

1.0 Beam Phase Cavity

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Note: FS is Full Scale, fS is femtosecond, F is Fahrenheit - context supercedes preceding definitions.

Purpose

The beam phase cavity will be used to measure the timing of the beam compared to the RF phase. The beam timing is dominated by the timing of the laser onto the cathode of the RF gun. The results of the measurement will be used to correct timing of the laser, which is expected to drift outside LCLS tolerances.

Specification

The specifications for the beam timing measurement are as follows.

Short term (2 second) timing jitter: 100fS rms

Long term (4 day) timing jitter: ± 1 pS

Data available at 120Hz

Control interface will be EPICS running on RTEMS

System Topology

The system will use a copper cavity at 2805MHz. The frequency is chosen to be 25.5MHz below the LO frequency of 2830.5MHz. The cavity frequency is 51MHz below the RF frequency of 2856MHz. This will suppress the signal from dark current off the gun by 47dB, see section on Dark Current Errors. The cavity signal of 2805MHz will mix with the LO frequency to give an IF frequency of 25.5MHz this frequency will be amplified, filtered and digitized as outlined in figure 1 below.

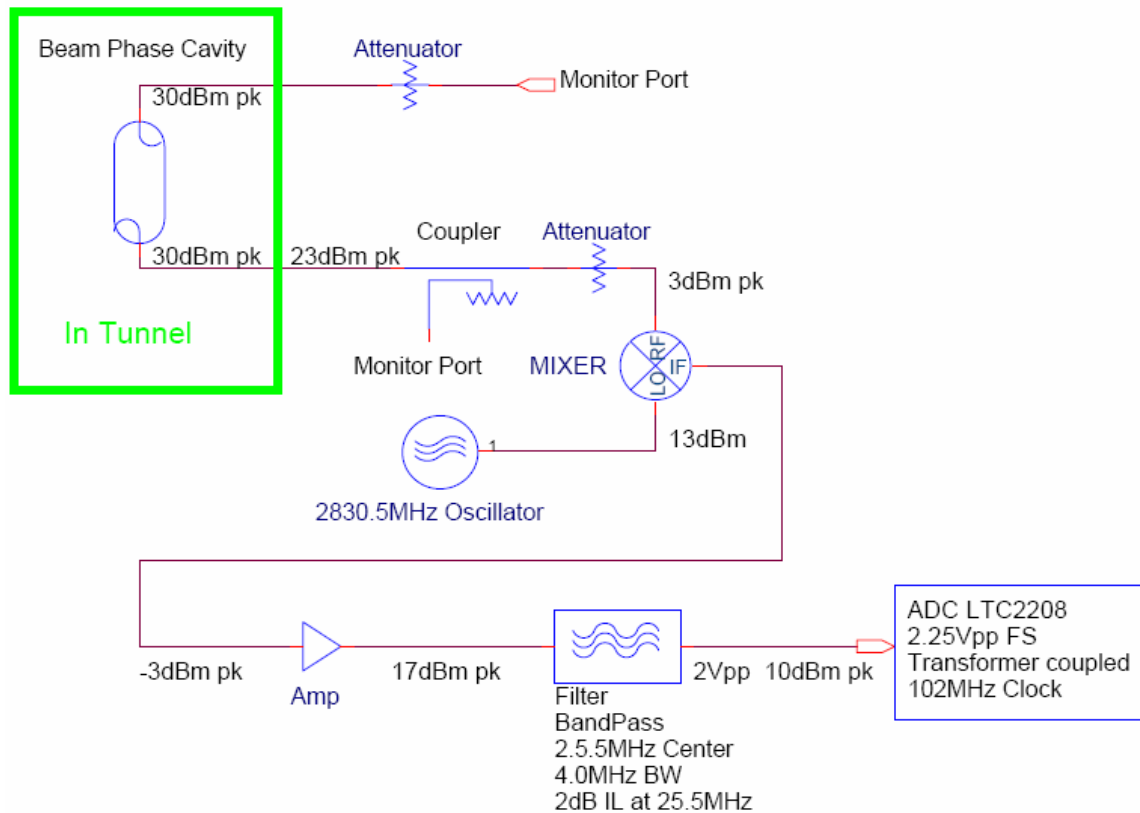


Figure 1. Phasing Cavity System Outline

2.0 Electronics

SNR

The ADC (LTC2208) has a SNR of about 76dB, which is the dominating noise source in the system. The SNR of 76dB has been measured in on a demo board in the lab.

The noise floor of the RF signal from the cavity is about -174dBm/Hz . The noise floor in the reference system, LO, is predicted to be $< -143\text{dBc/Hz}$ at 2830.5MHz. At 13dBm this is -130dBm/Hz . The LO noise floor is 44dB higher than the cavity signal noise floor and thus dominates the RF signal noise levels.

A 4MHz IF bandpass filter will give a SNR of 77dB on a signal with a noise floor of -143dBc/Hz . By use of a 4MHz IF bandpass filter the noise level from the signal will be about the same as the ADC noise level. The filter will reduce the rise time of the beam induced cavity signal to the order of 80nS. The noise levels of the system outlined in figure 2.

ADC Clock Jitter

The ADC clock allowable jitter is calculated from:

$$\text{SNR} = -20\text{Log}(2\pi \times 25.5\text{MHz} \times T_{\text{jitter}})$$

From this equation 1pS rms clock jitter gives a SNR of 76dB. This amount of clock jitter will double the noise level at the input to the ADC. The 102MHz clock jitter is expected to be on the order of 100fS rms jitter. This corresponds to a SNR of 96dB. The clock jitter will not add significantly to the noise levels.

