ABSTRACT

This report describes a LLRF system developed at Lancaster University and an experiment undertaken at the STFC Daresbury Laboratory by Cockcroft Institute Staff to validate an approach to phasing the crab cavities for the International Linear Collider. The work has involved the manufacture and processing of a pair superconducting cavities, the development of a vertical cryostat facility, the development of a digital control system for each cavity and the development of an interferometer.

In a test conducted during August 2008 it was demonstrated that an RF interferometer with digital phase detection and digital cavity controllers can lock a pair of superconducting cavities having realistic levels of microphonics such that r.m.s. phase errors are less than 120 milli-degrees at 3.9 GHz.

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1. **Introduction**

This report describes a LLRF system and an experiment undertaken at the Cockcroft Institute, Daresbury by Lancaster University and ASTeC to validate an approach to phasing the crab cavities for the International Linear Collider. The work has involved the manufacture and processing of a pair superconducting cavities, the development of a vertical cryostat facility, the development of a digital control system for each cavity and the development of an interferometer.

A major focus of the overall project and not covered in this report has been the design of a nine cell crab cavity with couplers appropriate to ILC parameters. This design is summarised in a separate EuroTeV report.1

The relative phasing of the RF fields in accelerator cavities determine acceleration and longitudinal bunch size. For the satisfactory operation of a linear accelerator (linac) the acceleration fields must be precisely phased with the particle source, bunch compressors and pulsed deflection magnets. For a collider it is also necessary for the two opposing linacs to be synchronised so that collisions occur at a precise location. If the linacs in a collider are inclined at an angle to each other then the bunches need to be aligned at the interaction point (IP) for optimum luminosity. For the International Linear Collider (ILC) this alignment will be performed with deflection cavities phased such that the centre of each bunch passes through the cavity at a time when it sees zero net field whilst the front and back of the bunch gets an equal and opposite kick. A deflection cavity phased in this manner is called a crab cavity. After passage through a crab cavity a bunch should start to rotate about its centre. Any error in the phasing of the crab cavities results in bunches getting a transverse kick and hence transverse deflection at the IP. If this transverse kick is small, but identical for the crab cavities on the opposing linacs then the bunches still collide with maximum luminosity but with small transverse offset.

The correct phasing of a complex accelerator system is typically achieved by a timing distribution system whereby pulses are distributed at regular intervals along dispersion less transmission lines to individual RF systems such as cavities. For simple RF systems a fixed frequency RF signal at the cavity frequency or some fixed fraction of the cavity frequency might be distributed. Phasing of individual components is then achieved by precise variation of path length with manual phase shifters.

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2. ILC Requirement

The phasing accuracy requirement for the ILC linacs is about 0.1 degrees r.m.s. at 1.3 GHz, which corresponds to a r.m.s. timing accuracy of 214 fs. The phasing accuracy for the ILC crab cavities with respect to each other is 0.125 degrees r.m.s. at 3.9 GHz which corresponds to a timing accuracy of 90 fs. The current proposal for the ILC linac timing system is to use mode locked lasers to stabilise the master clock and to use optical pulses on fibre cables to distribute the timing reference. Claims for the performance of optical distribution systems reach accuracies of 10 fs. Due to the extent and complexity of the ILC timing distribution system it is uncertain whether the accuracy of the linac timing distribution system in the vicinity of the IP will be significantly better than 200 fs.

Where an optical timing distribution system is utilised a fixed frequency RF signal must be synchronised to the pulses at the point of use. This synchronisation procedure invariably involves semi-conductor devices such as photodiodes and consequently the accuracy of RF phasing is ultimately limited by the noise in the semi-conductor devices. When driving cavities, the RF reference signal is used both to provide the input for the cavity power amplifier and also as the local oscillator for the cavity phase detection system.

There are two required synchronisations, the local RF timing signal to the optical signal and the cavity RF phase to the local RF timing signal. The accuracy to which a multi-cell crab cavity can be synchronised to an RF reference signal has been analysed elsewhere. (As a first stage to the validation process the tests undertaken and described in this report used single cell cavities rather than multi-cell cavities.)

Traditionally RF timing distribution systems were used before the development of optical timing distribution systems. Where the systems to be synchronised are far apart temperature fluctuations and vibration on the RF cables between the systems can give significant phase jitter. In order to eliminate this phase jitter the RF links need to be operated as interferometers.

The complication with an RF interferometer is the level of reflection one gets with RF components and phase shifters and the consequential corrections than need to be applied. Given that both the optical timing system and the RF timing system depend primarily on jitter in a number of semi-conductor devices the ultimate performance over relatively short distances where attenuation is not an issue is unlikely to be dissimilar. Given that laser timing systems are being extensively researched elsewhere for the ILC it was decided that our effort should focus on the on performance possible with an RF interferometer.


5 A.C.Dexter and G.C.Burt, EuroTeV report 2008-064, “Phase and Amplitude Control of Dipole Crabbing Modes in Multi-Cell Cavities”
3. Coupler Assembly and Test

Single cell cavities as shown in figure 3.1 were used for proof of principle testing of the phase control system. The cell dimensions were adapted from cell dimensions for the nine cell cavity being proposed for the ILC crab cavity $^1, ^6, ^7$. Three cavities were manufactured by Niowave in the US for the project. A full description of the manufacturing process was provided by Niowave and is given here as Appendix 3. Project staff from the CI were involved with cleaning and conditioning of the cavities in the US before delivery to Daresbury.

![Figure 3.1 Niobium Single Cell Cavity with Dipole Mode at 3.9 GHz](image)

New clean room and ultra clean water rinse facilities have also been established at Daresbury for cavity assembly operations. Initial tests at Niowave measured the peak field and intrinsic Qs for each of the three cavities. For these tests the cavities were run in a frequency lock loop. The external Q for the input couplers were typically $5 \times 10^8$ and the external Q for the output couplers were typically $5 \times 10^{10}$.

The phase control tests require phase lock rather than frequency and with a loaded Q much closer to the planned operation loaded Q of $3 \times 10^6$. Niowave provided probe couplers to reach from the cavity flange to the cavity centre which could be cut at Daresbury to the required length. Antenna (input) and probe couplers (output) were assembled onto the cavity in the clean room facility as shown in figure 3.2. The cavity output coupler was mounted first and its external Q was measured using a the transmission through a probe coupler who’s external Q was known. Then the probe was replaced with the input coupler and its external Q was measured using the transmission from the input to output couplers.

Extending the analysis of appendix 1 one can show that

$$Q_{co}Q_{ct} = 4Q_L^2 \frac{P_f}{P_{out}}$$

where $P_f$ is the forward power to the input port and $P_o$ is the output power from the probe port. The result can also be inferred from (15.40). The ratio $P_f/P_{out}$ is determined by Network Analyser transmission measurement in dB.

Unfortunately due to the high sensitivity to the coupler positions meant that the external Q factors drifted significantly during assembly and cool down. This meant that the external Qs for the input couplers were near to $2 \times 10^7$ rather than the $3 \times 10^6$ measured. For validation of

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the LLRF system this was not thought to be a problem. However the output coupler also increased by an order of magnitude, significantly decreasing the output power levels.

Figure 3.2  Coupler assembly
4. Vertical Cryostat Description

A new vertical cryostat facility shown in figure 4.1 has been established at STFC’s Daresbury laboratory. Figure 4.2 is a schematic of the facility seen from above. Part of the justification for this facility was for the development of superconducting crab cavities for the ILC. The facility has allowed proof of principle tests for phase control system under development.

![Figure 4.1: Vertical Cryostat facility at Daresbury](image1)

The phase control experiments were conducted at 4.2 K. The facility can pump to 2 K but operation at the associated higher intrinsic Q for this temperature was not required.

![Figure 4.2: Vertical cryostat showing top platform and staff displacement robots](image2)
A realistic test required cavity operation with an external Q close to $3 \times 10^6$, that planned for ILC operation. At 4.2 K the intrinsic cavity Q of was expected to be $3 \times 10^7$ which is somewhat above $3 \times 10^6$. For reasons explained in the next section the external Q for the tests ended up near to $1 \times 10^7$.

Locations of the cavity high vacuum pump and the cryostat pump for 2K operation are shown in figure 4.2.

A test of the LLRF system was its ability to handle microphonics. Whilst the positioning of the cavity vacuum pump was regarded as a temporary measure it did provide realistic microphonics.

Figure 4.3 shows the cryostat lid assembly. All components in the cavity hang from the lid. The lid seals to the cryostat vessel with a rubber ring.
5. Cavity Support and Tuner Description

In order to phase lock two cavities it is necessary for their natural frequencies to be relatively close with respect to their bandwidth. The planned bandwidth for the tests was 1 kHz but as the coupler probe external Q had varied while mounting the cavity the actual cryogenic bandwidths were 400 Hz for cavity one and 240 Hz for cavity three. An accurate prediction of the cavity frequency to within a few tens of Hertz at 4.2 K is not possible from measurements made while the cavities are at room temperature. At room temperature the cavity bandwidths are about 2 MHz. The resonance frequency also shifts as the cavity cools. In order tune the cavities for phase locking at least one cavity needed a tuner with a range of several MHz. In order to shift the cavity dipole mode frequency of 3.9 GHz by 4 MHz an axial squashing movement of about 0.23 mm was required.

The tuner design adopted is shown in figure 5.1. Routing of the cables is shown figure 4.3.

![Figure 5.1 Tuner mechanism](image)

The tuner was developed in a short period of time and the concept was untested. The central pillar provides an anchored length against which the tuner mechanism can act to squash the cavity. Squashing is achieved by a lever action. A limitation of this type of tuner is that it can only squash the cavity and not stretch it. Replacing the cable with a rod does not help as the rod would be so long that it would buckle in compression.

Initially the cavity frequencies were 7 MHz apart. During final assembly it was realised that the tension needed in the cable to get the two cavities to the same frequency was excessive. The cavity with the highest frequency, already less than 3.9 GHz, was inelastically squashed to bring its frequency to within approximately 1 MHz of the other. However due to alignment errors on the tuners it was only possible to mount a tuner on cavity one.
This became problematic as during the tests cavity one, which started with the higher frequency was initial squashed to have the same frequency the cavity three. During the tests a flaw with the tuner assembly became apparent as it was causing the centre frequency of cavity one to drift by 5 kHz over a few minutes as much as 15 kHz over a day. Throughout the test the tuner needed continual adjustment and no convenient actuator had been installed. After many cycles of adjustment, at one stage, cavity one attained a frequency 10 kHz below cavity three when there was no tension in the cable. At this stage bringing the cavities to the same frequency was impossible. Overnight cavity one did however relaxed slightly so tuning to the same frequency became possible again.

The longitudinal tuning sensitivity of our cavity was 17.4 MHz mm$^{-1}$. This means that 17 kHz corresponds to a movement of 1 $\mu$m. Ideally the tuner must give tuning stability at the level of 40 Hz for a 400 Hz bandwidth. This means that the tuning mechanism must be smooth at the movement level of 2 nm. Tuning to better than 1 kHz was hit and miss and one concludes that the tuning mechanism developed only operated smoothly for movements greater than about 100 nm. For movements less than 100 nm great attention on bearing detail is needed.

Consideration of tuning drift from cable expansion is of interest. The linear expansivity of the steel cable at 4.2 K is likely to be extremely small ($<< 1 \times 10^{-6}$ K$^{-1}$). There is however about 0.5 metres of cable in a transition zone between 100 K and 300 K. The average expansivity of steel in this range is about $8 \times 10^{-6}$ K$^{-1}$ hence if the average temperature of this part of the cable fluctuates by 0.5 K then the length change is about 2$\mu$m. The leverage ratio between cable movement and cavity movement is 1:3 hence for this 0.5 K fluctuation we might get a movement of 600 nm. Using the tuning sensitivity of 17.4 MHz mm$^{-1}$ the 0.5 K fluctuation gives a tune shift of 10 kHz.
6. LLRF Cavity Control Circuit Description

The phase of an RF cavity is controlled by varying the phase of the input. The response of the cavity with respect to an input is described by the theory of the driven oscillator set out in Appendix 1. The overall system to synchronise the two crab cavities is comprised of two cavity control systems and an interferometer. This section describes the digital cavity control system being developed by the Cockcroft Institute for control of one cavity to a local reference. The arrangement shown in figure 6.1 permits amplitude and phase control and corresponds to that used in the tests described here.

