Characterizing SerDes with a Real Time Oscilloscope

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Overview

1. Introduction
2. Equalization
3. Measurements
4. Conclusions

Introduction

For specifying and testing Transmitters for copper links we need to cover:

1. Signal level for each digital state transmitted (including effects of Tx FFE)
2. SNDR and SNDR statistical distribution
3. Jitter including jitter statistical distribution out to some BER
4. Effective channel loss to the test point (makes all other measurements much harder)
5. Return loss (not covered by method I describe below.)

The first 4 can be determined from a single long captured waveform.

All of these except return loss could be measured with a single waveform capture if all data is used, not just a a few points per repeat of a PRBS pattern. If all transitions are used a single block of just a bit over 600us would be enough to measure $1*10^6$ jitter values of each of the 12 different kinds of transitions. Similarly, if all data points can be used then $4*10^6$ data values can be measured for each state.

Equalization

To use all transitions the signal must be well enough equalized to eliminate ISI jitter and ISI at the data sampling point. This can be done by post processing. The process of determining the correct equalization function will give the effective channel loss.

The biggest technical problem with a real time scope is the high noise level. To minimize this problem the equalization should be done in such a way as to minimize noise enhancement. This implies minimizing noise bandwidth of the equalization function.

I propose to equalize to a data response of a “raised cosine” which I consider to be

$$raised\,cosine = \cos^2\left(\frac{1}{2}\pi * f * UI\right) \quad if \quad f * UI < 1$$

$$0 \qquad otherwise$$

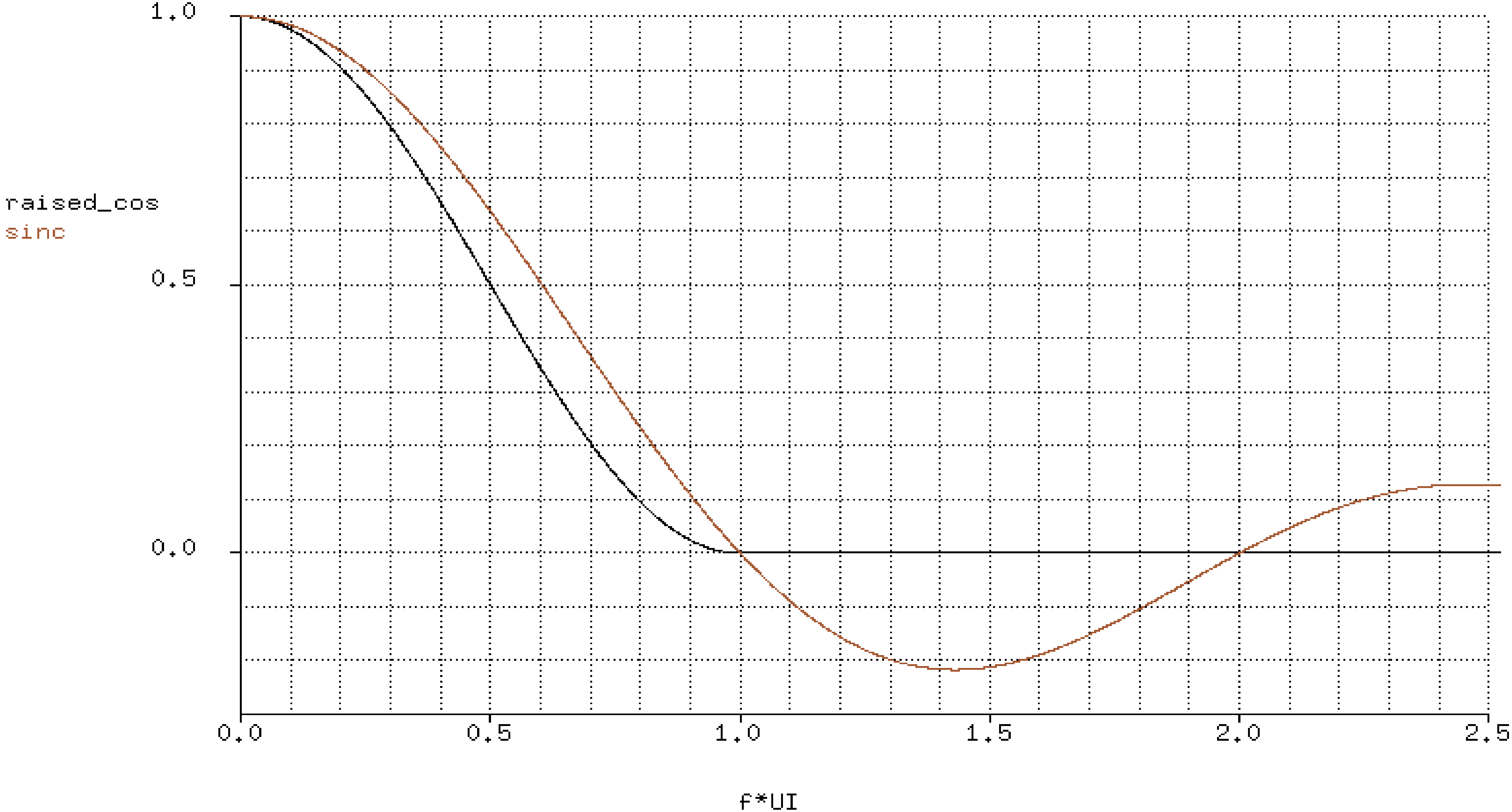
With an impulse response:

