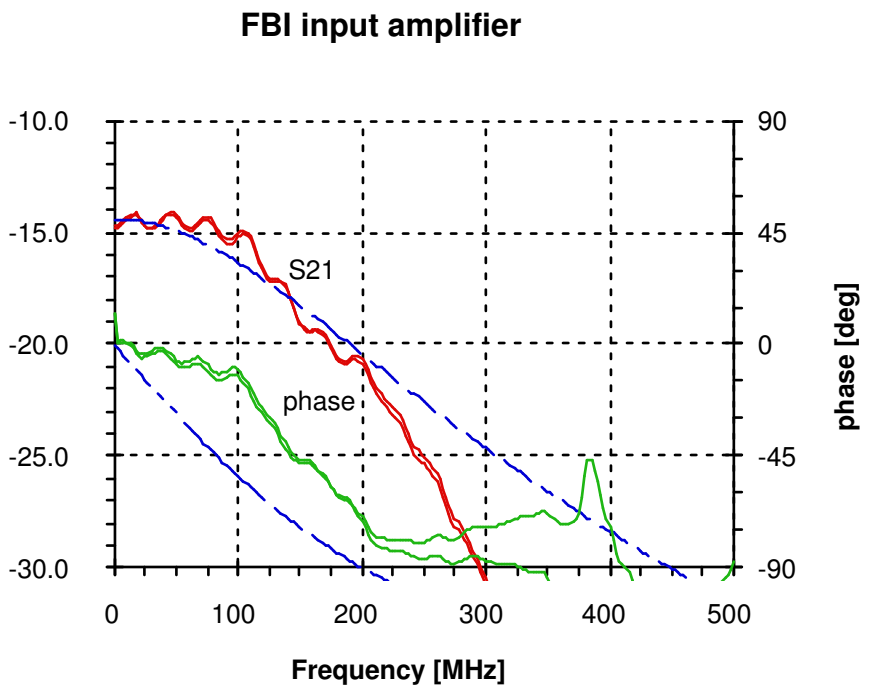



**Figure 4**. Fast integrator response (Signal Monitor/Input Signal) for the p and pbar fast integrator modules . The dotted traces indicate values for a simple 2nd order 200MHz low pass filter.

The most tenuous part of the FBI calibration is the actual integrator section. The 45Ω switch impedance, the 200Ω 1% resistor, and the 82pf 2.5% polystyrene capacitor set the integrator time constant to about 20nsec. Careful analysis of the AD8055 op-amp suggests an accurate response with a 500MHz bandwidth is obtainable, figure 5. The switches require 2nsec to switch into and out of the integrate mode. An output buffer following the integrator has variable gain to obtain the desired response. The 2KHz low frequency limit of the integrator is below the 3KHz ac coupling in the RWM.

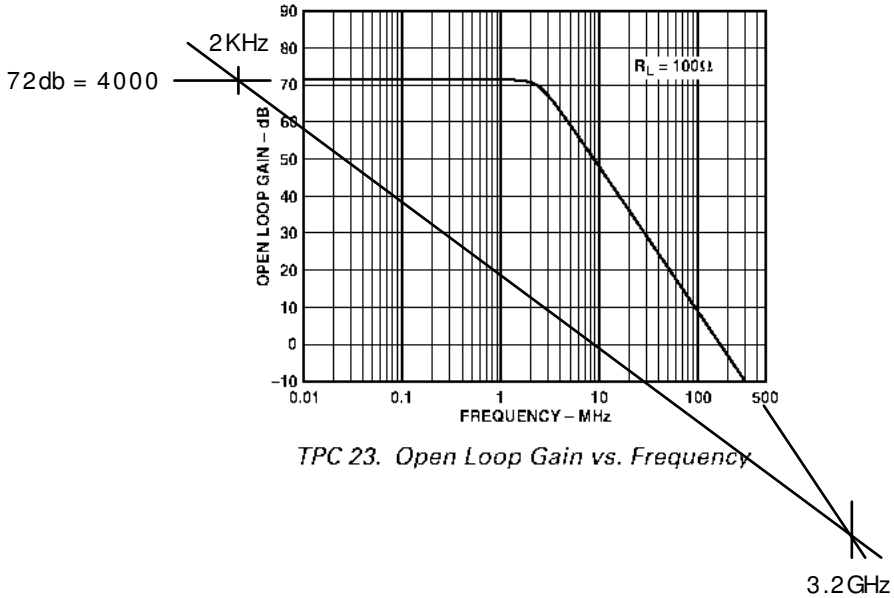



**Figure 5**. The AD8055 open loop gain with the desired closed loop integrator response superimposed (gain=1 at 8MHz). The open loop gain must be larger than the desired closed loop gain.