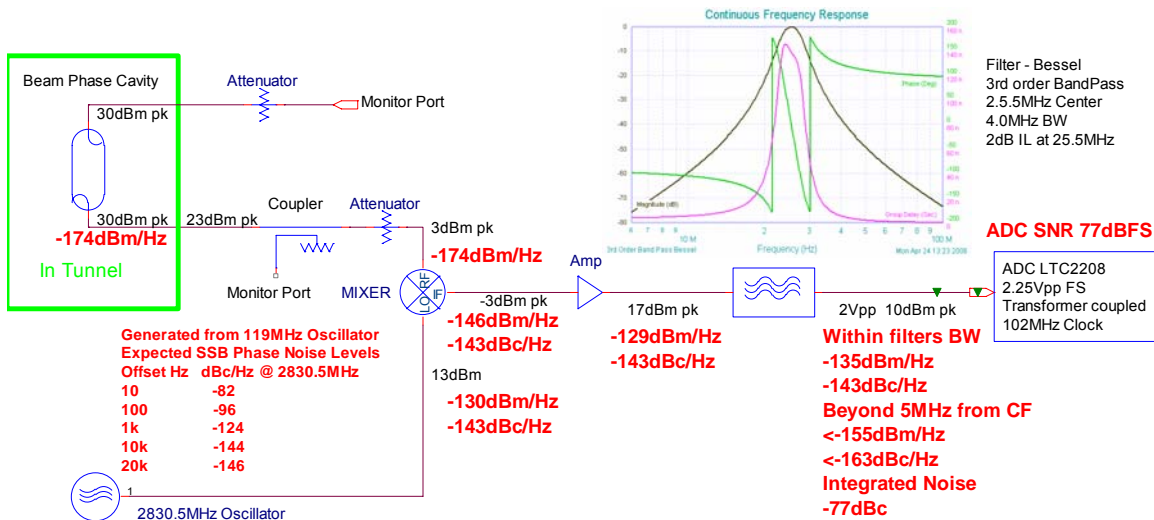


Figure 2. System Noise Levels

Temperature Drifts

Temperature Coefficient of the cavity is 28kHz/°F (50kHz/°C). The specified water temperature stability is $\pm 0.1^\circ\text{F}$. The frequency change for this temperature variation is $\pm 2.8\text{kHz}$. A 2.8kHz frequency shift will cause a phase change of 1 degree in 1uS.

Phase Stable 1/2 inch superflex Heliac cable will be used from the electronics to the cavity. The cable is 80 feet in length and runs from the temperature stabilized RF Hut, down a penetration, through a shield wall, and to the cavity in the injector tunnel, Figure 3. The temperature coefficient of the cable is listed as -1 to $+3\text{ppm}/^\circ\text{F}$. For this analysis a temperature coefficient of $4\text{ppm}/^\circ\text{F}$ is used. The cable will be run with water stabilized lines for most of the run. The specification for the water temperature is $\pm 0.1^\circ\text{F}$. About 10% of the cable might be exposed to temperatures in the RF Hut and injector area of $\pm 1^\circ\text{F}$. Eighty feet of cable is $1\text{e}6$ degrees S-band and a $\pm 0.1^\circ\text{F}$ temperature variation will cause a $\pm 0.4^\circ$ phase change, or $\pm 400\text{fS}$. If the cables near the connectors are not stabilized well this number could easily double to $\pm 800\text{fS}$.

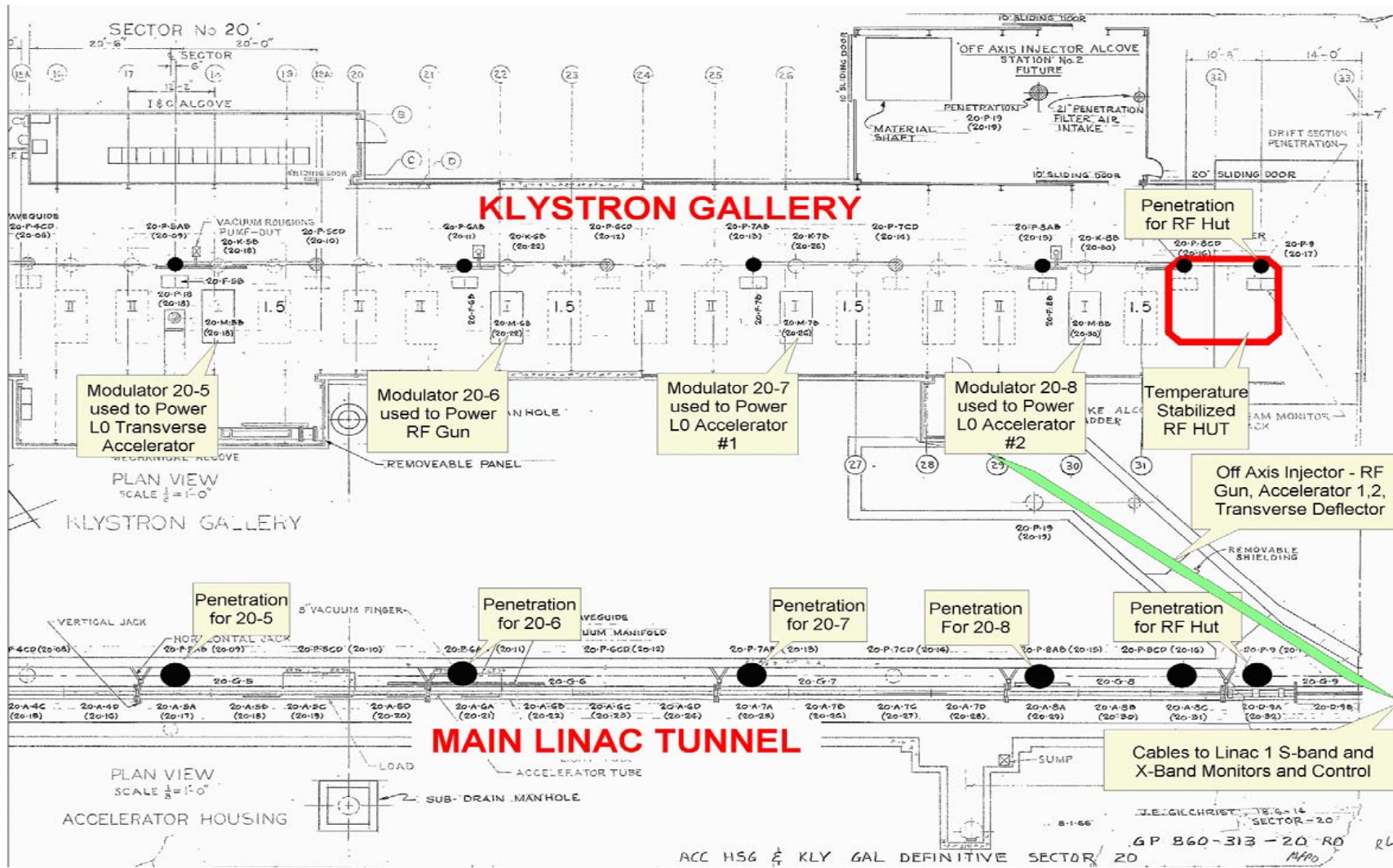


Figure 3. RF HUT with respect to LCLS Injector

Digital Data Processing

The 25.5MHz signal will be digitized at 102MHz. The 4MHz bandwidth filter before the ADC will dominate the rise time and the cavity Q will dominate the fall time of the signal. Figure 4 is an example of the digitized waveform.