$$impulse\,response = \frac{\sin(\pi \cdot x)}{\pi \cdot (x - x^3)}$$

$$where \quad x = 2\frac{time}{UI}$$

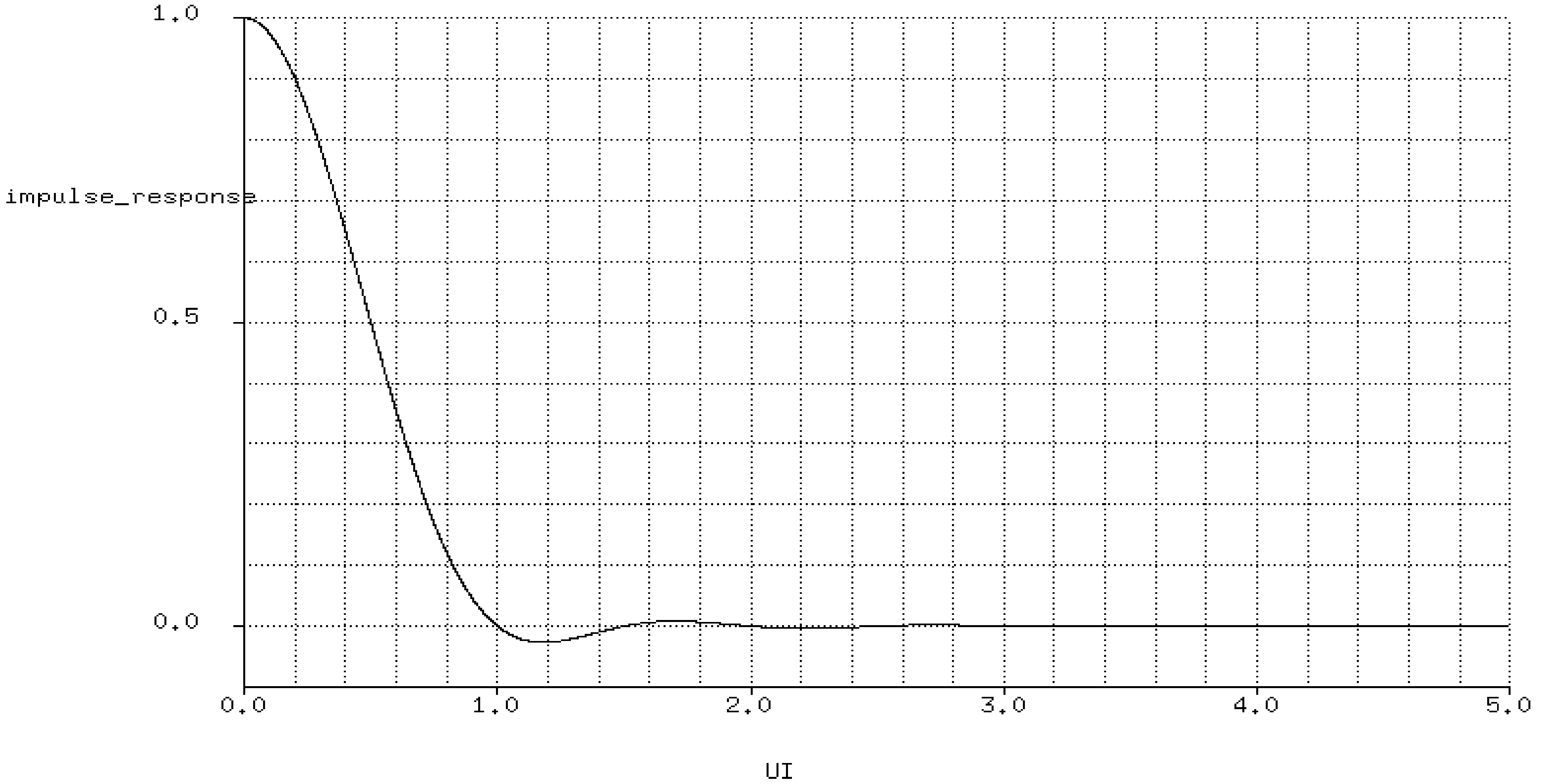
The impulse response of the raised cosine is 0 at every integer and half integer value of time/UI except 0 and UI/2.