| **Table 2**. Tev FBI system gain: | | db@1GHz | Ω at dc |
|---|---|---|---|
| Rgap | 1.288Ω | | |
| output cable loss 319.4ns | | -.78 | .197 |
| 4 feet 1/2" heliax 4ns | | -.09 | .005 |
| resistive splitter 1:3 | 1/3 | | |
| 8 feet 1/4" superflex 9ns | | -.33 | .090 |
| resistive splitter 1:2 | 1/2 | | |
| fast integrator | 2 | | |
| integrator time constant | 20ns | | |
| e | 1.60210e-19 | | |
| N | 1e9 protons | | (.291Ω total) |

Tev

$$\frac{Volt}{e9} = 1.288\Omega \frac{Ne}{20ns}\frac{1}{3}\frac{1}{2}2$$
$$= .003439$$
$$or$$
$$291\frac{e9}{Volt} \quad at\ the\ integrator\ output$$

| **Table 3**. Tev FBI system gain: | | db@1GHz | Ω at dc |
|---|---|---|---|
| Rgap | .9727Ω | | |
| output cable loss 131ns | | -1.69 | .092 |
| resistive splitter 1:2 | 1/2 | | |
| 3db attenuator | .7079 | | |
| 8 feet RG58 | | -1.60 | .119 |
| resistive splitter 1:2 | 1/2 | | |
| integrator time constant | 20ns | | |
| e | 1.60210e-19 | | |
| N | 1e9 protons | | (.291Ω total) |

MI

$$\frac{Volt}{e9} = .9727\Omega \frac{Ne}{20ns}\frac{1}{2}(.7079)\frac{1}{2}2$$
$$= .002754$$
$$or$$
$$363\frac{e9}{Volt} \quad at\ the\ integrator\ output$$

$$for\ a\ 2ns\ \sigma_t\ gaussian\ bunch$$
$$I(t) = \frac{1}{\sigma_t\sqrt{2\pi}} e^{-\frac{1}{2}\left(\frac{t}{\sigma_t}\right)^2}$$
$$Ipeak = \frac{1}{\sigma_t\sqrt{2\pi}}\int_{-\infty}^{\infty} I\,dt = \frac{Ne}{\sigma_t\sqrt{2\pi}} = 32\frac{mA}{e9}$$
$$at\ the\ fast\ integrator\ input$$
$$Vpeak = Ipeak 1.288\Omega\frac{1}{3}\frac{1}{2} = 6.9\frac{mV}{e9}$$

The dc cable resistance forms a voltage divider with the 50Ω load and will contribute a 0.5% effect. It's important to note that frequency dependent cable losses contribute to dispersion which can make significant errors, discussed later. For now, if the integrating window is sufficiently long, all of the signal current entering the cable at one end will come out the other.

The CLC409 input amplifiers used in the fast integrator module have a ±3.5V maximum output. For a 2nsec sigma t bunch this corresponds to 255e9 maximum intensity. The AD8055 amplifier used for the integrator section has a ±3.1V output range. The Comet digitizers have a ±1V full scale range or about 300e9 maximum. AC coupling and background subtraction extend the measurable range by up to 36% depending on the number of bunches in the machine.

The 12 bit Comet digitizers use offset binary scaling with the number left shifted by 3 bits (multiplied by 8). The input voltage for a binary number bn is:

$$\begin{aligned} Vin &= \frac{bn - 8\times 2048}{8\times 2048} \quad 0 \le bn \le 8\times 4096 \\ &= bn(.000061035) - 1 \end{aligned}$$

There are no gain and offset adjustments on the comet digitizer cards. An offset in the front end program is used to make each of the comet cards read zero with a 50Ω terminator on their input. The gain of the fast integrator is adjusted with a calibration pulse to compensate for the comet card gain. Final adjustments are made with beam to make the wide gate sum (C:FBIPWG) with a 9 bucket gate width agree with the dc beam current (T:IBEAM). The narrow gate reading (C:FBIPNG) is then made to agree with the wide gate with the gate width set to 1 bucket.

note: The offsets and gains cannot be corrected in the ACNET data base because the summation is done in the front end. The summation and individual readings are returned in a common array whose elements all use the same ACNET scaling.

