Some measurements at IUCF with the new preamp board

Gerard Visser, IUCF, 2/6/08

Characterization of gain and linearity, part I

For this measurement one channel was used, terminated into various load impedances. The unused channel outputs were left open, i.e., saturated. The control lines for the GAS-1 are as Fernando set them (I haven't looked into the settings or tried to change them).

The new preamp board and the interposer board were installed in a shielded box. The test pulse was from a Tektronix AFG3252 generator, and the output was observed with a Tektronix P6247 differential probe (set for 200 MHz BWL) and TDS784D oscilloscope (set for 250 MHz BWL). The test circuit is documented in Figure 1.

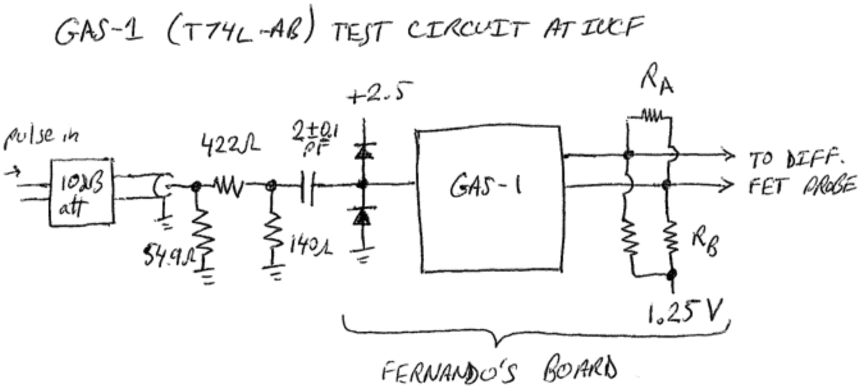

Figure 1. Test circuit, simplified. See preamp board schematics for further details. Effective charge injection capacitance is 157.6 fF ±5%.

A typical response to a voltage step input through the charge injection circuit above is shown in Figure 2. Response to a range of amplitudes is shown in Figure 3, showing the gradual nonlinear response above roughly 250 fC. Note that the peaking time is shifted as well in the nonlinear region, probably indicating some stage of the signal processing is slew rate limited. It could be the output stage, I should come back to check if the peaking time shift is dependent on the load impedance. A quick look shows it is not much dependent on the load impedance, so (I presume) is happening at an earlier stage.

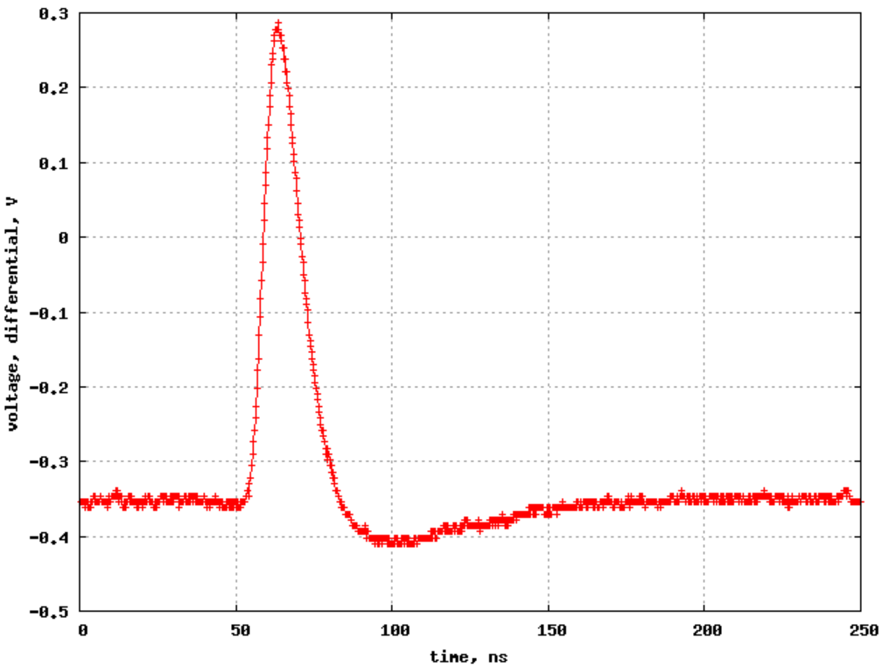

Figure 2. Response GAS-1 to 252 fC ±5% charge impulse. Output load impedance is 100 Ω. The noise here is dominated by probe and oscilloscope noise.