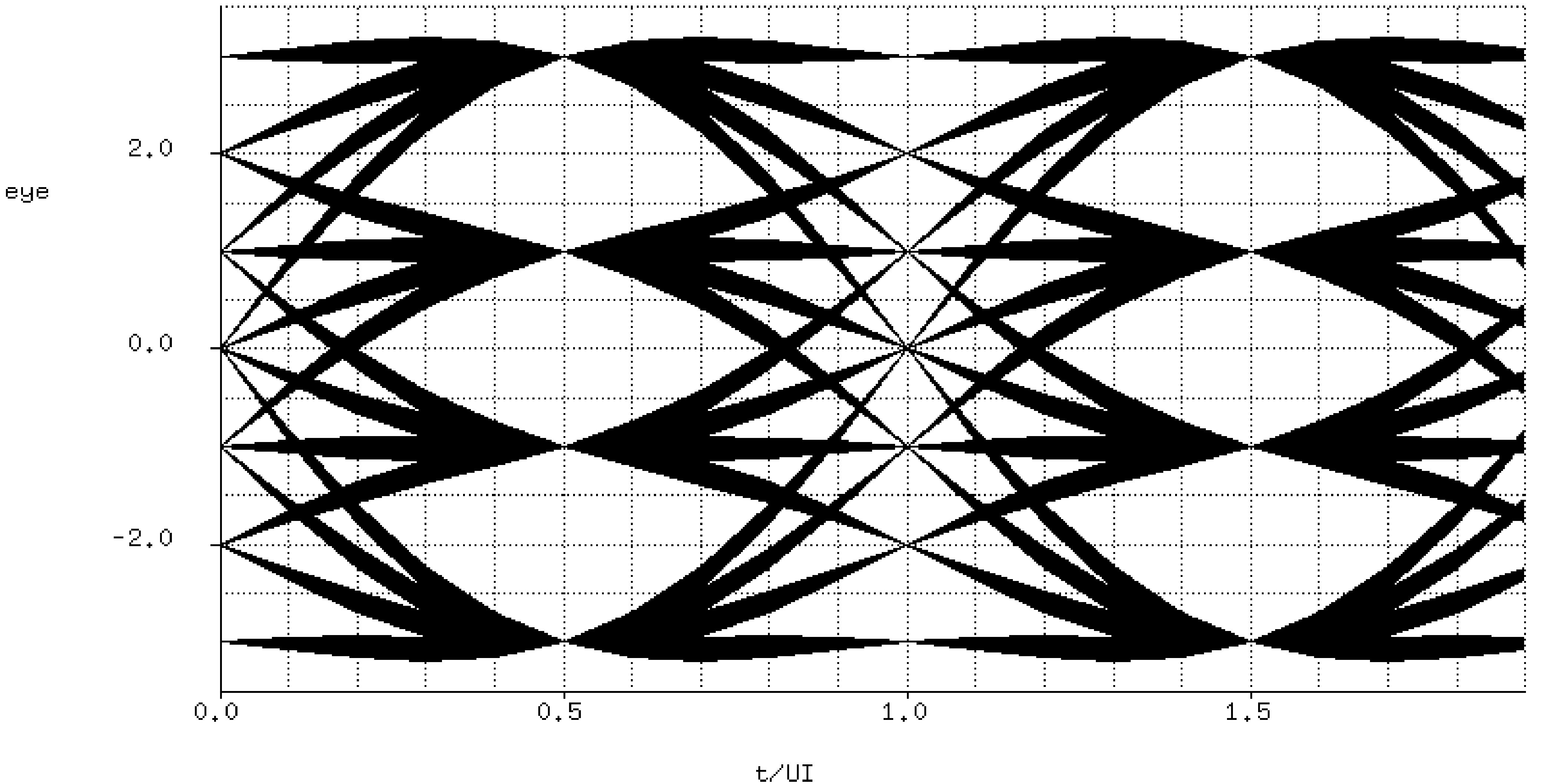
Frequency response



Note that the raised cosine's magnitude is always equal to or smaller than the sinc's magnitude and zero above the data rate frequency. Raised cosine is an excellent noise limiting filter.

Impulse response





QPRBS13 eye pattern with raised cosine filtering

For transmitter jitter measurement purposes I define the transition time as the time when the signal crosses the level half way between the end data points, not the point where the data becomes valid at its new state.