The FBI can be accurately calibrated without beam using the AWG. After going through this procedure we used the Tektronix Arbitrary Waveform Generator (AWG) to calibrate the Tev proton FBI. The AWG duplicated the exact bunch spacing with 18.8 nsec wide square pulses. The amplitude was varied to simulate changing intensity. Excellent agreement (0.1%) between the beam derived scale factor and the AWG calibration demonstrates the accuracy of this approach.

The FBI scale factor used is 306 e9/Volt which is close to the estimated 291 e9/Volt. The 4.8% difference is likely required to compensate for a gain errors in the comet digitizer card and/or the fast integrator module.

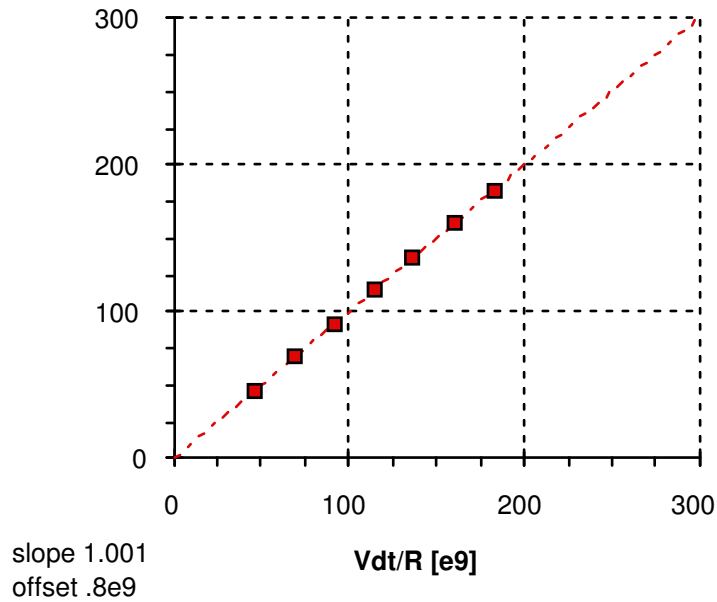



**Figure 6.** FBI measurements compared to Arbitrary Waveform Generator. R=1.288/3

**LIMITED BANDWIDTH**

Frequency dependent effects such as limited bandwidth and cable dispersion will make the FBI or SBD measurements depend on bunch shape. The RWM itself is very flat (±1db) from 3.0KHz to 3GHz as measured in the lab. The resistive splitters used to divide the signal are equally good. The output cable, however, has 4.26db of attenuation at 1GHz and changes with the square root of frequency.

| Tev cable | ft | db | MI cable | ft | db |
|---|---|---|---|---|---|
| 7/8" heliax | 278 | 3.64 | 7/8" heliax | 129 | 1.69 |
| 1/2" heliax | 4 | .09 | RG58 | 8 | 1.60 |
| 1/4" superflex | 8 | .53 | | | |
| db@1GHz | | 4.26 | | | 3.29 |

$$Tevatron\ [db] = 4.26\sqrt{\frac{f}{1GHz}}$$

This attenuation leads to a frequency dependent phase shift that in cables is referred to as dispersion. It will attenuate, stretch, and add tails to fast transient signals such as gaussian shaped bunches.

An important characteristic of cable dispersion is that the resistance between the center and outer conductors is very large. As a consequence, any charge injected at one end can be extracted from the other. An accurate bunch intensity can be obtained provided the integration window is sufficiently long so as to include the tail caused by dispersion. The same line of reasoning applies to any linear system. The equation below demonstrates the relevant property of fourier transforms where G(f) is the fourier transform of g(t).

$$\int_{-\infty}^{\infty} g(t)dt = G(0)$$

The "wide gate" integration window is adjustable in rf clock ticks (or integer multiples of 18.83nsec) and is typically set to 9 buckets. An RC time constant on the board is used to set the "narrow gate" integration window to 18.8nsec. They are tuned to within 100psec and appear to be quite stable. Figures 7 through 10 below indicate cable dispersion is a 5% effect for the narrow gate measurement and a 2% effect for the 9 bucket wide measurement. The 200MHz bandwidth of the fast integrator module is a 0.03% effect.