![Diagram of LLRF Cavity Control Circuit](image)

**Figure 6.1** Control layout for one cavity as used in August 08 tests
The distinctive feature of the control system under development is the use of Hititite HMC439QS16G digital phase detectors. These detectors are being investigated as their linearity offer advantages with respect to system calibration. Their phase jitter performance however is significantly worse than double balanced mixers. From figure 6.2 it can be seen that the phase noise at 1280 MHz is about -135 dBc/Hz and is relatively flat with offset hence phase noise in 1 MHz bandwidth is about -80 dBc. Using the conversion of appendix 2, table 15.1 the r.m.s. phase jitter \( = 1.41 \times 10^{-4} \) radians \( = 8 \) milli-degrees \( = 17 \) fs. This is quite large but still significantly less than the timing requirement of 90 fs. Frequencies greater than 1 MHz have virtually no effect on the cavity phase jitter performance where a superconducting cavity with a bandwidth close to or less than 1 kHz is used. Digital phase detectors only operate up to a frequency of 1.3 GHz hence they must be used with frequency dividers (8 milli-degrees of phase jitter at 1.3 GHz implies 24 milli-degrees of phase jitter at 3.9 GHz). The 3.9 GHz signal is frequency divided by 3 using HMC437MS dividers, these generate an additional 2 milli-degrees r.m.s. phase jitter at 1.3 GHz.

An alternative to measuring phase with dedicated phase detectors is to down convert the cavity RF output to a lower frequency whilst preserving phase, digitally sample the wave and compare its phase to the internal clock of a digital processor – usually an FPGA. The program in the FPGA then determines the required correction to the input phase.

With respect to the control system under development as shown in figure 6.1 the control box has an RF input and an RF output marked in red. All lines marked in red carry 3.9 GHz. The bus in the cavity control box can hold eight input-output cards, see figure 6.3. This configuration uses three 16 bit 105 MBPS ADC inputs (latency = 130 ns). The top ADC input gives reflected power from the cavity, the second gives cavity amplitude and the third gives cavity phase. All black lines carry low frequency signals. This configuration also uses three 16 bit 40 MBPS DAC outputs (latency = 10 ns). Two of the DACs provide I and Q signals for the vector modulator that is housed in the control box. The vector modulator is an AD8341, it has an output noise floor of 150 dbm/Hz, and gives 2 milli-degrees r.m.s. phase jitter for an input level of 0 dBm.

The 3.9 GHz system oscillator is locked a 10 MHz quartz oscillator. The figure shows one external DAC output being used to perturb the frequency of the 10 MHz quartz master oscillator. A variation of about +/-100 kHz from the 3.9 GHz VCO was possible, however this was inadequate to drive the cavities which had an unplanned offset of 14 MHz. The output

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frequency can also be changed by ramping the phase on the vector modulator. This can be achieved by applying sine and cosine signals to the I and Q inputs. The maximum frequency variation provided by the vector modulator will be limited here by the clock speed of the DSP (if for instance the sine and cosine waves are generated with 12 points per cycle and the programme take 3 μs to cycle, the frequency range is 30 kHz).

Output from the system oscillator is split with an isolated tee to provide reference signals for the phase detectors and also a drive signal for the cavity amplifier. The phase of the signal to the cavity is adjusted by the vector modulator in the control box.

The cavity only has one input coupler and one output coupler. The 3.9 GHz output from the cavity is split with an isolated tee to provide separate signals for amplitude and phase measurement. Amplitude is measured with a diode detector followed by a low noise amplifier.

In the figure 6.1 the 1.3 GHz signals are marked in light blue. Each divider provides two differential output signals both of which are then amplified by differential amplifiers. One signal is taken to an ADC input on the control box and the other is amplified for display on an oscilloscope.

Figure 6.3 shows the boards which have been developed to mount the ADC and DAC chips and their associated amplifiers. Having chips on separate boards aids, scalability, fault location, reparability and cooling. The horizontal boards are DSP and FPGA evaluation boards. In section 7, figure 7.3 there is a photo of boards and vector modulator in their boxes.

The DSP shown a Texas TMS320C6713 and the FPGA shown is a Xilinx Spartan III. There are several options for sharing processing between the devices. One option is to let the FPGA accept samples from the ADC at the maximum sampling rate, perform filtering and run a simple control loop. The DSP would adapt control coefficients, manage calibration, control automatic tuners and handle faults. For the current tests all the processing was performed by the DSP. The evaluation board limits the DSP clock speed to 225 MHz. At this highest clock speed we had difficulty with a spurious oscillation in the cavity output spectrum at an offset of 900 Hz. The oscillation disappeared at a DSP clock speed of 200 MHz.

As the 16 bit ADC has only 65536 levels then if the amplifier’s gain allows 360° to be mapped onto this full range then, the smallest angle that might be resolved is 6 milli-degrees.
bit ADC the last 3 bits are only significant when averaged over a large number of samples. This means that in practice the resolution is nearer to eight times 6 milli-degrees i.e. 50 milli-degrees. Figure 6.4 shows an alternative arrangement where two phase signals are available to the processor, one which spans $360^\circ$ with a resolution of 50 milli-degrees and the other with a span of $3.6^\circ$ about zero and has a resolution better than 1 milli-degree. The figure also shows an option for better control of the output modulation by modulating the phase at full range in the first vector modulator before attenuating the output in the second vector modulator.

![Figure 6.4 Envisaged Circuit Layout for Data Handling](image)

Cavity control loop latency limits the useable gain and hence the performance. It is preferable to have the high power amplifiers within 10 m of the cavity ($2 \times 10 \text{ m} \sim 66 \text{ ns}$ from a budget of about 1000 ns).
7. LLRF Interferometer Description

If a wave on a transmission line has a non-zero standing wave ratio then the standing wave is not perfect and the phase at any two points is typically different. Conversely for equal forward and reflected waves \( \cos(\omega t - kz) + \cos(\omega t + kz) = 2\cos(\omega t)\cos(kz) \) so the phase at every point is the same. When the waves have unequal magnitudes i.e. \( \cos(\omega t - kz) + R\cos(\omega t + kz) \) then \( \cos(\omega t) \) is never a factor. At RF frequencies transmission lines always have loss and hence the standing wave ratio is always non zero.

For phase synchronisation one needs to have the same phase at two points. For accelerator operation one often needs a fixed phase difference between two points.

If the forward wave is \( \cos(\omega t - kz) \) then the phase at \( z_1 \) is \( kz_1 \) and the phase at \( z_2 \) is \( kz_2 \) etc. Typically the positions \( z_1 \) and \( z_2 \) vary by small amounts and hence the phase difference is unknown even when \( k \) is known. If the forward wave is sent back with unknown phase shift \( \phi \) then the backward (reflected wave) varies as \( \cos(\omega t + kz + \phi) \), hence the phase at \( z_1 \) is \( -kz_1 + \phi \) and the phase at \( z_2 \) is \( -kz_2 + \phi \). By varying \( \phi \) and the separation \( z_1 - z_2 \) it is possible to independently fix the phase difference between forward and reflected wave at both \( z_1 \) and \( z_2 \), taking the set phase differences as \( \alpha \) and \( \beta \) we have that

\[
2kz_1 + \phi = \alpha \quad \text{and} \quad 2kz_2 + \phi = \beta
\]  

(7.1)

Solving to eliminate \( \phi \) gives

\[
kz_1 - kz_2 = \frac{\alpha - \beta}{2}
\]  

(7.2)

Since \( kz_1 \) was the phase at \( z_1 \) and \( kz_2 \) was the phase at \( z_2 \) then the phase difference between the forward signals (or indeed the return signals) is determined by (7.2) where \( \alpha \) and \( \beta \) can chosen and fixed with an active control loop (\( \alpha \) is the phase difference between forward and reflected waves at \( z_1 \) and \( \beta \) is the phase difference between forward and reflected waves at \( z_2 \)). It is therefore possible to have a known phase shift between two points when the transmission line between them is imperfect and has a slowly fluctuating length.

A difficulty with the technique described is that reflection on the line gives systematic errors. Our proposal is to correct these errors with a DSP from knowledge of line length corrections being made to maintain \( \alpha \) and \( \beta \). An alternative to using a CW wave is to use pulses. When pulses are used unwanted reflections appear as signals at the wrong amplitude and time and hence can be rejected. Figure 7.1 illustrates how synchronisation with pulses or wave peaks and troughs works.

![Figure 7.1 Conceptual schematic for interferometer – sequential crests, troughs or pulses in red and black, forward wave is solid, reflected wave is dashed, end locations are green](image-url)

Figure 7.1 Conceptual schematic for interferometer – sequential crests, troughs or pulses in red and black, forward wave is solid, reflected wave is dashed, end locations are green
Figure 7.1 also illustrates that synchronisation will have a phase uncertainty of 180°. This can be resolved by approximate knowledge of the line length. Figure 7.2 shows how the interferometer can be implemented for a continuous wave.

The implementation of figure 7.2 fixes the phase shifts $\alpha$ and $\beta$ at zero and information for computing systematic errors is not available for processing. For effective operation only one master oscillator should be used and it needs to be very stable over the time period that it takes for a signal to traverse the coaxial link and return.

Knowing the exact points of synchronisation is useful as transporting the signal on to a new location gives a new phase shift and it helps to know where one starts. For the layout shown the points of synchronisation are the centres of the two digital phase detectors. The digital phase detectors give d.c. output voltages when the RF inputs have identical frequencies. The loop filters drive both phase differences to zero by varying on the LHS the coaxial line length and on the RHS the length of the loop that returns the input signal. The digital phase detectors have limited input range between -10 dBm and +10 dBm. Whilst this is sufficient for the return signal to be measured back at the LHS phase detector, one needs the return signal to be large with respect to unwanted reflections so that systematic corrections stay small. For this reason an amplifier is used on the return loop. Figure 7.3 shows how the full system can be implemented in a way that corrections can be applied and calibrations made.

Figure 7.3 Full Layout for Cavity RF Synchronisation
In figure 7.3 the loop filters are replaced with a digital controller composed of a ADC, an FPGA and a DAC. The voltage being applied to the phase shifter on the coaxial line determines the correction required to correct for reflection from the various RF components. A phase shifter is required on the coaxial line as waves must pass in both directions. On the precision reflector loop ideally there should only be a forward wave. A vector modulator can now replace the phase shifter in figure 7.2 as the FPGA can produce appropriate I and Q inputs. Use of a vector modulator allows the return signal to be amplitude modulated. This is useful in checking calibration because the part of the backward wave arriving at the phase detector adjacent to the oscillator that is amplitude modulated must come from the far end rather than from reflections at the phase shifter or the directional coupler.

The reference point for phase synchronisation is the centre of the phase detectors. The distance between the phase detectors and the centre of the cavity is an uncontrolled length. To avoid additional phase jitter or error in absolute calibration these lengths should be minimised. Preferably all the components in each yellow box of figure 7.3 should be mounted on a single board in a temperature controlled and vibration free environment. The board should be mounted on the end of the cavity output coupler. Such a PCB with directional couplers, Wilkinson dividers, phase detectors, low noise differential amplifiers and the divider to reduce the 3.9 GHz signal to 1.3 GHz is under development. A picture of the first prototype is shown in figure 7.4. It was not completed in time for use in the August tests.

![Prototype integrated interferometer termination](image)

An important aspect of maintaining accurate control of the cavity is to correct for gain and offset in the amplifiers between the digital phase detectors and the ADC in the cavity control box shown in figure 6.4 but not figure 7.3. This can be done by running the interferometer at a frequency offset to the cavity for a few cycles. The ADC records a maximum voltage for +180° and a minimum voltage for -180°. Corrections for offset and gain can then be made by the FPGA. Calibration of the amplifiers on the interferometer can be achieved by giving the backward wave a frequency offset to the forward wave.
Completion of the development LLRF system set out in figure 7.4 requires considerable effort and testing at an intermediate stage to demonstrate proof of principle was considered essential. Figure 7.5 shows a schematic of the layout used for the proof of principle experiments undertaken in August 2008 with two single cell superconducting cavities in a vertical cryostat.