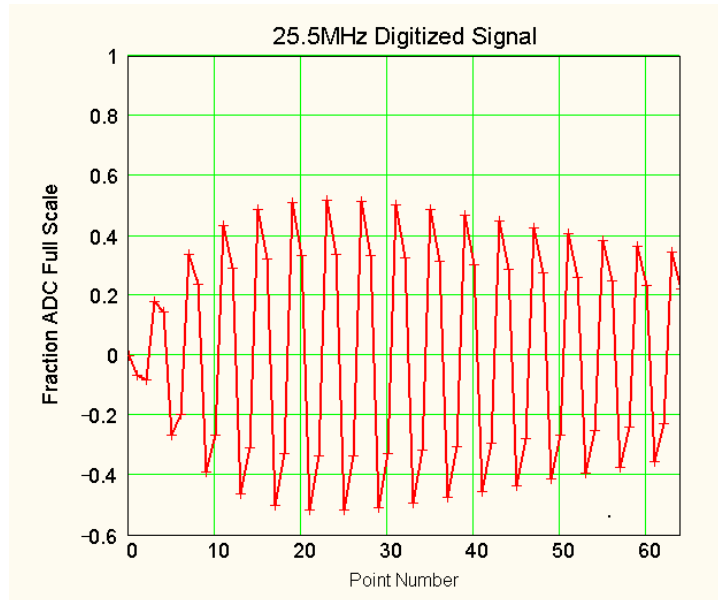


Figure 4. Digitized IF Signal

The signal will then be digitally down converted and normalized by division of the measured/calculated function. The normalization vector will be calculated during calibration. The values of the normalization vector will be such that when multiplied by the I and Q vectors, all I values will be equal and all Q values will be equal. The equation 1 below will be used in down conversion and normalization. The resultant I and Q vectors are shown in figure 5.

Equation 1:

$$I_N := \cos\left[\frac{N \cdot (2\pi)}{4}\right] \cdot \frac{\text{Signal}_N}{\text{Normalization}_N}$$
$$Q_N := \sin\left[\frac{N \cdot (2\pi)}{4}\right] \cdot \frac{\text{Signal}_N}{\text{Normalization}_N}$$

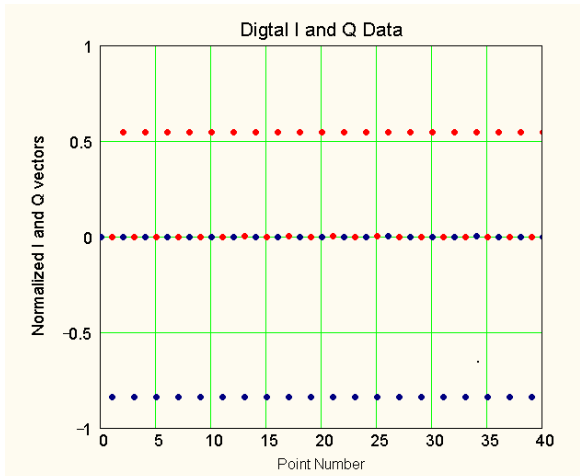


Figure 5. Normalized IQ vectors

The Cosine and Sine functions are a vector of 1s, -1s, and 0s. The normalization vector is a vector of 16bit integers to multiply the signal by.

The beam phase will be calculated from two averaged values of I and Q measurements. The two values of I and Q will determine a slope and offset of the phase over a TBD time period. Errors in the measurement will be evaluated to determine the number of points to average.

