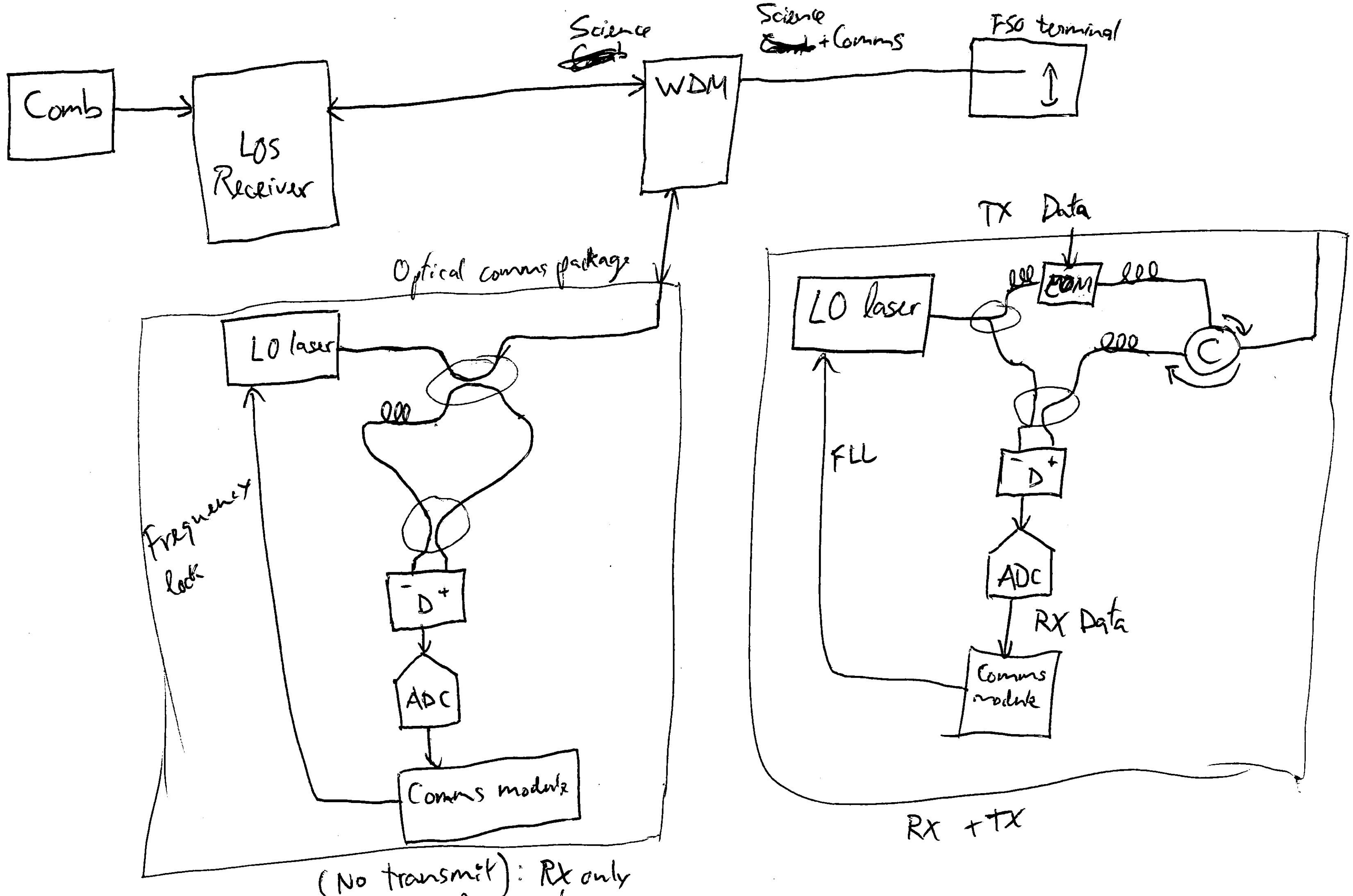
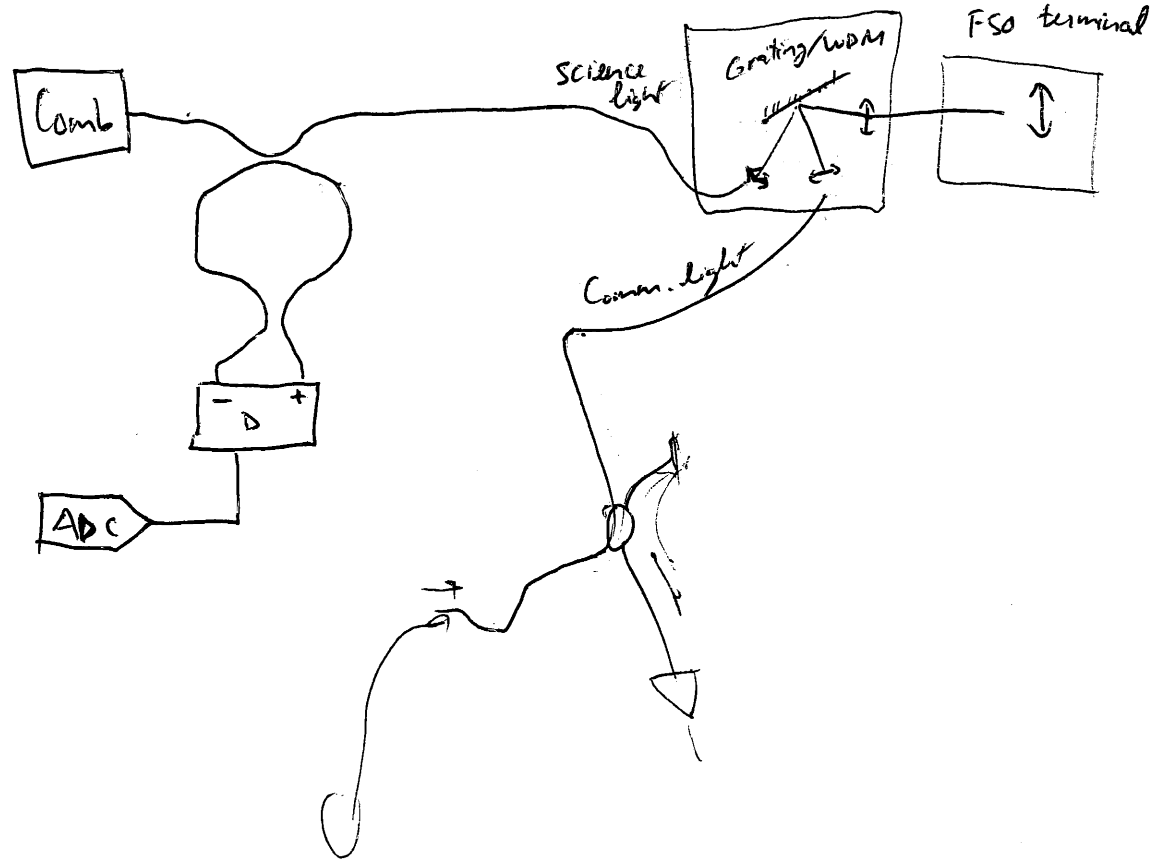
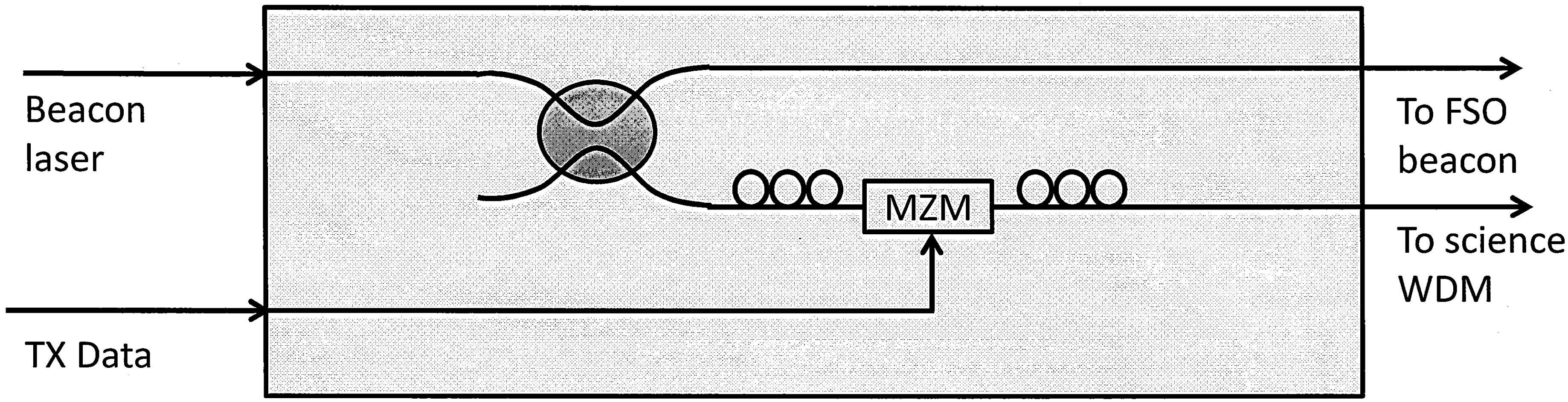


Bad design - uses first coupler inefficiently

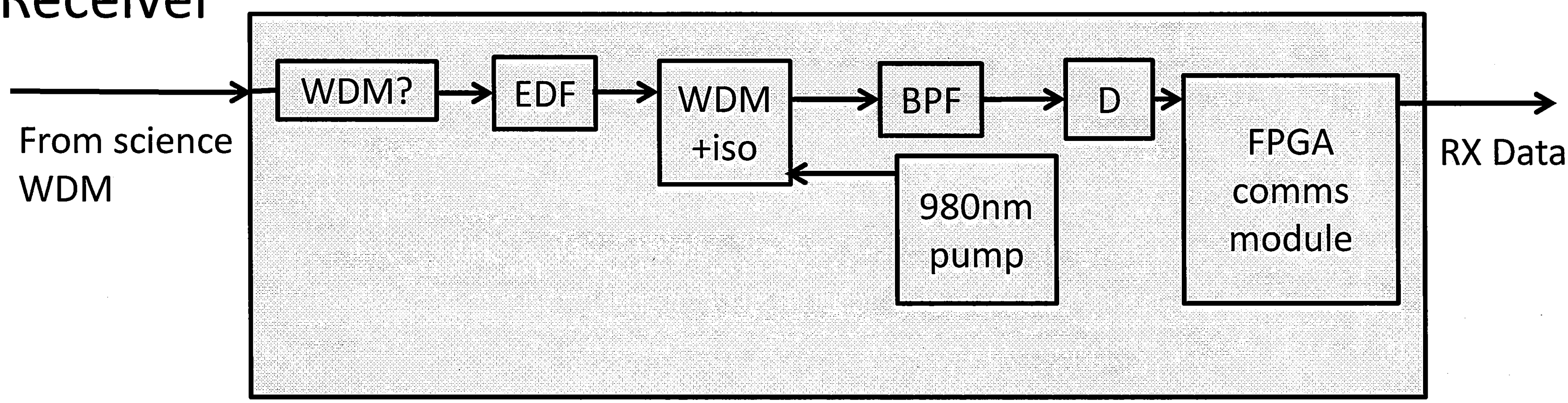

Comb
Science light
Grating/WDM
FSO terminal
Comm. light
ADC

Incoherent comms, separate RX and TX

Transmitter

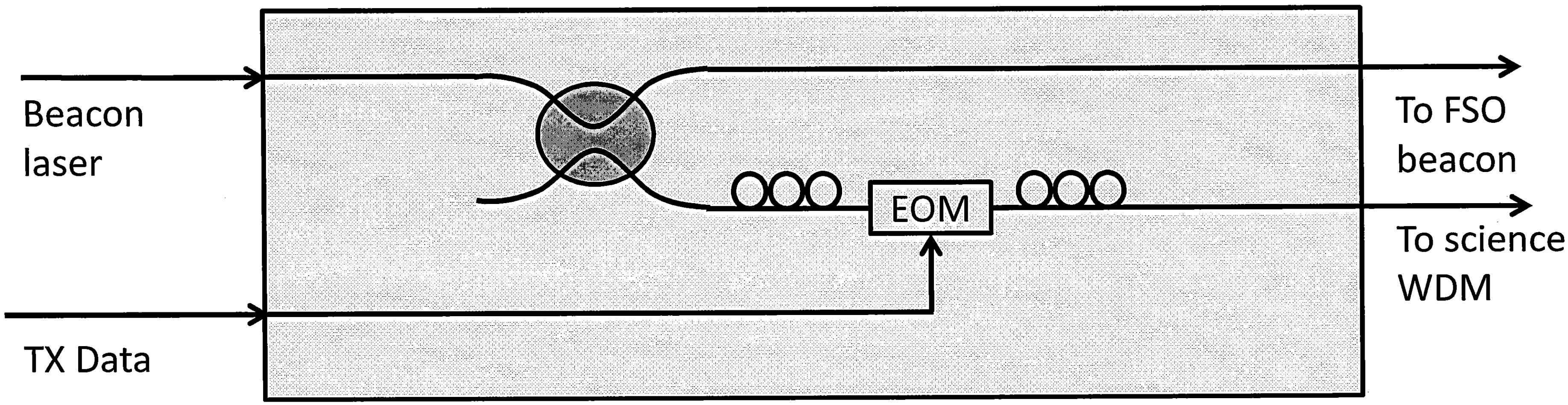


Receiver

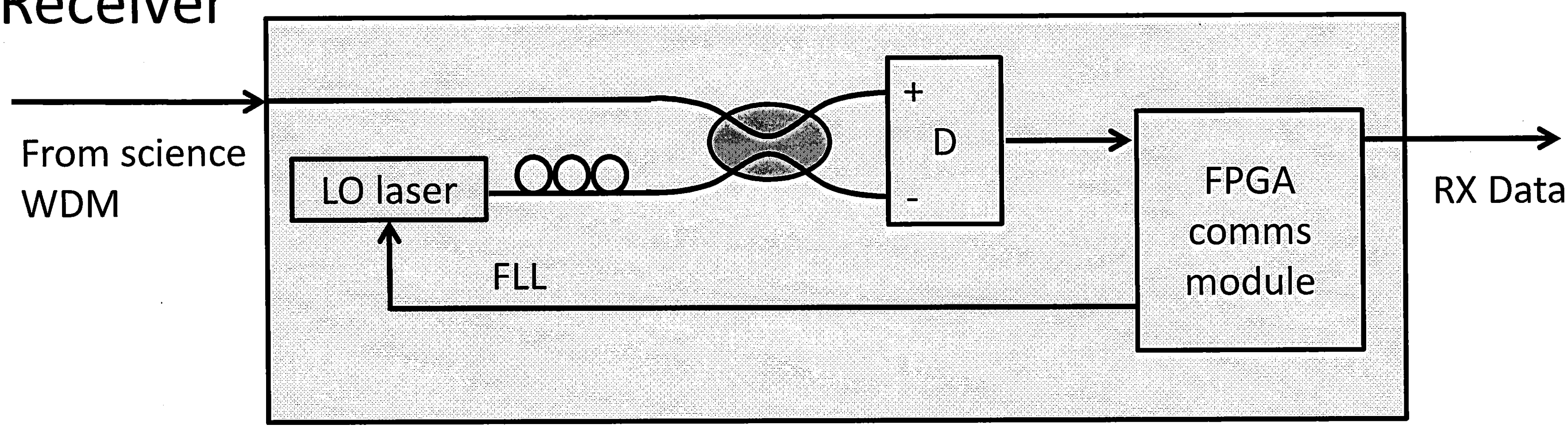


Coherent comms, separate RX and TX

Transmitter



Receiver



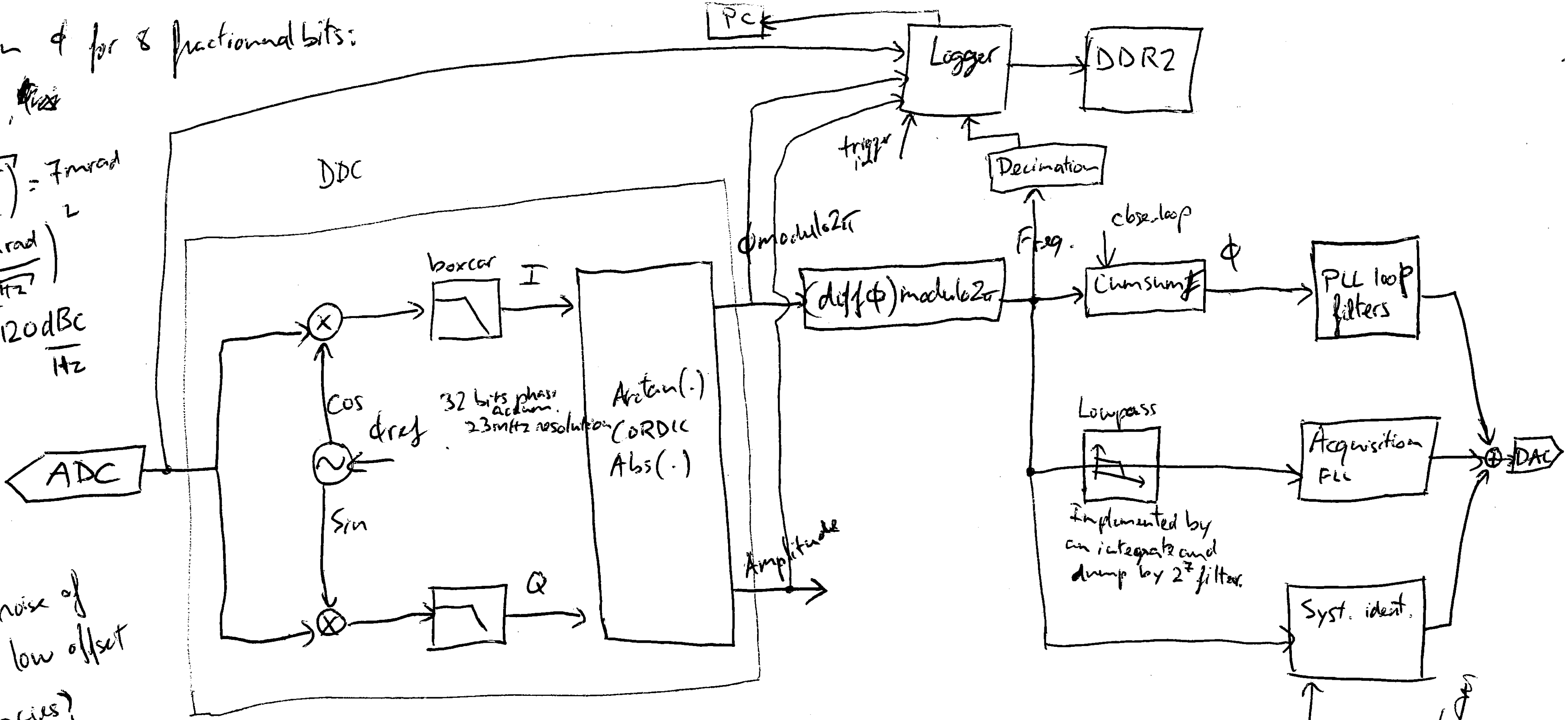
Noise on ϕ for 8 fractional bits:

$\Delta = \frac{2\pi}{2^8}$,

$\sigma_\phi = \sqrt{\frac{\Delta^2}{12}} = 7\,\text{mrad}$

$S_\phi = \left(\frac{1\,\mu\text{rad}}{\sqrt{\text{Hz}}}\right)^2$

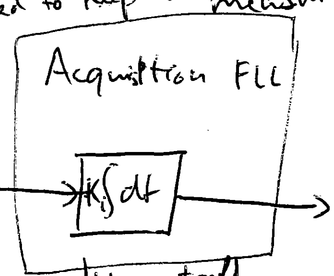
$S_\phi = -120\,\frac{\text{dBc}}{\text{Hz}}$



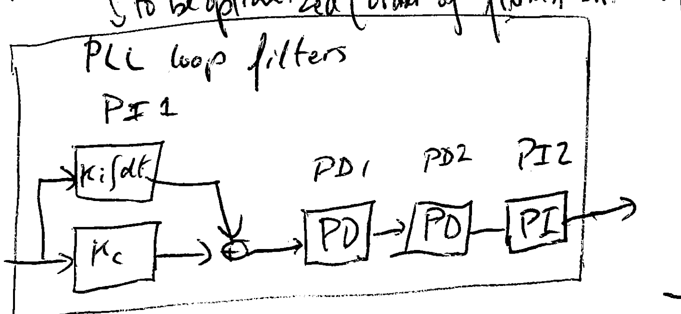
-> Phase noise of ADC at low offset frequencies?