Figure 7.5 Schematic of layout used for August 08 tests

Figure 7.6 shows actual components of the LLRF system used in the August 2008 tests, together with the multi-cell aluminium cavities used for tests not requiring superconducting cavities.

A significant drawback of an RF interferometer is that components along the interferometer and especially the electronic phase shifter give unwanted reflections. Worse still the reflection coefficient of the electronic phase shifter depends on the phase shift it is providing at any instant. Fortunately all the reflection coefficients do not vary with time and the phase shift provided by the phase shifter is known at any instant. This means that a correction curve can be determined experimentally. Figure 7.7 shows the phase difference between the cavities as the interferometer line length is varied with the manual phase shifter. The curve was measured with aluminium cavities rather than superconducting cavities. Ideally the curve should be flat. Every point on the curve has a one to one correspondence with the voltage applied to the electronic phase shifter hence a correction can be applied by the DSP. If the line between the cavities varies by a couple of degrees then the correction at 3.9 GHz is about 400 milli-degrees at worst.
Figure 7.6  Components used for August 08 tests

Figure 7.7  Phase difference between cavities as a function of line phase shifter
The oscilloscope measurements of phase as a function of time used to construct figure 7.7 shows a small amount of phase jitter. This jitter determines the ultimate performance. Figure 7.8 shows the level of jitter over a time scale of 50 ms for optimum controller gain. The calibration at 3.9 GHz is 15 milli-degrees per mV. The peak to peak oscilloscope voltage jitter in the trace is 15 mV, this converts to a phase jitter 225 milli-degrees pk to peak which is 75 milli-degrees r.m.s.. The target performance is 125 milli-degrees r.m.s.. Once the controller has been optimised for superconducting cavities one would not expect the performance for the superconducting cavities to be any worse than warm cavities.

Figure 7.8  Cavity to cavity jitter for warm operation
8. Cavity Parameters after Cooling

Cavity parameters were determined by measuring S parameters with a Network Analyser once the cavities had been cooled to 4.2 K. Three cavities were manufactured and cavities 1 and 3 were used for the tests. Results after running for several hours under vacuum for cavities 1 and 3 together with computed Q factors are given in tables 8.1 and 8.2. Note the Q factor of the output coupler can be computed either from its reflection coefficient or from the Q factor of the input coupler and S12.

<table>
<thead>
<tr>
<th>Cavity 1 (Tunable)</th>
<th>Value</th>
<th>Unit</th>
</tr>
</thead>
<tbody>
<tr>
<td>f</td>
<td>3.8857 GHz</td>
<td></td>
</tr>
<tr>
<td>S12 including cables</td>
<td>-36.0 dB</td>
<td></td>
</tr>
<tr>
<td>S12 cavity only</td>
<td>-21.0821 dB</td>
<td></td>
</tr>
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<td>S12 cavity only</td>
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<td></td>
</tr>
<tr>
<td>Bandwidth</td>
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<td>QL</td>
<td>9.714E+06</td>
<td></td>
</tr>
<tr>
<td>S11 off resonance (output)</td>
<td>-9.9029 dB</td>
<td></td>
</tr>
<tr>
<td>S11 on resonance (output)</td>
<td>-9.9587 dB</td>
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</tr>
<tr>
<td>S22 off resonance (input)</td>
<td>-11.015 dB</td>
<td></td>
</tr>
<tr>
<td>S22 on resonance (input)</td>
<td>-20.183 dB</td>
<td></td>
</tr>
<tr>
<td>S11 cavity (output)</td>
<td>-0.0558 dB</td>
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<td>S11 cavity (output)</td>
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<td>S22 cavity (input)</td>
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<td>S22 cavity (input)</td>
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<tr>
<td>Qe (output)</td>
<td>3.034E+09</td>
<td>calc from S11</td>
</tr>
<tr>
<td>Qo</td>
<td>3.009E+07</td>
<td></td>
</tr>
<tr>
<td>Qe (output)</td>
<td>3.360E+09</td>
<td>calc from S12</td>
</tr>
</tbody>
</table>

Table 8.1  Cavity 1 parameters measured with network analyser after operation
Cavity 3 parameters measured with network analyser after operation

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
<th>Unit</th>
</tr>
</thead>
<tbody>
<tr>
<td>f</td>
<td>3.885557000</td>
<td>GHz</td>
</tr>
<tr>
<td>S12 including cables</td>
<td>-34.77</td>
<td>dB</td>
</tr>
<tr>
<td>S12 cavity only</td>
<td>-21.45</td>
<td>dB</td>
</tr>
<tr>
<td>S12 cavity only</td>
<td>0.0072</td>
<td></td>
</tr>
<tr>
<td>Bandwidth</td>
<td>256.1</td>
<td>Hz</td>
</tr>
<tr>
<td>QL</td>
<td>1.517E+07</td>
<td></td>
</tr>
<tr>
<td>S11 off resonance (output)</td>
<td>-10.0628</td>
<td>dB</td>
</tr>
<tr>
<td>S11 on resonance (output)</td>
<td>-10.0902</td>
<td>dB</td>
</tr>
<tr>
<td>S22 off resonance (input)</td>
<td>-9.2587</td>
<td>dB</td>
</tr>
<tr>
<td>S22 on resonance (input)</td>
<td>-24.1662</td>
<td>dB</td>
</tr>
<tr>
<td>S11 cavity only (output)</td>
<td>-0.0274</td>
<td>dB</td>
</tr>
<tr>
<td>S11 cavity only (output)</td>
<td>0.9937</td>
<td></td>
</tr>
<tr>
<td>S22 cavity only (input)</td>
<td>-14.91</td>
<td>dB</td>
</tr>
<tr>
<td>S22 cavity only (input)</td>
<td>0.0323</td>
<td></td>
</tr>
<tr>
<td>QL/Qe (input)</td>
<td>0.5899</td>
<td></td>
</tr>
<tr>
<td>Qe (input)</td>
<td>2.572E+07</td>
<td></td>
</tr>
<tr>
<td>QL/Qe (output)</td>
<td>0.9984</td>
<td></td>
</tr>
<tr>
<td>Qe (output)</td>
<td>9.633E+09</td>
<td>calc from S11</td>
</tr>
<tr>
<td>Qo</td>
<td>3.713E+07</td>
<td></td>
</tr>
<tr>
<td>Qe (output)</td>
<td>4.996E+09</td>
<td>calc from S12</td>
</tr>
</tbody>
</table>

Table 8.2  Cavity 3 parameters measured with network analyser after operation

Cavity parameters are summarised in table 8.3

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Cavity 1</th>
<th>Cavity 3</th>
</tr>
</thead>
<tbody>
<tr>
<td>Qo</td>
<td>3.009E+07</td>
<td>3.713E+07</td>
</tr>
<tr>
<td>Qe input</td>
<td>1.441E+07</td>
<td>2.572E+07</td>
</tr>
<tr>
<td>Qe output</td>
<td>3.034E+09</td>
<td>9.63E+09</td>
</tr>
</tbody>
</table>

Table 8.3  Summary
9. Cavity Fields and Output

For a dipole cavity the shunt impedance is defined as

\[ R_d = \frac{V_d^2(a)}{P_o} \left( \frac{ao}{c} \right)^2 \]  

(9.1)

Where \( \omega \) is the cavity angular frequency, \( P_o \) is power dissipated in the cavity and \( V_d(a) \) is the longitudinal voltage at distance \( a \) from the axis where \( a \) is small (note that on axis a dipole cavity has zero longitudinal field).

For the purpose of estimating peak voltages in the cavity we shall suppose that the peak voltage occurs roughly when \( \frac{ao}{c} = \frac{\pi}{2} \) hence using (9.1) we define

\[ V_{peak} \approx 0.5 \pi \sqrt{R_d P_o} \]  

(9.2)

From appendix 1 equation (15.41) the relationship between forward power and power dissipated in the cavity when driven at its centre frequency is given as

\[ P_f = \frac{Q_o Q_{ei}}{4Q_L^2} P_o \]  

(9.3)

where \( Q_{ei} \) is the external Q of the input port. Hence from (9.2) and (9.3) one obtains

\[ V_{peak} \approx \pi \sqrt{Q_L^2 \left( \frac{R_d}{Q_o} \right) \frac{P_f}{Q_{ei}}} \]  

(9.4)

The set point amplitude was chosen so that 10 W amplifiers in use delivered 7 W. Cable losses were -4 dB. Using these values table 9.1 estimates peak voltages in the cavities.

Table 9.1 also gives power output from the probes determined as

\[ P_{out} = \frac{Q_o}{Q_{eo}} P_o \]  

(9.5)

where \( Q_{eo} \) is the external Q of the output port.

<table>
<thead>
<tr>
<th>Unit</th>
<th>Cavity 1</th>
<th>Cavity 3</th>
<th>ILC per cell</th>
</tr>
</thead>
<tbody>
<tr>
<td>Frequency</td>
<td>Hz</td>
<td>3.886E+09</td>
<td>3.886E+09</td>
</tr>
<tr>
<td>QL</td>
<td></td>
<td>9.71E+06</td>
<td>1.52E+07</td>
</tr>
<tr>
<td>Qe (input)</td>
<td></td>
<td>1.44E+07</td>
<td>2.57E+07</td>
</tr>
<tr>
<td>Qe (output)</td>
<td></td>
<td>3.03E+09</td>
<td>9.63E+09</td>
</tr>
<tr>
<td>Qo (4K for tests but 2K for ILC)</td>
<td></td>
<td>3.01E+07</td>
<td>3.72E+07</td>
</tr>
<tr>
<td>R/Q (Ohms)</td>
<td>Ω</td>
<td>53</td>
<td>53</td>
</tr>
<tr>
<td>Amplifier Power</td>
<td>W</td>
<td>7</td>
<td>7</td>
</tr>
<tr>
<td>Cable losses</td>
<td>dB</td>
<td>-4</td>
<td>-4</td>
</tr>
<tr>
<td>Forward Power (W)</td>
<td>W</td>
<td>2.79</td>
<td>2.79</td>
</tr>
<tr>
<td>Peak cavity voltage</td>
<td>V</td>
<td>97702</td>
<td>114250</td>
</tr>
<tr>
<td>Power dissipated in cavity</td>
<td>W</td>
<td>2.42</td>
<td>2.69</td>
</tr>
<tr>
<td>Energy stored in Cavity</td>
<td>J</td>
<td>0.0030</td>
<td>0.0041</td>
</tr>
<tr>
<td>Output to probe</td>
<td>W</td>
<td>0.0241</td>
<td>0.0104</td>
</tr>
</tbody>
</table>

Table 9.1 Cavity Parameters
With a maximum cavity voltage of 115 kV in cavity 3 for the tests, the vertical cryostat did not need radiation shielding. For a nine cell cavity at ILC operating voltage the maximum field becomes 2.7 MV.

For operation at 4.2 K with both cavities energies power dissipated to the liquid helium was just over 5 Watts. This was comparable with static losses for the vertical cryostat.
10. Double Balanced Mixer Setup and Calibration

Figure 10.1 shows the configuration of a double balanced mixer. For phase detection typically the LO oscillator should be at a level that it saturates a pair of diodes, either s & t or v & u depending on its polarity at any instant. Depending on which pair of diodes is conducting, the path between the IF output and earth E uses opposite halves of the RF secondary. Consequently the RF gets multiplied by 1 or -1 when the LO changes polarity. 

![Double Balanced Mixer Setup](image)

Taking the RF and the LO oscillators to have the same frequency except for a time dependent phase \( \phi(t) \) and the RF to have amplitude \( V_{RF} \) which is well below diode saturation voltage then the IF output is given as

\[
V_{IF} = \frac{4}{\pi} V_{RF} \cos(\omega t + \phi) \left\{ \cos(\omega t) - \frac{1}{3} \cos(3\omega t) + \frac{1}{5} \cos(5\omega t) + \cdots \right\}
\]

Hence

\[
V_{IF} = \frac{4}{3\pi} V_{RF} \cos(\phi) + \text{very high frequencies}
\]

If the signal level of the LO is insufficient to saturate the diodes, output falls and the noise figure rises. If the RF signal saturates the diodes the output is determined by the product of two square waves and there is a small enhancement of the output coming from products of higher terms.

Equation (10.2) immediately reveals that for maximum sensitivity of phase measurement the phases of the local oscillator and the RF must be close to quadrature.