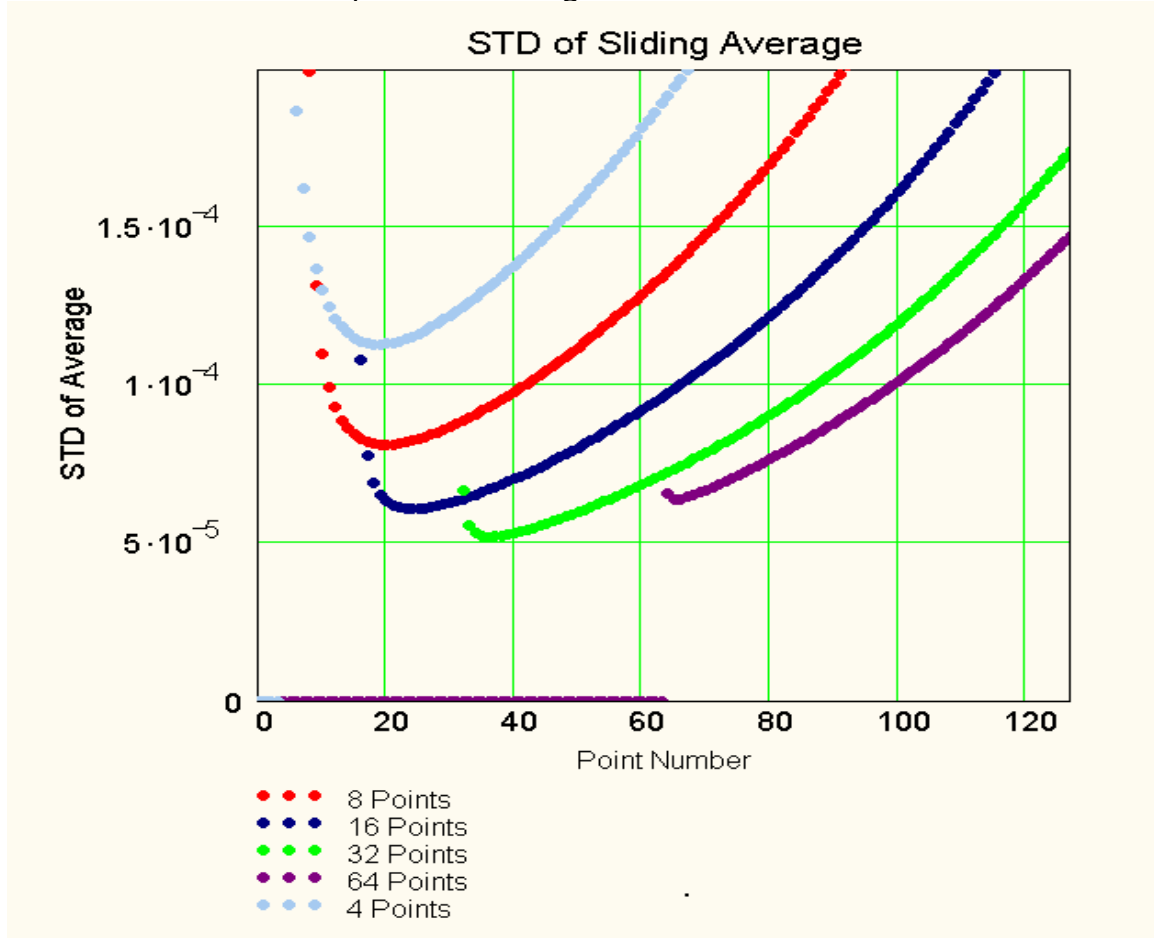


Figure 6. STD of Sliding Averages

Figure 6 is a plot of the standard deviation of a sliding average of the specified number of points along the normalized 25.5MHz signal sampled at 102MHz. The point number on the horizontal axis is the number of the point in the center of the integral. Integrating over several data points will increase the accuracy. The plot shows the lowest error, or STD, to be that of a 32 point average.

The bunch phase will be calculated from two points, which form the equation of a line, in the phase vs time domain, figure 7.

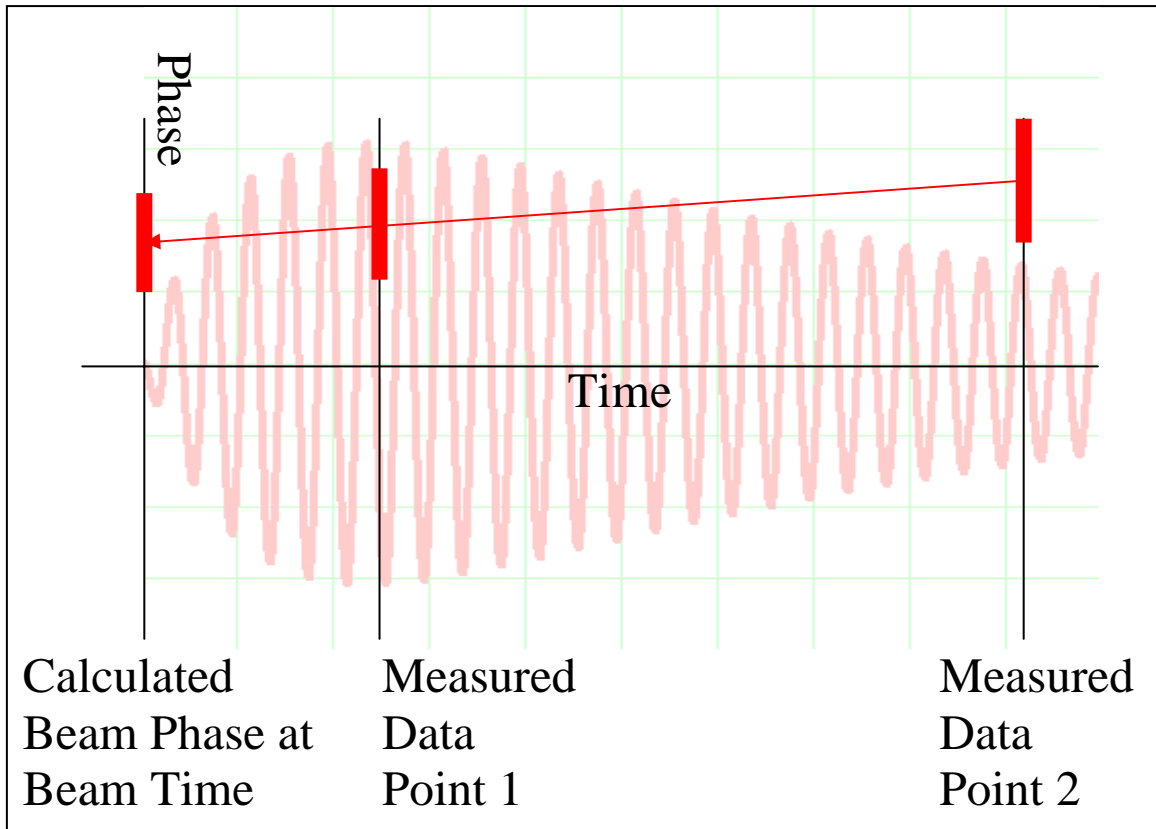


Figure 7. Calculation of frequency and phase from a line through 2 points.

The below equation and figure 8 were used to determine the two points to use and the number of points to average, where n91 is an index for point 1 and pnt2 was varied manually to minimize a point on the plot. The plot is for 8, 16, and 32 point averages. A filter delay of 120nS was also included in the equation. These numbers are expected to change slightly for the real data set. From the plot below, the minimum error is found with an average of 16 data points at point 18 and point 120. The value here is 1.1e-4.

$$\text{PhasErr}_{n91} := \sqrt{\left[\left[\sqrt{(\text{STD}_{n91})^2 + (\text{STD}_{\text{pnt2}})^2} \right] \cdot \left(\frac{n91 + \text{FilterDelay}}{n91 + \text{FilterDelay} - \text{pnt2} + \text{FilterDelay}} \right) \right]^2 + (\text{STD}_{n91})^2}$$

n91 is an index for point 1

FilterDelay is the delay for the filter, 12 points, 120nS

pnt2 is the index at point 2

The index is in 9.8nS steps, 102MHz.