-> Online removal by measuring reference tone (f_r?) and feeding back to f_{ref}. (Later)

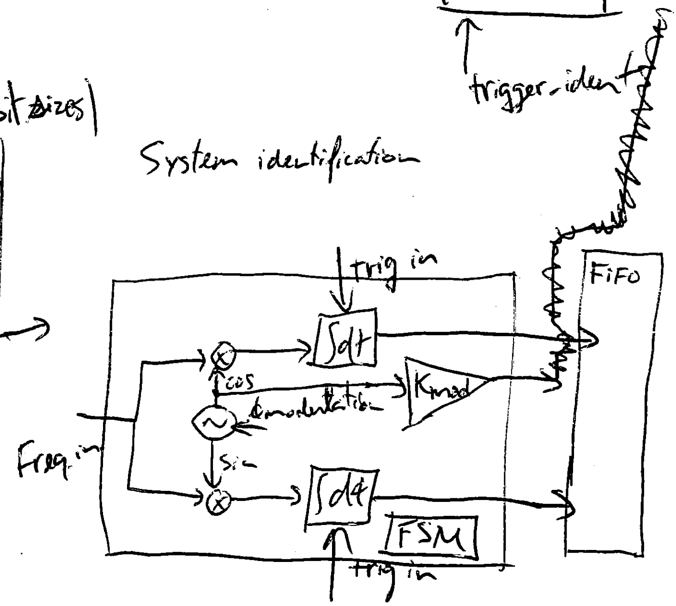
Used to keep beat note within measurable range



↓ to be optimized (order of filters, bit sizes)

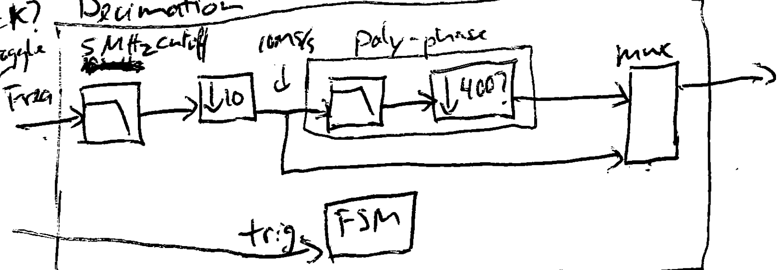


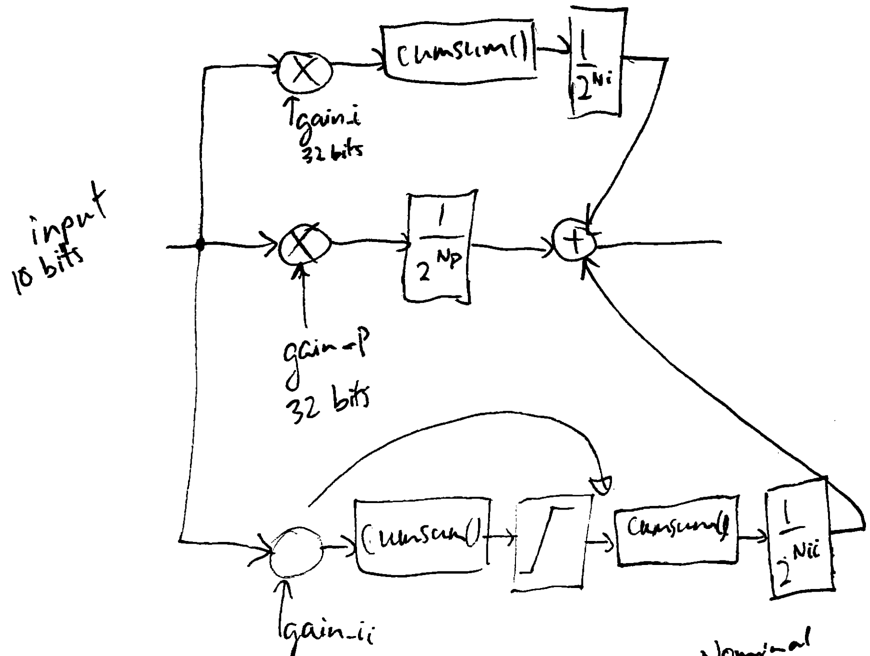
System identification



-> Can we enable a smooth change in PLL parameters with breaking lock?

-> Twice the block, toggle between the two after convergence?





$$H(f) = \frac{K_p}{2^{N_p}} + \frac{K_i}{2^{N_i}} \cdot H_{cumsum}(f)$$

$$H_{cumsum}(f) \simeq \frac{f_s}{j2\pi f}$$

$$H_i = \frac{f_s \cdot K_i}{j2\pi f}$$

$$H_{ii} = \frac{f_s^2 \cdot K_{ii}}{j^2(2\pi f)^2}$$

$$H_i = H_{ii} = K_i = \frac{f_s \cdot K_{ii}}{j2\pi f}$$

$$K_{ii} = \frac{2\pi f}{f_s} \cdot K_i$$

$$\boxed{\frac{1}{2^{N_p}}} \quad \text{vs} \quad \boxed{\frac{1}{2^{N_i}} \cdot \frac{f_s}{j2\pi f}}$$

$$\boxed{\frac{2^{31}}{2^{N_p}}} \quad \text{vs} \quad \boxed{\frac{2^{31}}{2^{N_i}} \cdot \frac{f_s}{j2\pi f}}$$

P branch:

$$\text{min gain} = H_{pmin} = \frac{1}{2^{N_p}} = \frac{1}{2^{16}}$$

$$\text{max gain} = H_{pmax} = \frac{2^{31}-1}{2^{N_p}} \approx 2^{15}$$

I branch:

$$\text{min gain} = H_{imin} = \frac{1}{2^{N_i}} \cdot \frac{f_s}{j2\pi f}$$

$$N_i = 16 + 12 = 28$$

$$\text{max gain} = H_{imax} = \frac{2^{31}}{2^{N_i}} \cdot \frac{f_s}{j2\pi f} = \frac{2^3 \cdot f_s}{j2\pi f}$$

ex. at $f = 1\text{MHz}$, min gain =

$$\text{max gain} = \frac{2^3 \cdot 100e6}{j \cdot 2\pi \cdot 1e6} = \frac{800}{j2\pi} \simeq \frac{100}{j}$$

$$K_p = \frac{gain\_p}{2^{N_p}}$$

Nominal $K_p = 1$

$gain\_p = 2^{24}$

$$2^{N_p} = \frac{2^{24}}{1}$$

$$K_i = \frac{gain\_i \cdot f_s}{2^{N_i}}$$

Nominal $K_i = 95e6$

$gain\_i = 2^{24}$

$$2^{N_i} = 2^{24} \cdot \frac{100e6}{95e6} \sim 2^{24}$$

$$K_{ii} = \frac{gain\_ii \cdot f_s^2}{2^{N_{ii}}}$$

Nominal $K_{ii} = 4e12$

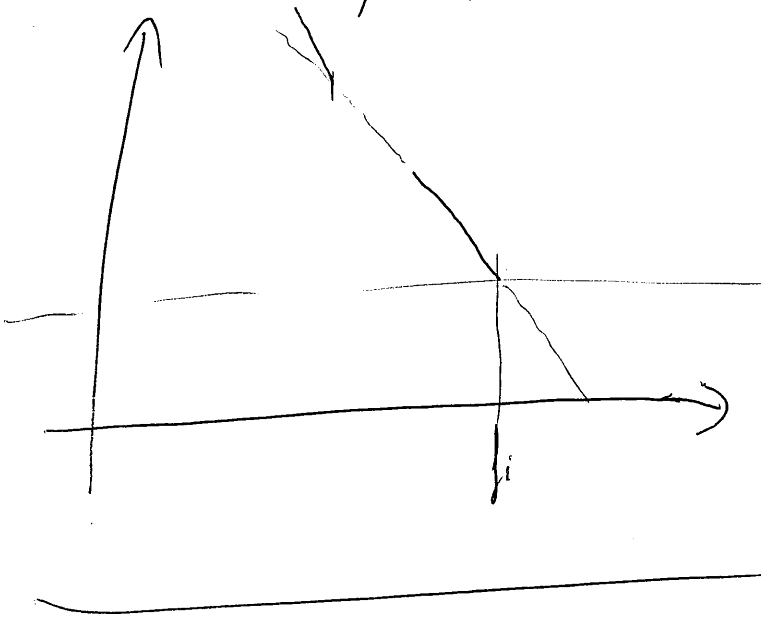
$gain\_ii = 2^{24}$

$$2^{N_{ii}} = 2^{24} \cdot 2500 = 2^{24+11} = 2^{35}$$

I^2 branch:

$$H_{ii} = \frac{K_{ii}}{2^{N_{ii}}} \cdot \left(H_{cumsum}(f)\right)^2 = \frac{K_{ii}}{2^{N_{ii}}} \cdot \frac{f_s^2}{j^2(2\pi f)^2}$$

Case 1: f_i relative to 0dB.



P gain = ~~Kp~~ Kp

integrator gain = $\frac{2\pi f_i}{2\pi f}$

I^2 gain = $\frac{2\pi f_i}{2\pi f} \cdot \frac{2\pi f_{ii}}{2\pi f}$

Critical points:

Actual transfer function is:

I: $H_i = gain\_i \cdot \frac{f_s}{2\pi f}$

I^2: $H_{ii} = gain\_ii \cdot \left(\frac{f}{2\pi f_s}\right)^2$

We want:

$H_i = \frac{1}{K_c} \cdot \frac{\cancel{2\pi} f_i}{\cancel{2\pi} f} \Rightarrow gain\_i = \frac{1}{K_c} \cdot \frac{2\pi f_i}{f_s}$

$H_{ii} = \frac{1}{K_c} \cdot \frac{\cancel{2\pi} f_i}{\cancel{2\pi} f} \cdot \frac{\cancel{2\pi} f_{ii}}{\cancel{2\pi} f} \Rightarrow gain\_ii = \frac{1}{K_c} \cdot f_i \cdot f_{ii} \cdot \left(\frac{2\pi}{f_s}\right)^2$

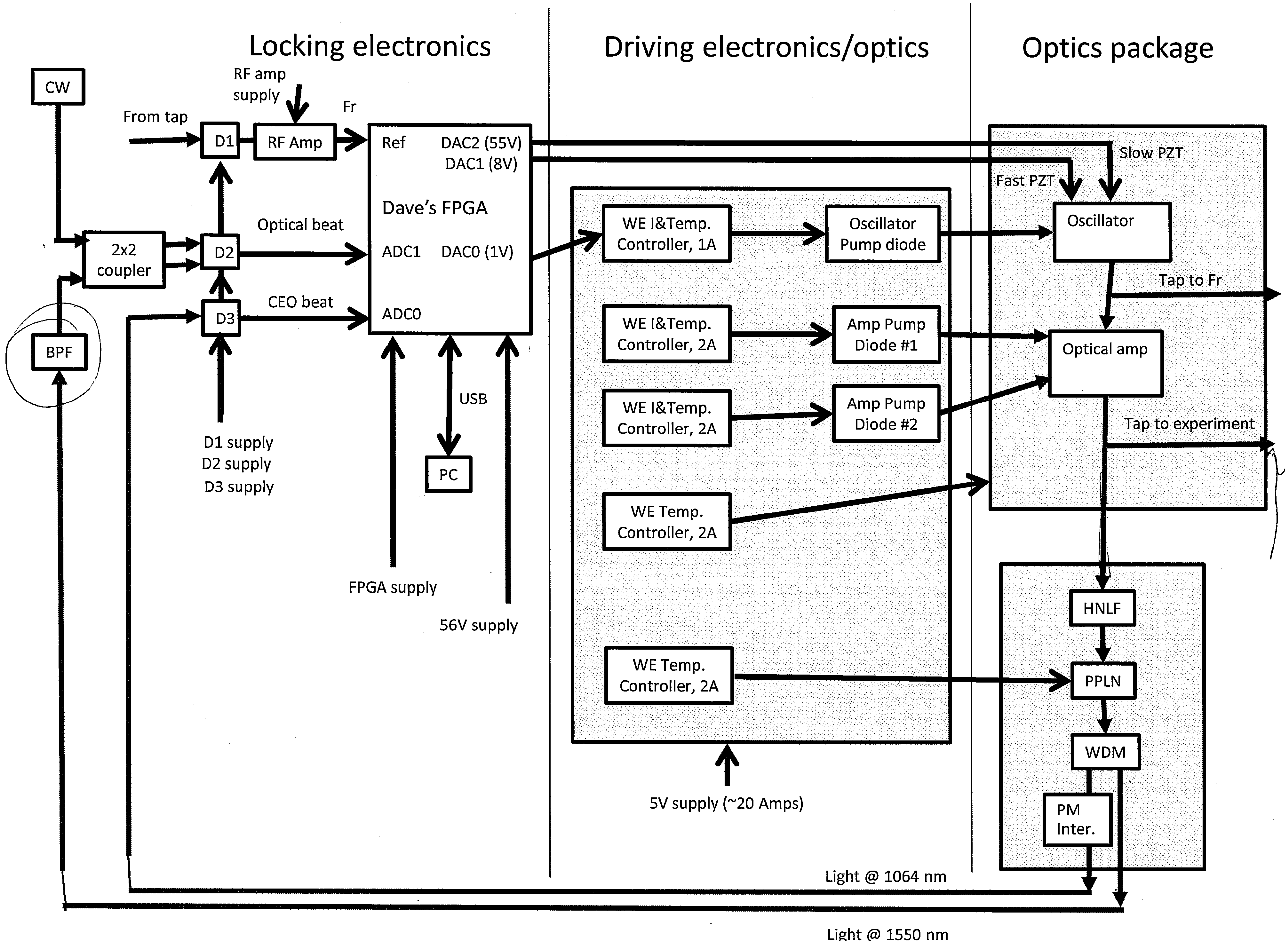
We want

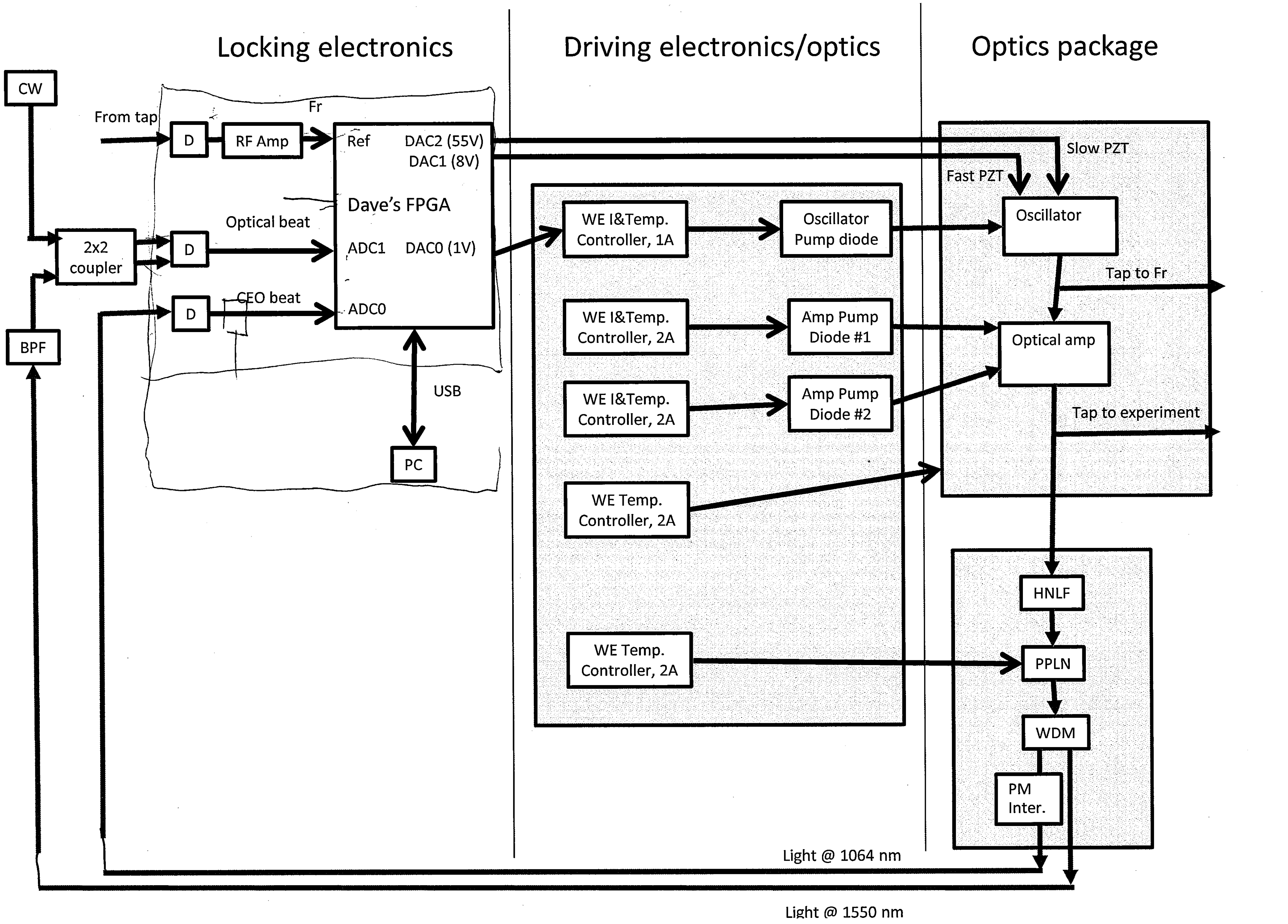
main

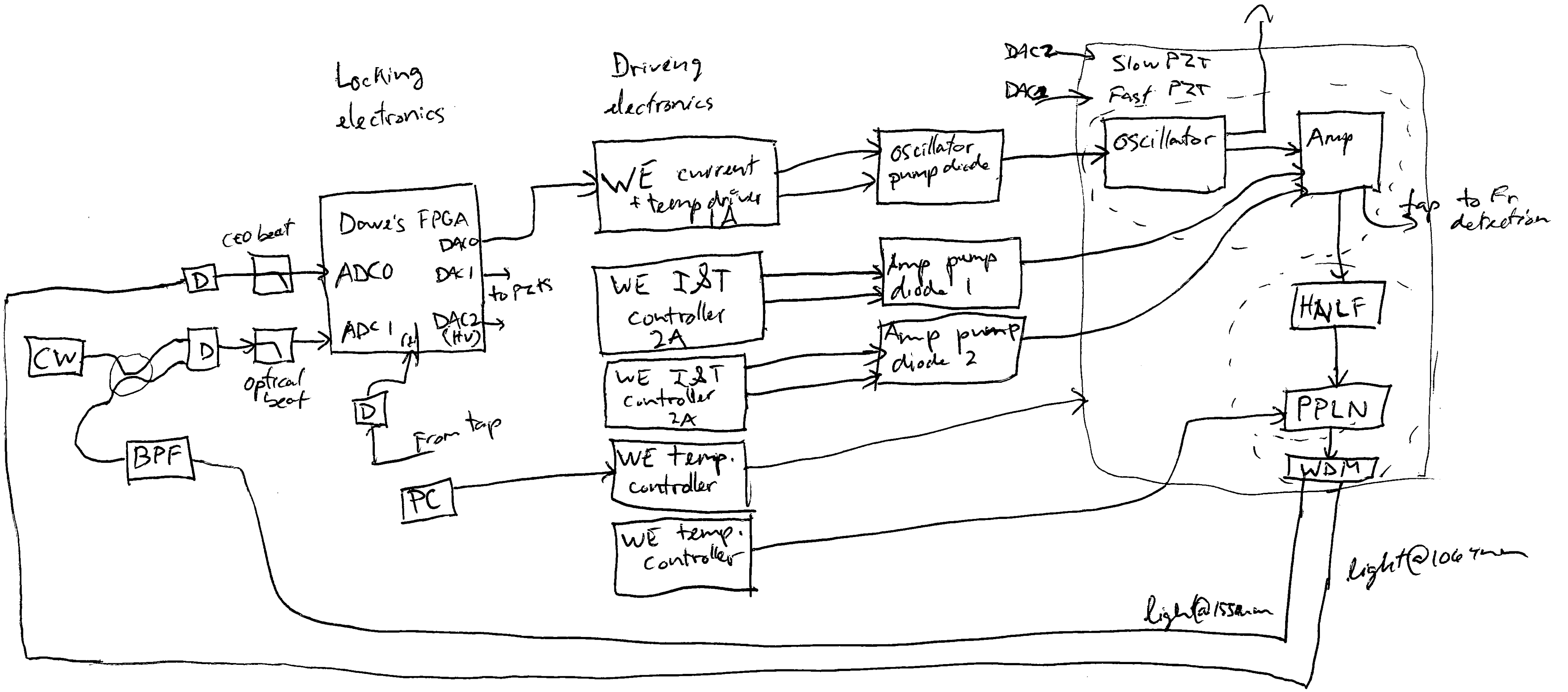
sl = SuperlaserLand() (~~[illegible]~~ creates the object. first ref

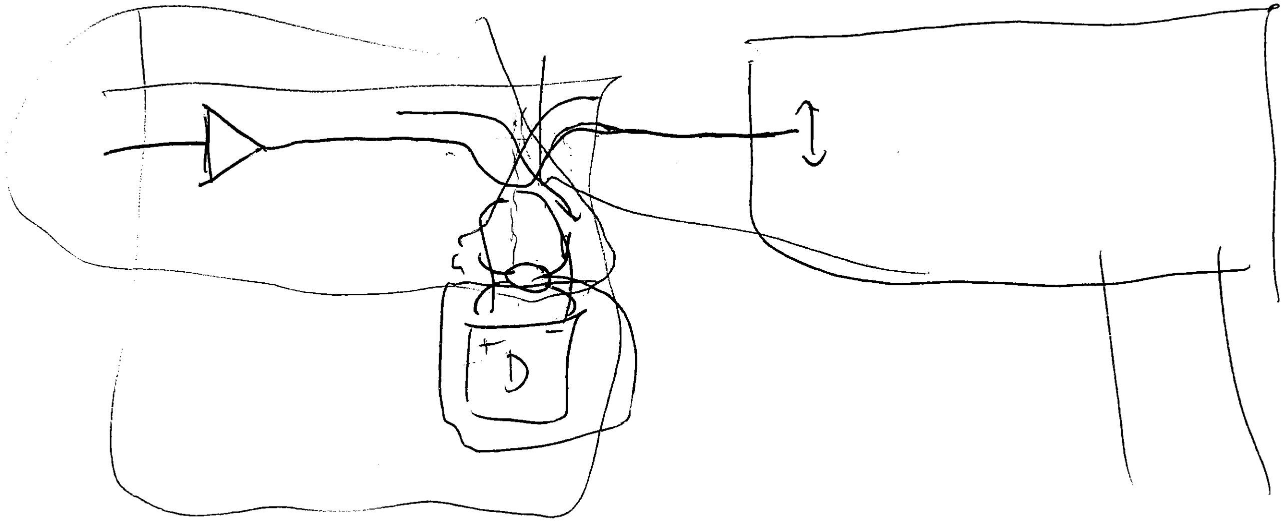
xem = XEM_main_window (sl) keeps a reference to sl

child widget: Loopfilters UI keeps a reference to sl

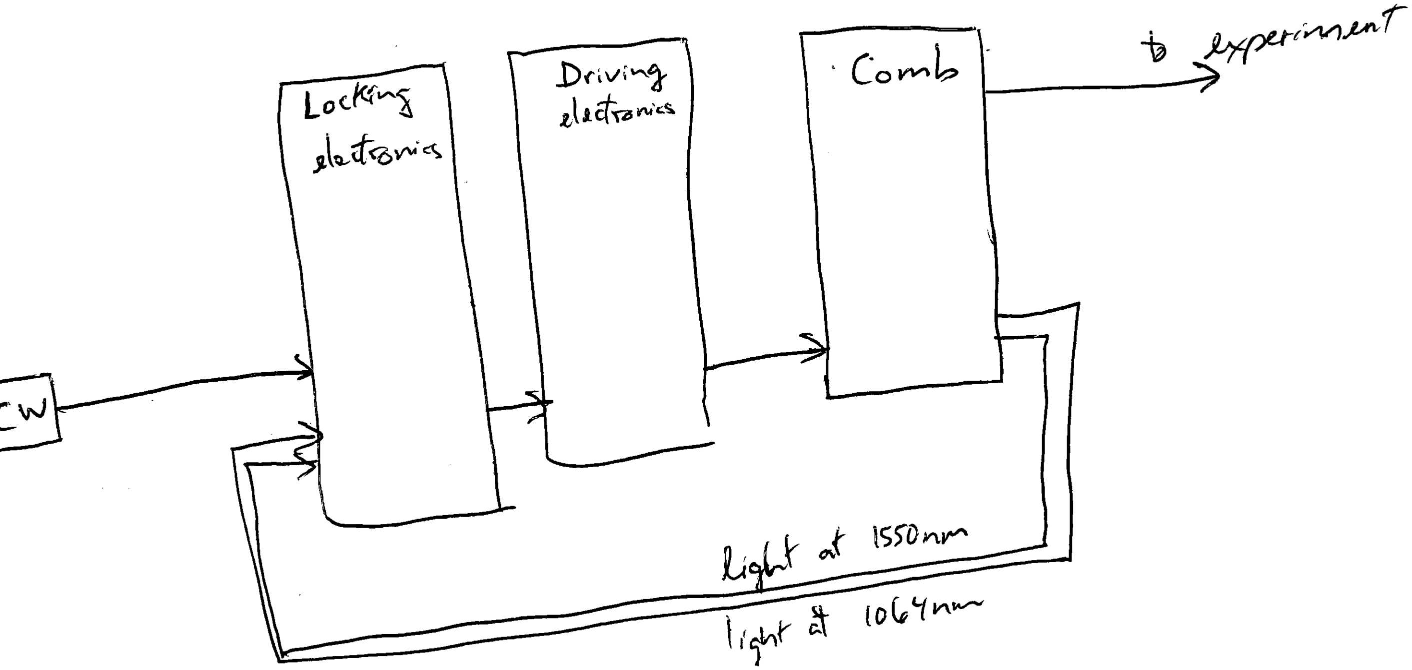

Locking electronics
Driving electronics/optics
Optics package
CW
RF amp supply
From tap
D1
RF Amp
Fr
Ref
DAC2 (55V)
DAC1 (8V)
Dave's FPGA
2x2 coupler
D2
Optical beat
ADC1
DAC0 (1V)
D3
CEO beat
ADC0
BPF
D1 supply
D2 supply
D3 supply
USB
PC
FPGA supply
56V supply
Slow PZT
Fast PZT
Oscillator
WE I&Temp. Controller, 1A
Oscillator Pump diode
Tap to Fr
WE I&Temp. Controller, 2A
Amp Pump Diode #1
Optical amp
WE I&Temp. Controller, 2A
Amp Pump Diode #2
Tap to experiment
WE Temp. Controller, 2A
HNLF
WE Temp. Controller, 2A
PPLN
WDM
PM Inter.
5V supply (~20 Amps)
Light @ 1064 nm
Light @ 1550 nm