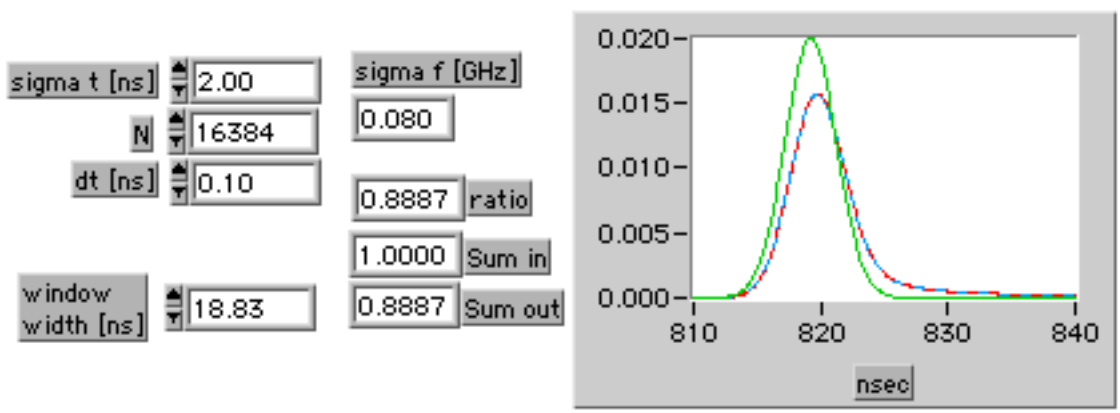



**Figure 7**. Output from LabView program used to calculate the effect of dispersion. Total cable attenuation used was 9.18db at 1GHz.

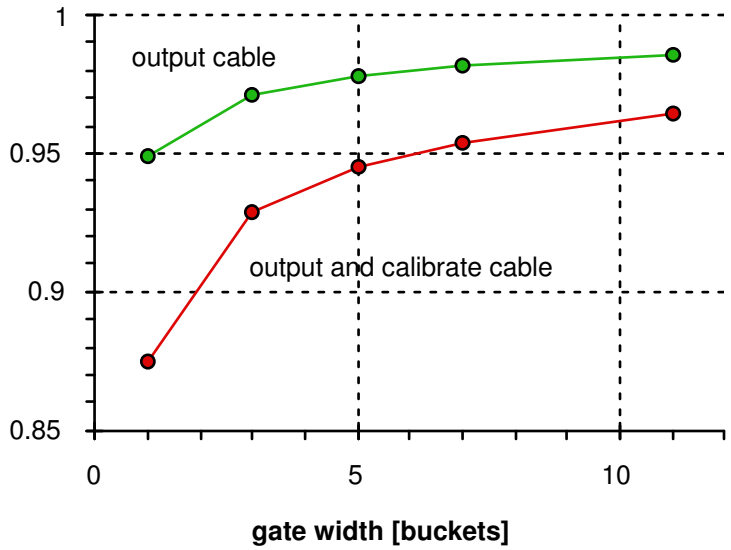



**Figure 8**. Calculated effect of cable dispersion on a 2nsec sigma t bunch for different integration gate widths. The effect is calculated for the output cable alone and for the output and calibrate cables in series.