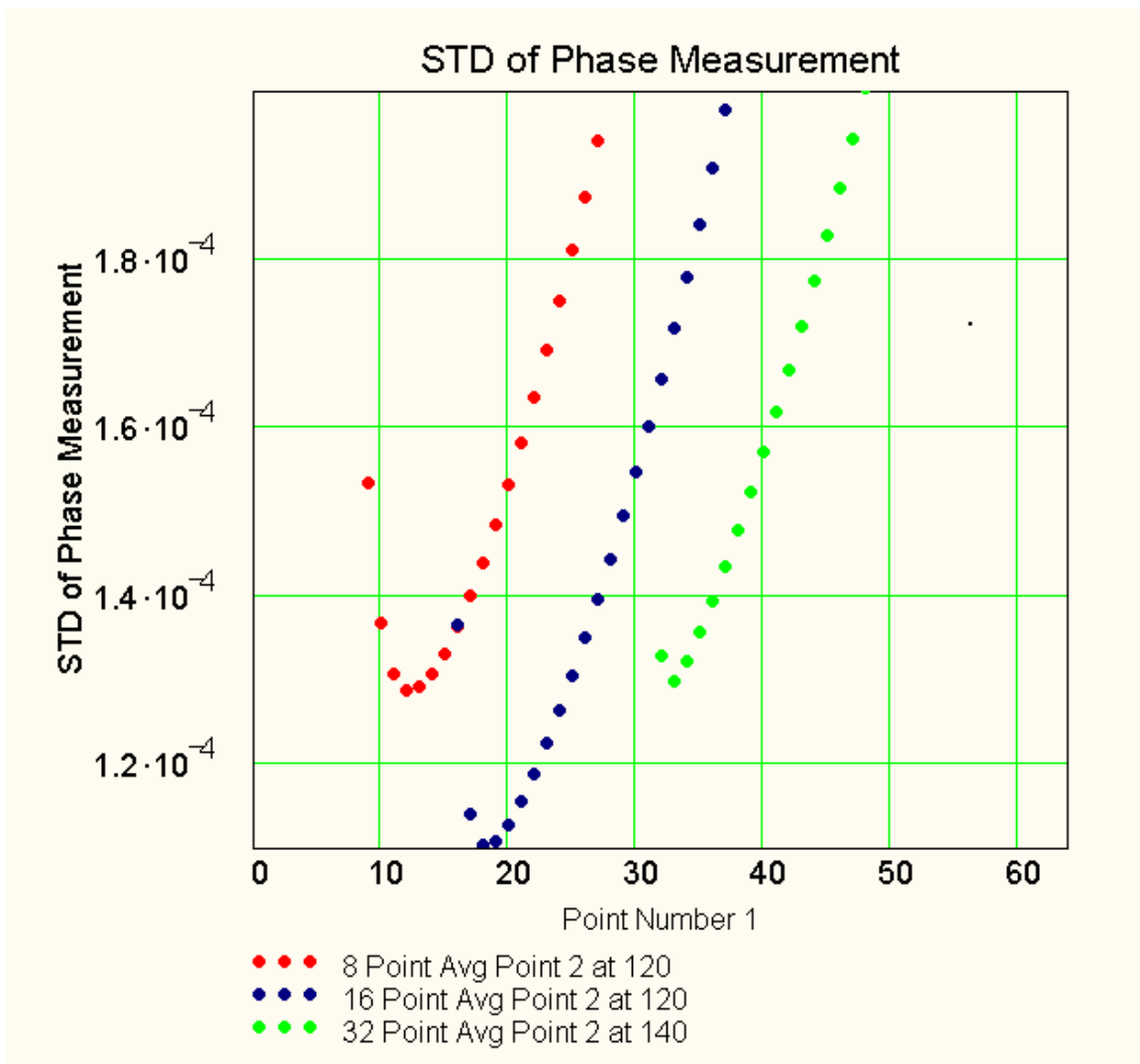


Figure 8. Error in Phase calculation using a 16 point average centered at point 18 for point #1.

Result sensitivities to changes in the data

The above analysis was done with changes to the frequency of the input signal, or cavity frequency, and phase of the signal, actual bunch timing changes, to determine the sensitivity of the result to input changes.

Sensitivity on phase vs cavity frequency is $(0.0006^\circ)/(\text{ppm freq change})$.

Looked at over $\pm 10\text{ppm}$

Sensitivity to beam phase change to calculated phase is 1:1 within STD of measurement.

Looked at over $\pm 10^\circ$

Summary of analysis

STD of measurement from the plot in the last section is 1.1×10^{-4} on a signal normalized to FS. The phase error would then be 0.0063° , or 6.3 fS rms .

If the noise levels at the ADC are four times, an SNR of 70dBc, phase error would then be **12.6fS rms**

Long term drifts due to temperature variations dominated by the cable can be as much as **$\pm 800 \text{ fS}$** .

Comments

Due to the relatively low noise levels it is necessary to keep cables and electronics away from low impedance fields.

3.0 LCLS Phase Detector Cavity Model Measurement and Calculation of the Dimensions for 2805 MHz Mechanical Design

3.1) Measurement on the Model

To design the LCLS bunch phase detector cavity a 1:1 model has been built to measure the basic cavity parameters and estimate the discrepancy between the measured values and “Superfish” simulation results. During the process of cavity design, the resonant frequency initially set for 2856.0 MHz has been changed to 2805.0 MHz in order to suppress the effect of the so called dark current – a 2856 MHz noise signal that would negatively affect the phase measurement precision. The dimensions of the built model are given in the FIG. 9.

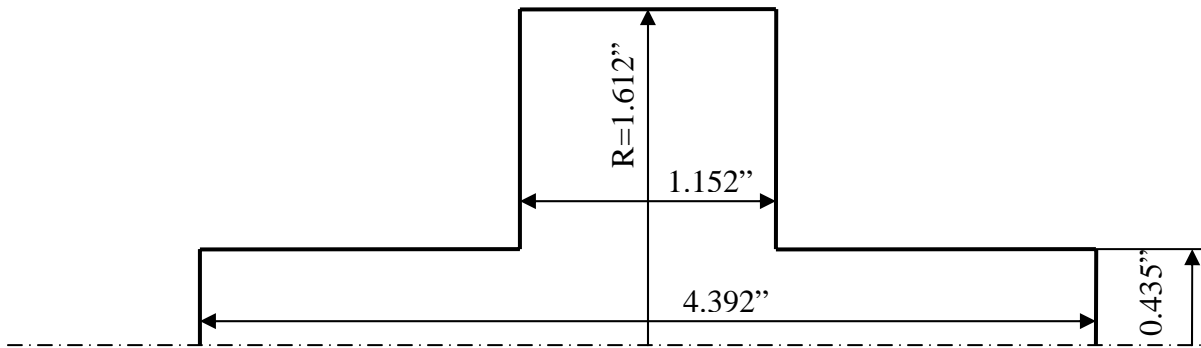


Figure 9. Dimensions of the built 1:1 model

The resonant frequency and unloaded Q of the cavity calculated with the Superfish simulation program and those actually measured on the model are given in the table 1.