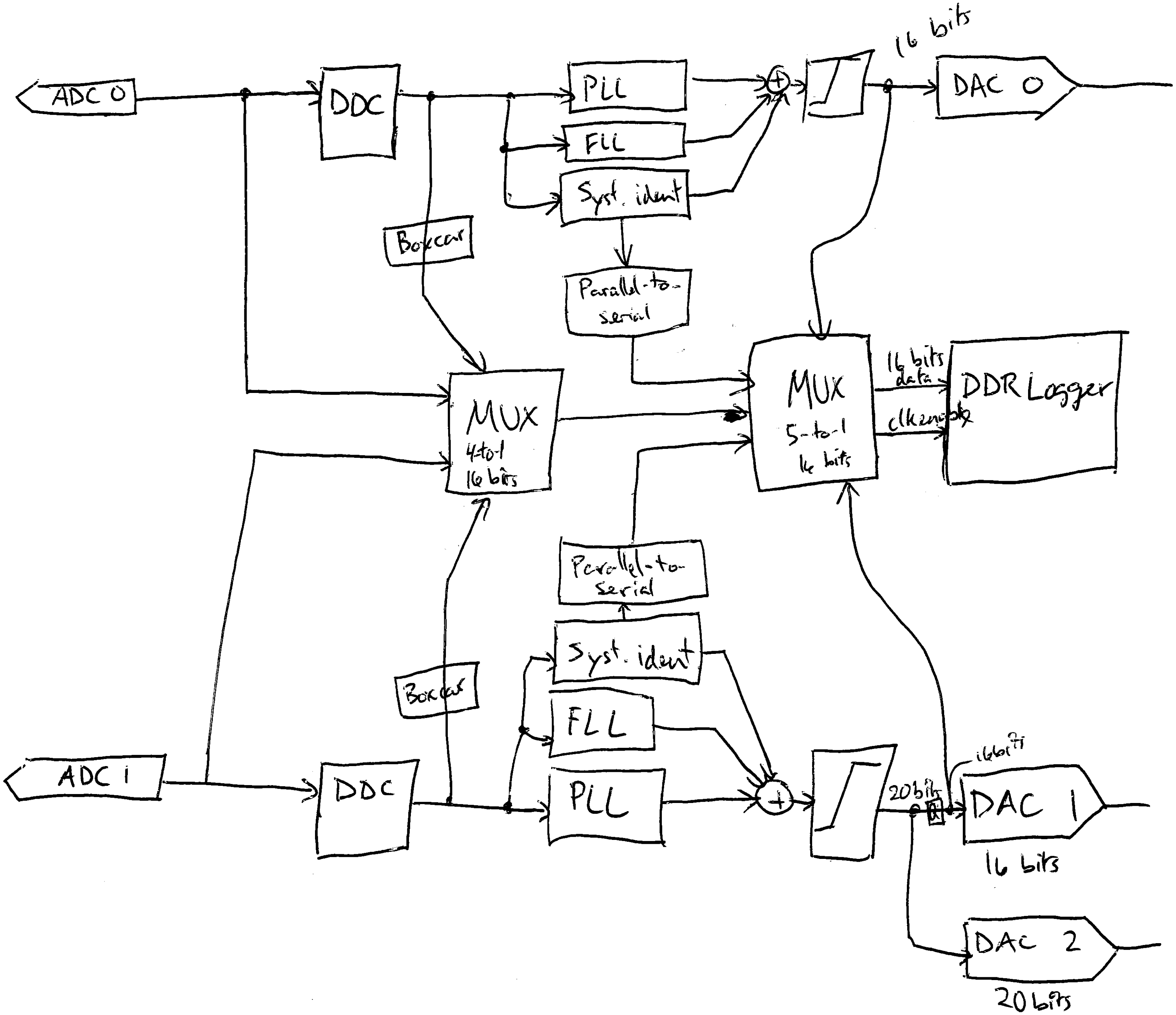

Locking electronics
Driving electronics/optics
Optics package
CW
From tap
Fr
D
RF Amp
Ref
DAC2 (55V)
DAC1 (8V)
Dave's FPGA
Optical beat
ADC1
DAC0 (1V)
CEO beat
ADC0
2x2 coupler
D
D
BPF
USB
PC
WE I&Temp. Controller, 1A
Oscillator Pump diode
WE I&Temp. Controller, 2A
Amp Pump Diode #1
WE I&Temp. Controller, 2A
Amp Pump Diode #2
WE Temp. Controller, 2A
WE Temp. Controller, 2A
Slow PZT
Fast PZT
Oscillator
Tap to Fr
Optical amp
Tap to experiment
HNLF
PPLN
WDM
PM Inter.
Light @ 1064 nm
Light @ 1550 nm

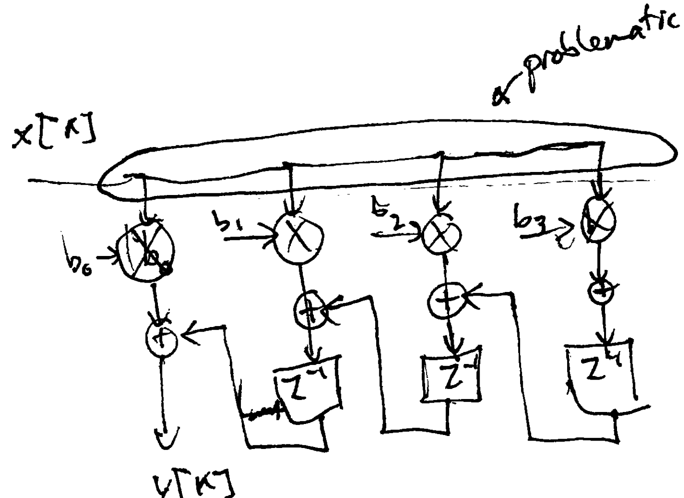

Locking electronics
Driving electronics
Dave's FPGA
ADC0
ADC1
DAC0
DAC1
DAC2 (HV)
to PZTs
CEO beat
optical beat
D
CW
BPF
From tap
PC
WE current + temp driver 1A
WE I&T controller 2A
WE I&T controller 2A
WE temp. controller
WE temp. controller
oscillator pump diode
Amp pump diode 1
Amp pump diode 2
DAC2
DAC1
Slow PZT
Fast PZT
Oscillator
Amp
tap to fr detection
HNLF
PPLN
WDM
light@1064nm
light@1550nm



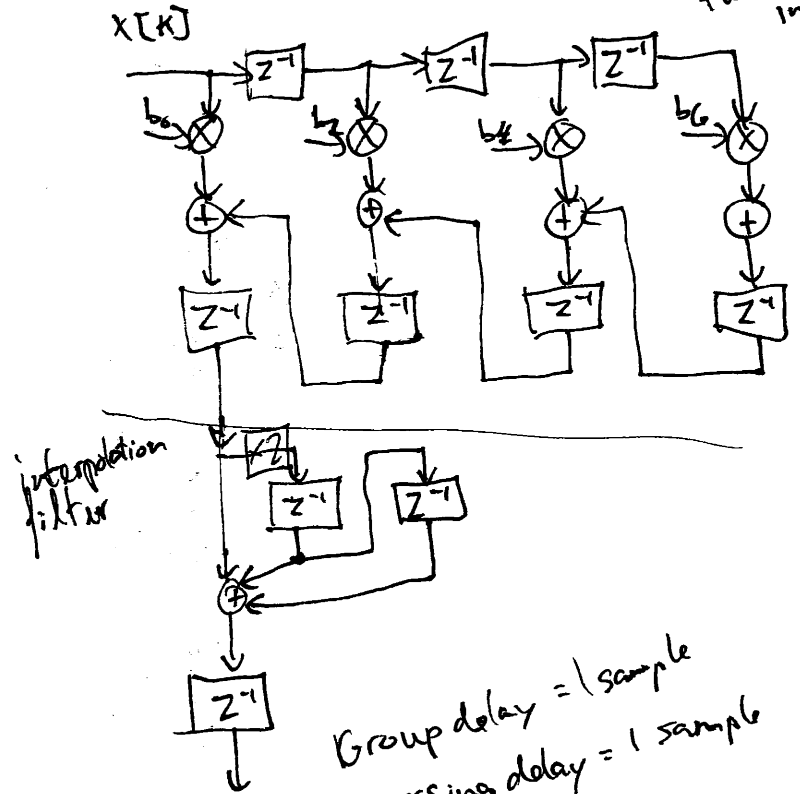
High-level diagram




ADC 0
DDC
PLL
FLL
Syst. ident
16 bits
DAC 0
Boxcar
Parallel-to-serial
MUX
4-to-1
16 bits
MUX
5-to-1
16 bits
16 bits
data
clkenable
DDR Logger
Parallel-to-serial
Syst. ident
Boxcar
FLL
ADC 1
DDC
PLL
20 bits
16 bits
DAC 1
16 bits
DAC 2
20 bits



- Needs interpolated FIR (every other sample is 0)



- The FPGA doesn't have two carry chains that operate in ~~[illegible]~~ opposite directions like this so the top shift register would need to be implemented in fabric which is OK.

- This gives only one extra delay in addition to the filter's group delay.

Group delay = 1 sample
+ processing delay = 1 sample

Need also to take into account the processing + group delay of the interpolation filter

- AM-to-PM conversion in the presence of out-of-band phase noise
Conversion factor is given by rms relative ~~phase~~ jitter over the full bandwidth and slope of tone
$\propto \Delta\theta_{RMS} f$ (maybe with 2π)
Probably almost same as static phase error.

First-order filter:

$$H(f) = \frac{1}{1 + jf/f_c}$$

Phase below cut-off:

$$\phi(f) \simeq -f/f_c$$

Equivalent delay : $T_{eq} \simeq \frac{1}{2\pi f_c}$

for $f_c = 200\text{kHz}$,

$$T_{eq} \simeq 800\text{ns}$$

Gain increase as double integrator for freqs below $(2\pi T_{ii})^{-1}$

$$H_c(s) = \left(\frac{T_1 s + 1}{T_1 s}\right) K \underbrace{\left(1 + \frac{1}{T_{ii} s}\right)}$$

$$= K\left(1 + (T_1 s)^{-1}\right)\left(1 + (T_{ii} s)^{-1}\right)$$

$$= K\left(1 + (T_1^{-1} + T_{ii}^{-1})\, s^{-1} + T_1 T_{ii} s^{-2}\right)$$

$K_p = K$

$K_i = \frac{K}{(T_1^{-1} + T_{ii}^{-1})}$

$K_{ii} = \frac{K}{T_1 T_{ii}}$

Pure delay:

$$H(f) = \exp(-j 2\pi f T)$$

Phase:

$$\phi(f) = -2\pi f T$$

Our current processing delay $T = 570\text{ns}$
(includes filters group delay)

$$H_p(f) = \frac{1}{1 + jf/f_1} \cdot \frac{1}{1 + jf/f_2}$$

$$H_c(f) = \frac{1 + jf/f_2}{1}$$

$$H_p(s) = \frac{1}{T_1 s + 1} \cdot \frac{1}{T_2 s + 1}$$

$$H_c(s) = \underbrace{\frac{T_2 s + 1}{1}}_{\text{Compensation for 2nd pole}} \cdot \underbrace{\frac{T_1 s + 1}{T_1 s}}_{\text{PI}} \cdot K$$

$$G(s) = H_p(s) H_c(s) = \frac{K}{T_1 s}$$

$$H_{cl}(s) = \frac{G(s)}{1 + G(s)} = \frac{(T_1 s)^{-1} K}{1 + (T_1 s)^{-1} K} = \frac{1}{\left(\frac{T_1}{K}\right) s + 1}$$

$$H_c(s) = K_p + \frac{K_i}{s} + K_d s$$

$$H_c(s) = (T_2 s + 1)\left(1 + (T_1 s)^{-1}\right) K$$

$$= \left(T_2 s + 1 + \frac{T_1}{T_2} + \frac{1}{T_1 s}\right) K$$

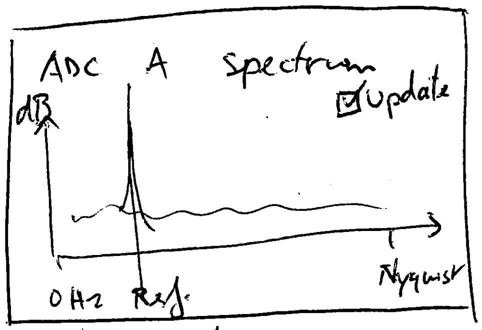
$$= \left(1 + \frac{T_1}{T_2}\right) K + \frac{K}{T_1 s} + K T_2 s$$

$K_d = K T_2$

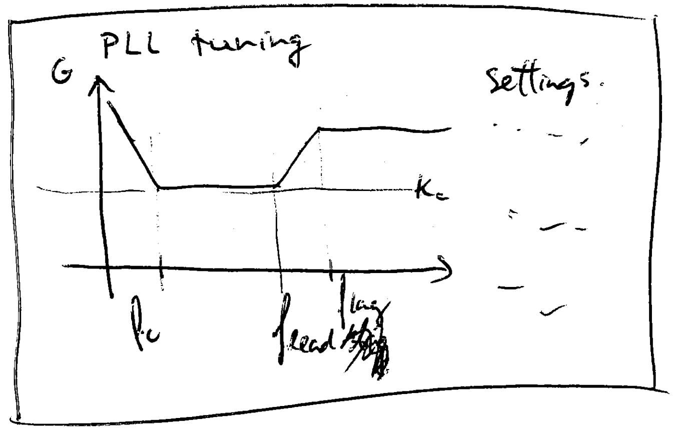
$K_i = \frac{K}{T_1}$

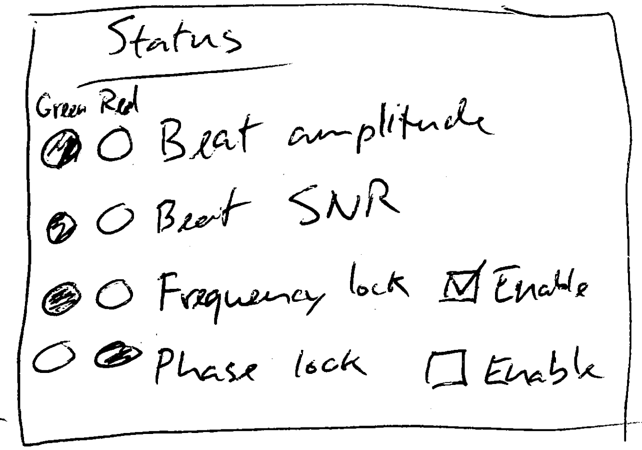
$K_p = \left(1 + \frac{T_1}{T_2}\right) K_c$

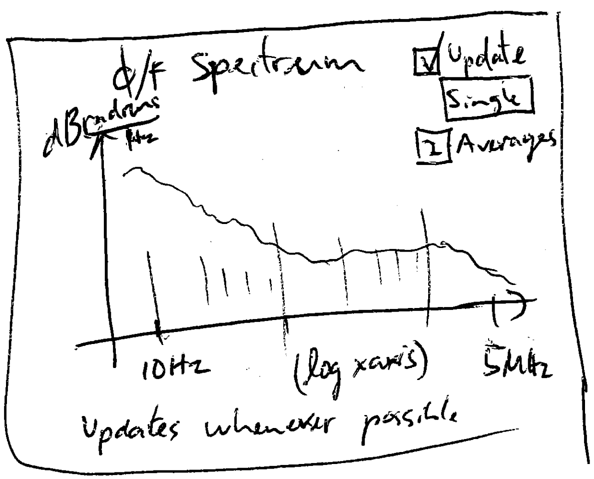
User interface (This works for a single lock)



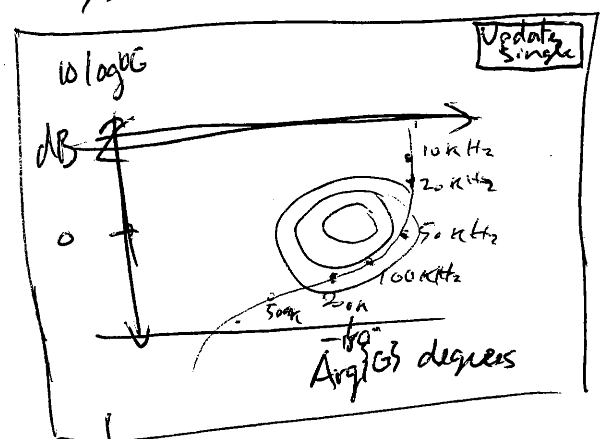
Updates @ 1 fps or more
FFT over 100-1k samples