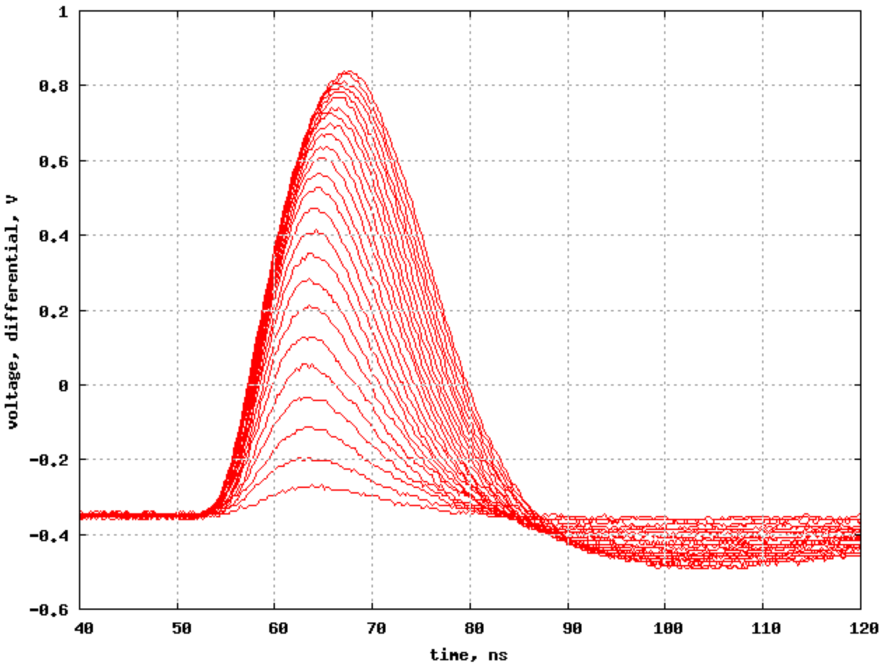

Figure 3. Response to charge impulses *N31.5 fC for *N*=1 to 25, output load 100 Ω.**

The 25 input pulse amplitudes mentioned in Figure 3 were tested with 4 different external output load resistances (see Figure 1):

1. $R_A = \infty$, R_B = 100 Ω – so 200 Ω differential
2. $R_A = \infty$, R_B = 50 Ω – so 100 Ω differential
3. R_A = 100 Ω, R_B = 50 Ω – so 50 Ω differential
4. R_A = 50 Ω, R_B = 50 Ω – so 33.3 Ω differential

Note that case 2 is roughly equivalent to using the cable with no termination resistors on the preamp board, and case 3 is roughly equivalent to using the cable with termination resistors on the preamp board, "double termination". I say roughly because the resistor values would in actuality be slightly different.

For each of the combination of amplitude & load resistance one event was stored, and the (leading part of the) waveform was fit to determine the baseline and the peak amplitude of the pulse. The pulse model was $V(t) = t^p e^{-t/\tau}$, which is a good fit to just the leading part of the pulse except for the highest amplitudes where the pulse shape distortion (mentioned above) was too large; these points are therefore discounted below. Results of this are shown in Figure 4, where the measured output voltage pulse height is translated into nominal output current by using the nominal differential load resistance, the parallel combination of R_A, $2R_B$, and I assume 1 kΩ (internal to the chip).

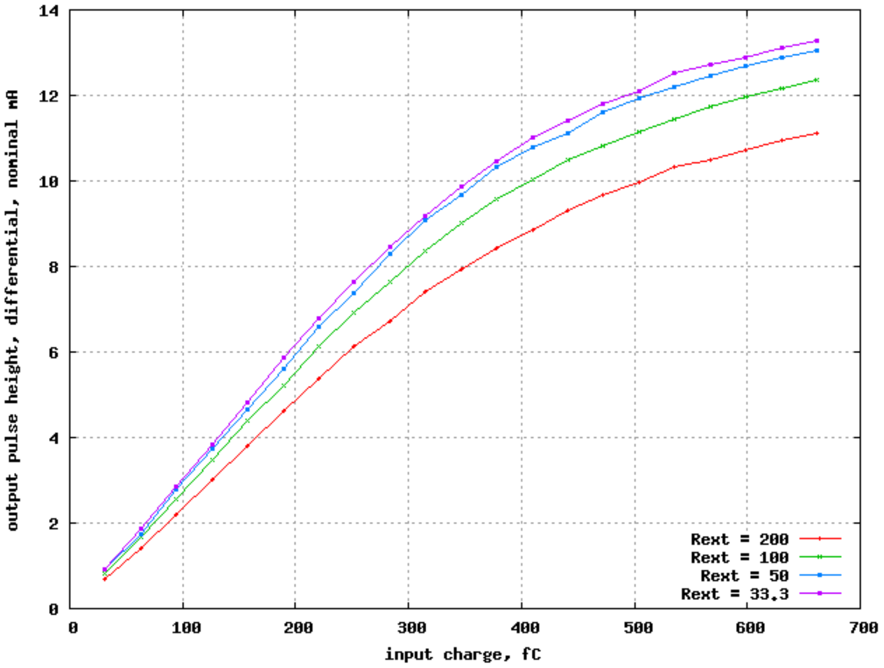

Figure 4. Output differential current pulse height vs. input pulse amplitude, calculated assuming nominal 1 kΩ (differential) internal termination.