TABLE 1 Calculated and measured resonant frequency and unloaded Q

	Superfish simulation	Measured on the model
Resonant frequency [MHz]	2848.810	2847.411
Unloaded Q	13700	12700

There is a difference of ~ 1.4 MHz between the Superfish simulation and the model measurement.

We used the copper cavity model to measure also the frequency vs. temperature dependence. The result is shown in FIG 10. The frequency-temperature dependence was confirmed to be about 50 kHz/°C that is in excellent agreement with similar design cavities.

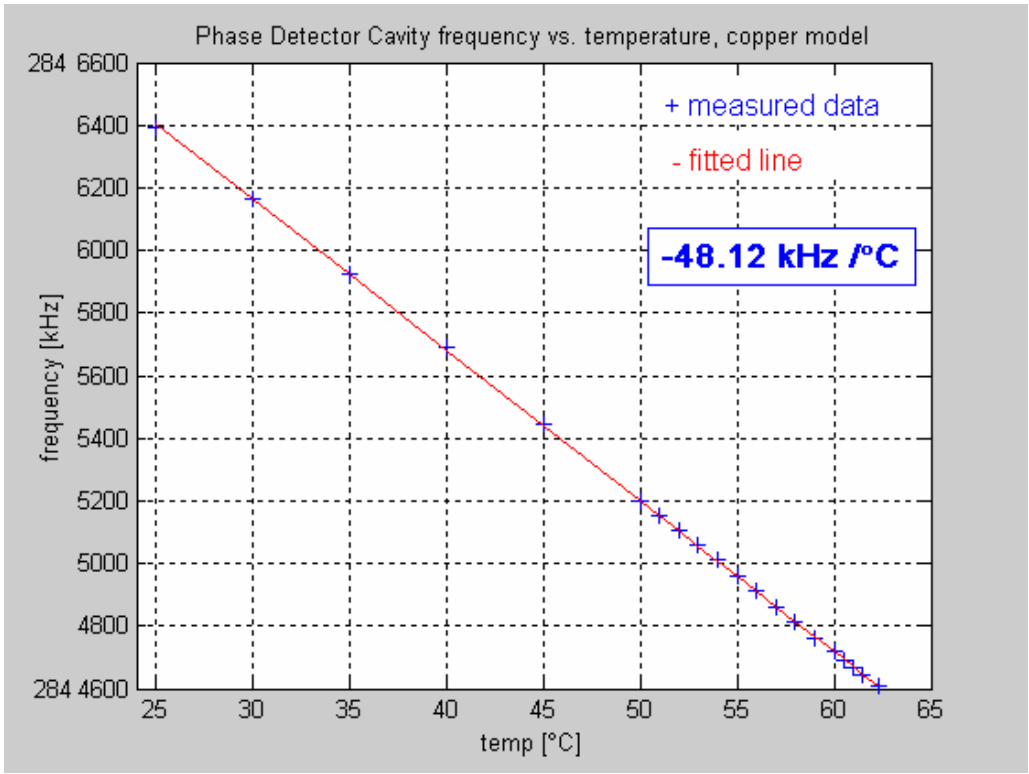


Figure 10. Frequency dependence on the cavity temperature

The final look of the model is in Fig. 11.

Rotatable Coupling Loop, (CL)

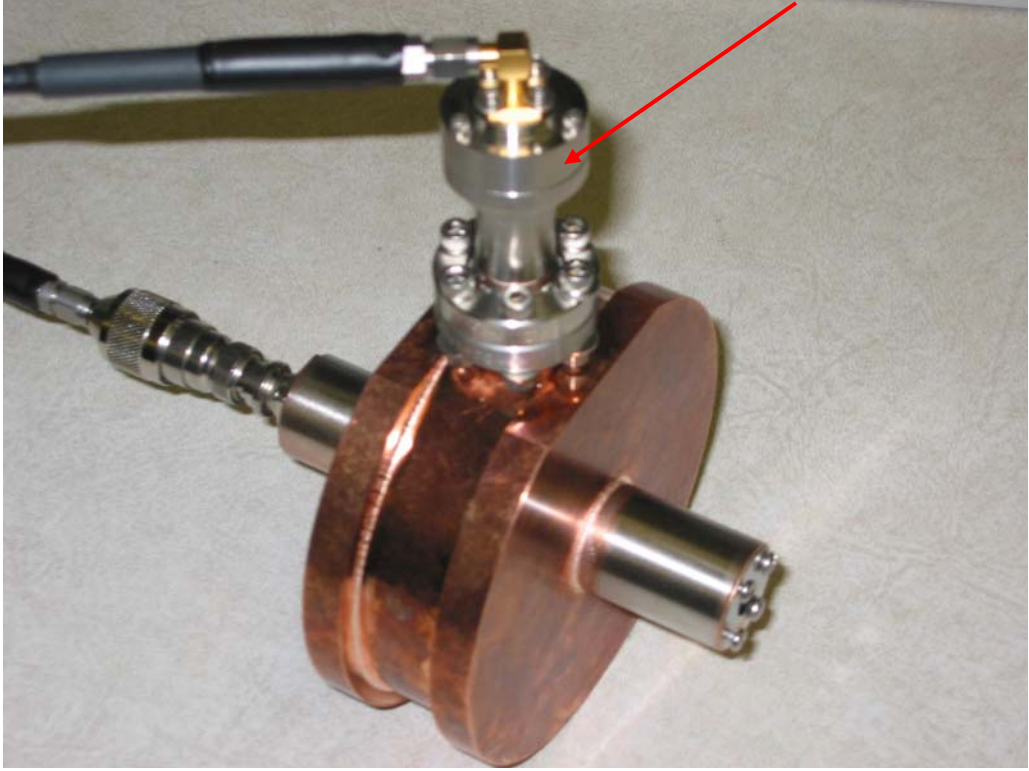


Figure 11. Photograph of the phase detector cavity model

3.2) Detuning Effect of the Coupling Loop

Superfish simulation and the measurement results presented in the first section are valid for the cavity without a coupling loop, (CL). To get the model resonant frequency that can be directly compared with the Superfish calculation, we inserted “a plug” into the CL port to eliminate the effect of the “hole” in the cavity cylinder wall. FIG. 12 shows the frequency detuning effect of the CL when changing coupling coefficient β . The coupling coefficient was changed by rotating the CL assembly. A single undercoupled loop, ($\beta \rightarrow 0.1$), lowers the frequency of initially unperturbed cavity by as much as 1.4 MHz. The design asks for using two coupling loops with $\beta = 0.12$, doubling the magnitude of this frequency shift. The frequency target for Superfish calculation to design a cavity resonating at 2805 MHz is discussed in the next section.