The double balanced mixer used for measurements was a level seven Mini-circuits ZX05-43+. Level seven implies 7 dBm input and the coefficient in (10.2) implies an attenuated output. Including losses the output for a 7 dBm input is about 1 dBm. The output impedance is 50 \( \Omega \) hence the output voltage for a phase error \( \phi \) of zero degrees is about 0.251 Volts. The output voltage for a phase error \( \Delta \phi \) near \( \phi = -\frac{\pi}{2} \) degrees is given as

\[
V_{out} = 0.251 \cos\left(-\frac{\pi}{2} + \Delta \phi\right) = 0.251 \sin(\Delta \phi)
\]

for \( \Delta \phi = 1 \) milli-degree \( V_{out} = 4.4 \mu V \) (-80 dBm)

for \( \Delta \phi = 20 \) micro-degrees \( V_{out} = 0.088 \mu V \) (-114 dBm)
The noise flow for 1 MHz bandwidth is about $kT\Delta f = -114$ dBm hence in principle the double balanced mixer can measure down to tens of micro-degrees in a bandwidth of 1 MHz. In practice when amplifier noise and drift is taken into account then without special effort the practical limit is hundreds of micro-degrees. Between the mixer and the point of measurement there will be cables which also contribute to jitter.

Our initial plan was to measure the phase jitter between the cavities by placing a double balanced mixer in the vertical cryostat so as to minimise cable lengths. As each cavity only had one output there was a requirement for splitters or directional couplers on the cavity outputs so that signals for the phase control system were also available. It was not clear as to the lowest temperature at which the mixers would continue to operate hence it was assumed that that mixer in the cryostat would need some sort of heater to bring its temperature to at least 77K and possibly higher. Consequently it was assumed that any container placed in the cryostat to hold the mixer and its associated components needed, to be continually evacuated to remove helium gas seeping through RF feed-throughs and hence providing insulation. The effort needed to complete this development was not available prior to the August 2008 tests hence the mixer was placed outside the cryostat. The mixer was finally placed along side the control system hence the cable lengths became 3 m inside the cryostat plus 3 m from the top of the cryostat to the control system. One of the mixer input lines also needed a manual phase shifter so that inputs could be brought into quadrature.

As a consequence of the external Q factors of the cavity couplers being high, the signals that were available to the double balanced mixer inputs after attenuation in the cable and splitting the signal reduced the signal level well below the 7 dBm required to saturate the mixer. Whilst the mixer still measures phase with low input signals, the noise figure is greatly increased and calibration depends on the amplitude of both input signals. Low input also meant low output and a high gain low noise amplifier was needed on the mixer output to make useful measurements.

Due to excessive frequency drift of cavity one, lock was not sustained for sufficient periods to get accurate calibration of the double balanced mixer during superconducting operation. (Note that cavity 1 was tuneable by means of a cable and the excessive drift was that associated with movements of tens of nanometres).
11. Locking Results
The locking performance was examined first using an Agilent Spectrum Analyser Model E4443A.

11.1 Source Characterisation
The master oscillator used throughout the tests was a Rhode & Schwarz Signal Generator. For jitter measurements on the cavity made with the spectrum analyser the noise from the source makes a contribution. Source noise at frequencies several cavity bandwidths from the drive frequency gets filtered by the cavity hence the output noise can be below the source noise. The most important region is less than 2.5kHz from the drive.

Figure 11.1 shows a screen shot of the source jitter to +/- 50Hz. Jitter at 10 Hz from the carrier is about -60 dB. Using table 15.1 in appendix 2 this level corresponds to about 80 milli-degrees r.m.s.

Figure 11.2 shows a similar measurement of phase noise on the source recorded a minute later as a data file rather than a screen shot. Precise measurements are slightly easier to determine from the data plot. In this figure there are peaks at +/- 3Hz with a level of -43 dB corresponding to 574 milli-degrees r.m.s.. The plot confirms a noise level of -60dB at 10 Hz.
Figure 11.2  Source noise from data file to +/-50Hz

Figures 11.3 shows noise on a span of 200 Hz taken at different instants. Peaks not visible on the 100 Hz span show up at +/- 60 Hz.

Figure 11.3  Source noise from data file to +/-100Hz

Figure 11.4 shows two more traces of the noise taken as screen shots. They illustrate that the magnitude of the jitter varies near 60 Hz.
Figure 11.4  source noise in 200Hz span at two instants

Figure 11.5 gives screen shots of source noise to 100 kHz offset.

Figure 11.5  source noise in 2kHz and 200kHz spans

Table 11.1 gives integrated values of the source noise to 1 MHz. As the minimum resolution bandwidth for the measurement was 1 Hz then we should probably discard the last two entries in the table.

<table>
<thead>
<tr>
<th>Range</th>
<th>Milli-degrees r.m.s.</th>
</tr>
</thead>
<tbody>
<tr>
<td>2.5kHz to 1 MHz</td>
<td></td>
</tr>
<tr>
<td>250 Hz to 1 MHz</td>
<td>36</td>
</tr>
<tr>
<td>25 Hz to 1 MHz</td>
<td>107</td>
</tr>
<tr>
<td>5 Hz to 1 MHz</td>
<td>129</td>
</tr>
<tr>
<td>4 Hz to 1 MHz</td>
<td>148</td>
</tr>
<tr>
<td>3 Hz to 1 MHz</td>
<td>403</td>
</tr>
<tr>
<td>2 Hz to 1 MHz</td>
<td>723</td>
</tr>
</tbody>
</table>

Table 11.1  Integrated source noise

In the next section it will be seen that the noise performance of the Rhode and Schwarz broadband signal generator was at a level that it masked the ultimate performance of the phase control system under test. We were unable to use the high quality narrow band 3.9 GHz source we had planned to use as its bandwidth was only 200 kHz and the cavities could not be tuned...
into the range. The cavities were about 6 MHz below 3.9 GHz and they would have needed stretching to bring them into the range.

11.2 Locking Results for Cavity 3

Cavity 3 was easiest to study as its centre frequency was relatively stable. The vacuum pump on the platform provided a significant level of microphonics. Figure 11.6 shows the output spectrum from the cavity with the control system off.

![Figure 11.6 Unlocked Output from Cavity 3 on 100 Hz span](image)

The peak at +/-29 Hz is 30 dB below the source and hence corresponds to a nominal phase jitter of 2.6 degrees r.m.s.. The peak at +/- 17 Hz is 35 dB below the source and hence corresponds to a nominal phase jitter of 1.44 degrees r.m.s.. Careful inspection of the peaks in figure 11.6 suggests that they could be broader than the 1 Hz resolution bandwidth setting for the spectrum analyser. If they are it means that the jitter they contribute could be enhanced by their width. Total jitter is determined by integrating the noise curve and using the RHS of equation (15.9). Below the peaks the level in the figure is about -65dB or $3.162 \times 10^{-7}$, integrating over a 50 Hz range, multiplying by 2 and taking the square root gives 0.0056 radians or 0.322 milli-degrees. Over this 100 Hz span it is apparent in this case that most of the jitter comes from the peaks. A proper integration of the data in figure 11.6 from -50 Hz to -2 Hz and then from 2 Hz to 50 Hz gives 3.95 degrees indicating that the peaks did in fact have a width equal to the resolution bandwidth.

Figure 11.7 shows the phase jitter on a span of 5 kHz. When the area under this curve is integrated from -2.5 kHz to -200 kHz and 200 Hz to 2.5 kHz it gives 3.1 degrees of phase jitter. This adds to the 4.4 degrees in the range to 50 Hz. If the data in figure 11.7 is integrated from -2.5 kHz to -50 kHz and 50 Hz to 2.5 kHz it gives 61 degrees of phase jitter which is a spurious result. One needs to terminate the integration several bandwidths away from zero offset. Unfortunately insufficient data was collect to be sure of the phase noise contribution between 50 Hz and 200 Hz, ideally a span of 1 kHz should also have been recorded. Terminating the
integration of figure 11.7 at +/- 100 Hz gives an r.m.s. jitter of 16 degrees and this is our best estimate.

The LLRF system varies the phase and amplitude of the input to cancel microphonics. The controller was implemented with a DSP and it ran a simple PI controller. Figure 11.8 shows how the microphonic spectrum is reduced as the gain is increased from 0.2 to 10.
The performance of the LLRF system depends on the proportional and integral terms of the controller. The gain used in figure 11.8 and subsequent figures is a multiplier for both the integral and proportional terms. In figure 11.8 the green curve is the microphonic level when the controller is off and is the same as that shown in figure 11.6. A gain of 0.2 reduces the principle microphonic peaks by about 15 dB which is a factor of 30. A gain of 1.0 reduces the largest peak by a factor of 25 dB which is a factor of 300. The response is not linear as the PI controller has an integral term. At a gain of 10.0 the microphonic spectrum has disappeared and the jitter appears comparable with that of the source as shown in figure 11.2.

The DSP clock speed could be varied between 50 MHz and 225 MHz. The ratio of the integral term to the proportional term had been optimised for a 50 MHz DSP clock speed but had not been normalised against the DSP clock speed. Figure 11.9 shows the enhanced performance when the DSP clock speed is increased to 200 MHz. A gain of 0.2 at 200 MHz is approximately equivalent to a gain of 1 at 50 MHz.

Increasing the clock speed “per se”, increased the integral term with respect to the proportional term but also reduces control system delay. Reducing delay moves the stability limit and hence the gain can be increased.

Figure 11.10 shows the cavity phase noise spectrum at maximum controller gain and DSP speed against the noise spectrum and the un-controlled spectrum. The cavity spectrum is indistinguishable from the source noise. From figure 11.9 it is apparent that the cavity spectrum was indistinguishable from the source spectrum at a gain of 1 let alone a gain of 10. The stability limit for the gain at a DSP clock speed of 200 MHz was somewhere between 10 and 20. Lock was never lost at a gain for 10 but was frequently lost at a gain of 20. This result was determined by ramping the gain over a period of a few seconds. The fact that a factor of 10 increase in gain was available after the cavity noise spectrum was identical to the source spectrum one supposes that the jitter associated with the cavity output in this case was ten
times better than the source. The source noise could not be eliminated in this experiment as the source was the reference for the phase detectors measuring the cavity phase and the cavity delays its output with respect to the source by the reciprocal of the bandwidth = 2.5 ms. There is an additional contribution from the noise floor of the spectrum analyser.

![Source vs Output Spectrum for Cavity 3, Unlocked and Locked at High Gain](image1)

**Figure 11.10** Spectral Output of Cavity 3 when phase locked with 200 MHz DSP clock

![Spectrum of Cavity 3 vs Gain at DSP Clock Speed of 200MHz](image2)
Figure 11.11 Spectral output of cavity 3 vs gain for 5 kHz bandwidth

For removal of phase jitter at higher frequencies beyond 50 Hz higher DSP clock speed are important. Figure 11.11 shows performance for a bandwidth of 5 kHz.

Interestingly figure shows at low gains the LLRF system can create additional unwanted phase noise above 500 Hz with respect to the unlocked cavity. When the gain is increased to 10 the phase noise is again comparable with the source.

Figure 11.12 shows the spectrum of forward power to the cavity when it is locked as determined by the controller plotted against the microphonic spectrum when it is unlocked. As one would expect the plots are strongly correlated.

11.3 Phase detector measurements for cavity 3 during lock

As well as inferring phase jitter from spectral output it was measured directly with respect to the source using frequency dividers and a digital phase detector. The phase detector was calibrated by splitting a signal from the signal generator, shifting one leg with a calibrated manual phase shift and comparing the phases of the two legs in the phase detector after division of each. Calibration was dependent on the gain of the differential amplifiers used with the digital phase detectors. In this instance the calibration gave 7.5 mV per degree.

Measuring the phase jitter between the cavity and the source for the unlocked cavity and with a bandwidth of 500 kHz gave a peak to peak signal output of 225 mV corresponding to 30° peak to peak which is about 10° r.m.s..

When the cavity was locked the peak to peak jitter was better than 3 mV on the same bandwidth. This implies a peak to peak jitter of 400 milli-degrees peak to peak or 140 milli-degrees r.m.s.. This jitter includes source noise, ADC noise and some oscilloscope noise.
Inspection of table 11.1 indicated that the source noise was typically 140 milli-volts r.m.s. hence all we can say is that the locking performance was substantially better than 140 milli-degrees r.m.s..