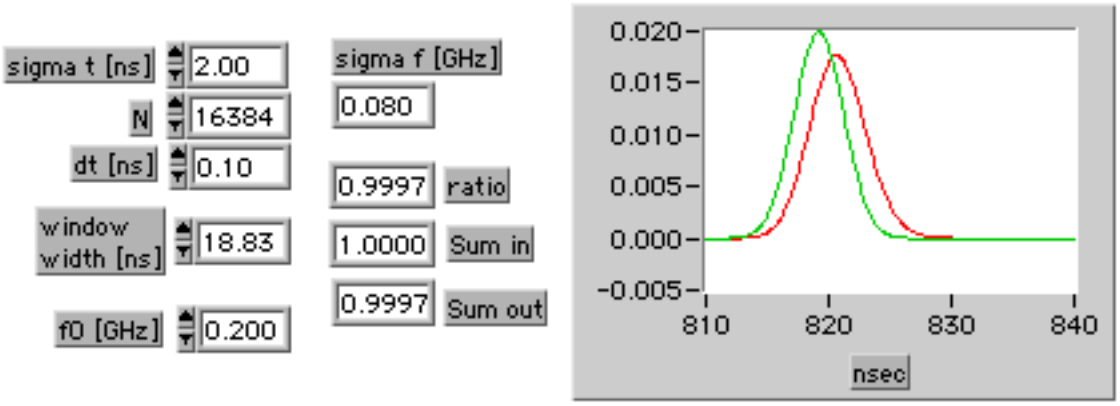



**Figure 9**. Output from a LabView program used to calculate the effect of a 200MHz 2nd order low pass filter. The bandwidth was chosen to match the fast integrator. The ratio of output to input area is .9997.

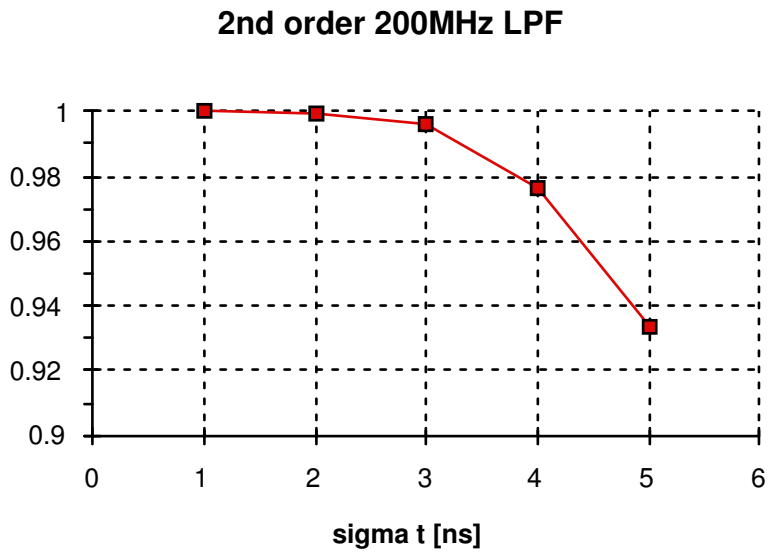



**Figure 10**. Effect of a 2nd order 200MHz low pass filter versus bunch length using a gate width of 1 bucket, 18.83nsec. The error was negligible for longer gate widths.

**BACKGROUND SUBTRACTION**

Because the RWM is ac coupled to the beam, the signal must average to zero under steady state conditions. To correct the measurement, the background level between bunches is integrated and subtracted from the measured bunch intensity. The ac coupling frequency is determined by the inductance of the ferrite loaded RWM and the gap resistance. The 3.0KHz ac coupling frequency has a time constant of 53usec or 2.5 turns. The accuracy of background subtraction depends on the loading pattern and timing of the background measurement. With a single bunch in the Tevatron, the background level will change by 33% in one turn.

For the 36 on 36 mode of operation in the Tev, the FBI samples odd bunches (1 through 35) and even bunches (2 through 36) on alternate turns. The background (bunch 37) is measured 42 buckets after bunch 36. The background level will decay by 1.5% in 42 buckets but the background level is only 3% (36/1113) of the bunch level. Pbar bunches are sampled simultaneously with a separate fast integrator module.

The effects of background are estimated below for the ideal case: square bunches and a low frequency time constant much longer than one turn.

| | |
|---|---|
| A | amplitude above background |
| dt | bunch width |
| Adt | true bunch intensity |
| MI | measured bunch intensity without background subtraction |
| BG | background level |
| GW | integration gate width |
| BGGW | measured "intensity" of background sample |
| N | number of bunches |
| T | rotation period (1113 buckets) |

$$BG = -N\frac{Adt}{T}$$

$$BGGW = -N\frac{Adt}{T}GW$$

$$MI = Adt\left[1 - N\frac{GW}{T}\right]$$

$$Adt = MI - BGGW$$

The error without background correction will be .090% for a single bunch measured with a one bucket gate width. However, the error would be 36% for 36 bunches using an 11 bucket gate width. The presence of pbar bunches will increase the error in proportion to total charge.

The accuracy with which background effects can be corrected depends on the low frequency corner and when background is actually sampled. Three groups of 12 proton bunches are loaded into the Tevatron one bunch at a time. The bunch spacing is 21 buckets between bunches and 140 buckets between groups of 12. The largest error during loading with background subtraction enabled is .32% and occurs when only the first 24 bunches are loaded. PBar bunches are subsequently loaded into the same pattern 4 bunches at a time.