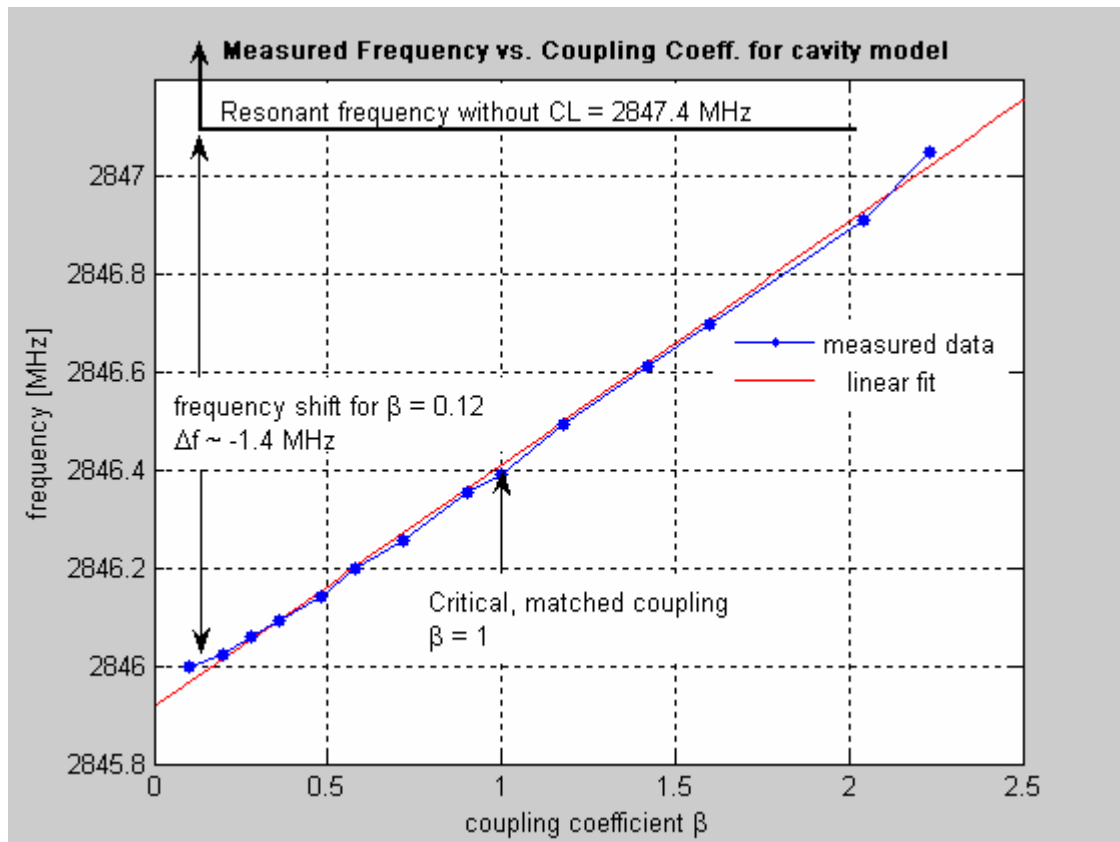


Figure 12. Effect of a single coupling loop on the resonant frequency

3.3) Calculation of the dimensions for 2805 MHz cavity

Comparing the measured and calculated results at ~ 2850 MHz it was possible to extrapolate the cavity design to a different frequency of 2805 MHz. From Table 1 and Figure 4 the target frequency for the Superfish calculation at 2850 MHz is about $1.4 + 2 \cdot 1.4 = 4.2$ MHz higher than the actual measured resonant frequency of the cavity model. The first component is the difference between the measured and calculated frequency, the second one describes the detuning effect of the two coupling loops. The difference between calculated and actual frequency will be slightly different at 2805 MHz. To estimate this difference, the same cavity

dimension was changed by the same amount and the resulting frequency shift was calculated at both frequencies, 2850 and 2805 MHz. The frequency shift at 2805 MHz was about ~4% smaller compared the shift at 2850 MHz. The difference of target frequency was therefore chosen to be only 4.0 MHz higher, i.e. ~2809.0 MHz. The Superfish calculated dimensions for this target frequency are shown in FIG. 13.

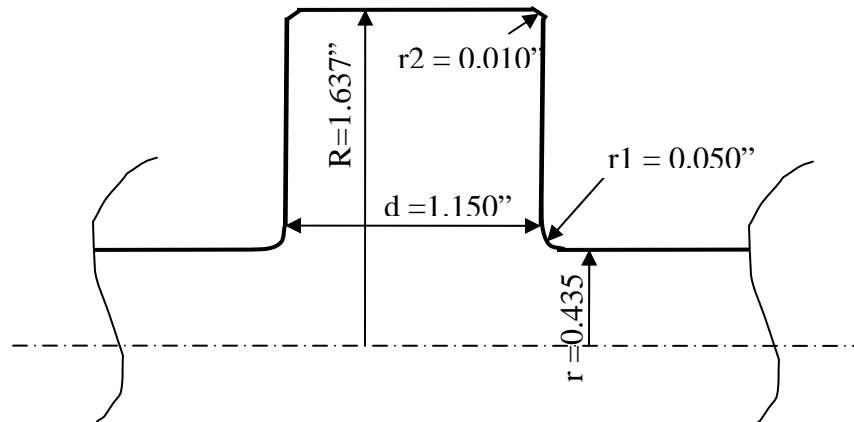


Figure 13. Dimensions used for final mechanical design

These dimensions are used for the final mechanical design of the cavity. To correct for the frequency difference that may be caused by the error in simulation and by the manufacturing tolerances, there will be 4 “dimple” tuners on the cylinder wall each with $\sim \pm 1$ MHz tuning range.