11.4 Locking Results for Cavity 1

Figure 11.14 shows a screen shot of cavity 1 in lock. At first sight it seems similar to cavity 3.

Figure 11.14  Spectral Output of Cavity 1 when gain10 and DSP clock at 50 MHz

Figure 11.15 plots the output spectrum of cavity 1 taken as data a minute after the screen shot in figure 11.15. The spectrum is compared it the spectrum of cavity 3.
Figure 11.15 Locked spectrum of cavity 1 vs cavity 3 at high gain

In figure 11.15 the spectrum for cavity 3 shows a peak at 3 Hz not present for cavity 1. This peak could well be associated with the cavity drift. In most other respects cavity 1 locked just as easily as cavity 3. By continual adjustment of the tuner to minimise the required input power for set point amplitude operation it was possible to maintain lock for a minute or some times longer until the tuner mechanism invoked an abrupt jump (measured in hundreds of nanometres). Without automatic and smooth tune-ability locking both cavities simultaneously was quite challenging.
12. Synchronisation Results

12.1 August 2008 tests
The target phase control performance at 3.9 GHz was 120 milli-degrees r.m.s.. Having demonstrated lock with each of the cavities separately it was necessary to bring them to the same natural frequency to lock them. Because of the large drift of cavity one this proved to be difficult. There was also some drift on cavity three. As both cavities were driven from the same source they would always have the same output frequency. The problem we had is that when cavity one’s natural frequency was a few kHz from the drive frequency the output was too low for the phase detectors to function correctly. Moreover the amplifier could not supply enough power to get a useful amplitude level. The strategy for locking was firstly to adjust the signal generator to the frequency of cavity three and then activate its DSP controller. The next step was to pull by hand on the tuner lever for cavity one to get its output to a high level (typically – 5 dBm) at this point its natural frequency is close to the drive (and hopefully cavity three). Once the natural frequency of cavity one was correct its DSP controller was activated. At this instant there was a tendency for Lorentz detuning to knock cavity three off lock. If and when lock was achieved on both cavities simultaneously one would need to adjust the pull on the tuning lever so that power to cavity three was minimised. If during this time cavity three had drifted we had to start again. Whilst simultaneous lock was achieved on a small number of occasions our skill at watching the power meter and adjusting the lever was inadequate to hold the lock for a sufficient period to perform careful calibration of the double balanced mixer.

Figure 12.1 shows an instance when both cavities were locked.

![Figure 12.1 Simultaneous lock](image)

In this instance the voltage jitter on the oscilloscope measuring output from the double balanced mixer was 50 mV peak to peak. Prior to the locking cavity one but with cavity three locked the double balance mixer was measuring 2 volts peak to peak. The system had reduced
the jitter by a factor of 40. The microphonic level was similar to that measured previously and would have been about 4 degrees r.m.s. Dividing 4 degrees by 40 means that 50 mV peak to peak corresponds to 100 milli-degrees rms. This suggests that the target of 120 milli-degrees had been met. Note that for warm cavities we achieved 75 milli-degrees. Given that the effective jitter (translated to 3.9 GHz) from the digital phase detectors was probably at least 25 milli-degrees and four were in use then these alone contribute 50 milli-degrees. Adding contributions from dividers and ADCs then 75 milli-degrees is about the limit on what one might expect to achieve with this system.

During the course of the tests considerable time was consumed dealing with the following issues:-

- finding enough couplers and cables for all the instruments that were introduced to the test
- adjustment of output levels
- understanding the drift and retuning cavity one
- waiting for cavity one to relax after compression
- eliminating a spurious resonance from a faulty amplifier
- investigating a spurious resonance at an offset of 900 Hz when the DSP clock was at its highest speed of 225 MHz.
- resolving an issue of poor lock when the I and Q cables had been apparently interchanged on the vector modulator. This we realised was due to the fact the control algorithm we had implemented would only lock over a phase range of about 120° and our modified set up to the warm tests had gained phase on the return loop.

By the final day of tests we had mastered a technique for getting the cavities to lock but unfortunately the helium level had dropped to the cavity neck and microphonics has become exceptionally high.

12.2 November 2008 Tests

The experiment was repeated somewhat unsuccessfully in November 2008. For this test we had made slight modifications to the tuners in the hope of improved performance. We wrote a new control algorithm that could pull the lock from a full 360°. We mounted the double balance mixer in a heated chamber within the cryostat. We also used embedded interferometer terminations as shown in figure 7.4.

Unfortunately during cool down the heater in the double mixer can turned itself off due to a LabView error and the component and its associated 3 dB splitters were cooled to 4 K. The mixer and splitters were irreversibly damaged.

On energisation, both cavity outputs were about 30dB below outputs for the August tests, this was mainly due to external Q variation during the cavity mounting. As a consequence considerable amplification was needed prior to the phase detectors. Control loop noise in a band to 1 MHz reduced the maximum gain that could be used in the control loop by a factor of at least ten.

In these tests the repaired tuners still had a tendency to jump during adjustment but had far less drift when left alone. Once the two cavities locked they sometimes remained locked for a hour or so without any tuning adjustments. Due to the reduced gain of the controller the best synchronisation achieved was about 136 milli-degrees r.m.s. cavity to cavity compared to something better than 100 milli-degrees for the August tests.

Figure 12.2 shows a short time slice of double balanced mixer output recorded on an oscilloscope for the phase difference between the two cavities for the unlocked and locked cases. Note that the amplitude scale in (b) is 20 times larger than in (a).
By performing an integration of ten 40ms long traces taken over a period of half an hour, r.m.s. phase jitter with both cavities locked was calculated to be 136 milli degrees. Most of the jitter is introduced by 1kHz component, the natural frequency of the loop when it oscillates. This is visible in the Fourier transform of the oscilloscope trace shown figure 12.3.

Figure 12.2  Phase jitter between Cavity 1 and Cavity 3, 1mV = 20 milli degrees
13. Conclusions

In a first test conducted during August 2008 we have demonstrated that an RF interferometer used with digital control and digital phase detection can lock a pair of superconducting cavities having realistic levels of microphonics such that r.m.s. phase errors are less than 120 milli-degrees at 3.9 GHz and probably down to 75 milli-degrees r.m.s.. Due to problems with the cavity tuners the experimental tests failed to gather data over long periods of time thereby permitting accurate calibration.

The result is also limited in the sense that single cell cavities were used, no beam was present and the environment was that of a vertical test facility. The locking of a nine cell cavity as proposed for the ILC has been studied and reported in a separate project document\(^9\). In that study the additional phase error associated with one’s ability to estimate excitation of the pi mode in the presence of other modes adds an additional 25 milli-degrees r.m.s. of uncertainty. Even with this addition the RF system under test quite probably satisfies the ILC specification with a little to spare.

13.1 Related Crab Cavity Studies

The study within this report is specific to the LLRF control and synchronisation of crab cavities for the ILC. A major focus of the overall EuroTeV crab cavity task and not covered in this report has been the design of a nine cell cavity with couplers appropriate to ILC parameters\(^{10}\). An important part of this work has been the computation of wakefields\(^{11}\) and the design of LOM, SOM and HOM couplers\(^{12}\). The base cavity design adopted was essential that of a Fermi-lab cavity developed for Kaon separation\(^{13}\).

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\(^9\) EuroTeV report 2008-064 already sited as reference 5.

\(^{10}\) EuroTeV report 2007-10 already sited as reference 1.


\(^{12}\) G. Burt, P. K. Ambattu, A. Dexter, P. Goudket, P. McIntosh, L. Bellantoni, Z. Li, L. Xiao, “Copper prototype measurements of the HOM, LOM and SOM couplers for the ILC crab cavity”, “, MOPP005, EPAC 2008, Genoa, Italy

14. Future LLRF system work

The LLRF system approach investigated by here was deliberately chosen to be different systems a DESY and CERN so that work was not being duplicated. There are two obvious modifications to the system investigated here which maintain the RF interferometer but offer better ultimate phase performance. The first is to replace the digital phase detectors with double balanced mixers, having done this the interferometer can be operated at 3.9 GHz and the frequency dividers are no longer needed. Before this is done we wish to fully investigate then benefits of digital phase detection which might arise from their linearity and insensitivity to amplitude. The second is to measure phase by down conversion and digital sampling. In the short term future work is expected include.

1. Continued develop monolithic phase detector board.
2. Interface FPGA to DSP on cavity control board
3. Replace analogue. loop filter on interferometer with FPGA controllers
4. Develop automatic calibration of interferometer.
5. Develop interface with cavity tuners.
6. More performance test on pairs of superconducting cavities at 1.3 GHz and 3.9 GHz.
7. Investigate the applicability of advance control algorithms beyond PI.
8. Develop RF controller for active damping of the SOM.

15. Acknowledgements

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16. Appendix 1  Cavity Theory

16.1 Two Port - Multi-Mode Equations

After diagonalisation of any equivalent circuit representation of a cavity driven from a current source, the resulting simplified equivalent circuit is composed of a number parallel resonators stacked in series. Each resonator models a discrete mode. Figure 15.1 shows the equivalent circuit of a cavity with two modes of interest and two ports.

At the terminals the voltage in each waveguide must equal the voltage in the lumped circuit. Along each waveguide the voltage and current satisfies the equations

\[ \frac{\partial V}{\partial z} = -L_{wg} \frac{\partial I}{\partial t} \quad \text{and} \quad \frac{\partial I}{\partial z} = -C_{wg} \frac{\partial V}{\partial t} \]

hence

\[ \frac{\partial^2 V}{\partial z^2} = L_{wg} C_{wg} \frac{\partial^2 V}{\partial t^2} \]

where \( C_{wg} \) is the capacitance per unit length and \( L_{wg} \) is the inductance per unit length. In this analysis the two waveguide will have differing impedances as required to get the appropriate Q factors for the two ports. Subsequent matching sections to match the port impedances to standard waveguide are not shown or included in the analysis as their functionality is transparent.

For a fixed frequency source of angular frequency \( \omega \) the voltage along each waveguide section is given as

\[ V(z, t) = \mathcal{F} \exp \left\{ j(kz - \omega t) \right\} + \mathcal{R} \exp \left\{ -j(kz + \omega t) \right\} \]  

(15.1)

where \( k = \omega \sqrt{L_{wg} C_{wg}} \)

and \( \mathcal{F} \) is the amplitude of the forward wave and \( \mathcal{R} \) is the amplitude of the reflected wave, both different for each section.

The current on the waveguides are therefore given as

\[ I(z, t) = \mathcal{F} \frac{\omega C_{wg}}{k} \exp \left\{ j(kz - \omega t) \right\} - \mathcal{R} \frac{\omega C_{wg}}{k} \exp \left\{ -j(kz + \omega t) \right\} \]

which can be written as

\[ I(z, t) = \frac{1}{Z_{wg}} \left[ \mathcal{F} \exp \left\{ j(kz - \omega t) \right\} - \mathcal{R} \exp \left\{ -j(kz + \omega t) \right\} \right] \]  

(15.2)

where \( Z_{wg} = \sqrt{\frac{L_{wg}}{C_{wg}}} \)  

(15.3)
If the terminal between the cavity and the waveguides is at $z = 0$ then the sum of the currents from the two waveguides equals the sum of the currents through the equivalent circuit components of each series resonator (i.e. we get an equation for each resonator / mode) hence
\[
\frac{1}{L_i} \int V_i \, dt + C_i \frac{dV_i}{dt} + \frac{V_i}{R_i} = \frac{1}{Z_{wg1}} \{F_i - R_{d1}\} \exp(\omega_j t) + \frac{1}{Z_{wg2}} \{F_2 - R_{d2}\} \exp(\omega_j t)
\]  
where $F_i$, $R_{d1}$, $F_2$ and $R_{d2}$ are forward and reflected amplitudes in the two waveguides.