One would expect these curves to lie on top of each other, at least for small signals where the output stage may be modeled as a (signal-dependent) current source with a parallel output impedance and a load impedance (independent of the signal). For larger signals the output impedance will obviously become signal dependent so the curves would diverge. The fact that they do not have the same slope at low signal amplitudes, is simply due to the fact that (above) we have an incorrect model of the output impedance of the chip – it is not just 1 kΩ plus or minus some uncertainty, but also of course has in parallel the output impedance of the transistors. It would seem, from this measurement, that about 320 Ω is the actual effective differential output impedance (Figure 5).

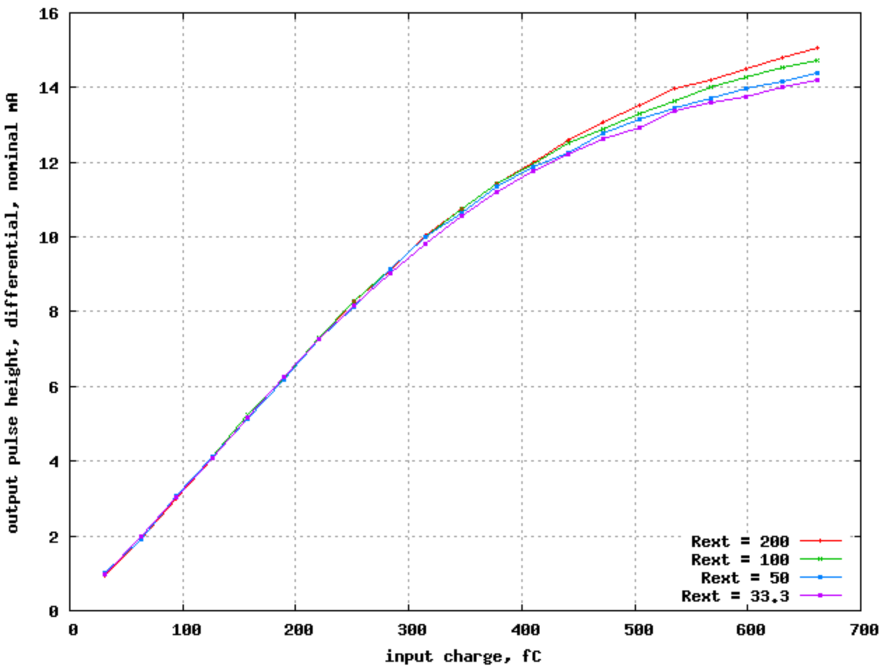

Figure 5. Output differential current pulse height vs. input pulse amplitude, calculation adjusted to assume 320 Ω (differential) internal termination.

Naturally this suggests making a direct measurement of the output impedance. I did this over a band 500 kHz to 50 MHz using a network analyzer with a Mini-Circuits ADT3-6T (50 Ω : 150 Ω) transformer and blocking capacitors to connect to the output of the GAS-1. Each side had a 100 Ω resistor to +1.25 V to ensure a reasonable DC bias point. Unfortunately, or perhaps in hindsight better to say fortunately, the results do not obviously confirm the above. Up to roughly 8 MHz the differential output impedance of the GAS-1 on the preamp board is equivalent to a model of 750 Ω in parallel with 10 pF. I believe this value (750 Ω) is more in line with what would be expected for the output resistance of the transistors in parallel with the internal resistors. Mitch – comments? At higher frequencies such a simple model is just not valid, there are clearly inductive effects coming into play, which might well make it look like 320 Ω for the one particular pulse shape shown above. So, the main conclusions are not to pay too much attention to the above output impedance discussion, and probably to use a termination resistor of 59.0 Ω on each leg of the output to +1.25 V if a good termination is desired at the preamp board. That sets it a little high to help cancel some of reflection from the capacitive loading. Time-domain reflection measurements with the long cable might indicate an adjustment to this value.