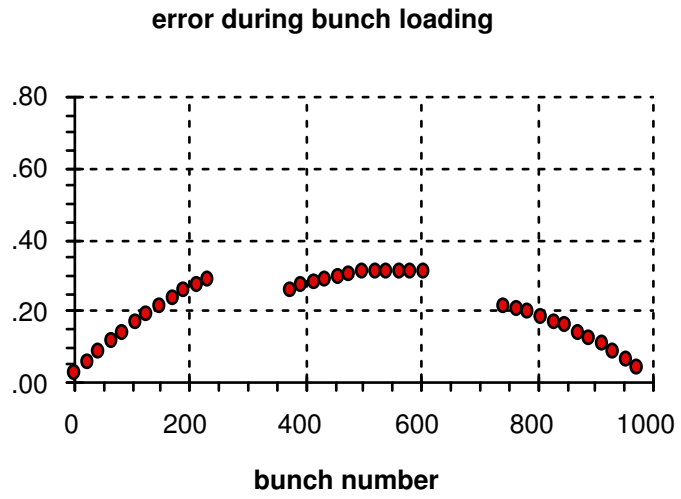



**Figure 11**. Calculated percentage error during Tevatron loading caused by ac coupling with background subtraction enabled. The graph indicates the error while the bunches to the left are the only ones in the machine.

**NOISE**

The front end of the fast integrator module consists of two CLC409 op-amps in series. The first one is configured with a gain of two and the second has unity gain. Together they form a 200MHz 2nd order low pass filter. The AD8055 is used to make an integrator with a unity gain frequency of 8MHz or 20nsec time constant. The low frequency gain of the integrator is limited by the open loop dc gain of the AD8055 to about 4000. The equivalent noise density at the integrator input from the 50Ω resistor, two CLC409's, the AD8055, and five 250Ω resistors is shown below with the estimated thermal noise from the RWM.

| device | nV/srHz |
|---|---|
| 50Ω | .9 |
| CLC409 | 2.2 |
| AD8055 | 6 |
| 250Ω | 2.0 |

*at fast integrator input*

$$V_{module} = \left[ \left( \frac{.9nV}{\sqrt{Hz}} \right)^2 + \left( \frac{2.2nV}{\sqrt{Hz}} \right)^2 + \left( \frac{1}{2} \frac{2.2nV}{\sqrt{Hz}} \right)^2 + \left( \frac{1}{2} \frac{6nV}{\sqrt{Hz}} \right)^2 + 2\left( \frac{2nV}{\sqrt{Hz}} \right)^2 + 3\left( \frac{1}{2} \frac{2nV}{\sqrt{Hz}} \right)^2 \right]^{\frac{1}{2}} = 5.2 \frac{nV}{\sqrt{Hz}}$$

*RWM thermal noise*

$$V_{RWM} = \sqrt{4KTBZ} = .15 \frac{nV}{\sqrt{Hz}}$$

$$K = 1.38 \times 10^{-23}$$

$$T = 300°$$

$$Z = 1.288\Omega$$

The total noise at the output of the integrator is calculated below. The integrator has a gain of 4000 from dc to 2KHz. Above 2KHz, the gain will be fo/f where fo=8MHz. The output bandwidth will be limited somewhat by the sin(x)/x response imposed by the integrator gate width but most of the noise of the integrator will not be affected. Noise from the unity gain output buffer is insignificant.

*at fast integrator module output*

$$V_{Nout} = 2\frac{5.2nV}{\sqrt{Hz}} \left[ \int_0^{f_0/4000} 4000^2 \, df + \int_{f_0/4000}^{\infty} \left( \frac{f_0}{f} \right)^2 df \right]^{\frac{1}{2}}$$

$$= 2\frac{5.2nV}{\sqrt{Hz}} \left[ 4000 f_0 + 4000 f_0 \right]^{\frac{1}{2}} \left( f_0 = 8MHz \right)$$

$$= 2.6mV\,RMS$$

$$for 292 \frac{e9}{Volt}$$

$$= .77e9RMS$$

The measured noise at the output of the fast integrator module does not change when the input is disconnected. A 20db (gain of 10) amplifier with a noise figure of 1.2db is used in front of the pbar fast integrator module. The output noise is reduced as is the full scale output, figure 12.