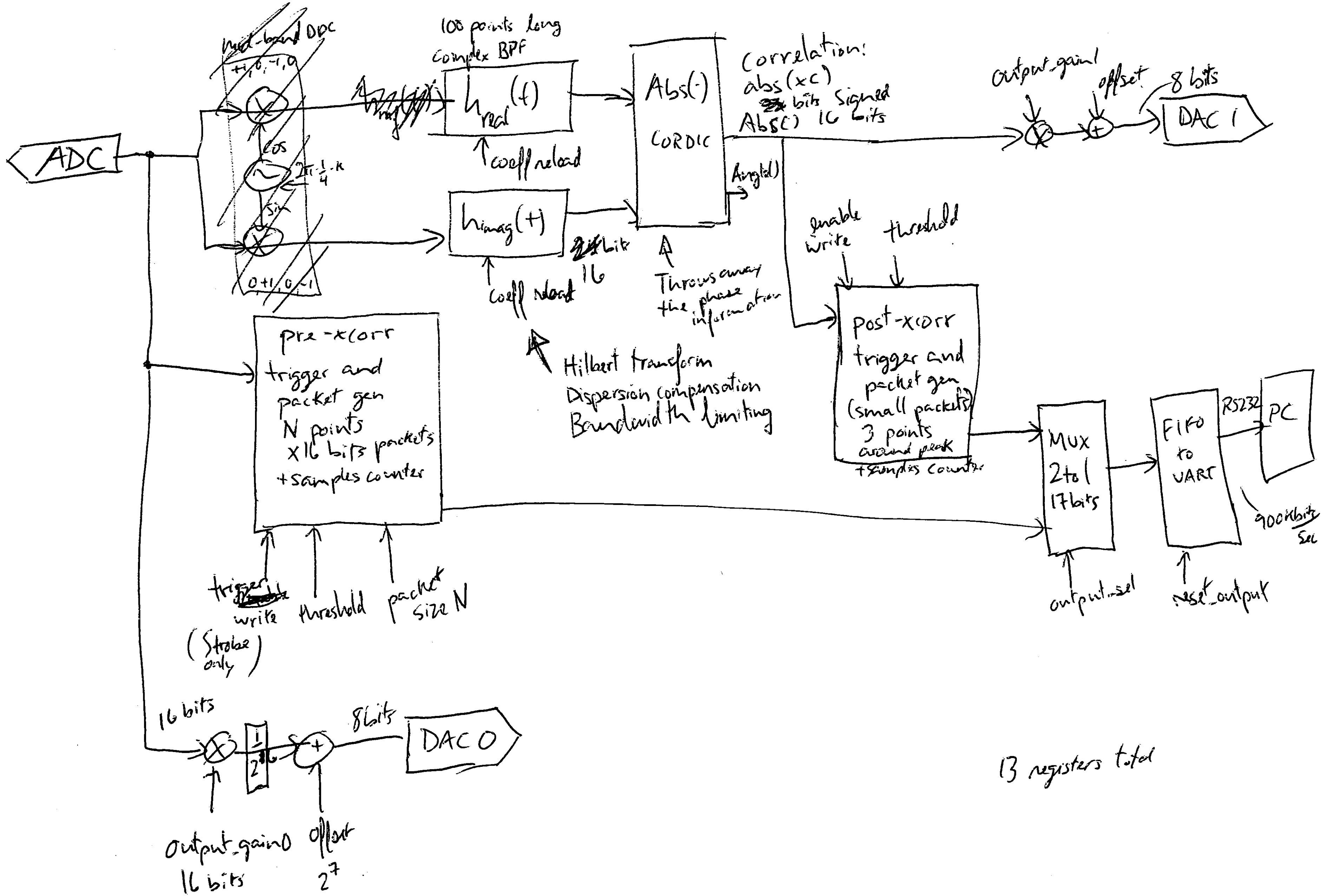




Syst. identification

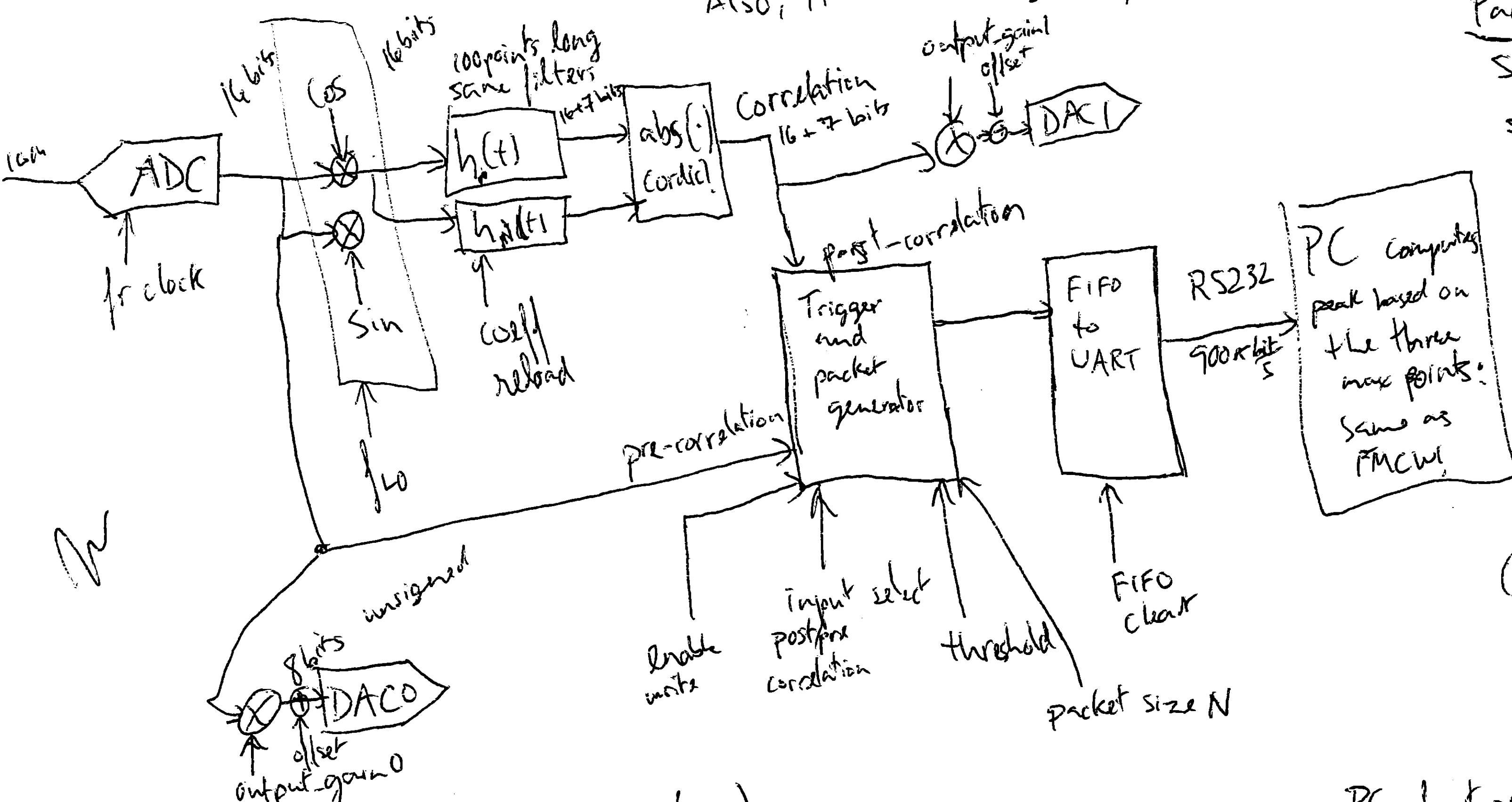


f_{min}
f_{step}
Number of freq.
Amplitude



13 registers total

$abs(\cdot)^2$ is simpler but is it correct?
Also, it has a larger dyn range



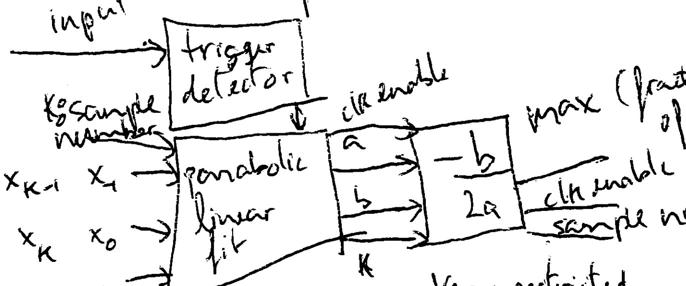
Packets must consist of:

Sample # (K)
sample $K-\left(\frac{N-1}{2}\right)$
Sample $K-\left(\frac{N-1}{2}\right)+1$
(...)
sample $K+1$
sample $K+2$
(...)
Sample $K+\left(\frac{N-1}{2}\right)$

N must be odd

(same code as in FMCW?)
Need rolling buffer

One step further: (see Julien's code...)



(see Julien's thesis for coefficients)

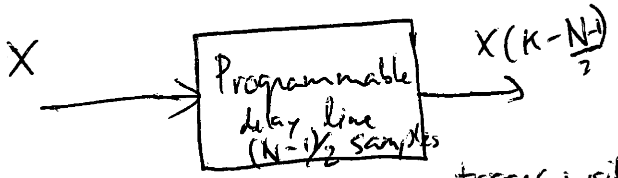
Very restricted output range needed: Implement pre-scaling of both inputs, then lookup table

needs $x_0 \geq x_{-1}$, $x_0 \geq x_1$	How to ensure?

Output of this goes to loop filter for PLL...

PC first sets registers to record one IGM, computes the filter coefficients to de-chirp the IGM, ~~then~~ sends the coefficients back, sets the trigger selection to post-correlation, a small number of points for realtime operation, then starts reading continuously.

pre-xcorr, large buffer packet gen

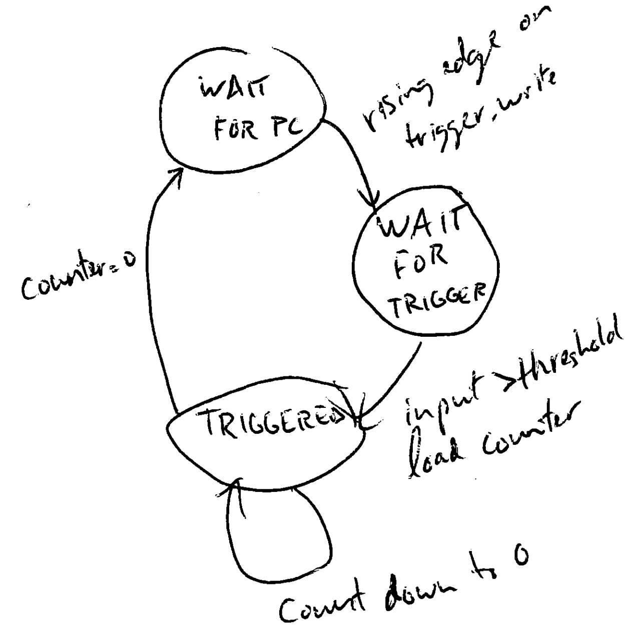


if rising edge on ~~output_is_ready~~ trigger_write output_is_ready <= '1'

if input > threshold and input(n-1) < threshold and output_is_ready
~~start output-counter~~ is_triggered <= '1'
end if

output_clk_enable <= '1' when output_counter > 0
'0' when else

if rising-edge on is_triggered
output <= samples_counter
else
output <= $x(k-\frac{N-1}{2})$



---

post-xcorr
trigger and packet gen

trigger if:
$x_0 \geq x_{-1}$ and $x_0 \geq x_{+1}$ and x_0 > threshold

~~if not triggered:~~
run shift register: $x_{-1} <= x_0$
$x_0 <= x_1$
$x_1 <=$ input

~~else~~ if triggered:
copy current shift register to internal registers.
start serializing output to output bus
4 x 24 bits to 6 x 16 bits
samples_counter

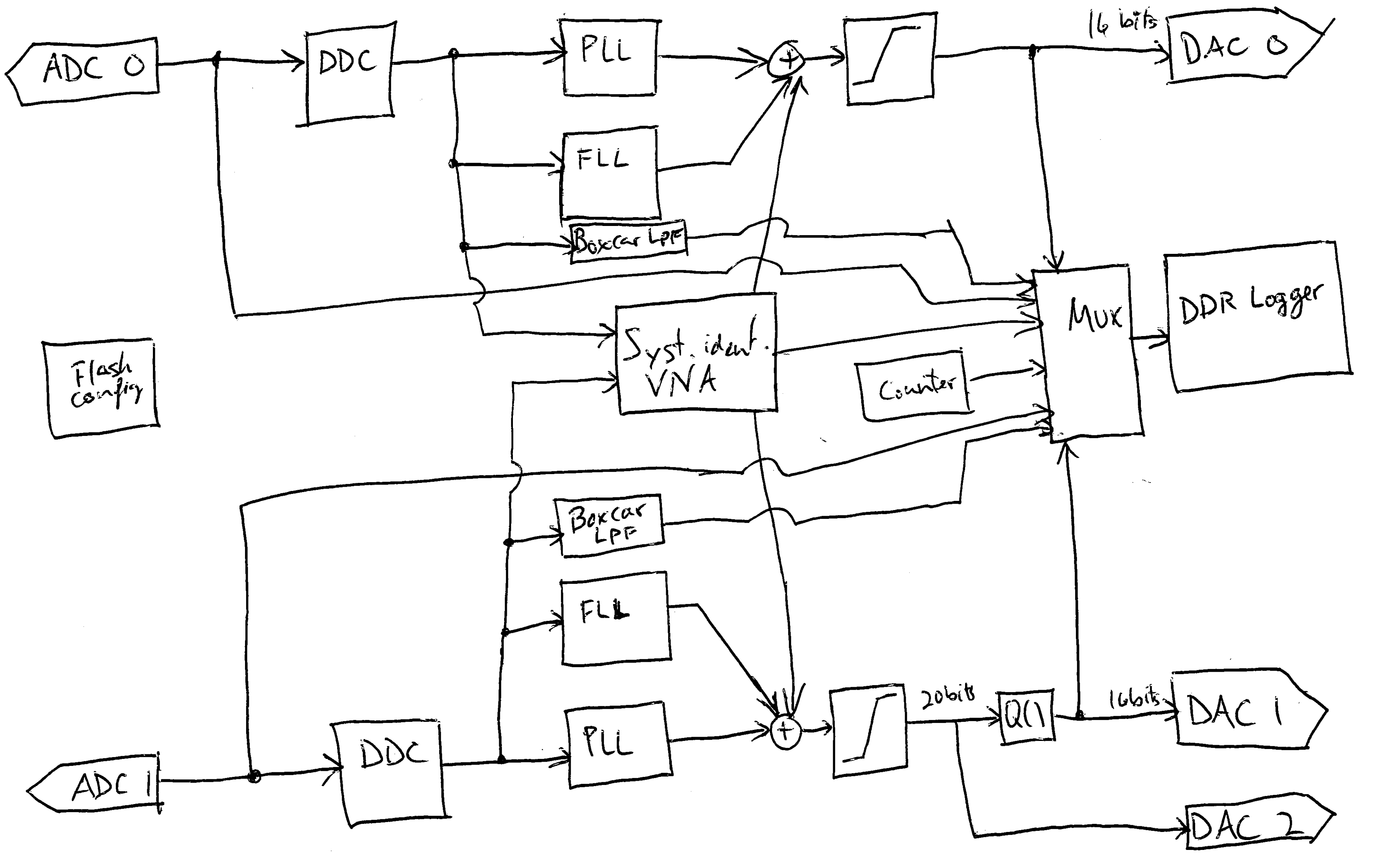
Implement as a single state machine.

States: IDLE if enable-write
WAITING
OUT 0
OUT 1 always
...
OUT 5

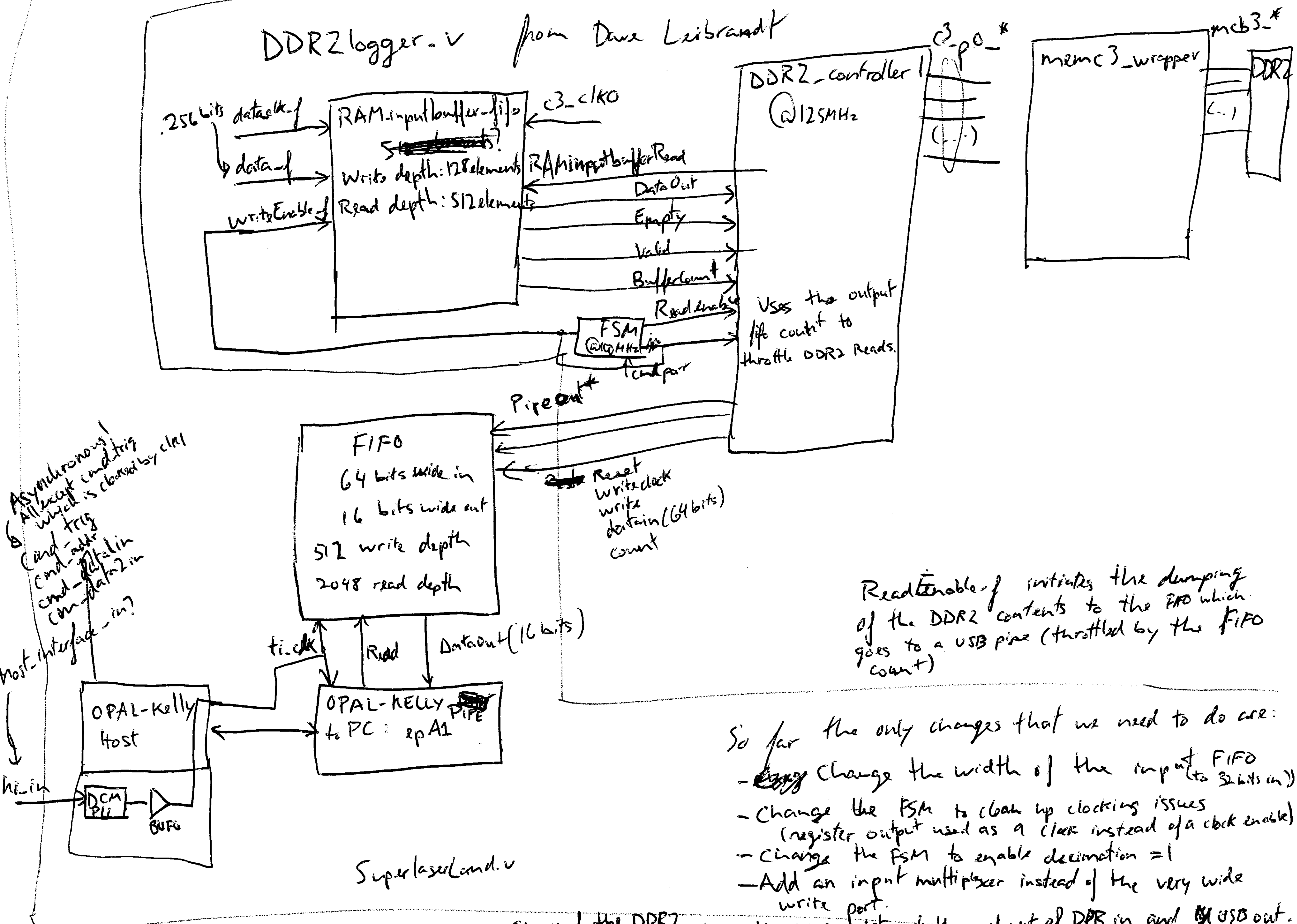
Implement as two modules
- One does the triggering and outputs three points that satisfy: $x_0 \geq x_{-1}$, $x_0 \geq x_1$, $x_0 \geq$ threshold

- One serializes the samples to 16 bits ~~and adds a sample counter~~.

- This way we can add the parabolic peak-finding in between.



ADC 0
DDC
PLL
FLL
Boxcar LPF
16 bits
DAC 0
Mux
DDR Logger
Syst. ident.
VNA
Counter
Flash config
Boxcar LPF
FLL
PLL
DDC
ADC 1
20bits
Q()
16bits
DAC 1
DAC 2



ReadEnable_f initiates the dumping of the DDR2 contents to the FIFO which goes to a USB pipe (throttled by the FIFO count)

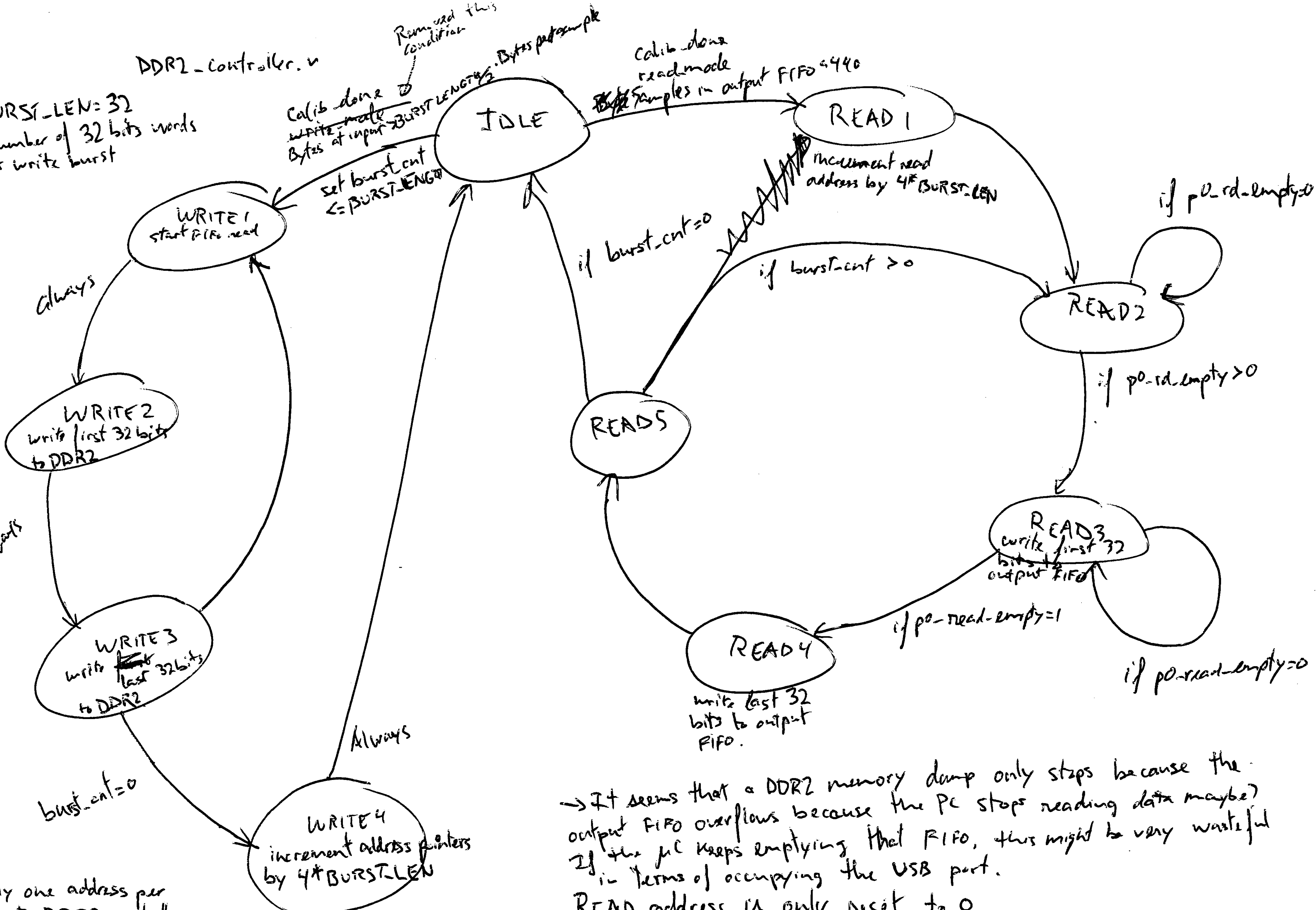
So far the only changes that we need to do are:

- Change the width of the input FIFO (to 32 bits in)
- Change the FSM to clean up clocking issues (register output used as a clock instead of a clock enable)
- Change the FSM to enable decimation = 1
- Add an input multiplexer instead of the very wide write port.
- Measure actual throughput of DDR in and USB out.
- Initiate dumping of RAM as soon as write is finished instead?

- Check if the DDR2 dump ever stops and maybe fix that.