DC balance of the outputs

Before installing any termination resistors I measured the common-mode DC imbalance at the output on 4 channels by shorting each outputs' positive and negative terminals. The voltage with respect to ground and current obtained by shorting the output pair to +1.25 V are recorded in the following table:

Common-mode voltage	Imbalance current
0.960 V	-1.431 mA
0.942 V	-1.526 mA

0.948 V	-1.491 mA
0.957 V	-1.435 mA

The GAS-1 supply voltage was 2.502 V and the reference voltage was 1.247 V. Secondarily this implies a DC common-mode output impedance of 203 Ω, which is in reasonable agreement with the 750 Ω differential measurement above.[*] Primarily, the conclusion from this measurement is that we may typically expect the common-mode line voltage in the absence of any external common-mode termination to vary by something of order 300 mV or more over temperature and process variations, *assuming these imbalance currents were really designed to come out zero*. Of course this is an unwarranted extrapolation, we have to measure it on a batch of chips to get some idea of the variation. That probably should be measured. I suspect that a 300 mV shift in the common-mode voltage could imply at least a few percent variation in the signal gain (to differential output) – this is a measurement to be done. A related conclusion is that the +1.25 V termination reference supply on the board, if used, should be capable of sourcing or sinking about 36 mA at least.

Termination and power discussion

I installed termination resistors on the preamp board using current best estimate of the proper value, that is 59.0 Ω. A short (1 m) cable was connected, with a 100 Ω differential termination at the far end. This is my proposed default configuration. At a later date we may go back to explore how much power is saved by omitting the termination resistors on the preamp, and how much the signals are degraded by doing so, and make an informed decision. I prefer to keep the best signal quality possible for the initial measurements.

We can easily make a rough estimate of the power savings from omitting the external terminations on the preamp: The baseline output current (differential) is about 5.3 mA, observed. With termination, the load is 50.5 Ω, voltage is 268 mV, power into cable and far termination resistor is 17 mW for 24 lines. Without termination, the load is 88.2 Ω, voltage is 468 mV, power into cable and far termination is 52 mW for 24 lines. In either case the supply current (and so the total power input) is absolutely the same. So, the power savings on the preamp board from using no termination resistors is 35 mW. That is approximately 2.5 %, I don't see that this is in any way significant. *Much more power could be saved* by addressing the output common mode current of -1.5 mA, which is consuming 115 mW[†], or by omitting the on-chip Thévenin termination (or replacing with bussed terminations to the external +1.25 V source), which are consuming 384 mW.

Characterization of gain and linearity, part II

Now I went back and examined the linearity in response to a step charge input in more detail over the really usefully linear lower part of the range. Results are shown in Figure 6, the first quantitative plot of *nonlinearity*, at least that I have seen. The nonlinearity is better than 0.5 % over the range 0 to ~280 fC, i.e., 0 to ~425 mV. This level of nonlinearity is I suppose perfectly acceptable, although I would like to encourage Simon to introduce an extra nonlinearity, say 2%, in his analysis of cosmic ray data and see what

---

[*] With a split termination, the common-mode impedance is R/2 and the differential-mode impedance is 2R.

[†] I assume a +3.2 V input supply.

effect it has on the cathode position resolution. It must have some effect, but we may guess it is acceptable; best to really check it though. The linearity really starts to degrade above 500 mV, so I will take this as the design full scale range of the ADC system as seen at the preamp board outputs. That is, the combined system of cable, receiver/shaper, and Struck ADC or FADC250 now, and later the combined system of cable and FADC125, will have a 500 mV full scale signal range or just a bit more. Hopefully this will mesh well with the chamber operating gain and signal size, it must be tested.[‡] Revising the gain a little bit will of course be possible.