If the coupling to different modes is dissimilar then $Z_{wg}$ takes a different value for each mode, From (15.1) and adding series voltages for each mode the voltage at $z = 0$ is given as
\[
V = \sum_{\text{modes}} V_i = (F_i + R_{i}) \exp(-\omega_j t) = (F_2 + R_{d2}) \exp(-\omega_j t)
\]

For compactness we now introduce the suffix $wgij$ where $i$ refers to the mode and $j$ refers to the port. Eliminating the reflected power between (15.4) and (15.5) gives
\[
\frac{1}{L_i} \int V_i \, dt + C_i \frac{dV_i}{dt} + \frac{V_i}{R_i} + \sum_{j=1}^{2} \frac{1}{Z_{wgij}} \sum_{k=1}^{N} V_k = \sum_{j=1}^{2} \frac{2F_j}{Z_{wgij}} \exp(-\omega_j t)
\]

This equation determines the modal voltages in the cavity as a function of the amplitude of the forward waves in the two waveguides.

Now define the natural frequency of the $i$th mode as
\[
\omega_i = \frac{1}{\sqrt{L_i C_i}}
\]

To evaluate $Z_{wgij}$ we write
\[
Q_{ej} = \frac{\omega_i U_{\text{stored}}}{U_{\text{emitted}}} = \frac{1}{2} \frac{\omega_i C_i V_i^2}{V_i^2 / Z_{wgij}} = \omega_i Z_{wgij} C_i
\]
\[
Q_{oi} = \frac{\omega_i U_{\text{stored}}}{U_{\text{diss}}} = \frac{1}{2} \frac{\omega_i C_i V_i^2}{V_i^2 / R_i} = \omega_i R_i C_i
\]

Hence dividing (15.8) by (15.9) we have that
\[
\frac{Q_{ej}}{Q_{oi}} = \frac{Z_{wgij}}{R_i}
\]

which can be re-arranged as
\[
Z_{wgij} = \left(\frac{R_i}{Q_{oi}}\right) Q_{ej}
\]

i.e. $Z_{wgij}$ is the product of the external $Q$ with the $R/Q$ of the bare cavity. Note that $Z_{wgij}$ is not that of the physical waveguide from the RF generator as represented in figure 15.1 by the transmission line from the current source to the transformer. The transformer models the coupler which transforms the voltage.

Differentiation of (15.6) and division by $C_i$ gives
\[
\frac{d^2 V_i}{dt^2} + \frac{\omega_i}{\omega_j R_i C_i} \frac{dV_i}{dt} + \sum_{j=1}^{2} \frac{\omega_i}{\omega_j Z_{wgij} C_i} \sum_{k=1}^{N} \frac{dV_k}{dt} + \frac{1}{L_i C_i} V_i = \sum_{j=1}^{2} \frac{2\omega_i}{\omega_j Z_{wgij} C_i} \frac{d}{dt} \left(F_j \exp(-\omega_j t)\right)
\]
and using (15.7), (15.8) and (15.10) in (15.11) gives

$$\frac{d^2V_i}{dt^2} + \frac{\omega_i}{Q_{oi}} \frac{dV_i}{dt} + \sum_{j=1}^{2} \frac{\omega_i}{Q_{eij}} \sum_{k=1}^{N} \frac{dV_k}{dt} + \omega_i^2 V_i = \sum_{j=1}^{2} \frac{2\omega_i}{Q_{eij}} \frac{d}{dt} \left[ F_j \exp(-j\omega t) \right]$$

(15.13)

defining

$$\frac{1}{Q_{Li}} = \frac{1}{Q_{oi}} + \sum_{j=1}^{2} \frac{1}{Q_{eij}}$$

(15.14)

equation (15.12) becomes

$$\frac{d^2V_i}{dt^2} + \frac{\omega_i}{Q_{Li}} \frac{dV_i}{dt} + \sum_{j=1}^{2} \frac{\omega_i}{Q_{eij}} \sum_{k=1}^{N} \frac{dV_k}{dt} + \omega_i^2 V_i = \sum_{j=1}^{2} \frac{2\omega_i}{Q_{eij}} \frac{d}{dt} \left[ F_j \exp(-j\omega t) \right]$$

(15.15)

16.2 Single mode single port steady state solution

Specialising the result in (15.15) to a single mode and a single port and shifting the phase of the incoming wave by $\pi/2$ so that the applied field at $t = 0$ is zero gives

$$\ddot{V} + \frac{\omega_c}{Q_L} \dot{V} + \omega_c^2 V = -\frac{2\omega_c \omega}{Q_c} F \exp(-j\omega t)$$

(15.16)

where $\omega$ is the drive frequency and $\omega_c$ is the cavity frequency. As the voltage is real and the excitation is real then we discard the complex part to give

$$\ddot{V} + \frac{\omega_c}{Q_L} \dot{V} + \omega_c^2 V = -\frac{2\omega_c \omega}{Q_c} F \cos(\omega t)$$

(15.17)

A driven oscillator will oscillate at the driver frequency once transients have decayed however there will be a phase shift between the driver and the cavity that depends on the natural frequency of the cavity at any instant. Small changes in cavity size give rise to phase jitter.

A change in the relative phase of the driver implies that its frequency must have changed for a period of time.

It follows that one expects the steady state solution of (15.17) to be of the form

$$V = A \cos(\omega t) + B \sin(\omega t)$$

(15.18)

Substitute (15.18) into (15.17) and equate cosine terms gives

$$-\omega^2 A + \frac{\omega_c}{Q_L} \omega B + \omega_c^2 A = -\frac{2\omega_c \omega}{Q_c} F$$

(15.19)

and equating sine terms gives

$$-\omega^2 B - \frac{\omega_c}{Q_L} \omega A + \omega_c^2 B = 0$$

(15.20)

hence

$$B = \frac{1}{Q_L \left( \omega_c^2 - \omega^2 \right)} A = \frac{1}{Q_L \left( \frac{\omega_{eij}}{Q_{eij}} \right)} A$$

(15.21)

substitution of (15.21) in (15.19) gives
\[
\left(\omega_c^2 - \omega^2\right)A + \frac{1}{Q_L} \frac{\omega_c^2 \omega^2}{\left(\omega_c^2 - \omega^2\right)} A = -\frac{2\omega_c \omega}{Q_c} \mathcal{F}
\]

hence

\[
A = -\frac{2\omega_c \omega}{Q_c} \mathcal{F} \frac{Q_L^2 \left(\omega_c^2 - \omega^2\right)}{Q_L^2 \left(\omega_c^2 - \omega^2\right)^2 + \omega_c^2 \omega^2} = -\frac{\mathcal{F} Q_L}{Q_c} \frac{Q_L \left(\frac{\omega_c}{\omega} - \frac{\omega}{\omega_c}\right)}{1 + Q_L^2 \left(\frac{\omega_c}{\omega} - \frac{\omega}{\omega_c}\right)^2}
\]

Hence from (15.21)

\[
B = -\frac{\mathcal{F} Q_L}{Q_c} \frac{1}{\left\{1 + Q_L^2 \left(\frac{\omega_c}{\omega} - \frac{\omega}{\omega_c}\right)^2\right\}}
\]

Hence after transients have died away

\[
V(t) = -\frac{2 Q_L}{Q_c} \mathcal{F} \frac{Q_L \left(\frac{\omega_c}{\omega} - \frac{\omega}{\omega_c}\right) \cos(\omega t) + \sin(\omega t)}{1 + Q_L^2 \left(\frac{\omega_c}{\omega} - \frac{\omega}{\omega_c}\right)^2}
\]

\[
= -\frac{2 Q_L}{Q_c} \mathcal{F} \frac{\sin(\omega t + \phi)}{\sqrt{1 + Q_L^2 \left(\frac{\omega_c}{\omega} - \frac{\omega}{\omega_c}\right)^2}}
\]

\[
= -\frac{2 Q_L}{Q_c} \mathcal{F} \cos \phi \sin(\omega_d t + \phi)
\]

where

\[
\cos(\phi) = \frac{1}{\sqrt{1 + Q_L^2 \left(\frac{\omega_c}{\omega} - \frac{\omega}{\omega_c}\right)^2}} \quad \text{and} \quad \sin(\phi) = \frac{Q_L \left(\frac{\omega_c}{\omega} - \frac{\omega}{\omega_c}\right)}{\sqrt{1 + Q_L^2 \left(\frac{\omega_c}{\omega} - \frac{\omega}{\omega_c}\right)^2}}
\]

Equation (15.27) tells us that when the phase shift between the applied current at the coupler and the cavity is \(\phi\), then the amplitude is reduced by a factor \(\cos \phi\). From (15.28) we have

\[
\tan(\phi) = Q_L \left(\frac{\omega_c}{\omega} - \frac{\omega}{\omega_c}\right)
\]

Defining

\[
\delta \omega = \omega_c - \omega
\]

and using (15.22)
\[ \tan(\phi) = Q_L \left( \frac{\omega_c - \omega}{\omega - \omega_c} \right) = Q_L \left( \frac{\omega_c - \delta \omega}{\omega - \delta \omega} \right) \sim \frac{2 Q_L \delta \omega}{\omega} \]  

(15.31)

hence

\[ \cos(\phi) \sim \frac{1}{\sqrt{\left(\frac{2 Q_L \delta \omega}{\omega_c}\right)^2 + 1}} \quad \text{and} \quad \sin(\phi) \sim \frac{\left(\frac{2 Q_L \delta \omega}{\omega_c}\right)}{\sqrt{\left(\frac{2 Q_L \delta \omega}{\omega_c}\right)^2 + 1}} \]  

(15.32)

Equation (15.32) tells us that when \( Q_L \) is low the phase shift \( \phi \) between the drive and the cavity response tends to zero, i.e. \( \phi \sim 0 \) and when \( Q_L \) is high \( \phi \sim \pi/2 \).

Equation (15.32) also shows that the sign of the phase determines whether that cavity natural frequency is above or below the drive frequency and hence equation (15.31) provides a tuning algorithm. It also allows the frequency deviation associated with microphonics to be determined from the phase jitter when the cavity is locked.

For \( \frac{\delta \omega}{\omega_c} >> 1 \) equation (15.27) tends to

\[ V(t) = \frac{1}{Q_c} F \omega_c \cos(\omega t) \]  

(15.33)

For \( \frac{\delta \omega}{\omega_c} << 1 \) equation (15.27) tends to

\[ V(t) = -2 \frac{Q_L}{Q_c} F \sin(\omega t) \]  

(15.34)

16.3 Power Requirement for parallel LCR

If the cavity is matched to the input wave guide then the impedance of the waveguide = \( R \) and the r.m.s. forward power is given as

\[ P_f = \frac{FF^*}{2Z_{wg}} = \frac{FF^*}{2Q_c \left( \frac{R}{Q_0} \right)} \]  

(15.35)

For an application where the voltage in the circuit must be maintained at a set point given as

\[ V(t) = V_{sp} \sin(\omega_s t + \phi) \]  

(15.36)

Then from (15.26)

\[ V_{sp} = -2 \frac{Q_L}{Q_c} F \frac{1}{\sqrt{1 + Q_L^2 \left( \frac{\omega_c - \omega}{\omega} \right)^2}} \]  

(15.37)

Hence the forward power \( P_f \) is given by

\[ P_f = \frac{V_{sp}^2}{2} \left( \frac{R}{Q_0} \right)^{-1} \frac{Q_c}{4Q_L^2} \left\{ 1 + Q_L^2 \left( \frac{\omega_c - \omega}{\omega} \right)^2 \right\} \]  

(15.38)

Power dissipated in the cavity \( P_o \) is given by

\[ P_o = \frac{V_{sp}^2}{2R} \]  

(15.39)
Hence (15.38) can also be written

$$\frac{P_f}{P_o} = \frac{Q_c Q_o}{4 Q_L^2} \left( 1 + Q_L^2 \left( \frac{\omega_c - \omega}{\omega \omega_c} \right)^2 \right)$$

(15.40)

Using (15.30) the forward power can therefore be accurately approximated as

$$\frac{P_f}{P_o} \approx \frac{Q_c Q_o}{4 Q_L^2} \left( 1 + Q_L^2 \left( \frac{2 \delta \omega}{\omega_c} \right)^2 \right)$$

(15.41)

Hence

$$\frac{P_f}{P_o} \approx \frac{Q_c Q_o}{4 Q_L^2} \left( \left( \frac{2 \delta \omega}{\omega f \text{loaded bandwidth}} \right)^2 + 1 \right)$$

where loaded bandwidth = $\frac{\omega_c}{Q_L}$. In practice one increases the cavity bandwidth as $\delta \omega$ increases, this means that

$$\frac{P_f}{P_o} \propto \frac{Q_c Q_o}{Q_L^2} \left( \frac{\omega_c}{\omega f \text{unloaded bandwidth}} \right)$$

(15.42)

i.e. power requirement increases in direct proportion to the frequency deviation caused by microphonics (where the bandwidth is adjusted to meet frequency deviation).