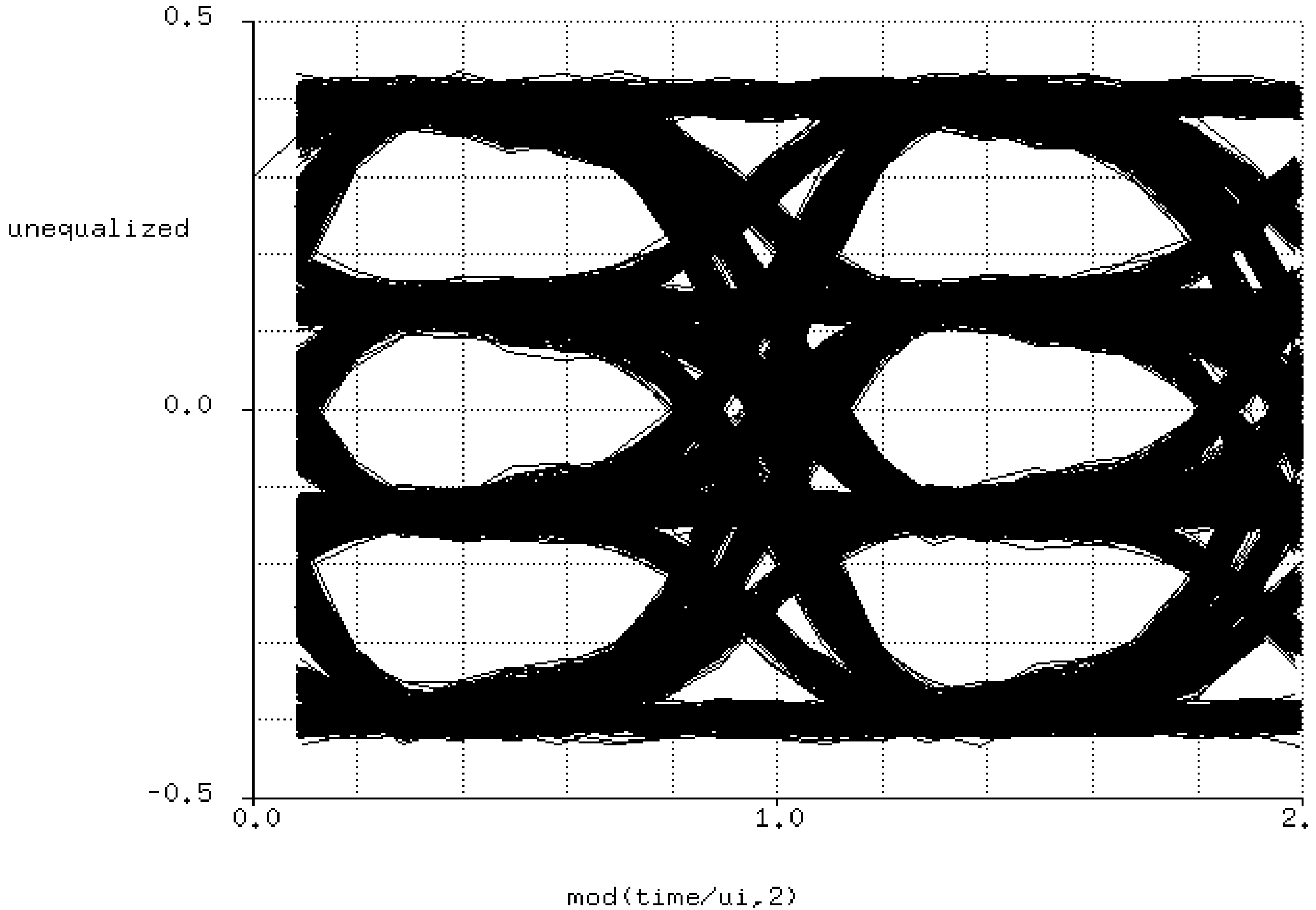
Equalization and measurement can be done in post processing.