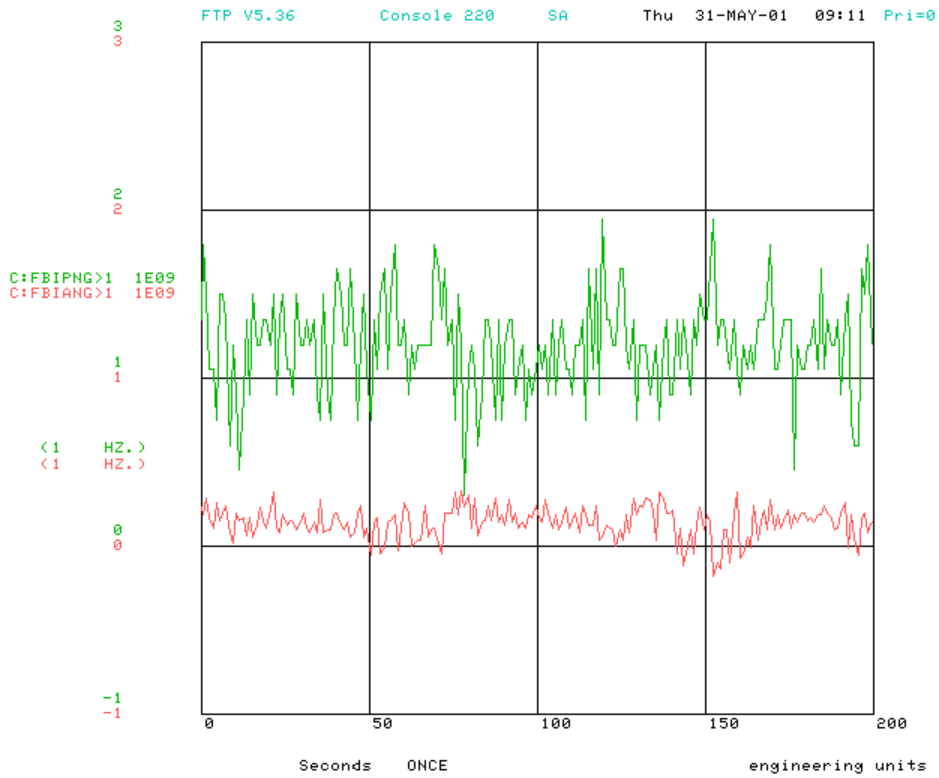



**Figure 12**. Typical FBI noise levels for proton and pbar channels (C:FBIPNG[1], C:FBIANG[1]). Proton channels have a full scale output of 306e9. Pbar channels have a 20db amplifier, a full scale output of 30.6e9, and correspondingly less noise.

**CONCLUSION**

The RWM response can be accurately measured through the calibrate and signal cables. An Arbitrary Waveform Generator is required to duplicate correct bunch spacing. The frequency dependent cable loss or dispersion can be calculated and matches the manufacturers specifications very well up to 1GHz. Cable dispersion has a significant effect on the calibration of the RWM by causing gain error that depends on integrating gate width. This could be erroneously interpreted as beam in neighboring buckets by the unwary user.

The best way to check the calibration of the FBI is to compare the wide gate sum with the Tev DCCT reading (T:IBEAM) as bunches are loaded. This can be automated with the "Lumber Jack" data logger. Lumber Jack supports least squares fitting between the two readings. Once the wide gate channel has been calibrated, the gate width can be reduced in steps watching for gate width dependence. The narrow gate reading can then be made to agree with the wide gate reading with a one bucket gate width.

The 200MHz bandwidth of the fast integrator module has very little effect on accuracy. The integrated amplifier output is surprisingly independent of bunch length provided all of the bunch charge fits within the gate width.

Background subtraction is required to accommodate ac coupling of the signal in the RWM. The accuracy of background subtraction depends on bunch spacing and timing of the background measurement but those effects are small.

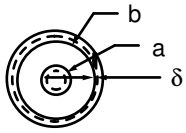
Keep in mind that the integration window is centered only for gate widths of odd numbers of buckets. Even gate widths are made by adding a bucket to the end of the gate.