DDR2_controller.v

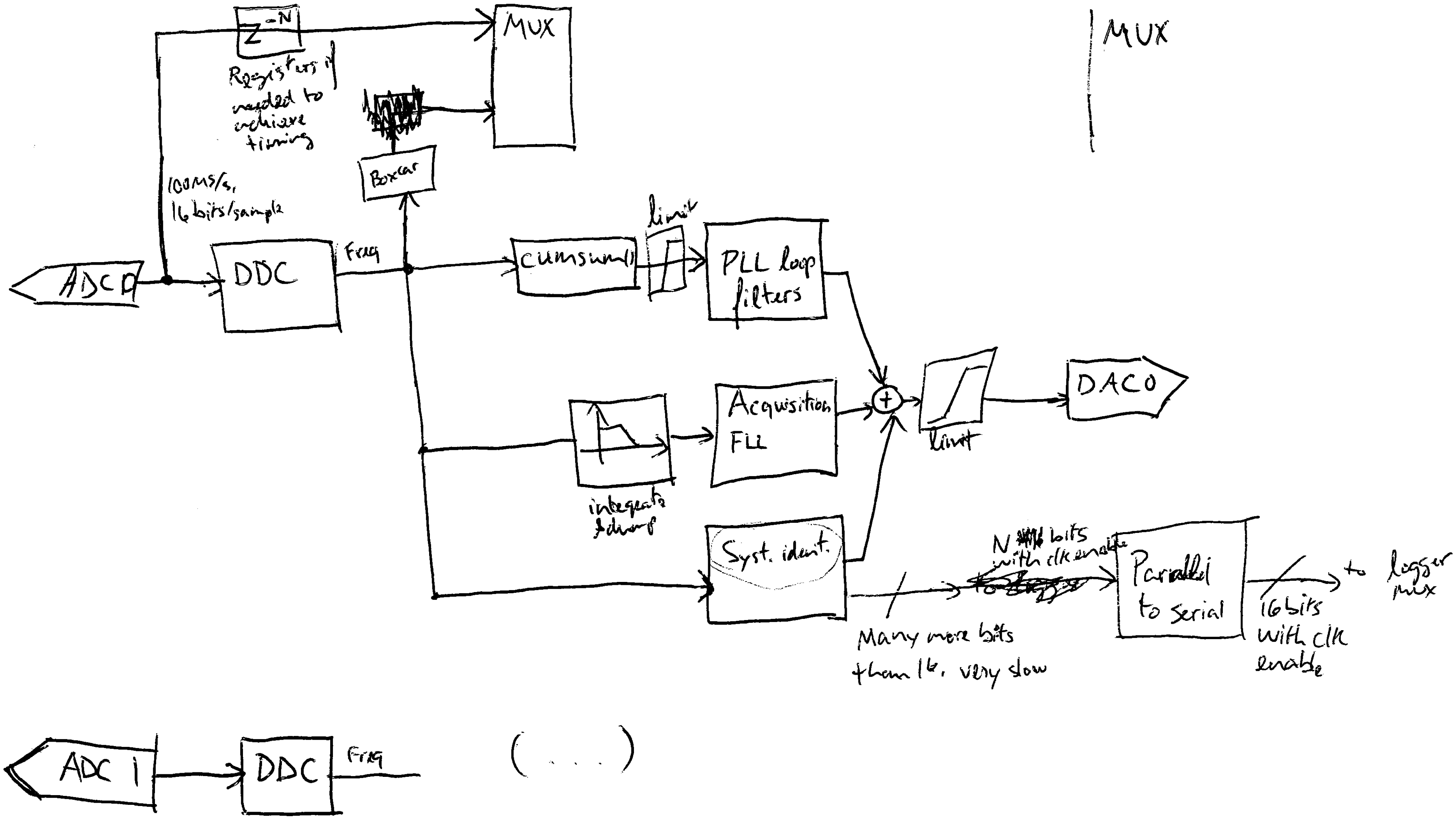
BURST_LEN: 32
Number of 32 bits words
per write burst



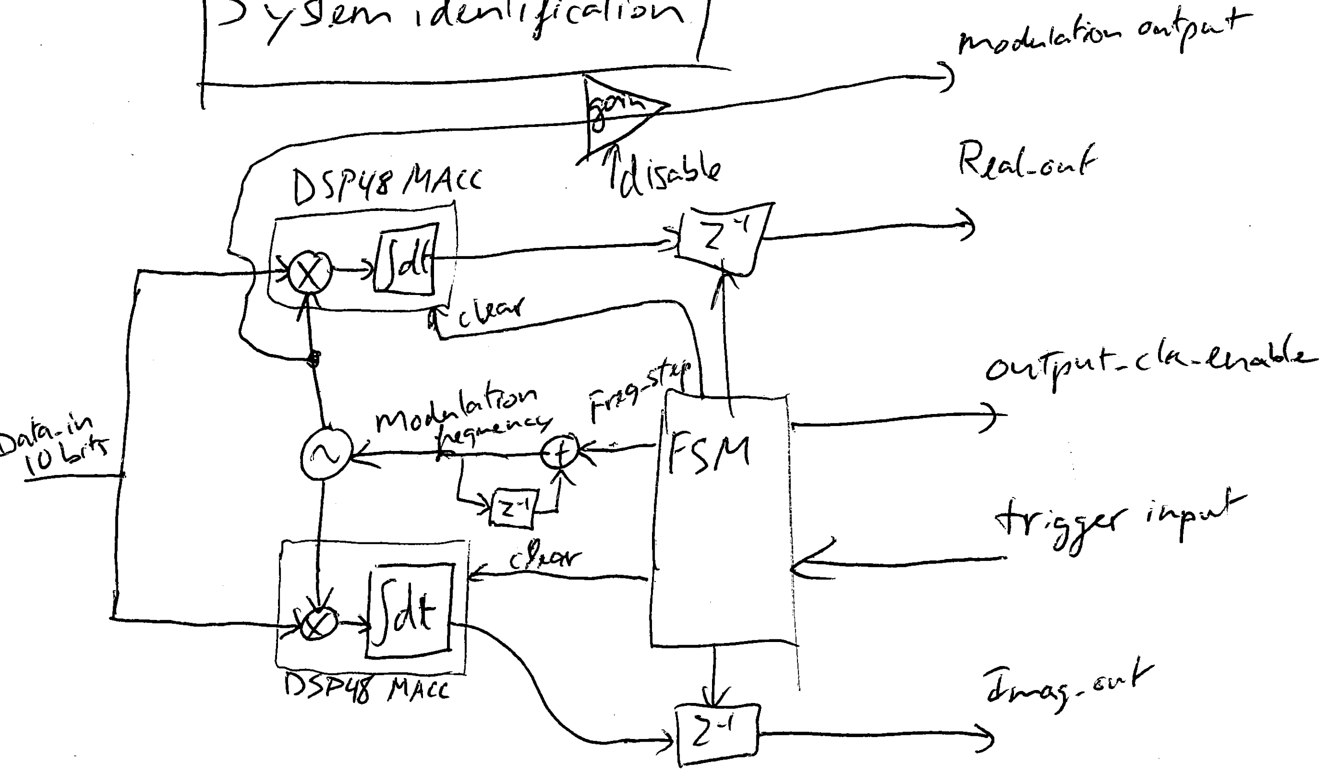
Only one address per burst: DDR2 controller (mcb3) probably takes care of addresses within a burst.

→ It seems that a DDR2 memory dump only stops because the output FIFO overflows because the PC stops reading data maybe? If the µC keeps emptying that FIFO, this might be very wasteful in terms of occupying the USB port.

READ address is only reset to 0 when reset is asserted, not when read is initiated. This reset is indeed initiated by DDR2logger.v at the ~~start of~~ reception of a read or write command.


z^{-N}
MUX
Registers if needed to achieve timing
Boxcar
100MS/s, 16 bits/sample
ADC 0
DDC
Freq
cumsum()
limit
PLL loop filters
Acquisition FLL
integrate & dump
+
limit
DAC 0
Syst. ident.
N bits with clk enable
Parallel to serial
to logger mux
16 bits with clk enable
Many more bits than 16, very slow
ADC 1
DDC
Freq
(. . .)
MUX

System identification



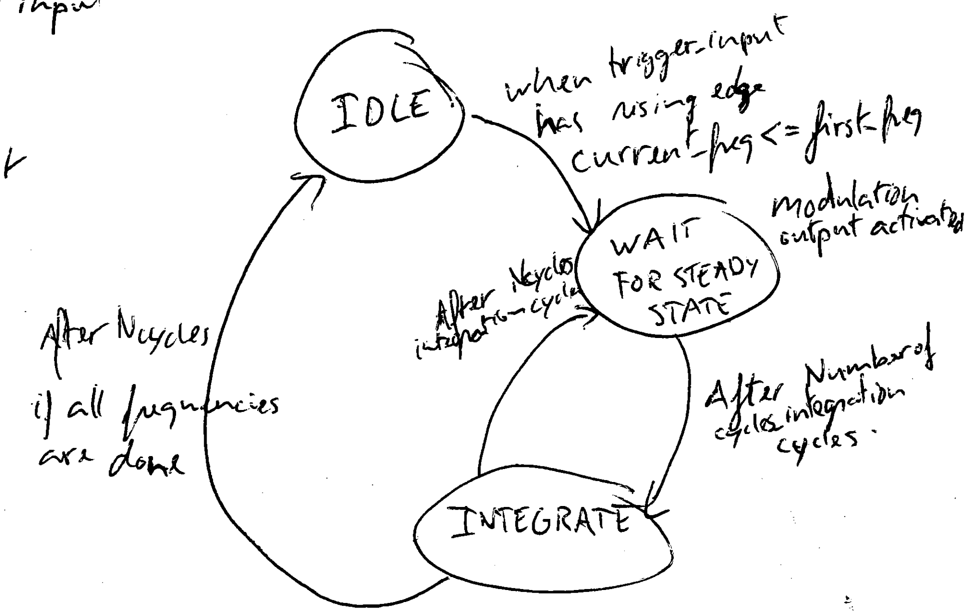
4 DSP48 slices

2 MACCs

1 for DDS (phase accum probably)

1 for output gain

FSM

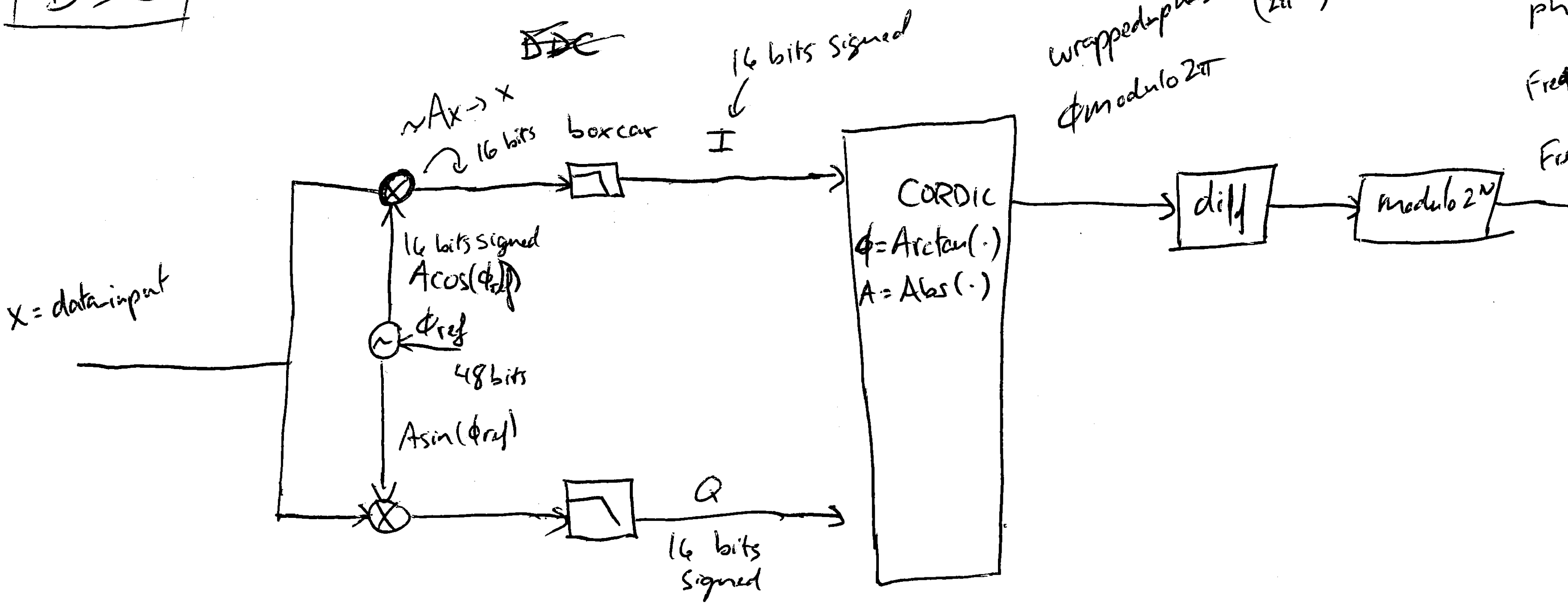


Expected value in the integrator is:

$N_{integration} \cdot \frac{A_{output}}{2}$ For a loop back system

$A_{output} = \text{output\_gain}$ (from 1 to 2^{15})

DDC



wrapped phase $= \left(\frac{\phi \cdot 2^N}{2\pi}\right) \bmod 2^N$

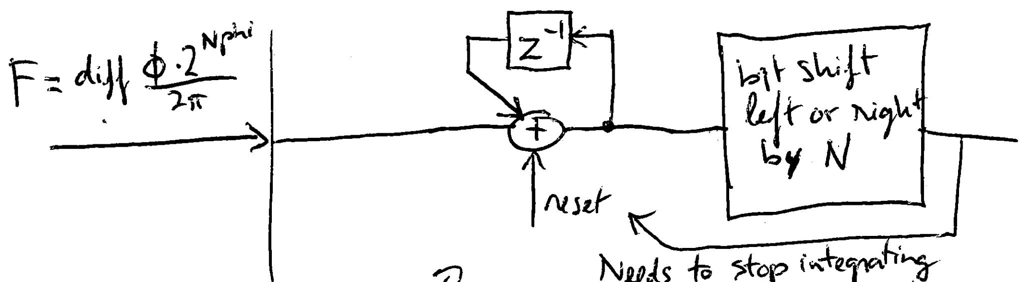
ϕ modulo 2π

$N = 10$ bits on phase

$Freq = diff\left(\left(\frac{2^N \phi}{2\pi}\right)\right) \bmod 2^N$

$Freq = \frac{2^N}{f_s} \cdot$ frequency

Acquisition FLL



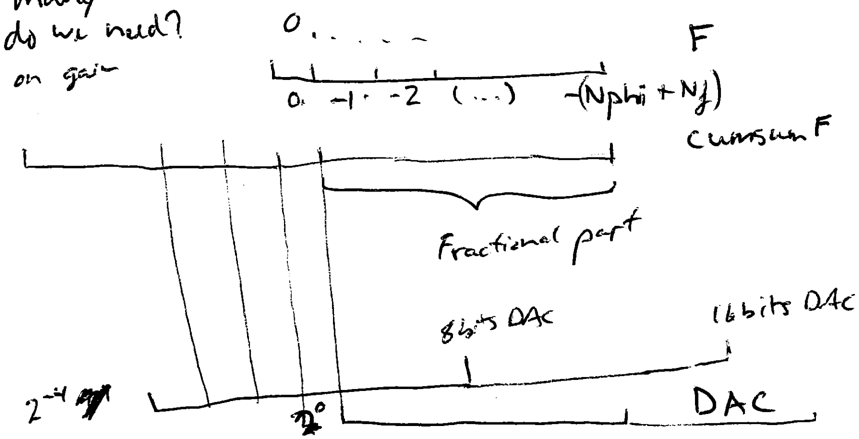
Needs to stop integrating if the DAC is railed.

integrator with reset and limits

Delay not really a problem as long the value does not wrap

The integrator output could easily update once every N clock cycles, with N a small number. This would allow the signal to propagate through the gain before checking for saturation. Would it be easier to just compare to limits? Possibly compute automatically from the gain value.

How many extra bits do we need? Depends on gain



Worst case is if DAC/vco gain is too high and thus we need to divide by a lot with K_i.

$$K_i = \frac{f_c}{2^{N_{phi}-N_{DAC}}} \cdot \frac{f_s}{f_{span}} \qquad f_s = 100\,\text{MHz}$$

Lowest K_i: low f_c, high f_{span}, high Nphi, low NDAC
1 Hz, 1 GHz, 10 bits, 8 bits

$K_i = \frac{1}{4} = 2^{-2}$

Highest K_i: High f_c, low f_{span}, low Nphi, High NDAC
1 kHz, 100 MHz, 8 bits, 16 bits

$K_i = 256e3 = 2^{18}$

Anything 2^8 and higher makes little sense as this means that every LSB flip on the phase will rail the DAC...

We have $Freq = \frac{\Delta\phi}{\Delta t} = \frac{Fs \cdot diff(\phi)}{2\pi}$ $\quad S_{Freq} = \left(\frac{Fs}{2\pi}\right)^2 4\sin^2\left(\frac{2\pi f}{2Fs}\right) S_\phi = \frac{Fs^2}{\pi^2}\sin^2\left(\frac{2\pi f}{2Fs}\right) S_\phi$

$Freq = \frac{Fs}{2\pi} \cdot diff\phi$

$Freq(f) = \frac{Fs}{2\pi} \cdot 2\sin\left(\frac{2\pi f}{2Fs}\right)\phi(f)$

$Freq(f) = \frac{Fs}{\pi}\sin\left(\frac{2\pi f}{2Fs}\right)\phi(f)$

$S_{Freq}(f) = \frac{Fs^2}{\pi^2}\sin^2\left(\frac{2\pi f}{2Fs}\right) S_\phi(f)$

S_f noise floor goes from 0 to $\frac{Fs^2}{\pi^2} S_\phi$, with $S_\phi = \left(\frac{1\,\mu rad}{\sqrt{Hz}}\right)^2$ for 8 fractional bits

Thus $S_f^{max} = \left(\frac{3\,Hz}{\sqrt{Hz}}\right)^2$

CORDIC $\sigma_\phi = \frac{\pi}{2\sqrt{2}\,2^{N_{iterations}}}$, for $N_{iterations} = 8$, $\sigma_\phi = 4{,}3$ mrad $\quad S_\phi = \frac{\sigma_\phi^2}{(fs/2)} = \frac{\pi^2}{4fs\,2^{2N_{iterations}}}$

How much frequency noise in 1 kHz BW? $\sigma_f|_{1kHz} = \sqrt{\int_0^{1kHz} S_f(f)\,df}$ $= \left(\int_0^{1kHz} \frac{Fs^2}{\pi^2}\sin^2\left(\frac{2\pi f}{2Fs}\right)\frac{\pi^2}{4fs\,2^{2N_{iter}}}\,df\right)^{1/2}$

$= \left(\frac{fs}{4 \cdot 2^{2N_{iter}}}\int_0^{1kHz}\sin^2\left(\frac{2\pi f}{2Fs}\right)df\right)^{1/2}$

$= \frac{fs}{2^{(N_{iter}+1)}}\left(\int_0^{1kHz}\sin^2\left(\frac{2\pi f}{2fs}\right)df\right)^{1/2}$

$= 1.2$ Hz.

Gain required for the acquisition PLL?

$F_{digital} = \frac{2\pi f_{inst}}{fs} \cdot 2^{N_{Phi}}$ $\left[\frac{Counts}{Hz}\right] = K_{Discriminator}$

DAC: $Gain = \frac{F_{span}}{2^{N_{DAC}}}$ $\left[\frac{Hz}{counts}\right] = K_{DAC}$ $\quad K_{DAC}K_{Discrim} = 2\pi\,2^{N_{phi}-N_{DAC}} \cdot \frac{F_{span}}{fs}$

$G(f) = K_{DAC} \cdot K_{DISCRIM} \cdot \frac{K_i}{2\pi f}$

$f_c = \frac{K_{DAC}K_{Discrim} \cdot K_i}{2\pi} \Rightarrow K_i = \frac{2\pi f_c}{K_{DAC}K_{Discrim}} = \frac{2\pi \cdot 1\,Hz}{2\pi\,2^{N_{phi}-N_{DAC}}}\frac{fs}{f_{span}} = \frac{f_c}{2^{N_{phi}-N_{DAC}}} \cdot \frac{fs}{f_{span}}$

We have added a low-pass filter with gain: $K_{filter} = 2^{N_f}$ with $N_f = 7$

$G(f) = K_{DAC} \cdot K_{Discrim} \cdot \frac{K_i}{2\pi f}$, $\quad K_{discrim} = \frac{2\pi f_{inst} \cdot 2^{N_{phi}+N_f}}{fs}$ $\left[\frac{Counts}{Hz}\right]$

Nichols Chart

X axis : phase in degrees of G : $\mathrm{Arg}(G(f))$

Y axis : gain in dB of G : $10\log_{10}(|G(f)|^2)$

We want to add a grid which indicates locii of constant ~~open~~ closed loop magnitude:

$H(f) = \frac{G(f)}{1+G(f)}$, solve: $\cancel{Ae^{j\phi} = \frac{Be^{j\phi}}{1+e^{j\phi}}}$ $Ae^{j\theta} = \frac{Be^{j\phi}}{1+Be^{j\phi}}$ for all A, θ

$H(1+G) = G$

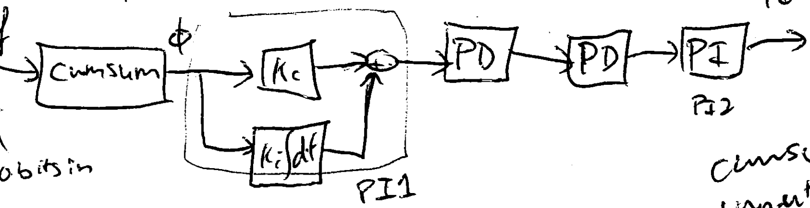
$H = G - HG = G(1-H)$

$\frac{H}{1-H} = G$: for all $H = Ae^{j\phi}$, lines of constant A.

PLL Loop filters

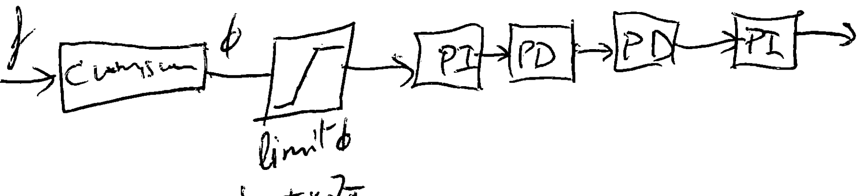
First design iteration

16 bits out

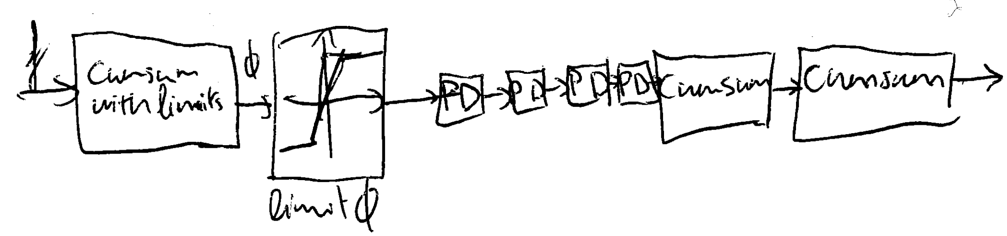


cumsum of the input ensures no quantization noise of DC

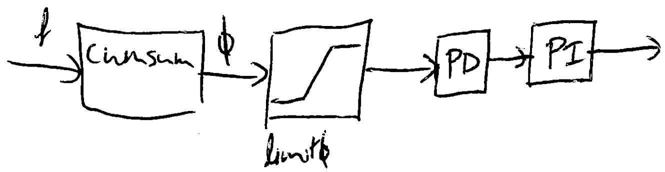
2nd design



3rd design



4th design



In normal operation (closed loop), the error signal ~~first~~ should be pretty small.