Measurement

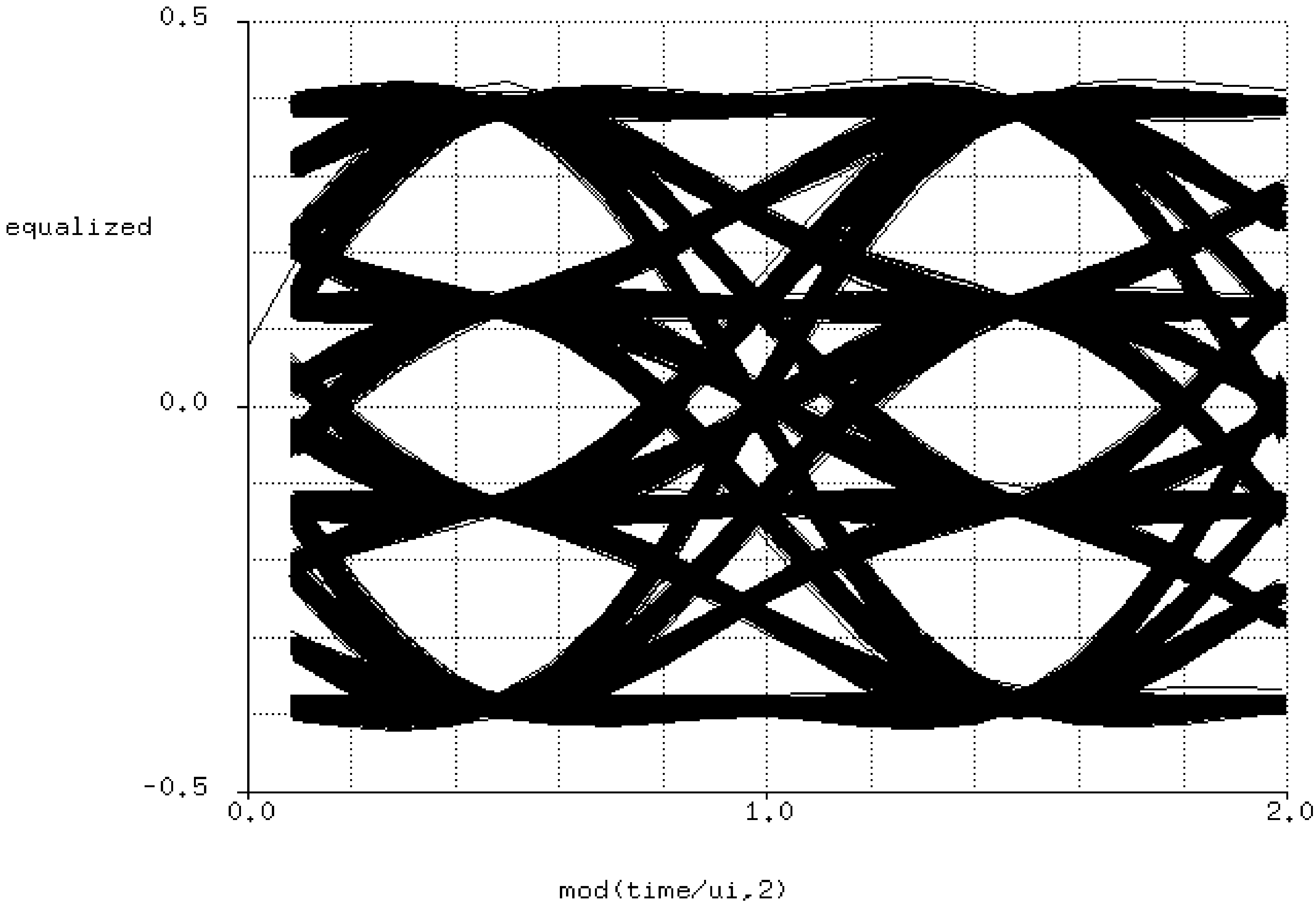
The first step was to characterize the real time scope. This we did by measuring a clean sine wave at near the Nyquist frequency. I then subtracted the best fitting sine and analyzed the residue. When I did this I got a good sine wave with harmonics down 50dB or more, indicating good linearity. Noise below 30GHz could be reasonably well approximated a constant noise power spectral density of $8*10^{-16}V^2/Hz$.

Otherwise the real time scope seemed close to being an ideal A-D.

We measured the output of a commercial BERT producing a PAM4 signal at 26.56 GBd. The path was pretty clean and we got this is the un-equalized EYE, re-sampled at 10 x the baud rate.



Once equalized to a raised cosine the EYE looked like this (re-sampled to Exactly 10 samples per baud. Note that at the sample points and 50% transition points the variation is much tighter.



Data levels, amplitude, linearity, and noise plus distortion

level	mean	dev	count
0	-0.387651	0.005366	2049359
1	-0.131373	0.005799	2050465
2	0.131220	0.005837	2050742
3	0.387738	0.005497	2049433

LSB= 256.40mV
MSB= 518.99mV = 2xLSB+ 6.196mV

The noise bandwidth for the equalizing function, which I define as

$$noise\,bandwidth = \int \frac{gain(f)^2}{gain(0)^2} df$$

is 17.8257GHz giving an expected noise is 3.8mV RMS. So the deviation of the levels is larger than expected from the scope noise alone. I will discuss the excess noise later.

Transition position and jitter

Transition time relative to average (ps)

level to \ level from ->	0	1	2	3
0	0.00	-0.90	0.18	-0.10
1	-1.04	–---	1.53	0.38
2	0.40	1.46	–----	-1.02
3	-0.12	0.19	-0.95	–---

LSB delay -0.981ps
MSB delay 0.289ps

Jitter rms (including offset) (ps)

level to V \ level from ->	0	1	2	3
0	1.02	1.12	0.48	0.42
1	1.22	–--	1.70	0.60
2	0.60	1.64	–---	1.21
3	0.40	0.48	1.15	–--

Note on these and later tables, the 0,0 point is replaced by the RMS of all the transitions.

The 12 types of transitions do not all occur at the same time within the UI. For this case it appears that the LSB transitions are about 1.3ps before the MSB. So the 0→1, 1→0, 2→3, and 3→2 transitions occur earlier than the 0→2, 2→0, 1→3, and 3→1 transitions. The 0→3, and 3→0 where both bits are switching in the same direction are intermediate while the 1→2, and 2→1 where the bits are fighting each other comes latest of all.

This could be a problem and lead to loss of eye opening especially if it is large. I think it would be a good idea to specify limits on transition time deviation from mean.