### 16.4 Fill Time for parallel LCR

Suppose at $t = 0$ the voltage is zero and has been zero for a long time. Also suppose that the power is switched on at $t = 0$ after being zero for a long time. The solution after $t = 0$ will have a transient as well as a steady state part. The transient can be determined from solutions of the auxiliary equation

$$m^2 + \frac{\omega_c}{Q_L} m + \omega_c^2 = 0$$

(15.43)

i.e. $m = -\omega_c/2Q_L \pm j \omega_p$

(15.44)

where $\omega_p = \omega_c \sqrt{1 - \frac{1}{4Q_L^2}}$

(15.45)

Combining the auxiliary solution with the particular integral (15.27) gives for $t > 0$

$$V(t) = \exp \left( -\frac{\omega_c t}{Q_L} \right) \left[ A \cos(\omega_p t) + B \sin(\omega_p t) \right] - 2Q_L F \frac{\cos \phi \sin (\omega t + \phi)}{Q_c}$$

(15.46)

Coefficients $A$ and $B$ are determined from the boundary conditions at $t = 0$ which are

$$V(0) = 0 \quad \text{and} \quad \int_{-\infty}^{0} V dt = 0$$

(15.47)

hence with the first boundary condition equation (15.46) gives

$$A = 2Q_L F \frac{\cos \phi \sin \phi}{Q_c}$$

(15.48)
Integrating equation (15.16) from $-\infty$ to 0, using both boundary conditions in (15.47) and the fact that $\dot{F}$ has been zero for the integration period gives

$$\frac{dV}{dt}_{t=0} = 0 \quad (15.49)$$

Differentiation of (15.46) and evaluating at $t = 0$ gives

$$\frac{dV}{dt}_{t=0} = -\left(\frac{\omega_c}{2Q_L}\right) A + \omega_c B - 2\omega \frac{Q_L}{Q_c} \dot{F} \cos \phi \cos \dot{\phi} \quad (15.50)$$

Combining (15.49), (15.50) and (15.48) gives

$$0 = -\frac{\omega^2_e}{Q_c} \dot{F} \cos \phi \sin \phi + \omega_c B - 2\omega \frac{Q_L}{Q_c} \dot{F} \cos^2 \phi$$

hence

$$B = \dot{F} \left[ \frac{1}{Q_c} \cos \phi \sin \phi + 2 \frac{\omega \frac{Q_L}{Q_c}}{\omega_e \frac{Q_L}{Q_c}} \cos^2 \phi \right] \quad (15.51)$$

If we fill on resonance then $\phi = 0$ and $\omega = \omega_e$ hence

$$V_{res}(t) = 2 \frac{Q_L}{Q_c} \dot{F} \left\{ \frac{\omega}{\omega_e} \sin(\omega_e t) \exp\left( -\frac{\omega_e t}{2Q_L} \right) - \sin(\omega t) \right\}$$

$$= 2 \frac{Q_L}{Q_c} \dot{F} \left\{ \exp\left( -\frac{\omega t}{2Q_L} \right) - 1 \right\} \sin(\omega t) \quad (15.52)$$

If we fill off resonance the form of voltage as a function of time as given by (15.46), (15.48) and (15.51) and depends on the steady state phase shift $\phi$ which in turn depends on the relative magnitude of $\frac{\omega^2_e}{Q_e} - \omega^2$ and $\frac{\omega_e \omega}{Q_L}$. These terms will become equal for a specific value of $Q$ which we will call $Q_{crit}$. Explicitly this condition is written

$$\omega^2_e - \omega^2 = \frac{\omega_e \omega}{Q_{crit}} \quad (15.53)$$

Using (15.30) the condition can also be written as

$$Q_{crit} = \frac{1 + \frac{\delta \omega_c}{\omega_c}}{\frac{2 \delta \omega_c}{\omega_c} \left( 1 + \frac{\delta \omega_c}{2 \omega_c} \right)} \quad (15.54)$$

The condition relates to whether the exponential term in (15.46) decays before the phase difference between the driven phase and phase by which the cavity would advance when oscillating freely, reaches $\pi/2$.

To a very good approximation

$$Q_{crit} \approx \frac{\omega_e}{2 \delta \omega_c} \quad (15.55)$$

It is useful to plot the magnitude and phase of (15.46) for values of $Q$ above and below $Q_{crit}$.
Figure 15.2 gives the development of the voltage for $Q = 10$ and $Q_{\text{crit}} = 100$. The frequency of the drive was 3.9 GHz and the natural frequency of the cavity was $+0.5\%$. For this case approximate fill time $\sim \frac{Q}{\omega_c} = 4 \times 10^{-9}$ s.

![current for Q critical = 100, Q = 10, psi = 0](image)

Figure 15.2

Figure 15.3a gives the envelope of figure 1, i.e. the magnitude of the current as a function of time. Figure 15.3b gives the corresponding phase as a function of time.

Figures 15.4a and 15.4b give the magnitude and phase for $Q = 100$ and $Q_{\text{crit}} = 100$. In this case the magnitude is 9 times bigger than for $Q = 10$ and the phase error was also larger.

Figures 15.5a and 15.5b give the magnitude and phase for $Q = 500$ and $Q_{\text{crit}} = 100$. In this case the final magnitude is similar to $Q = 100$ however the cavity gets overfilled initially and takes some time to settle to the final fill level.

![peak output for Q critical = 100, Q = 10](image)

Figure 15.3

![phase delay for Q critical = 100, Q = 10](image)

![peak output for Q critical = 100, Q = 100](image)

![phase delay for Q critical = 100, Q = 100](image)

Figure 15.4
Figure 15.5

peak output for $Q_{\text{critical}} = 100$, $Q = 500$

phase delay for $Q_{\text{critical}} = 100$, $Q = 500$
17. Appendix 2 RMS Phase Jitter from Spectrum Analyser

17.1 Single frequency phase jitter

Suppose we have a signal with phase jitter given as

\[ V = V_o \cos\left(\omega t + \phi_n \cos(\omega_n t)\right) \]  

(16.1)

where \( \omega \) is the carrier frequency

\( \omega_n \) is the frequency of the phase noise

\( \phi_n \) is the peak amplitude of the phase deviation

The voltage \( V \) can be decomposed into its Fourier components, which can be expressed in terms of Bessel functions. When \( \phi_n \) is small the decomposition is very simple and is made as follows:-

\[ V = V_o \{\cos(\omega t) \cos(\omega_n t) - \sin(\omega t) \sin(\omega_n t)\} \approx V_o \{\cos(\omega t) - \phi_n \cos(\omega_n t) \sin(\omega t)\} \]

Hence

\[ V \approx V_o \left\{\cos(\omega t) - \frac{\phi_n}{2} \sin[(\omega + \omega_n) t] + \frac{\phi_n}{2} \sin[(\omega - \omega_n) t]\right\} \]  

(16.2)

It can be seen therefore that the Fourier coefficients determine the maximum phase departure for any particular disturbance frequency.

17.2 Spectral Density

Typically spectrum analysers measure Power Spectral Density (PSD). The PSD is mathematically defined as the Fourier transform of the autocorrelation function. The PSD is equivalently determined from the squared modulus of coefficients in the Fourier Series.

Invariably a time varying voltage signal will be sampled over a finite time period \( T \); such a signal is illustrated in figure 1.

![Figure 1](image)

**Figure 1** Time signal with apparent auto-correlation.

The signal in figure 6 has an apparent pattern of repetition and hence has a degree of auto-correlation. From Fourier’s theorem it is known that this time signal can be expressed as a series of sine and cosine waves i.e.

\[ V(t) = \frac{a_0}{2} + \sum_{n=\pm1} a_n \cos \frac{2\pi nt}{T} + b_n \sin \frac{2\pi nt}{2T} \]  

(16.3)
where the coefficients $a_n$ and $b_n$ are determined by the integrals

$$a_n = \frac{2}{T} \int_0^T V(t) \cos \frac{2\pi n t}{T} \, dt$$  \hspace{1cm} (16.4a)$$

and

$$b_n = \frac{2}{T} \int_0^T V(t) \sin \frac{2\pi n t}{T} \, dt$$  \hspace{1cm} (16.4b)$$

The frequencies of the waves used to construct the time domain voltage signal are given by

$$f_n = \frac{n}{T}$$  \hspace{1cm} (16.5)$$

It is apparent therefore that the precise frequencies which are present in the Fourier series depend on the sampling interval.

Suppose for instance that the sampling interval is 1 ms and we are expecting a frequency of 100 MHz then $n = 10^5$. So if 100 MHz is present then the next frequency present in the series is

$$f_{n+1} = \frac{10^5 + 1}{10^{-3}} = 100.001 \text{MHz}$$

i.e. we can only resolve frequencies to within 1 kHz and it is not meaningful to talk about frequencies in between unless the sampling interval is increased. If the sampling period is very much longer than the period of the frequencies of interest then the spectral coefficients $a_n$ and $b_n$ in this vicinity can be regarded as continuous functions hence we write

$$a_n = a(f_n, T) = \tilde{a}(f)$$  \hspace{1cm} (16.6a)$$

$$b_n = b(f_n, T) = \tilde{b}(f)$$  \hspace{1cm} (16.6b)$$

A spectrum analyser displays $\tilde{a}^2(f) + \tilde{b}^2(f)$ with a normalisation that gives Watts per Hz. Normalisation is possible by virtue of Parseval’s identity that gives

$$\frac{2}{T} \int_0^T V^2(t) \, dt = \frac{a_0^2}{2} + \sum_{n=1}^{\infty} \left( a_n^2 + b_n^2 \right)$$  \hspace{1cm} (16.7)$$

i.e. the integral on the left is the voltage squared which is proportional to the total power and Parseval’s identity tells that the sum of the square of the Fourier coefficients is also proportional to the total power.

It follows from Parseval’s identity that as one goes to a longer sampling period the frequencies get closer together and hence the coefficients $a_n$ and $b_n$ get proportionally smaller. The spectrum analyzer averages the squares of these values in a frequency range i.e.

$$\tilde{a}^2(f) + \tilde{b}^2(f) = \frac{1}{\delta f} \sum_{f} \left( a_n^2 + b_n^2 \right)$$  \hspace{1cm} (16.8)$$

and hence the display has units of Watts per Hertz. In order to recover the value of $\left( a_n^2 + b_n^2 \right)$ for a particular frequency one must multiply $\tilde{a}^2(f) + \tilde{b}^2(f)$ by its normalization which is invariably 1 Hz.
17.3 Phase Jitter

If the spectral density has been measured, then at a given frequency offset from the carrier
equation (16.2) can be used to determine the phase deviation associated with that frequency
where those phase deviations are small.

Using (16.2) and where the carrier voltage has been normalized to one, the peak to peak phase
range will be given by $4a_n$ hence the r.m.s. phase range is $\sqrt{2}a_n$.