Same thing for the integral $\phi = \text{cumsum}(f)$, it should stay within $\pm 2\pi$ or better.

→ The 2nd design iteration has two operation modes: linear and saturated (non linear)

In linear mode, which is the desired mode, everything works as expected.

In saturated mode, the error signal has the correct sign but is clamped at some value. The ~~loop~~ output will ramp as a parabola in the correct direction until the error de-saturates.

It basically does the same thing as a PFD (Phase-frequency detector) except that it keeps track of how many cycles were lost.

→ Having the output stage as an integrator makes it easy to add anti-windup and hold capabilities.

→ Can we do the whole chain in exact arithmetic? No quantization noise anywhere except at the end? Probably not unless we stick to "nice" values for all the filters coefficients.

whats the transfer function that we want anyway?

$$K_\phi = \frac{2^{N_{phi}}}{2\pi} \left[\frac{counts}{rad}\right]$$

$N_{phi} = 10$ bits $\Rightarrow K_\phi = 162 \left[\frac{counts}{rad}\right]$

So $K_\phi K_{DAC} K_{VCO} = 988 \frac{kHz}{rad}$

$$K_{DAC} = \frac{V_{max}}{2^{N_{DAC}}} \left[\frac{V}{count}\right]$$

$N_{DAC} = 16$ bits.

What is F_{max}?

Let's assume $F_{max} = 400MHz$

$$K_{VCO} = \frac{F_{max}}{V_{max}} \left[\frac{Hz}{V}\right]$$

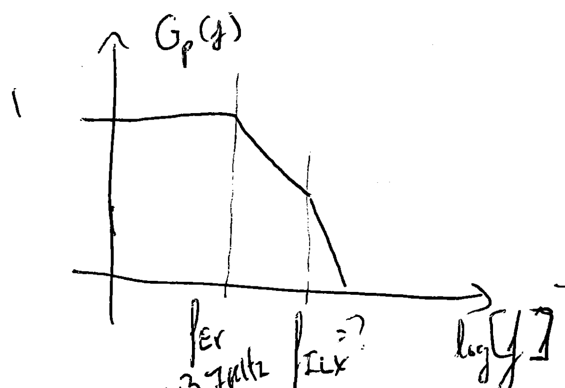
$$K_{DAC} K_{VCO} = \frac{F_{max}}{2^{N_{DAC}}} \left[\frac{Hz}{count}\right]$$

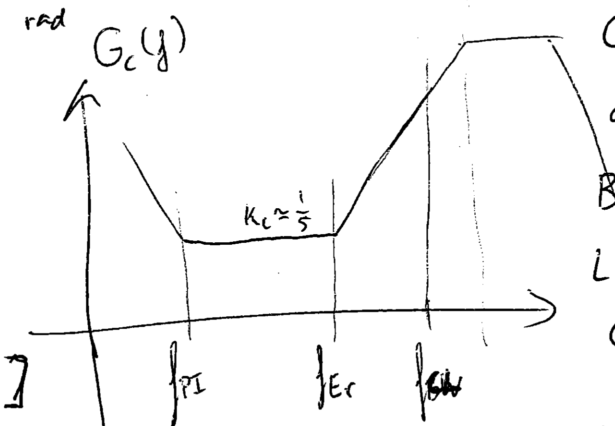
$K_{DAC} K_{VCO} = 6.1 \frac{kHz}{count}$

Loop gain $G(f) = \underbrace{\frac{1}{f}}_{\text{Freq to phase conversion}} \cdot \underbrace{K_\phi \cdot K_{DAC} K_{VCO}}_{988 \frac{kHz}{rad}} \cdot G_c(f) \cdot G_p(f)$

$G_c(f)$: Controller transfer function

$G_p(f)$: Actuator transfer function (normalized to DC gain = 1)





$G_c(f) G_p(f)$ needs to be proportional to $f \propto K_c$ around the cut-off frequency.

Below the loop BW, we can have ~~an~~ integrator

Let's assume that $G_c(f)$ is PI + PD

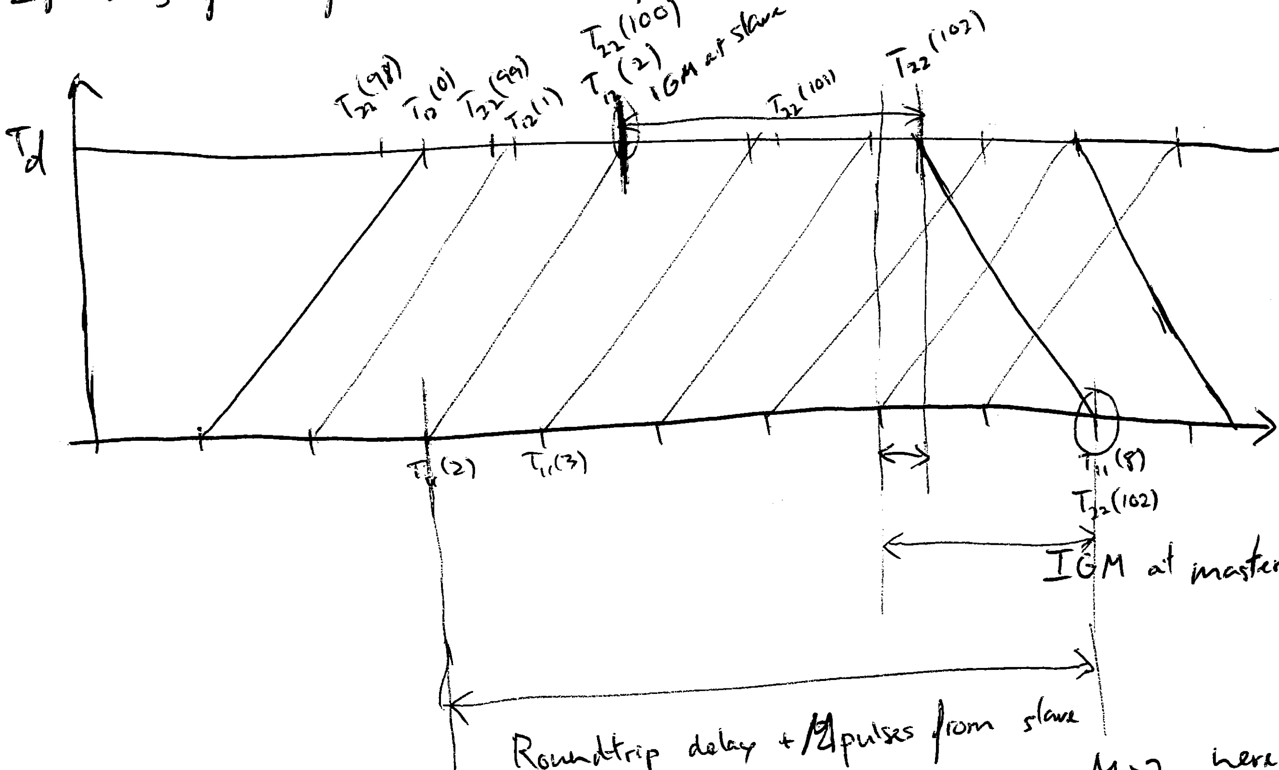
$G_c(f) = \left(K_c + \frac{K_i K_c}{2\pi f}\right) G_{lead\,lag}(f)$, where $G_{lead\,lag}(f)$ cancels the Er roll-off

For a closed loop BW of 200kHz, we want $K_c \simeq \frac{1}{5}$

$f_{CL} \simeq 200kHz$

f_{lag} must be well past the cut-off

If T_d is fixed for the duration of a pair of IGM:



Roundtrip delay + M pulses from slave comb $= N \cdot T_{r1}$

M=2 here

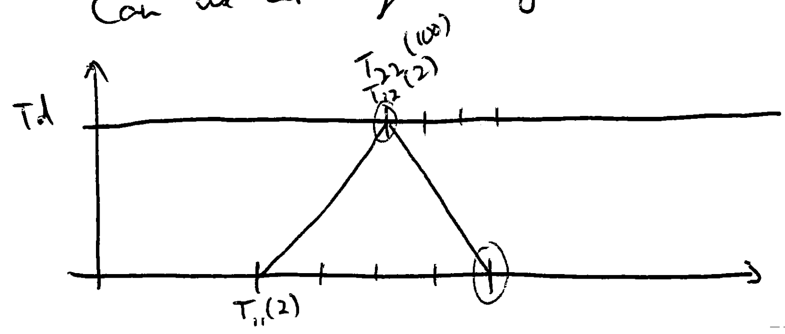
$2T_d + M T_{r2} = N \cdot T_{r1}$

T_{r2} can be known (on average) through delay between two IGMs at master site.

T_{r1} is assumed to be known perfectly.

How well can T_{r2} be known? Depends on how much it changes between two IGMs.

Can we arrange things so that $M=0$?



This implies $M T_{r1} = 2T_d$.
Can we expressely add a delay to T_d to make this equation match? Removes T_{r2} from the measurement of T_d. Same thing as looping back the master's pulses.
Actuator only needs $\sim \frac{2T_d}{M}$ range.

\\Avocet\Public\Landesk

Assume that T_d is slowly varying, thus constant over 2 adjacent icms:

(1) $T_{11}(N) = T_d + T_{22}(M)$

(2) $T_{22}(P) = T_d + T_{11}(Q)$

$T_{11}(\cdot)$ is known

$T_{22}(\cdot)$ is unknown, but $T_{11}(Q) = T_{11}(N) + (Q-N)T_{r1}$

T_d is unknown ~~that~~ $T_{22}(P) = T_{22}(M) + (P-M)T_{r2}$

} How well does this require knowledge of T_{r1}, T_{r2}?

(1) $T_{11}(N) = T_d + T_{22}(M)$

(2) $T_{22}(M) + (P-M)T_{r2} = T_d + T_{11}(N) + (Q-N)T_{r1}$

Solve for T_d: (1) + (2):

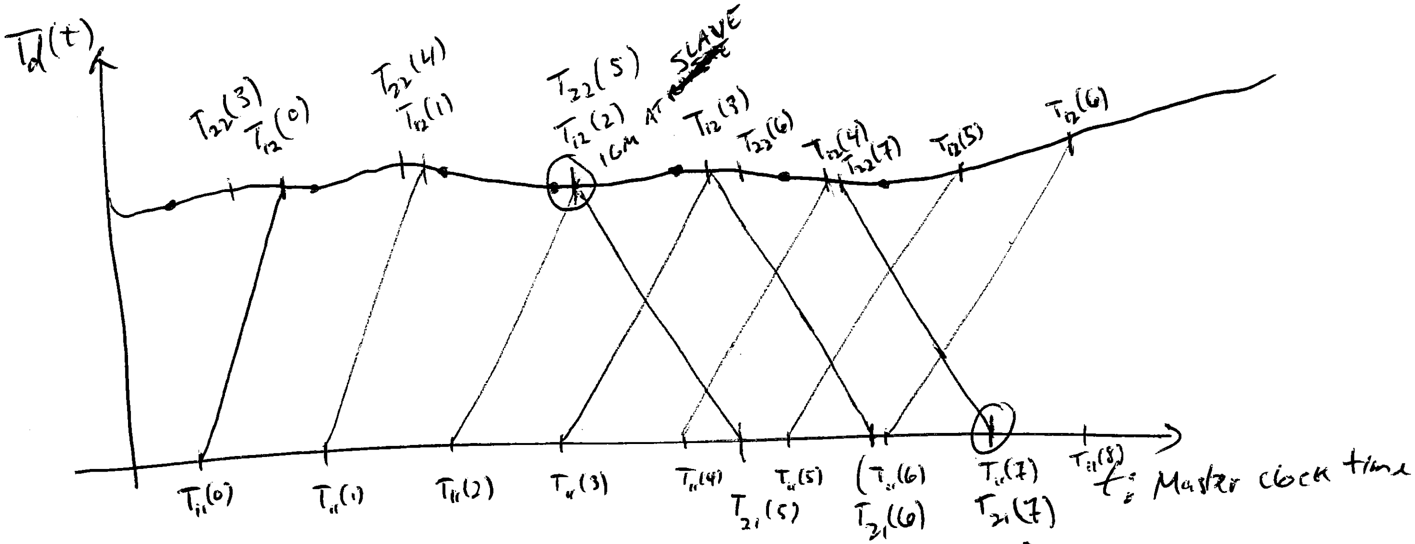
$\cancel{T_{11}}(N) + \cancel{T_{22}}(M) + (P-M)T_{r2} = 2T_d + \cancel{T_{11}}(N) + \cancel{T_{22}}(M) + (Q-N)T_{r1}$

$(P-M)T_{r2} - (Q-M)T_{r1} = 2T_d$ } Once again, requires T_{r1} and T_{r2}

Solve for T_{22}: (1) − (2):

$T_{11}(N) - T_{22}(M) - (P-M)T_{r2} = \cancel{T_d} + T_{22}(M) - \cancel{T_d} - T_{11}(N) + (Q-N)T_{r1}$

$2T_{11}(N) - (P-M)T_{r2} - (Q-N)T_{r1} = 2T_{22}(M)$ } T_{r1} and T_{r2}



IGM AT MASTER SITE

The pulse numbers are a coincidence.

Two equations to solve for two unknowns:

(1) $T_{11}(N) = T_d(T_{11}(N)) + T_{22}(M)$

(2) $T_{22}(P) = T_d(T_{22}(P)) + T_{11}(Q)$

$T_d(t)$ is slowly varying so we can replace ~~our~~ $T_{22}(P)$ by coarse estimate of $T_{22}(P)$: $T_{22}(P) \simeq T^*_{22}(P) = T_{21}(P) - T_d(T_{21}(P))$

$T_{21}(P) - T_d^*$

(2) $T_{22}(P) = T_d(T_{21}(P) - T_d^*) + T_{11}(Q)$

$T_{11}(Q) = T_{11}(N) + (Q-N)T_{r1}$

(2) $T_{22}(P) = T_d^* + T_{11}(N) + (Q-N)T_{r1}$

(2) and (1): $T_{22}(P) = T_d^* + T_d^* + T_{22}(M) + (Q-N)T_{r1}$

Interpolation: $T_{22}(M) = T_{22}(P) + (M-P)T_{r2}$

Assume that only one sideband affects the phase per each frequency.

$$\theta = \mathrm{Arg}\left\{ H_{\mathrm{sign}}(f) + \frac{2\pi f_c A_{\mathrm{timing}}}{2} \exp(j(\phi_0+\pi))\, H_{\mathrm{sign}}(f - f_{\mathrm{timing}}) \right\}$$

$$= \mathrm{Arg}\left\{ H_{\mathrm{sign}}(f) + \frac{2\pi f_c A_{\mathrm{timing}}}{2} H_{\mathrm{sign}}(f - f_{\mathrm{timing}}) \left(\cos(\phi_0+\pi) + j\sin(\phi_0+\pi)\right) \right\}$$

small deviations from 0 phase.

$$= \cancel{\mathrm{Arg}}\ \mathrm{Arctan}\left\{ \frac{\mathrm{Im}\{\}}{\mathrm{Re}\{\}} \right\} \simeq \frac{\mathrm{Im}\{\}}{\mathrm{Re}\{\}} = \frac{H_{\mathrm{sign}}(f - f_{\mathrm{timing}})}{H_{\mathrm{sign}}(f)} \cdot \frac{2\pi f_c A_{\mathrm{timing}}}{2} \cdot \sin(\phi_0+\pi)$$

choose worst case ϕ_0: $\sin(\phi_0+\pi) = 1$

$$= \frac{2\pi f_c A_{\mathrm{timing}}}{2} \cdot \frac{H_{\mathrm{sign}}(f - f_{\mathrm{timing}})}{H_{\mathrm{sign}}(f)}$$

Thus the shape of the phase error is given by the shape of the spectrum

$$F\left\{ A(t)\exp\left[j\left(2\pi f_c\left(t + A_{timing}\cos\left(2\pi f_{timing}t + \phi_0\right)\right)\right)\right]\right\}$$

$$j\exp(jx)$$

$$= \exp(j\pi)\exp(jx)$$

$$= \exp j(x+\pi)$$

$$= F\left\{ A(t)\exp(j2\pi f_c t)\exp\left(j2\pi f_c A_{timing}\cos(\ldots)\right)\right\}$$

$$= F\{A(t)\} * F\{\exp(j2\pi f_c t)\} * F\left\{\exp\left(j2\pi f_c A_{timing}\cos(\ldots)\right)\right\}$$

$$= H_{ign}(f) * F\left\{\exp\left(j2\pi f_c A_{timing}\cos(\ldots)\right)\right\}$$

$$\simeq H_{ign}(f) * F\left\{1 + j2\pi f_c A_{timing}\cos\left(2\pi f_{timing}t + \phi_0\right)\right\}$$

$$= H_{ign}(f) * \left[\delta(f) + \frac{j2\pi f_c A_{timing}}{2} F\left\{\exp\left(j\left(2\pi f_{timing}t + \phi_0\right)\right) + \exp\left(-j2\pi f_{timing}t - j\phi_0\right)\right\}\right]$$

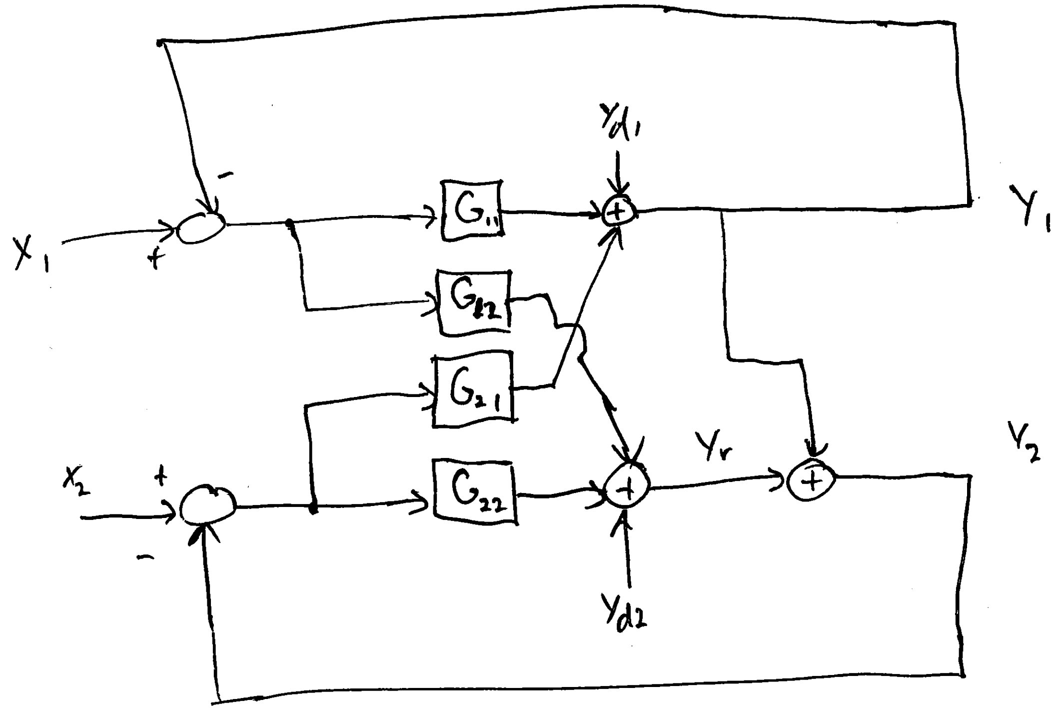
$$= H_{ign}(f) * \left[\delta(f) + j\pi f_c A_{timing}\left(\exp(j\phi_0)\,\delta(f - f_{timing}) + \exp(-j\phi_0)\,\delta(f + f_{timing})\right)\right]$$

$$= H_{ign}(f) + j\pi f_c A_{timing}\exp(j\phi_0)\,H_{ign}(f - f_{timing}) + j\pi f_c A_{timing}\exp(-j\phi_0)\,H_{ign}(f + f_{timing})$$

$$= H_{ign}(f) + \frac{2\pi f_c A_{timing}}{2}\exp\left(j(\phi_0 + \pi)\right) H_{ign}(f - f_{timing}) + \frac{2\pi f_c A_{timing}}{2}\exp\left(j(\pi - \phi_0)\right) H_{ign}(f + f_{timing})$$

Assume that $\operatorname{Arg} H_{ign}(f) = 0$ for all f.

Next: Include non-orthogonal actuators:



$$\frac{1}{sT+1} = G$$

$$H = \frac{G}{1+G}$$

$$= \frac{1}{\frac{1}{G}+1}$$

$$= \frac{1}{sT+1+1}$$

$$= \frac{0,5}{sT/_2 + 1}$$

$$Y_r = \frac{1}{1+G_2} Y_{d2} + \frac{G_2}{1+G_2}\left[(\alpha-1)Y_1 + \alpha Y_{m1} + X_2 - Y_{m2}\right]$$

$$= \frac{1}{1+G_2} Y_{d2} + \frac{G_2}{1+G_2}\left[(\alpha-1)\left(\frac{1}{1+G_1}Y_{d1} + \frac{G_1}{1+G_1}X_1 - Y_{m1}\frac{G_1}{1+G_1}\right) + \alpha Y_{m1} + X_2 - Y_{m2}\right]$$

We can drop X_1 and X_2 since they usually are 0.

$$Y_r = \underbrace{\frac{1}{1+G_2}}_{\substack{\cancel{\text{Rejected}}\\ \text{High-passed}}} Y_{d2} + \underbrace{\frac{G_2}{1+G_2}}_{\text{low-pass}}\left[(\alpha-1)\underbrace{\frac{1}{1+G_1}}_{\text{High-passed.}} Y_{d1} + \left(\alpha - (\alpha-1)\underbrace{\frac{G_1}{1+G_1}}_{\text{low-passed}}\right)Y_{m1} - Y_{m2}\right]$$

~~So~~ if

$$Y_r = \frac{1}{1+G_2} Y_{d2} + \frac{G_2}{1+G_2}\left[\overbrace{(\alpha-1)\cdot\frac{1}{1+G_1}\cdot Y_{d1}}^{\cancel{\text{(CEO)}}\text{ First loop (CEO loop) residuals}} + \alpha Y_{m1} - (\alpha-1)\frac{G_1}{1+G_1}Y_{m1} - Y_{m2}\right]$$

So that means that if $\alpha = 1$ (we decouple the two loops), instead of putting all the residuals of the first loop onto the rep rate, we leave them untouched ~~on~~ on the optical phase.