To give something more similar to what a sampling scope would give, subtract the mean delay from each of the 12 transitions. This leaves:

Jitter rms (ps)

level to V \ level from ->	0	1	2	3
0	0.57	0.67	0.45	0.40
1	0.63	*****	0.73	0.46
2	0.44	0.75	*****	0.64
3	0.38	0.44	0.64	*****

Much lower

Applying a 4MHz “Golden PLL” reduce the jitter somewhat

Jitter rms (ps)

level to V \ level from ->	0	1	2	3
0	0.49	0.60	0.34	0.27
1	0.56	----	0.67	0.35
2	0.33	0.69	----	0.57
3	0.24	0.33	0.57	----

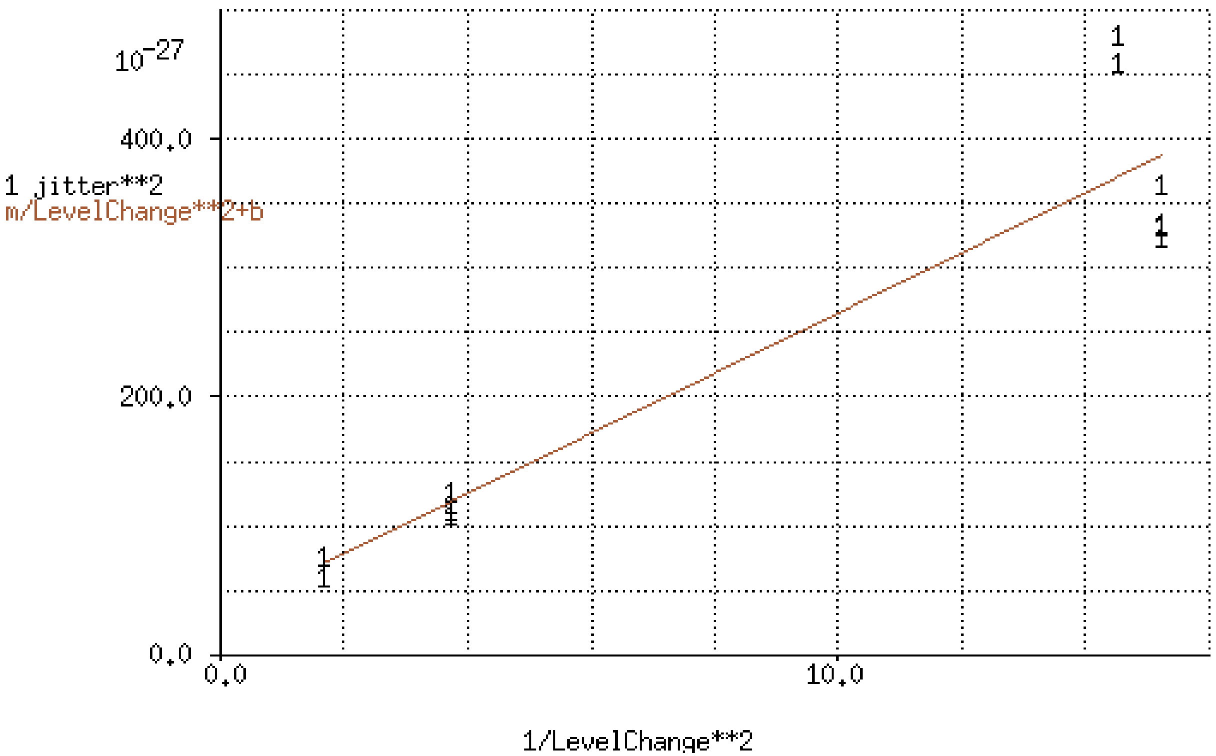
You will notice that the larger swing transition such as $0 \rightarrow 3$, and $3 \rightarrow 0$ have smaller jitter than the small swing transitions such as $0 \rightarrow 1$, $1 \rightarrow 0$, $2 \rightarrow 3$, $3 \rightarrow 2$, $1 \rightarrow 2$, and $2 \rightarrow 1$. This is because of the scope noise. If there is a voltage error ε at the transition point then, for a raised cosine equalized signal, there will be an error in the transition time of

$$time\ error = \frac{\epsilon}{\frac{dv}{dt}} = \epsilon \cdot \frac{2}{3} \cdot \frac{UI}{\Delta V}$$

Then assuming that there is some intrinsic jitter in the Tx which can be RSSed with the scope noise induced jitter then:

$$TJ^2 = TxJ^2 + \frac{4}{9} \cdot RMS\ Noise^2 \cdot \frac{UI^2}{Level\, change^2}$$

Making a linear fit between TJ2 and 1/Level change2 allow us to find values of Tx jitter and Scope noise which fit the data.