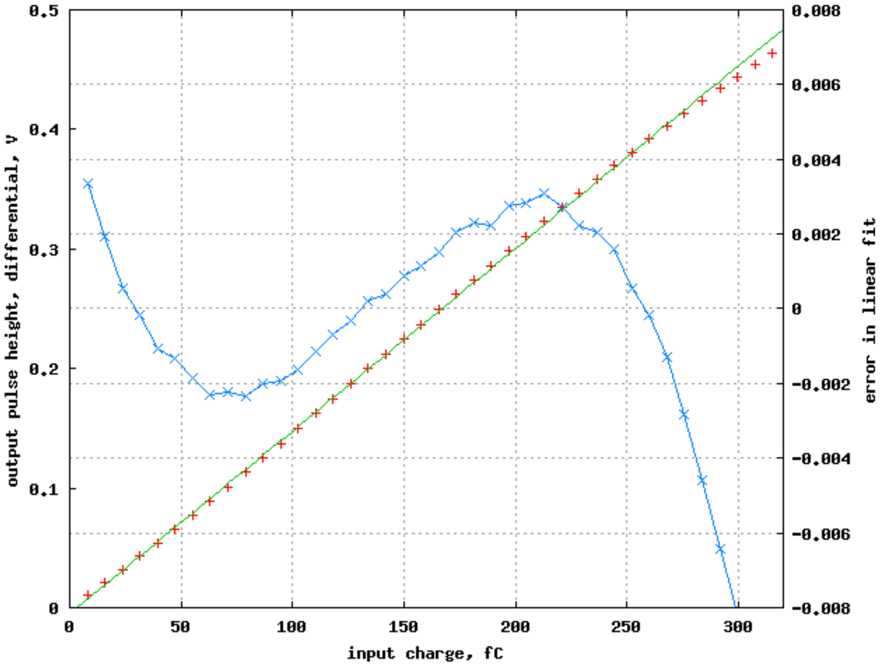

Figure 6. Output differential voltage pulse height vs. input pulse amplitude, for instantaneous (step) charge input. Nonlinearity, i.e., error in linear fit, is shown on the right-hand axis. Note that the fit function consists of a gain *and an offset*. Without the offset the results are significantly worse.

Response to chamber pulse shape

The response of the ASIC to real chamber pulses is of course often best investigated with a real chamber at hand. However, in one respect test pulser data is superior – one can easily control how much charge is being injected. (Of course testing with a low energy X-ray source such as ^{55}Fe is similarly useful.)

The total charge seen from a straw tube detector up to time t is $Q(t) = Q_0 \dfrac{\ln\left(1+\dfrac{t}{t_0}\right)}{\ln\left(1+\dfrac{t_f}{t_0}\right)}$,

where the characteristic time $t_0 = \dfrac{(a+\lambda)^2 \ln\dfrac{b}{a}}{2\mu V_A}$ and the final time $t_f = t_0\left(\dfrac{b^2}{(a+\lambda)^2} - 1\right)$.

[‡] Will we be able to really exercise the full range of path lengths in the CDC?

Here, with values for the CDC, $a = 10\ \mu\text{m}$ is the wire radius, $b = 8\ \text{mm}$ the cathode radius, $\lambda = 1\ \mu\text{m}$ the distance of mean of avalanche above the wire[§], $\mu = 1.54 \times 10^{-4}\ \text{m}^2/\text{V}\cdot\text{s}$ is the mobility of ions[**], $V_A = 1900\ \text{V}$ the bias voltage. Consequently $t_0 = 1.38\ \text{ns}$ and $t_f = 731\ \mu\text{s}$.

I have at hand an arbitrary waveform generator with a 8.192 µs record length. That is more than sufficient to simulate a chamber pulse, but the total charge must be scaled appropriately. To generate a signal that simulates a chamber pulse of total charge Q_0, the waveform amplitude should be set so that there is a total charge

$$Q_0 \frac{\ln\left(1+\frac{8.192\ \mu\text{s}}{t_0}\right)}{\ln\left(1+\frac{t_f}{t_0}\right)}$$

in the pulse. That is a simple matter.

By this means the "canonical" 94 electron (1 cm Ar) pulse with a gas gain of 5×10^4, that is 753 fC *total* charge, was generated for several values of t_0 bracketing the calculated value above. The waveform generator output, fed to the test circuit of Figure 1, is shown in Figure 7. Results are shown in Figure 8 through Figure 10.

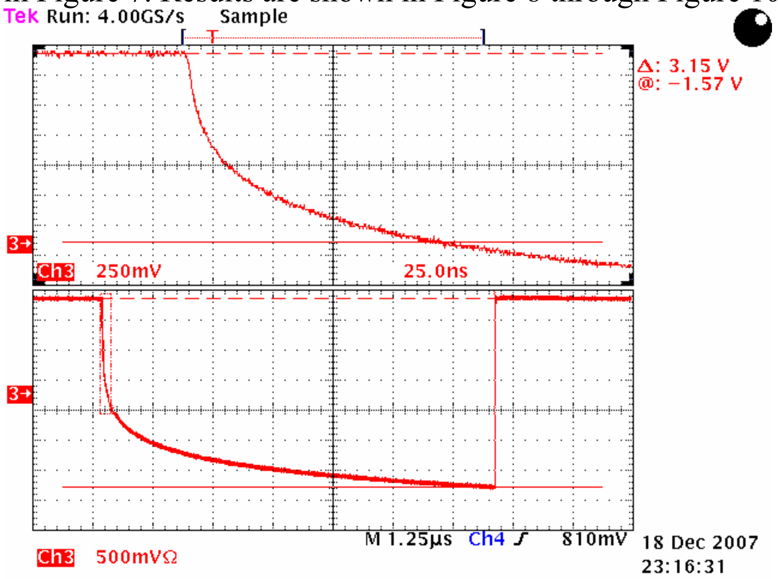

Figure 7. Voltage input to test circuit to simulate a 753 fC CDC signal with t_0 = 1.38 ns. Upper is zoom of lower as indicated by dashed-line box. Total charge in this pulse, i.e., to 8.192 µs, is 3.15 V × 157.6 fF = 496 fC.