If we have a large number of frequencies present in the signal and they are uncorrelated then
they each add a phase error. These phase errors are by definition sinusoidal and all have
differing frequencies. One can therefore add random phasors of appropriate magnitudes to
get the overall phase deviation. Random phases add as length vectors in a Brownian motion
random i.e.

\[
\begin{align*}
\text{Net distance travelled} = \sum_{n} \delta L_i & \approx \sqrt{\sum_{n} |\delta L_i|^2} = \sqrt{n} |\delta L_{\text{rms}}| \quad \text{where} \quad \delta L_i = \text{vectored step length.}
\end{align*}
\]

Accordingly r.m.s. phase deviation in radians is given by

\[
\begin{align*}
\text{RMS phase error} = \sqrt{2 \sum_{1}^{\infty} (a_n^2 + b_n^2)} = \sqrt{2 \int f \delta f \{ \bar{\alpha}^2(\delta f) + \bar{\beta}^2(\delta f) \}} \quad (16.9)
\end{align*}
\]
### 17.4 Estimation Table for Discrete Peaks in Spectrum

<table>
<thead>
<tr>
<th>dB below carried</th>
<th>fraction of carrier</th>
<th>RMS phase jitter (rads)</th>
<th>RMS phase jitter degrees</th>
</tr>
</thead>
<tbody>
<tr>
<td>-5</td>
<td>3.162E-01</td>
<td>0.795271</td>
<td>45.56566</td>
</tr>
<tr>
<td>-10</td>
<td>1.000E-01</td>
<td>0.447214</td>
<td>25.62345</td>
</tr>
<tr>
<td>-15</td>
<td>3.162E-02</td>
<td>0.251487</td>
<td>14.40913</td>
</tr>
<tr>
<td>-20</td>
<td>1.000E-02</td>
<td>0.141421</td>
<td>8.10285</td>
</tr>
<tr>
<td>-25</td>
<td>3.162E-03</td>
<td>0.079527</td>
<td>4.55657</td>
</tr>
<tr>
<td>-30</td>
<td>1.000E-03</td>
<td>0.044721</td>
<td>2.56235</td>
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<tr>
<td>-35</td>
<td>3.162E-04</td>
<td>0.025149</td>
<td>1.44091</td>
</tr>
<tr>
<td>-40</td>
<td>1.000E-04</td>
<td>0.014142</td>
<td>0.81028</td>
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<tr>
<td>-45</td>
<td>3.162E-05</td>
<td>0.007953</td>
<td>0.45566</td>
</tr>
<tr>
<td>-50</td>
<td>1.000E-05</td>
<td>0.004472</td>
<td>0.25623</td>
</tr>
<tr>
<td>-55</td>
<td>3.162E-06</td>
<td>0.002515</td>
<td>0.14409</td>
</tr>
<tr>
<td>-60</td>
<td>1.000E-06</td>
<td>0.001414</td>
<td>0.08103</td>
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<td>-65</td>
<td>3.162E-07</td>
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<td>-70</td>
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<td>0.000447</td>
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<td>-75</td>
<td>3.162E-08</td>
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<tr>
<td>-80</td>
<td>1.000E-08</td>
<td>0.000141</td>
<td>0.00810</td>
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<td>-85</td>
<td>3.162E-09</td>
<td>0.000080</td>
<td>0.00456</td>
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<td>-90</td>
<td>1.000E-09</td>
<td>0.000045</td>
<td>0.00256</td>
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<tr>
<td>-95</td>
<td>3.162E-10</td>
<td>0.000025</td>
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<tr>
<td>-100</td>
<td>1.000E-10</td>
<td>0.000014</td>
<td>0.00081</td>
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</tbody>
</table>

*Table 16.1*
18. Fabrication of 3.9 GHz Single-Cell SRF Niobium Cavities
Report from Niowave, Inc., 1012 N. Walnut St., Lansing, MI 48906 USA

18.1 Summary
Niowave has generated drawings for and fabricated three 3.9 GHz single-cell SRF niobium cavities; all dies and fixtures are also documented. Cavities were modelled from a Microwave Studio design, drawn in SolidWorks, and constructed using standard techniques using fine grain (> 50 μm) niobium with a RRR >300 and a material thickness of 2.8 mm. End flanges are Nb-Ti material using knife-edge conflat seals. Microwave measurements were done using a network analyzer and recorded. Cavity processing and electron beam welding are also documented.

18.2 Cavity Design and Material Properties
Niowave drew a 3.9 GHz single cell cavity using SolidWorks 2007 software, modeling the electromagnetic design provided by Daresbury in Figure 1.1(a). The electronic drawing file was sent in February to Carl Beard and Peter McIntosh (updated version posted in Appendix A). A 3D SolidWorks model of the fabricated cavity is shown in Figure 1.1(b). Universal drawing files and 3D models (.dxf and .stp) are also provided with this final report.

![Figure 17.1](image1.png)

Table 1.1 lists several key properties of the niobium used for fabrication. More extensive material specifications, including chemical composition and mechanical properties, are tabulated in Appendix B.

<table>
<thead>
<tr>
<th>RRR</th>
<th>Grain Size</th>
<th>Nb Wall Thickness</th>
</tr>
</thead>
<tbody>
<tr>
<td>&gt;300</td>
<td>ASTM #6</td>
<td>2.8 mm (±0.15)</td>
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</tbody>
</table>

18.3 Dies and Fixtures
Dies necessary for stamping and forming the cavity tubes and cells were fabricated by Niowave and are shown in Figure 1.2(a). Weld fixtures, Figure 1.2(b), necessary for electron beam welding were also assembled. All dies and machining fixtures were made of 7075 or 6061 aluminum; welding fixtures were constructed with 304 stainless steel.
18.4 Cavity Components and RF Check

Three sets of cells, tubes, and niobium-titanium knife-edge conflat flanges were stamped and machined for assembly of three 3.9 GHz cavities, as shown in Figure 1.3(a). A copper cavity was first fabricated to confirm a frequency of 3.9 GHz, shown in Figure 1.3(c).

Measured frequencies of the Nb cavities for each polarity of the TM 110 mode are recorded in Table 1.2. The *pre-weld* frequencies were measured with a network analyzer on half-cells and beam tubes held in place with conductive tape. Each half-cell had an additional 0.0875" of material length in the equator region. From the frequency measurements and microwave simulations, it was determined to remove 0.0530", with the remaining 0.0345" left to accommodate for weld shrinkage and cool-down frequency shifts.

The *post-weld* frequencies that were measured with a network analyzer on all the cavities in air at room temperature prior to chemical etching are recorded in Table 1.2. The dipole degeneracy was split without the need for flats on the equator, which most likely is due to the axial tube welds.

<table>
<thead>
<tr>
<th>Cavity</th>
<th>Measured Frequency (MHz)</th>
<th>Pre-weld</th>
<th>Post-weld</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>Dipole 1: 3865.06</td>
<td>Dipole 2: 3871.09</td>
<td></td>
</tr>
<tr>
<td></td>
<td>Dipole 1: 3885.25</td>
<td>Dipole 2: 3887.74</td>
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</tr>
<tr>
<td></td>
<td>Dipole 1: 3885.75</td>
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</tr>
<tr>
<td>3</td>
<td>Dipole 1: 3889.75</td>
<td>Dipole 2: 3892.01</td>
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</tbody>
</table>
Figure 1.3  (a) Stamped Nb cells and tubes and Nb-Ti conflat flanges; (b) 3.9 GHz Nb single cell (pre-weld); (c) 3.9 GHz Cu single cell

18.5 Cleaning and Acid Etching

Each cavity component was cleaned in a class 100 certified cleanroom and the Nb cells and tubes were “buffered chemical polishing” (BCP) etched prior to electron beam welding. Components were cleaned following standard cleanroom procedures and BCP etched for 10 minutes at a rate of about 1.5 μm/min. Figure 1.4 shows photographs of the cavity processing.

Figure 1.4 Cavity component processing: (a) ultrasonic cleaning in ultra pure water; (b) BCP etching tubes
18.6 Electron Beam Welding (EBW)

The cavities were electron beam welded at Sciaky, Inc., a world-renowned manufacturer of electron beam systems, utilized also by SRF laboratories such as Argonne, Fermilab and Michigan State University. Under Niowave supervision, all seam and full penetration welds were successfully completed in a timely manner. Photographs of the vacuum chamber are shown in Figure 1.5 along with equator welds.

![Figure 1.5 Sciaky's EBW system: (a) vacuum chamber including diffusion pump and controls; (b) experienced EBW operator; (c) setup of full penetration equator welds](image)

The five cavity welds are labeled on the CAD drawing in Appendix A. They consist of three seam welds: 1) inner flange-to-tube; 2) inner iris; and 3) outer iris; and two full penetration welds: 1) tube seam; and 2) equator. Figure 1.6 chronologically illustrates each weld setup. Niobium shims were used to protect cavity components from direct contact with the stainless steel fixtures.

![Figure 1.6 Chronological order of each EB weld: (a) full penetration tube weld; (b) inner flange-to-tube weld; (c) inner iris weld; (d) full penetration equator weld and outer iris weld](image)
18.7 Completed Cavities

Photographs of each cavity can be seen in Figure 1.7. Flanges are ¾” thick rather than the standard ½” to safeguard against any warping of the knife edge during welding. Each knife edge was seal tested with a copper gasket to ensure a vacuum tight seal.
18.8 Cavity Processing

The cavity was processed through a series of steps to ensure a particle-free and smooth inner surface for testing. Once the cavity left the machine shop and electron beam welding was completed, it was thoroughly cleaned with acetone, methanol, and Micro90 solution, using ultra pure water (UPW) outside the cleanroom. It then entered the cleanroom where it was ultrasonic cleaned (USC), rinsed, and air dried.

The cavity was then etched with fresh BCP solution, consisting of nitric acid, hydrofluoric acid and phosphoric acid, a ratio of 1:1:2, respectively, for 75 minutes to etch ~150 µm. Etch rates typically range from 1.5-2 µm/min, depending on niobium content in the solution. Acid temperature was held between a safe level of 5 and 15 ºC using an ice bath and the cavity itself as the heat exchanger. Figure 4.1 shows the cavity being etched.

![Figure 4.1 Etching cavity under fume hood. Cavity was cooled in ice bath with a protective PP flange (protects the Nb-Ti flange from acid exposure) and PP plug on bottom](image)

After etching, the cavity was transported to the cleanroom in an UPW bath. It immediately entered the high pressure rinse (HPR) station and was rinsed for ½ hour at 1200 psi (82 bar) with UPW. HPR is the final stage of processing; therefore, special care and caution is always exercised (i.e. handling). During HPR the cavity is manually rotated along its axis and moved in the vertical direction at varying speeds. The cavity was then laid horizontally to air dry under a HEPA filter inside the cleanroom. Figure 4.2 shows HPR.

![Figure 4.2 HPR housing and setup. Cavity was HPR for ½ hr. at 62-82 bar using UPW](image)

After the cavity air dried, all components, including RF feedthroughs and antennas, were assembled and attached to the insert. Note that all components of the insert were thoroughly cleaned prior. Figure 4.3 depicts the cavity/insert assembly.
Figure 4.3 Cavity being attached to Dewar insert. Special care in handling is taken during this procedure

The cavity was pumped to a vacuum pressure in the $10^{-8}$ mbar range using a clean turbo pumping system. The insert was finally carried out of the cleanroom and inserted into the Dewar for testing. See Figure 4.4 for insert transporting.

18.9 Schedule
Dec. 7, 2006 Quote sent to Daresbury
Dec. 27, 2006 Order placed with Niowave
Jan. 22, 2007 2D SolidWorks drawings sent via email to Carl Beard and Peter McIntosh
February 2007 Niobium stock arrived and cells/tubes stamped
March 20, 2007 First set of EBW performed
April 11, 2007 Second set of EBW performed
April 23, 2007 Delivery of cavities and final report package to Peter McIntosh at the TTC Meeting at Fermilab
18.10 Drawings

Figure 1.8 2D SolidWorks drawing of 3.9 GHz single cell including dimensions and weld schedule
### 18.11 Material Data Sheets for Niobium

**Material Data Sheet for Niobium**

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<thead>
<tr>
<th>Element</th>
<th>Test</th>
<th>Result</th>
</tr>
</thead>
<tbody>
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<td></td>
<td></td>
</tr>
<tr>
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**Chemical Composition (wt%)**

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<th>Element</th>
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**Mechanical Properties**

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<thead>
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<tr>
<td>Strength</td>
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**Certificate**

**Manufacturer:**

TOYO DENKAI CO., LTD.

**Date:** 12/31/2008

**Certificate No.:** 20329

**Customer:**

(Stamp or signature)

**Material Test Results**

(Notes or comments)
# Chemical Composition (in wt%)

<table>
<thead>
<tr>
<th>65.9</th>
<th>65.9</th>
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</thead>
</table>

## Mechanical Properties

| 2.8 x 106 x 1.12 |
| 2.8 x 106 x 1.16 |

## Test Results

<table>
<thead>
<tr>
<th>Test</th>
<th>Spec min</th>
<th>Spec max</th>
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<tbody>
<tr>
<td></td>
<td>0.0294</td>
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</table>

## Remarks

- **Material Source:** A193 B8
- **Supplier:** TONO DINNER CO., LTD.
- **Date:** August 8, 2007
- **Customer:** MESSA MESSA, INC.