3.4) Calculation of Frequency Sensitivity on Cavity Dimensions

A mechanical design has to set the tolerance limits for the manufacturing. The deviation of some dimensions from their nominal value will have larger effect on the resonant frequency. Table 2 shows the effect of 0.001” change for each of the five cavity dimensions shown on FIG 13. The frequency change was calculated with Superfish program. The results are arranged in the order of decreasing sensitivity

Cavity dimension	$\Delta f / \Delta(\text{dimension})$	Note
Pillbox radius, R	1.770 MHz / 0.001”	Manufact. tolerance and temperature effect
Cavity gap, d	0.860 MHz / 0.001”	Manufact. tolerance, temper. effect and vacuum loading
Beam pipe radius, r	0.295 MHz / 0.001”	Manufact. tolerance and temperature effect
Cavity nose radius, r1	0.120 MHz / 0.001”	Manufact. tolerance
Radius of cavity corner, r2	0.065 MHz / 0.001”	Manufact. tolerance

4.0 Dark Current Errors

V_p is the peak voltage generated in the cavity at the instant the bunch passed by, or the peak of a sine wave in the case of a long ($\gg 700\text{nS}$) bunch train. The voltage will scale with current if it is a bunch or continuous bunch pattern like dark current. When a single 1nC bunch passes through the cavity, the induced voltage is 1350Volts . A steady state dark current of 200pC/uS will produce 2.6V peak fields as the cavity oscillates in a steady state. The ratio of dark current signal to signal from the bunch is 2.6 to 1350 . In signal processing we actually reduce the peak signal from the bunch by a factor of 2 . So the voltage ratio of interest is 2.6V to 675V . The phase error from this is maximized if the voltage vector of the 2.6V is 90 degrees off from the voltage vector of the 675V . This angle is then $2.6/675$ radians, or 0.22 degrees, or 220fS at 2805MHz .

In actuality, the phase of the dark current signal is not 90 degrees out from that of the beam so the error will not be so great. If the signals were in phase, the error from dark current would be an amplitude error and not a phase error. There is expected to be a factor of three greater reduction in the error from dark current due to this.

The attenuation is about -47.2 dB for $f=2856-51\text{MHz}$, i.e. cavity impedance 51 MHz away from resonance is $R_s * 10^{(-47.2/20)} = 6550$ Ohm. ($R_s \sim 1.5$ Mohm) Figure 14

The total DC dark current for $207\text{pC}/1\text{us}$ is $I_{dc} = 2e-4$ A with $2e-5$ A of jitter. From the value of DC current we can estimate the peak voltage at the bucket repetition frequency 2856 MHz as $2 * I_{dc} = 4e-4$ Apeak, @ 2856MHz .

This AC current will generate about 2.6 V_p signal in the cavity with 0.26 V_p jitter. That compares to about 1350 V_p induced by the 1 nC bunch charge passing through the cavity.

The ratio of the regular bunch generated signal to the dark current jitter then will be $0.26/675 \approx 1e-4$, or -70 dB. This signal to noise ratio will allow the system to meet specifications.

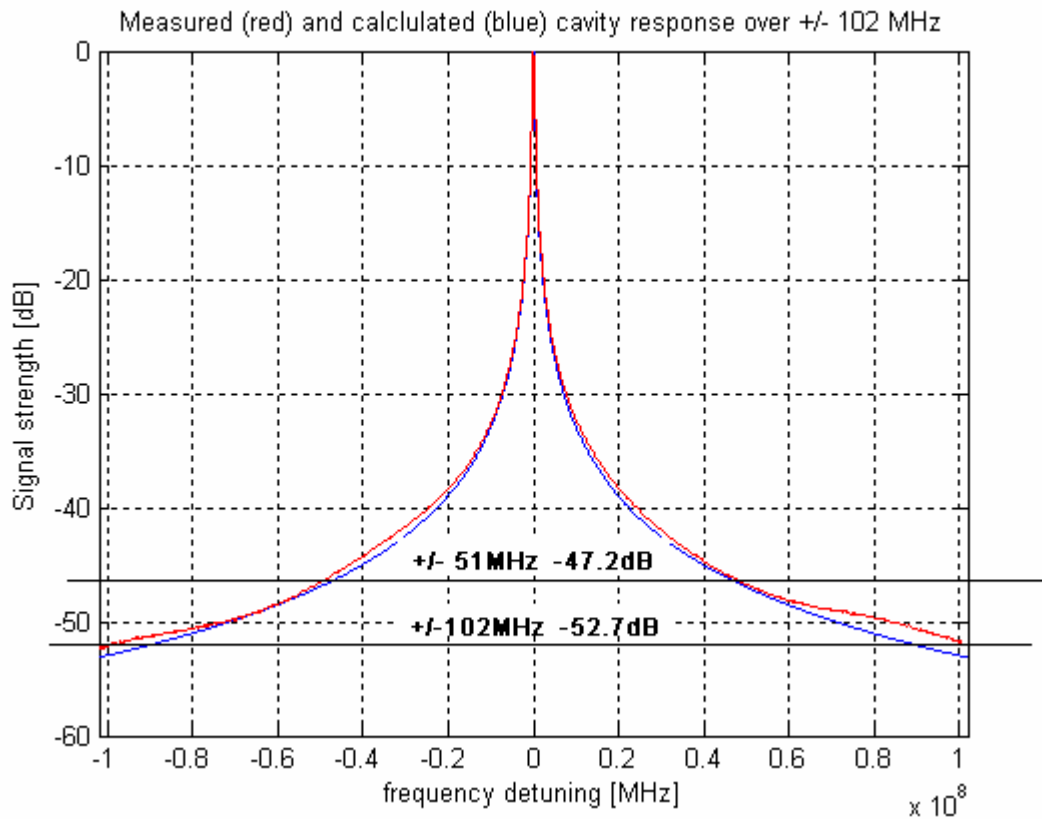


Figure 14. Cavity Signal Strength vs. Frequency

An offset of 102MHz was evaluated to further gain dark current signal suppression. The 102MHz frequency offset suppressed the signal by -52.7dB, only 5.5dB less than the 51MHz offset. The benefit of having the IF frequency the same as the RF out weighted the 5.5dB gain in suppression of dark current. If 102MHz were chosen either a different LO frequency would need to be generated, or the IF frequency to be digitized would be higher. The higher IF frequency would put more stringent requirements on the digitizer clock.