[§] Value is from Sauli paper.

[**] Again, Sauli paper, I assume 80/20 Ar/CO_2. I don't assume this is necessarily the right value, just ±50%.

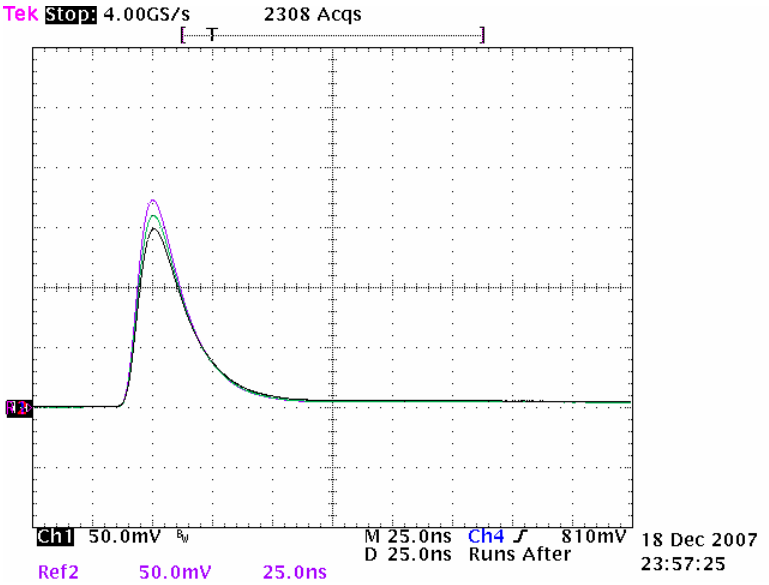

Figure 8. GAS-1 output from simulated 753 fC CDC signal with t_0 = 1.15, 1.38, 1.66 ns. Output load resistance is 59.0 Ω from each side to +1.25 V and 100 Ω differential. Averaged waveforms.

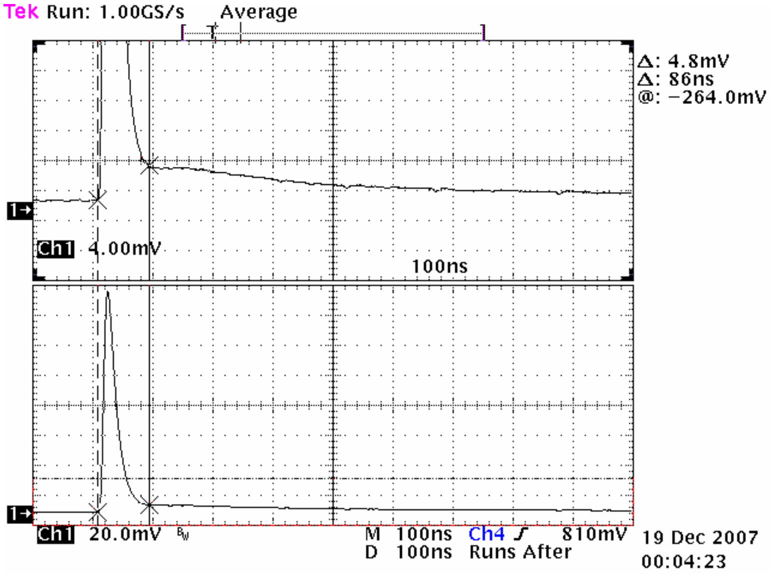

Figure 9. Detail of pulse tail, GAS-1 output from simulated 753 fC CDC signal with t_0 = 1.66 ns as above. Averaged waveform. Pulse height 147 mV, tail is within 3% in 86 ns. I think that's actually pretty good![††]

At this point, a linearity plot should be taken again using this pulse form. In principle the linearity may be different in some detail with a different input pulse shape, so it should be checked.

[††] Of course, this conclusion is contingent on my using the correct t_0 value to simulate the real CDC. We have to check it.