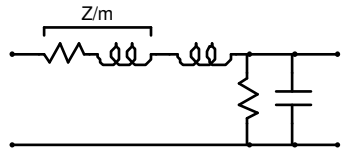
The Tev SBD system uses a LeCroy 9374L digital oscilloscope.

| | |
|---|---|
| 1GHz | 3db bandwidth |
| 2GHz | sample rate (with adapter) |
| 8bit | resolution |
| 1% | gain and offset accuracy |
| 10ppm | time base accuracy |

| | |
|---|---|
| HP 3 way resistive splitter | HP 11850C |
| Mini Circuits 2 way resistive splitter | ZFRSC-2050 |

## Appendix A. Skin effect losses in cables
**Jim Crisp**




$$Z_0 = \sqrt{\frac{R + j\omega L}{G + j\omega C}} \qquad \gamma = \alpha + j\beta = \sqrt{(R + j\omega L)(G + j\omega C)}$$

$R \ll \omega L$ *and* $G \ll \omega C$ (*high frequency approximation*)

$$Z_0 \approx \sqrt{\frac{L}{C}} \qquad \alpha \approx \frac{R}{2Z_0} + \frac{G}{2}Z_0 \qquad \beta \approx \omega\sqrt{LC} = \frac{\omega}{\upsilon_p}$$

skin depth in a good conductor: $\delta = \sqrt{\frac{2\rho}{\mu\omega}}$

The inductive impedance equals the real impedance in a good conductor.

$$Z/m \approx \left(1 + \frac{b}{a}\right)\frac{\rho}{\pi b \delta}(1 + j) = R/m\,(1 + j)$$

$$\omega L/m \approx R/m + \omega L_0/m \qquad L_0 = \frac{Z_0}{\upsilon_p} \qquad C/m \approx \frac{1}{Z_0 \upsilon_p} \qquad G/m \approx 0$$

or $\quad R/m \approx 2\alpha Z_0 \sqrt{f/f_0} \quad$ *for* $\alpha$ *measured at* $f_0$

$$\boxed{\begin{matrix}\alpha \\ \beta\end{matrix} = \left[\frac{1}{2}\left(\sqrt{(R^2 + \omega^2 L^2)\omega^2 C^2} \begin{matrix} - \\ + \end{matrix} \omega^2 LC\right)\right]^{1/2}}$$

$$\frac{Vout}{Vin} = e^{-\gamma d}\ mag = e^{-\alpha d}\ ang = \left(\frac{\omega}{\upsilon_p} - \beta\right)d$$

This model works very well at frequencies that have a skin depth less than the thickness of the material used to make the transmission line. In copper, the skin depth at 7KHz is 1/32". Below this frequency limit, the losses are dominated by the voltage divider formed with the dc resistance of the conductors and the impedance of the load. The cable loss will never be less than the loss from the dc resistance.

## Appendix B. Tev and MI RWM parameters

**Tev**

Rgap 1.288Ω
output cable 319 nsec 7/8" heliax
calibrate cable 319 nsec 1/2" heliax
signal path: 4.26 db @1GHz
RWM
319ns 7/8" heliax HP 11850C dc-3GHz
4ft 1/2" heliax
1:3 resistive splitter (FBI, MDAT FBI, SBD) ZFRSC-2050
8ft 1/4" superflex
1:2 resistive splitter (p and pbar FBI)
20db amplifier before pbar integrator

Tev FBI

$$\frac{Volt}{e9} = 1.288\Omega \frac{Ne}{20ns}\frac{1}{3}\frac{1}{2}2$$
$$= .003439$$
$$or$$
$$291\frac{e9}{Volt} \quad at\ the\ integrator\ output$$

**MI**

Rgap .9727Ω
output cable 131 nsec 7/8" heliax
calibrate cable 131 nsec 1/2" heliax
signal path: 3.29 db @1GHz
RWM
1:2 resistive splitter (FBI, SBD) Weinschel WA1506A SN 455
3ns RG58
3db Texscan attenuator
4ns RG58
1:2 resistive splitter (p and pbar FBI) Minicircuits ZFRSC-2050
1ns RG58

MI FBI

$$\frac{Volt}{e9} = .9727\Omega \frac{Ne}{20ns}\frac{1}{2}(.7079)\frac{1}{2}2$$
$$= .002754$$
$$or$$
$$363\frac{e9}{Volt} \quad at\ the\ integrator\ output$$