→ What happens if the first loop has a "servo bump": that means that $\frac{1}{1+G_2}$ and $\frac{G_2}{1+G_2}$ are resonating, ie G_2 is close to -1, yielding something like $\frac{1}{1+\text{"close to 1"}} \simeq \frac{1}{\text{small}} = \text{big}$. Same thing for $\frac{G_2}{1+G_2} \simeq \frac{-\text{close to }1}{1-\text{close to }1} \simeq \frac{\text{close to }1}{\text{small}} = \text{big}$.

In this case, the residuals of the CEO loop are amplified by <u>both</u> servo bumps, unless $\alpha = 1$.

The nice thing about putting $\alpha = 1$, is that the phase noise should now be more uniform across all the comb modes, instead of hitting a minimum at the optical lock point and increasing from there.

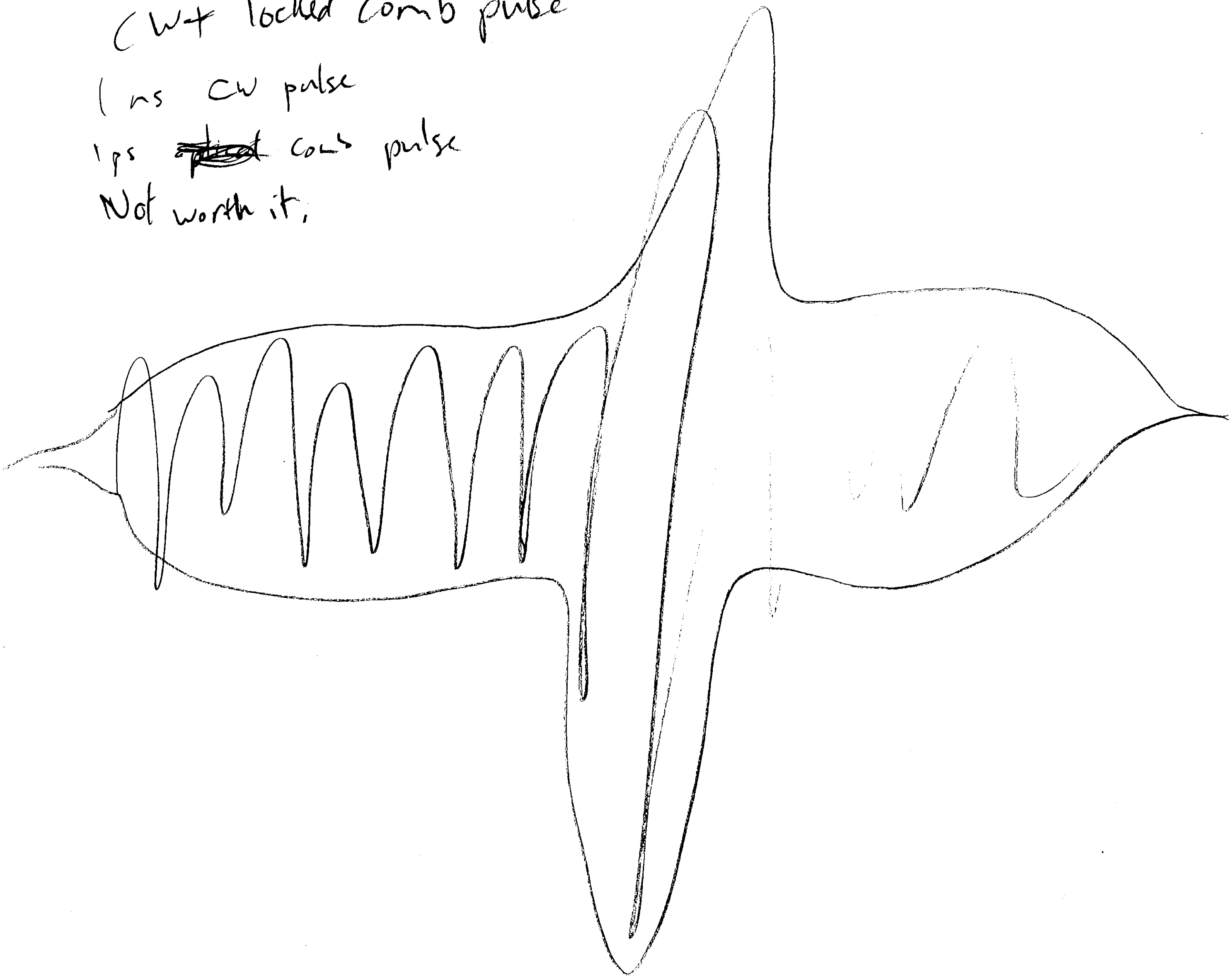
That is of course in addition to minimizing the timing jitter...

CW + locked comb pulse

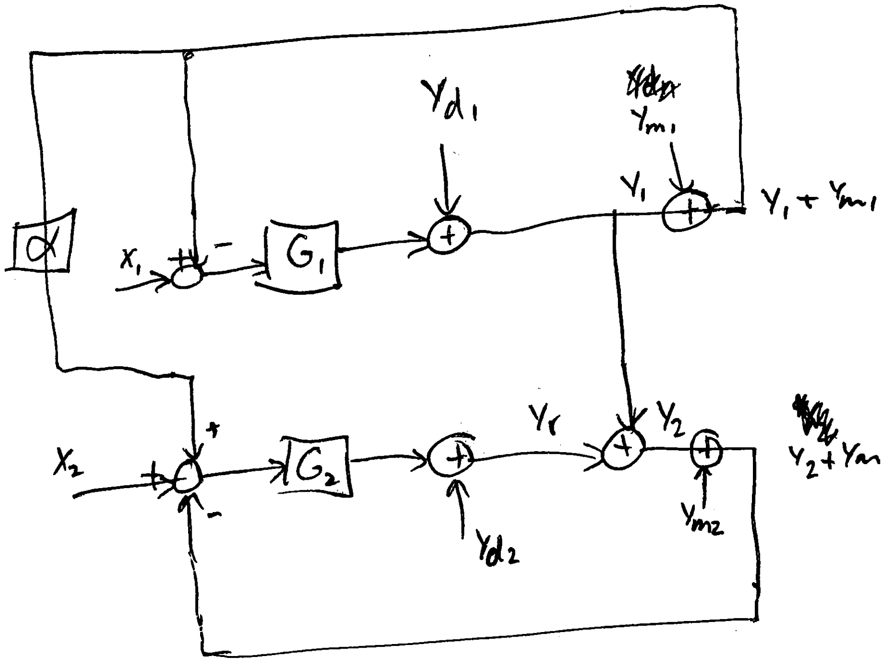
1 ns CW pulse

1ps ~~[illegible]~~ comb pulse

Not worth it.



How different is it if we explicitly remove the coupling:



$Y_{d_{11}}$ would be Y_1 measurement noise

Y_{d_1} and Y_{d_2} are the actual disturbances on the phases.

Y_{m_1} and Y_{m_2} are the measurement noises.

$$Y_1 = Y_{d_1} + G_1\left(X_1 - (Y_1 + Y_{m_1})\right)$$

$$Y_1(1 + G_1) = Y_{d_1} + G_1 X_1 - G_1 Y_{m_1}$$

$$Y_1 = \frac{1}{1+G_1} Y_{d_1} + \frac{G_1}{1+G_1} X_1 - \frac{G_1}{G_1+1} Y_{m_1}$$

$$Y_1 = \underbrace{\frac{1}{1+G_1} Y_{d_1}}_{\text{high-pass on disturbances (because } G \propto \frac{1}{j})} + \underbrace{\frac{G_1}{1+G_1} X_1}_{\text{low pass w/ unit gain on phaseref.}} - \underbrace{Y_{m_1} \frac{G_1}{1+G_1}}_{\text{measurement noise is added with low-pass TF.}}$$

So Y_r is completely decoupled from whatever happens in the second loop.

$$Y_2 = Y_1 + Y_{d_2} + G_2\left(X_2 + \alpha Y_1 + \alpha Y_{m_1} - Y_2 - Y_{m_2}\right)$$

$$Y_2(1 + G_2) = Y_1(1 + \alpha G_2) + Y_{d_2} + G_2 X_2 + G_2 \alpha Y_{m_1} - G_2 Y_{m_2}$$

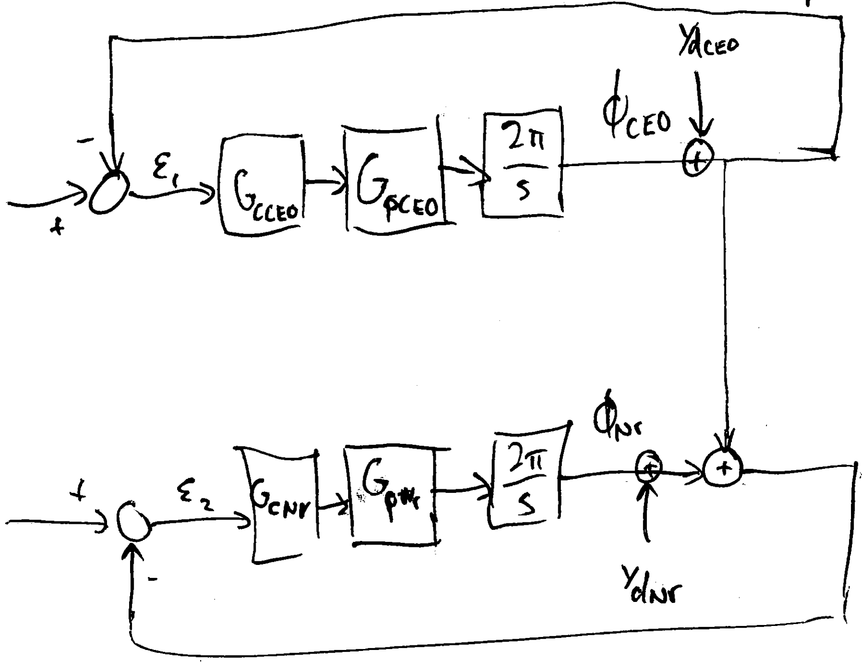
$$Y_2 = \frac{1+\alpha G_2}{1+G_2} Y_1 + \frac{1}{1+G_2} Y_{d_2} + \frac{G_2}{G_2+1} X_2 + \frac{\alpha G_2}{1+G_2} Y_{m_1} - \frac{G_2}{1+G_2} Y_{m_2}$$

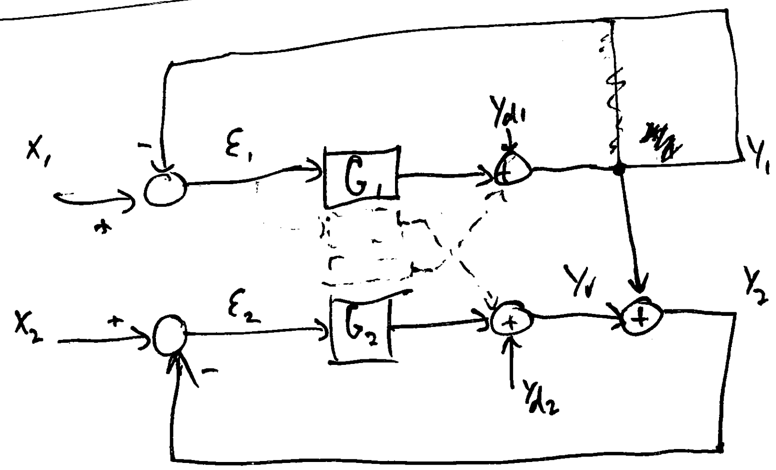
We care (sometimes) more about $Y_r = Y_2 - Y_1 = \frac{1+\alpha G_2 - 1 - G_2}{1+G_2} Y_1 + \frac{1}{1+G_2} Y_{d_2} + \frac{G_2}{1+G_2} X_2 + \frac{\alpha G_2}{1+G_2} Y_{m_1} - \frac{G_2}{1+G_2} Y_{m_2}$

$$Y_r = (\alpha - 1)\frac{G_2}{1+G_2} Y_1 + \frac{1}{1+G_2} Y_{d_2} + \frac{G_2}{1+G_2} X_2 + \frac{\alpha G_2}{1+G_2} Y_{m_1} - \frac{G_2}{1+G_2} Y_{m_2}$$

- Temp control, set point settable from PC
  Needs ~2A current output, thermistor input.

$\phi_{opt} = \phi_{CEO} + \phi_{NR}$ $\phi_{NR} = \frac{2\pi}{s} \cdot N \cdot f_r$





We want the TFs:

$\frac{Y_1}{X_1}$, $\frac{Y_2}{X_2}$, $\frac{Y_1}{Y_{d1}}$, $\frac{Y_2}{Y_{d2}}$

$\frac{Y_1}{X_2}$, $\frac{Y_2}{X_1}$, $\frac{Y_1}{Y_{d2}}$, $\frac{Y_2}{Y_{d1}}$

We are also interested in $Y_r = Y_2 - Y_1$ (Represents the phase due to timing jitter only)

$Y_1 = Y_{d1} + G_1\varepsilon_1 = Y_{d1} + G_1(X_1 - Y_1)$

$Y_2 = Y_{d2} + Y_1 + G_2\varepsilon_2 = Y_{d2} + Y_1 + G_2(X_2 - Y_2)$

$Y_1(1 + G_1) = Y_{d1} + G_1 X_1$

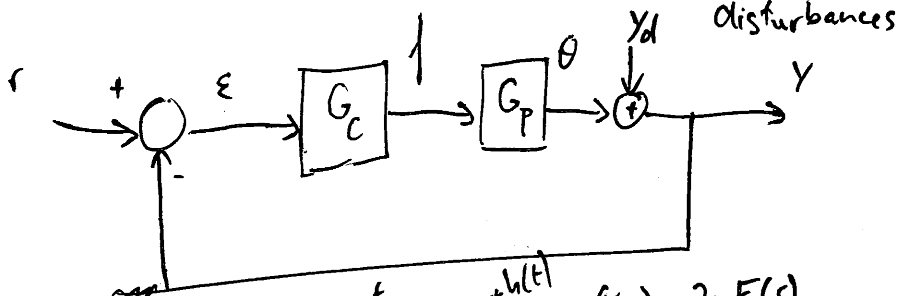
$Y_1 = \frac{Y_{d1}}{1+G_1} + \frac{G_1}{1+G_1} X_1$

$Y_2(1 + G_2) = Y_{d2} + Y_1 + G_2 X_2$

$Y_2 = \frac{Y_{d2}}{1+G_2} + \frac{Y_1}{1+G_2} + \frac{G_2}{1+G_2} X_2$

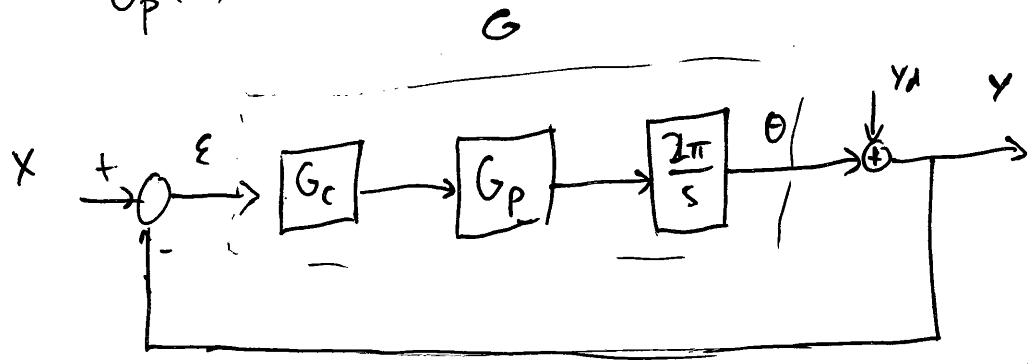
So Y_1 shows up as perturbations on the second loop and no TF is modified in any way (because there are no feedback paths. What about $Y_r = Y_2 - Y_1$?

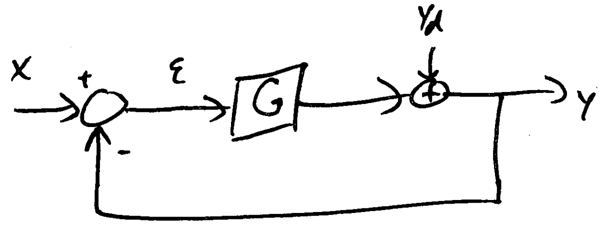
$Y_r = \frac{Y_{d2}}{1+G_2} + \frac{G_2}{1+G_2} X_2 + \left(\frac{1}{1+G_2} - 1\right) Y_1$



$$\theta = 2\pi \int_{-\infty}^{t} f(\tau)\, d\tau \;*\; h(t) \Rightarrow \theta(s) = \frac{2\pi F(s)}{s}$$

$G_p(s) =$





G: Open-loop transfer function

H: Closed loop transfer function

H_d: Closed loop disturbance rejection transfer function

$$H = \frac{Y(s)}{X(s)} \qquad H_d = \frac{Y(s)}{Y_d(s)}$$

$$Y = G\varepsilon = G(X - Y)$$

$$Y(1+G) = GX$$

$$\frac{Y}{X} = \frac{G}{1+G} = H$$

$$Y = Y_d + -GY$$

$$Y(1+G) = Y_d$$

$$\frac{Y}{Y_d} = \frac{1}{1+G}$$

In our case, G_p is hopefully a first-order lowpass $\frac{G_0}{1+\tau s}$, $\tau \approx 50\mu s$ or so (Er lifetime)

We always have an integrator due to phase being the integral of inst. freq.

$$H(1+G) = G$$

$$H + HG = G$$

$$H = G(1-H)$$

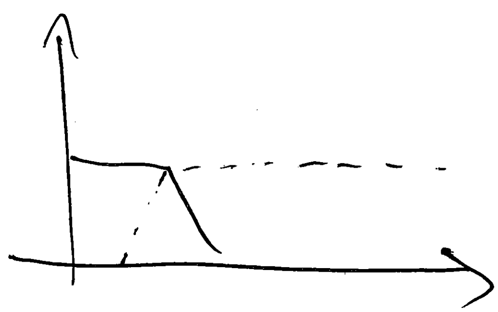
$$\frac{H}{1-H} = G$$

What if G_c contains ~~an integrator~~ a delayed P term plus another P term with less delay (can approximate to 0 delay for now).

$$G_c = K_1 e^{-jsT_1} + K_2 \qquad K_2 \text{ higher than } K_1.$$

The first term uses a ~~phx~~ high dynamic range phase detector for capture, while the second term uses a faster phase detector with only $\pm\pi$ of approximate range.

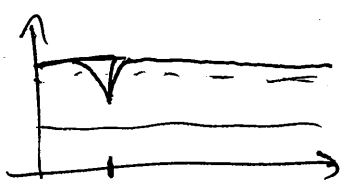
The important points are where the ~~clos~~ open-loop transfer function reaches 1 of gain, and also where the relative phase between the two P terms reaches π, where they will subtract instead of summing. If their amplitude is different, the larger one will still dominate. Can we make the faster be high-pass only? Some sort of cross-over design like for a loudspeaker?



Make the crossover before the delay accumulates too much phase.

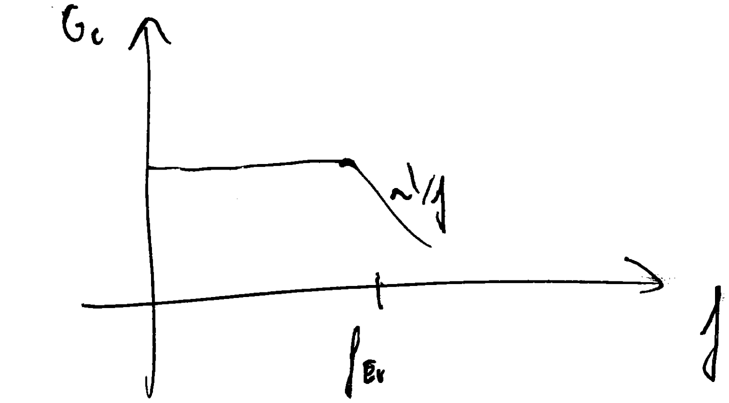
$$G_c = \frac{K_1 e^{-jsT_1}}{1 + sT_c} + \frac{sT_c}{1+sT_c} K_2$$

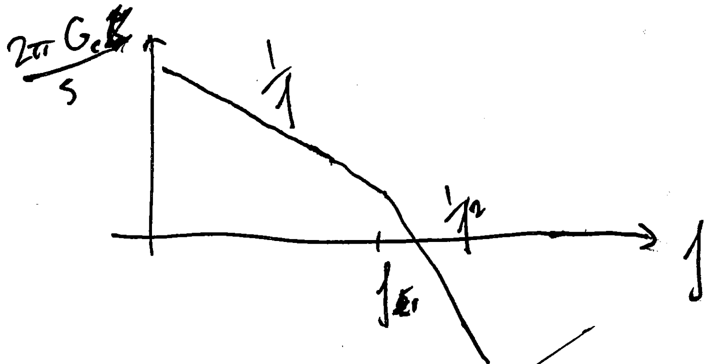
Ultimately we want the closed-loop BW to be higher than what we could get with just the slow term, which implies that we want $K_2 > K_1$. Is the cross-over design still workable? Do we even have to care to put filters in?