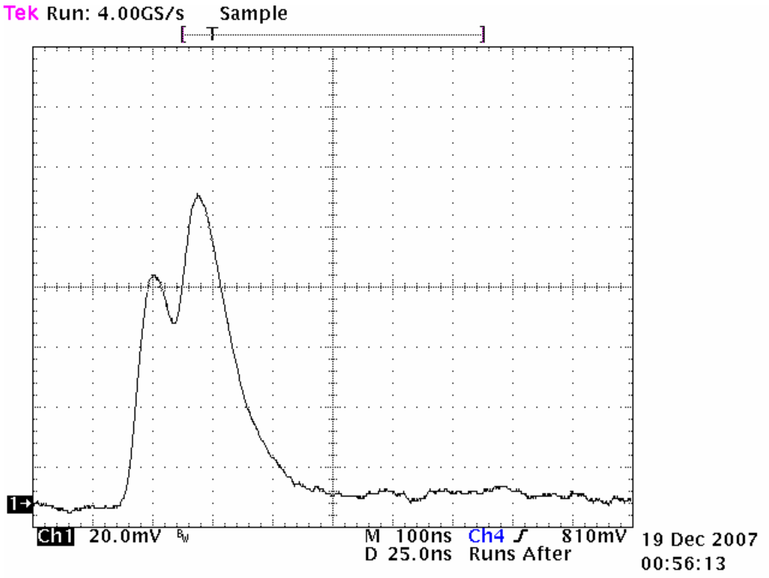

Figure 10. Illustration of a somewhat more realistic pulse shape. Two clusters of 47 electrons each in simulated CDC with t_0 = 1.38 ns. Noise level here certainly contains a significant contribution from the scope and test setup, i.e., GAS-1 noise is lower than this.

Also along the lines of linearity, we should have it that integral of pulse is independent of shape, e.g., number and time distribution of clusters, only depends on total signal charge. That should be true to the few % level, it needs checking, I will do this.

Noise and dependence on detector capacitance

To connect the GAS-1 output and our 100 Ω signal cables directly to 50 Ω test equipment, a matching network (Figure 11, but here used in the reverse orientation). A 1:2 impedance ratio transformer was not used, firstly to provide some flexibility for the cable characteristic impedance but also more importantly because the 1:3 transformer used here has superior bandwidth. Measured attenuation is about 5.9 dB within the passband of the transformer, nominally 60 kHz to 400 MHz. In other words, the voltage on the 50 Ω output is 0.359 times the voltage on the 100 Ω line coming in.

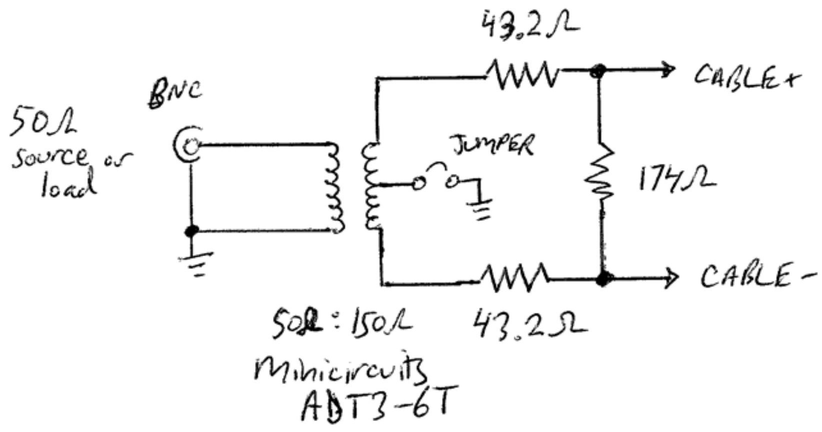

Figure 11. Matching network to connect the cable to 50 Ohm coax. The common-mode jumper would typically be set on source end of the line only, similar to proposed cable useage in GlueX.

The output noise spectrum from the GAS-1 was measured using an HP 8561B spectrum analyzer with the resolution bandwidth and video bandwidth both set to 1 kHz, reference level set to -80 dBm, signal attenuation off. The cable length was 1 m, preamp board is stuffed with 59.0 Ω termination resistors, and the transformer circuit of Figure 11 was used to connect the signal to the 50 Ω spectrum analyzer input. Note in particular the differential load resistance presented to the GAS-1 is 50 Ω (a double-terminated 100 Ω cable connection). Results are shown in Figure 12 and Figure 13; note that this is the voltage on the cable, i.e., the attenuation of Figure 11 circuit is accounted for.