$where$

$$m = 2.331 \cdot 10^{-26} \left(sec/volt \right)^2$$

$$b = 3.256 \cdot 10^{-26}\, sec^2$$

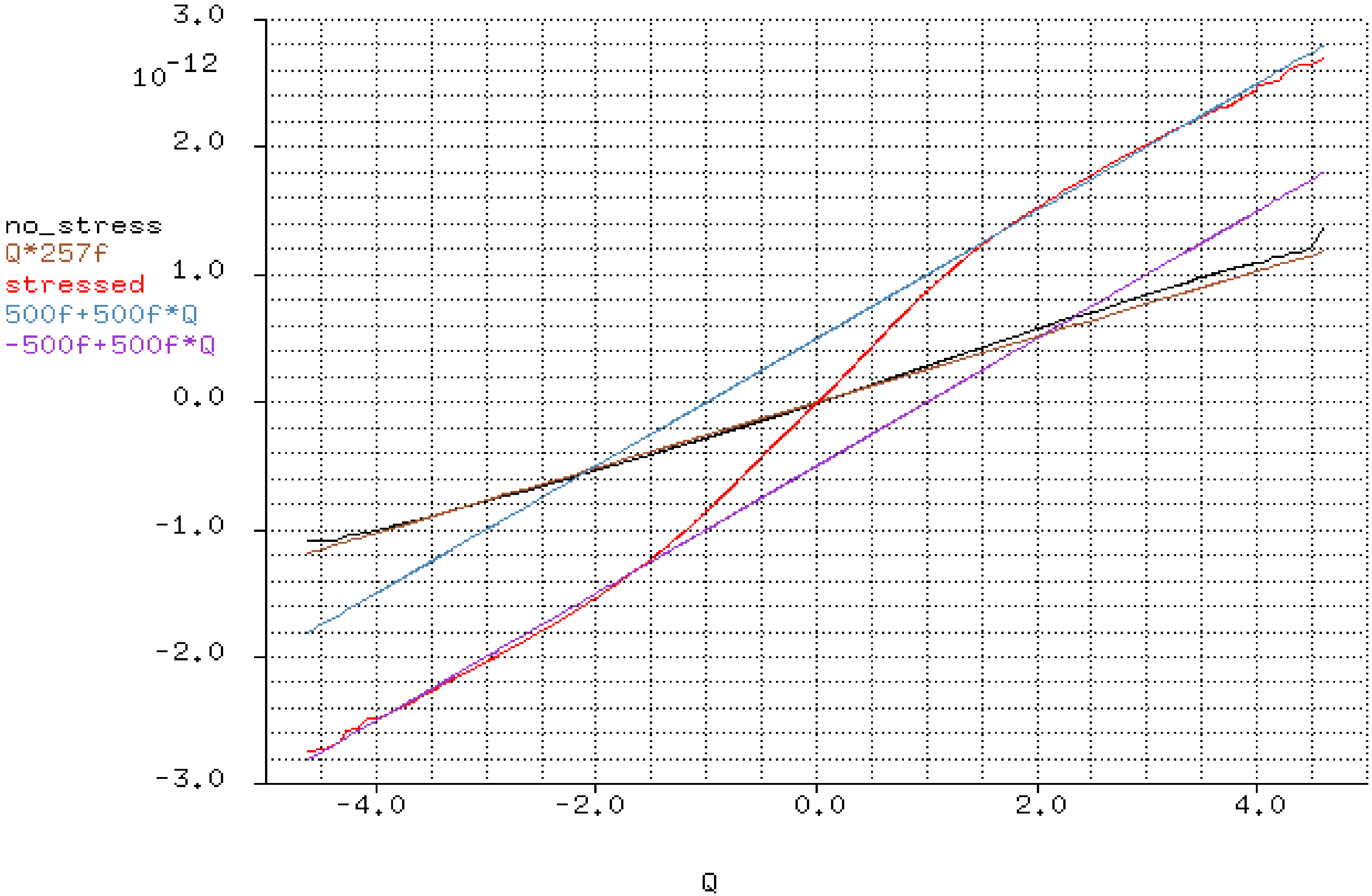
Decoding m and b gives:

$$TxJ = 180\, fs\, RMS$$

$$RMS\, noise = 6.1\, mV\ \ RMS$$

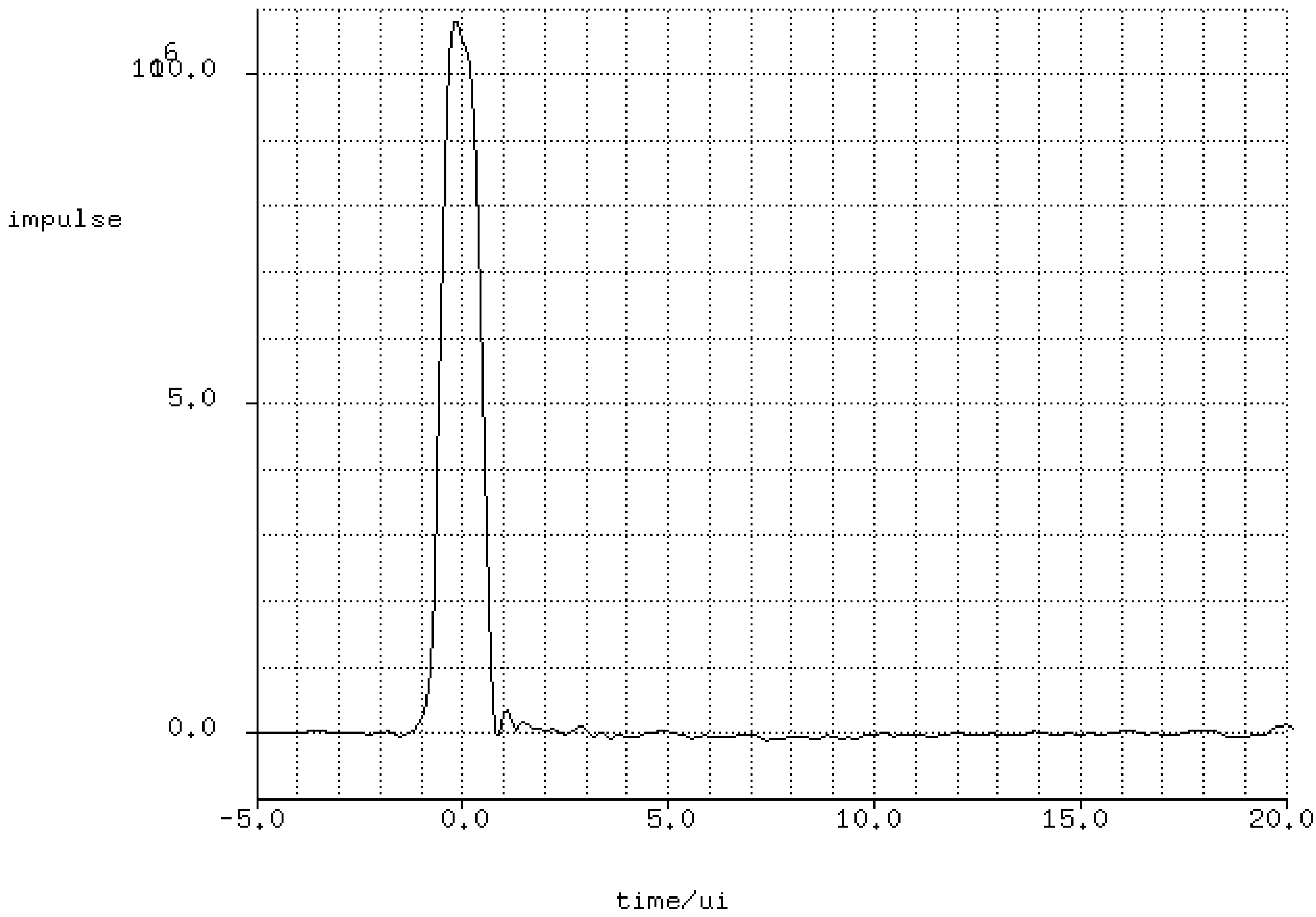
This RMS noise is quite a bit higher than the predicted 3.8mV RMS. It is more than 4dB higher. But it is close to the 5.6mV RMS average noise seen in the levels. I do not know where the additional noise comes from. Some of it may come from the BERT. Some of may come from using a non-repetitive pattern. The equalization uses a DFT which requires a repetitive pattern and the end effects may add some noise. When I analyzed a real SerDes with a repetitive pattern the predicted noise fit the data much better.

One of the engineers who helped with the measurement measured the same BERT output using the same setup but with some intentional jitter added in. Below is a Q plot of the clean BERT, impaired BERT, and some dual Dirac fitting curves. These are for the 0-3 transition.



The fit indicates 1ps p-p HPBJ and an added 430fs RMS added RJ After RSS subtracting the base RJ.
This gives p-p jitter of 128mUI p-p at J4 or 125mUI if the un-stressed RJ is RSSed out.

Here is the pulse response of the effective channel from the data generator to the scope



Conclusion

In the introduction I said that for specifying and testing Transmitters for copper links we need to cover:

1. Signal level for each digital state transmitted (including effects of Tx FFE)

The measured level for each of the digital states is show in slide 19.

In the introduction I said that for specifying and testing Transmitters for copper links we need to cover:

2. SNDR and SNDR statistical distribution

If

$$SNDR = dB\left(\frac{\text{voltage difference to nearest other level}}{RMS\,noise\ at\ sample\ point}\right)$$

Then SNDR for this case is 33dB based on the level information shown on slide 18, not correcting for scope noise.
If needed a Q plot or J plot could give the SNDR based on some Q or J metric.

In the introduction I said that for specifying and testing Transmitters for copper links we need to cover:

3. Jitter including jitter statistical distribution out to some BER

The RMS jitter for all 12 transition types is show in slides 19, 21 and 22 based on various assumptions. Slide 25 shows a Q plot for the 0,3 transition. Of course Q plots can be made for any and all transitions.

Slide 19 also shows deviation of various transitions from the mean.

In the introduction I said that for specifying and testing Transmitters for copper links we need to cover:

4. Effective channel loss to the test point (makes all other measurements much harder)

The pulse response shown on slide 26 is basically the same as the “linear fit pulse response” as defined in IEEE802.3 clause 85.8.3.3, which is the way we have been defining effective channel loss. The shown pulse response is derived differently but if truncated, will differ only insignificantly from linear fit pulse response. It can be used the same way for measuring equalization tap values.

In the introduction I said that for specifying and testing transmitters for copper links we need to cover:

5. Return loss

    Which I said I would not deal with and I have not dealt with it.