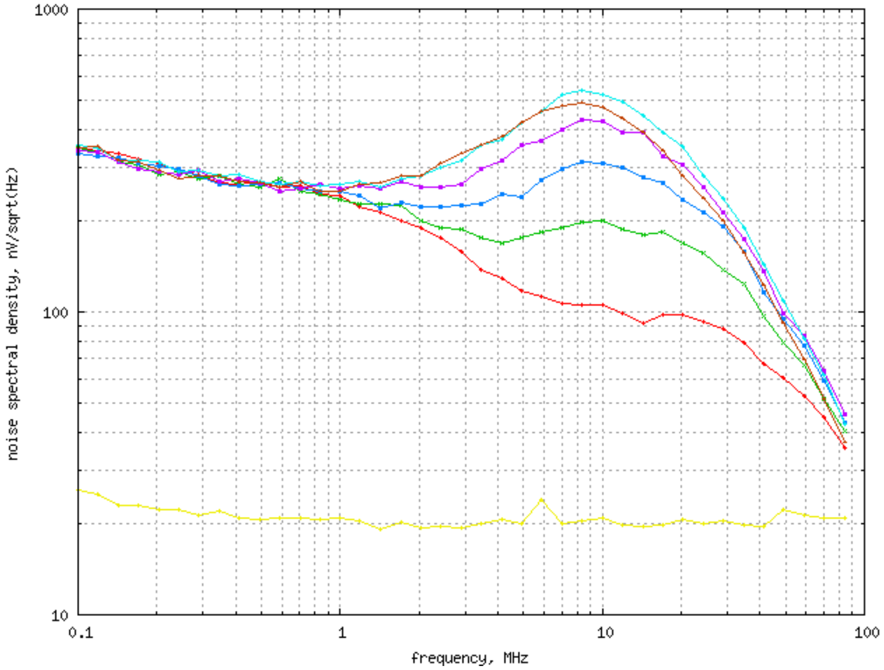

Figure 12. GAS-1 output noise voltage spectrum, with no added capacitance (red), 20 pF to ground (green), 40 pF (blue), 60 pF (purple), 80 pF (cyan), 60 pF to neighbor channel (brown), and test setup noise floor with GAS-1 power off (yellow).

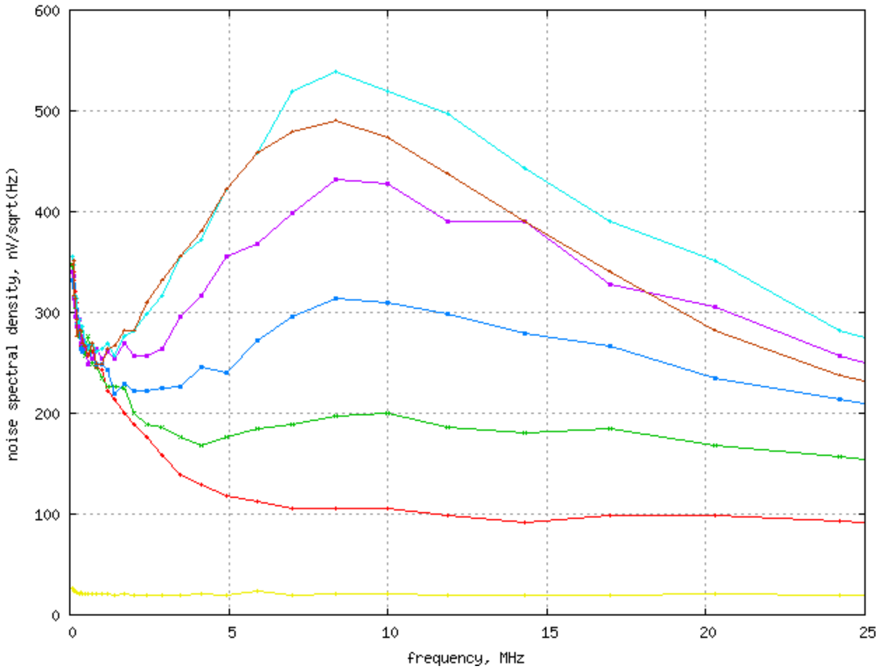

Figure 13. Same data as in Figure 12, on a linear scale.

As a reality check the rms noise voltage was directly measured with an oscilloscope vertical histogram. This is only an order-of-magnitude comparison since I do not have detailed info on the oscilloscope frequency response – but the results agree with the noise measurements above.

It may be important to point out that the noise contributed by a capacitance connected across channels is generally greater than that contributed by a capacitance from the channel to ground, except at the highest frequencies. For the FDC cathodes, of course, a considerable part of the capacitance is between adjactent strips. I suppose this effect is not a surprise, the equivalent input voltage noise of two channels contributes in the case of cross capacitance. But it means we have to be careful not to underestimate the FDC cathode noise level from a simple measurement of GAS-1 noise with capacitance to ground.

I also intend to measure the noise from an (AC coupled) 10 kΩ and a 200 kΩ resistor. The 10 kΩ value is currently proposed for use on the HV coupling boards for FDC – how much noise will this contribute, is a larger value motivated? It is worthwhile to check it, not done yet.

TDR measurements of reflection from preamp board

To be done… To confirm or modify the choice of 59.0 Ω resistor.

DC sweep of a forced input current

To be done… This will measure the DC gain of the chip, the bias voltage at the input, DC input impedance, estimate the DC open-loop gain of the input stage. Fairly easy to do, I have a Keithley “sourcemeter” to use.