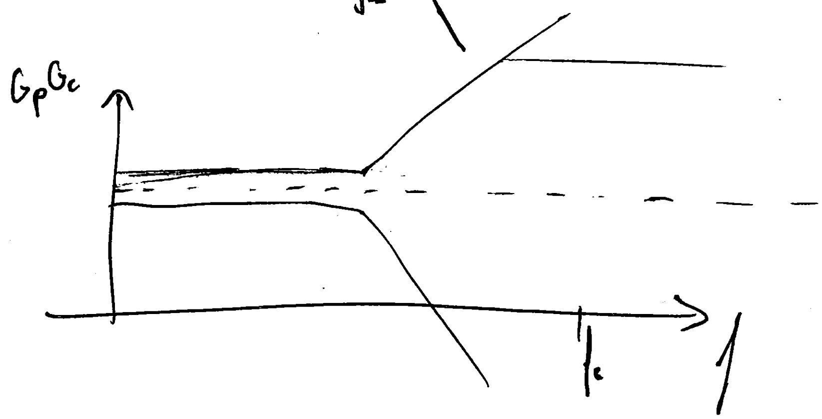
$2\pi f T_1 = \pi$ $\quad f$

This will make the total phase oscillate with $\approx \frac{K_2}{K_1}$ radians of amplitude.
Minimum at $2\pi f T_1 = \pi \Rightarrow f T_1 = \frac{1}{2} \Rightarrow f = \frac{1}{2T_1}$





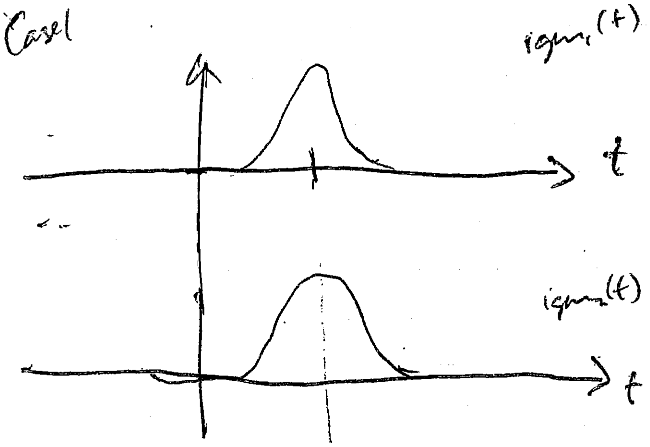
We want the total open-loop transfer function to be an integrator (around the critical frequency)



f_c is where total loop gain is equal to ~1. We want as much phase margin as possible to have low servo bump.

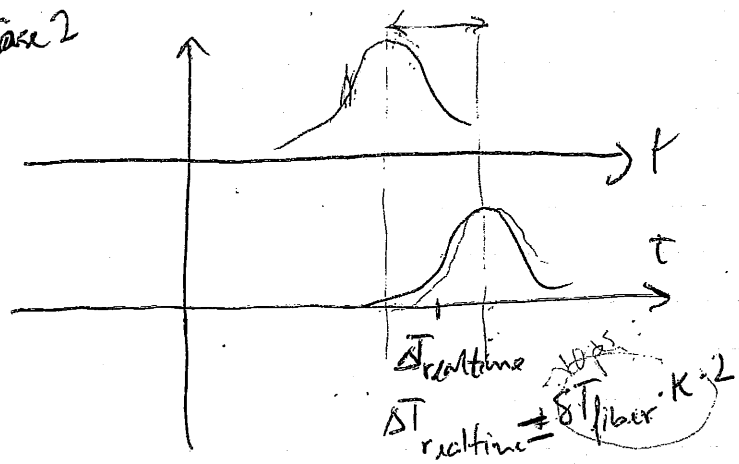
$$\left(1 + jf/f_{p2}\right)\frac{1}{\left(1 + jf/f_p\right)}$$

Case 1



Link shorted, common timebase, paths adjusted so that IGMs line up.

Case 2



Add link delay, common timebase.

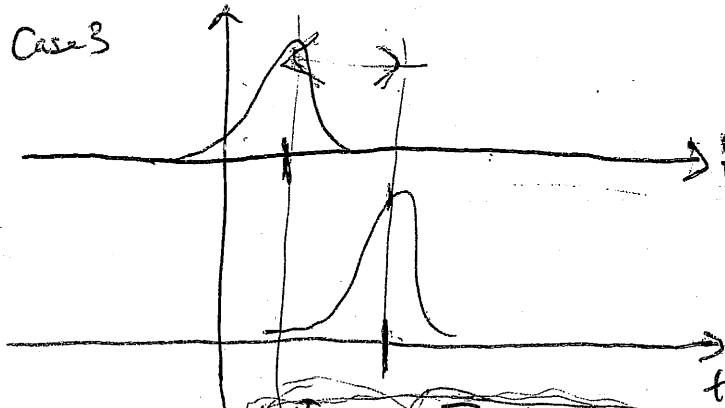
Assume link delay

$\Delta T_{link} = \Delta T_{realtime} \frac{\Delta f_r}{f_r}$

$\Delta T_{link} = \frac{\Delta T_{realtime}}{\text{Compression factor}}$

$K =$ Compression factor $= \frac{f_r}{\Delta f_r}$.

Case 3



Has link delay, two separate timebases.

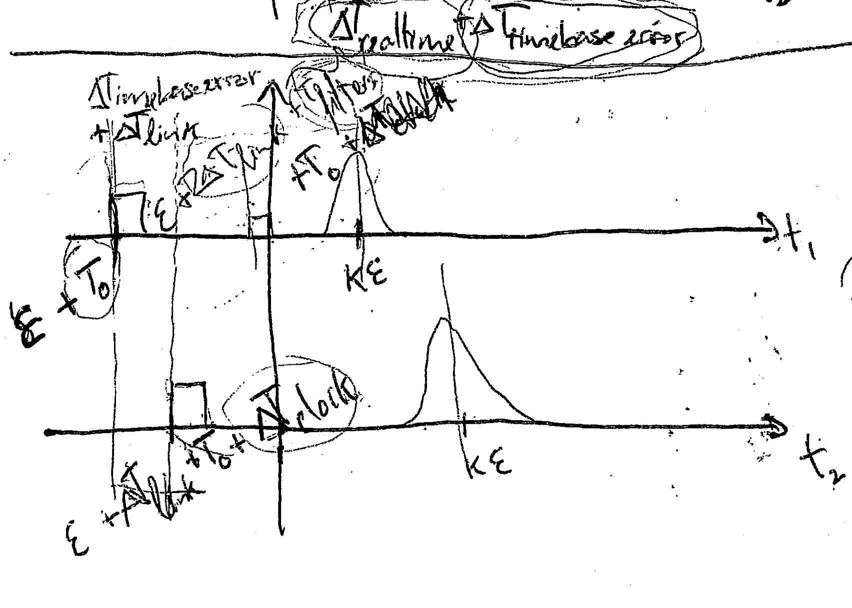
We make an error in link delay: $\Delta T_{link} - \Delta T_{link,measured}$

$= \frac{\Delta T_{timebase\ error}}{\text{Compression factor}}$

Assume $\Delta F_r = 2kHz$, $F_r = 200MHz$

$K = \frac{2e8}{2e3} = 1e5$

$\Delta T_{timebase} = 10ns$, $\Delta T_{link\ error} = \frac{10ns}{1e5} = 10fs$.



$t_2 = t_1 + \Delta T_{clock}$

How do we sync the timebases?

Simple fact: When we get an IGM at one site, we know that the pulses overlap on that detector to <100 fs. The time offset between the pulses at the remote site is ΔT_{link}.

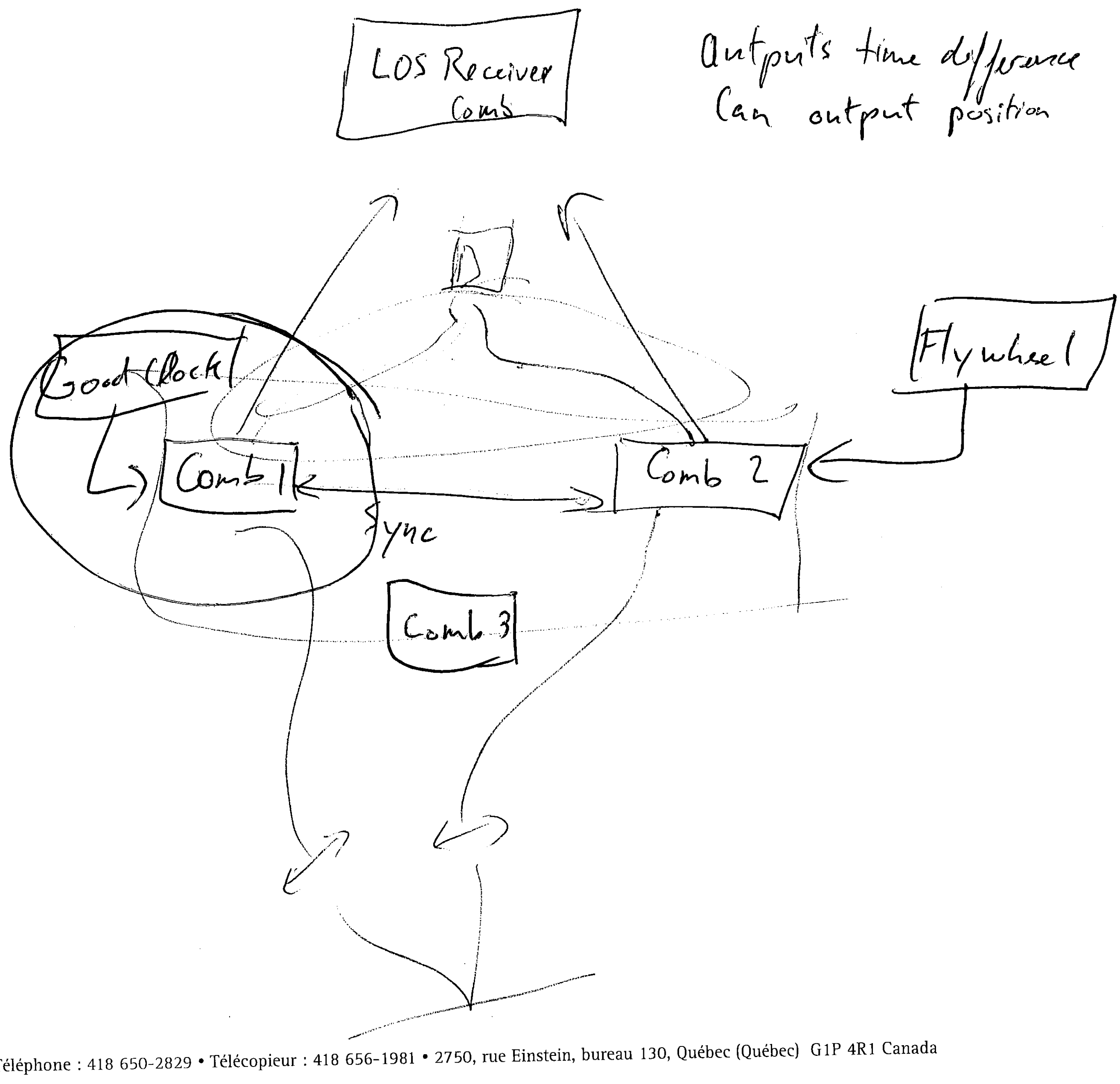
Link: 100km   300m → 1µs

100km → 1µs · 300 = 300µs

Plane at ~~1000kph~~ 300 $\frac{m}{s}$ (almost Mach 1)

$300\frac{m}{s} \cdot 300\mu s = 3e2 \cdot 3e2 \cdot 1e{-6} = 10 \cdot 1e4 \cdot 1e{-6}$

$= 1e5 \cdot 1e{-6} = 0{,}1m$



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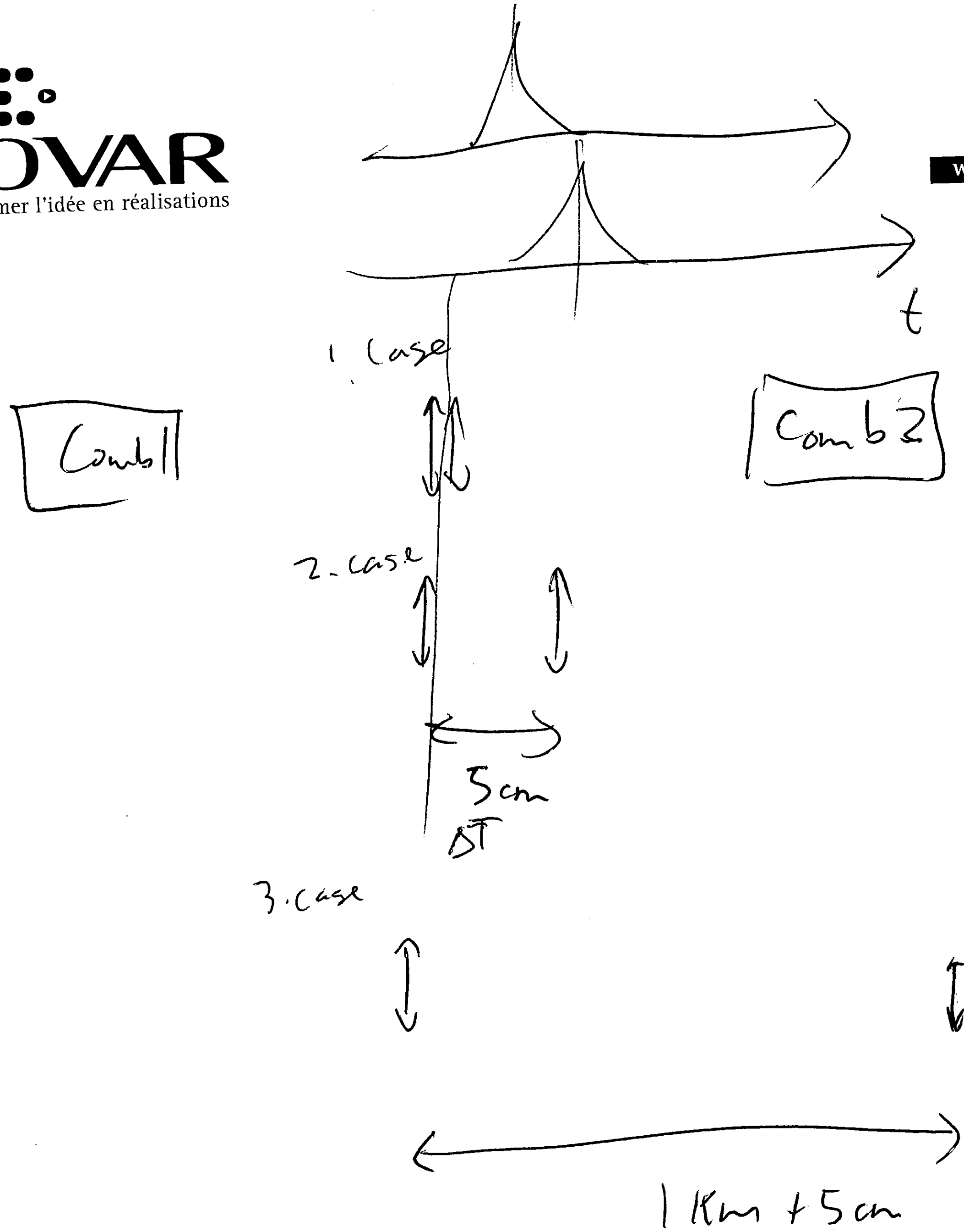

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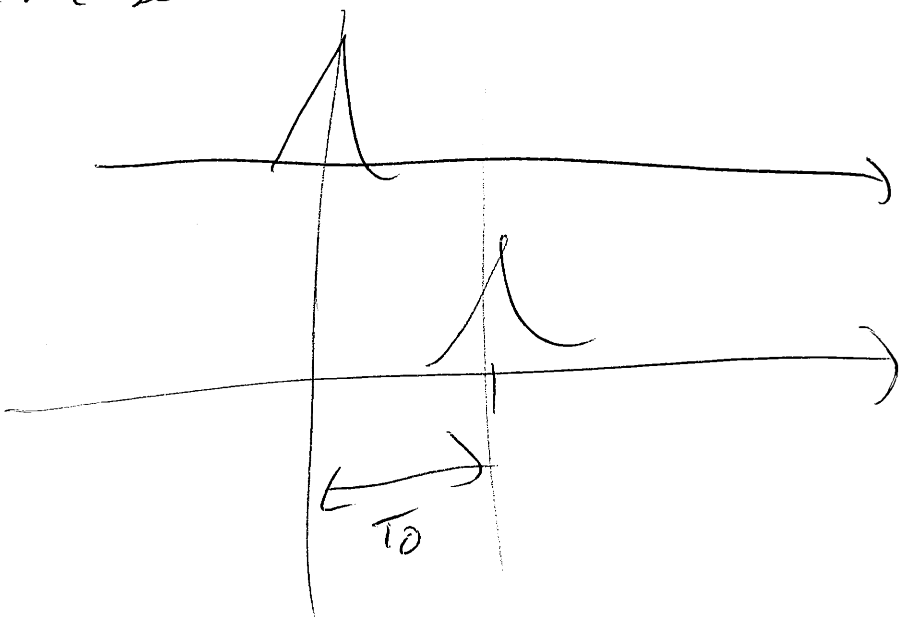
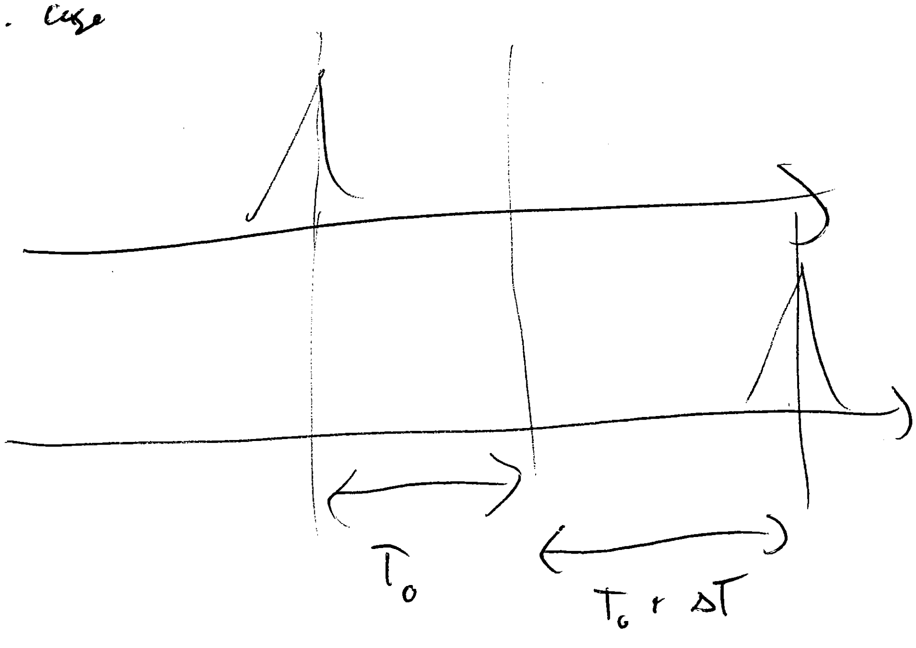
1. case



2. case



3. case

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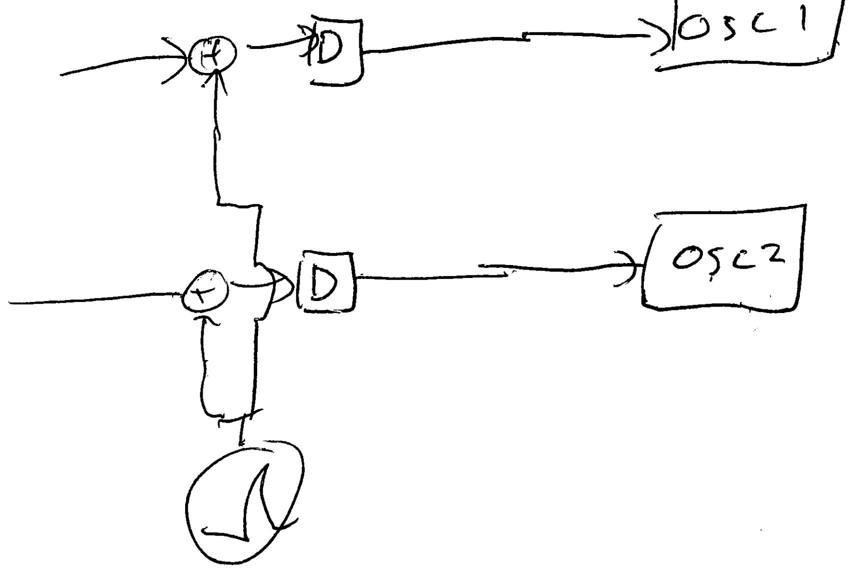
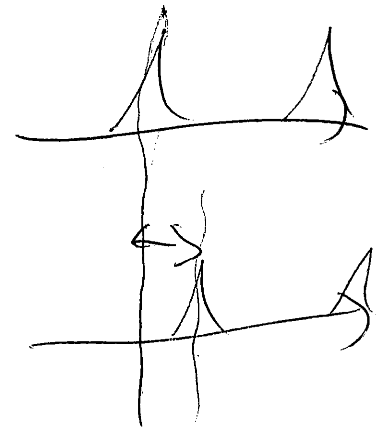
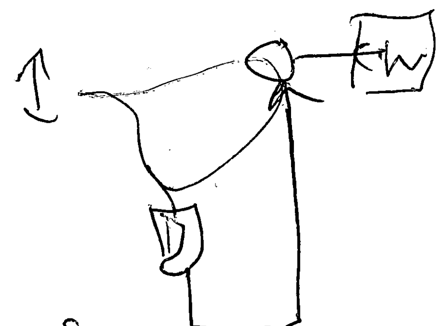




$T_0 + \Delta T + \Delta T_{osc}$



$\frac{1e-9}{100e3} = \frac{1e-9}{1e5} = 1e-14$

10 fs

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$$p(I_{obj},\theta_K) = \frac{\sqrt{I_{obj}}}{2\sqrt{\pi}\,\langle I_{obj}\rangle^{3/2}\,\sigma_z}\exp\left(\frac{-I_{obj}}{\langle I_{obj}\rangle}\left(1+\frac{\theta_K^2}{4\sigma_z^2}\right)\right), \quad I_{obj}>0$$

$$= b\sqrt{x}\,\exp(ax)$$

$$p(\theta_K) = \int_{-\infty}^{\infty} p(I_{obj},\theta_K)\,dI_{obj} = \int_0^{\infty} p(I_{obj},\theta_K)\,dI_{obj}$$

$$= b\int_0^{\infty}\sqrt{I_{obj}}\,\exp(a I_{obj})\,dI_{obj} = b\left[\frac{1}{a}\sqrt{I_{obj}}\,\exp(a I_{obj}) + \frac{j\sqrt{\pi}}{2a^{3/2}}\,\mathrm{erf}\left(j\sqrt{a I_{obj}}\right)\right]_0^{\infty}$$

$$a = -\frac{1}{\langle I_{obj}\rangle}\left(1+\frac{\theta_K^2}{4\sigma_z^2}\right) \qquad b = \frac{1}{2\sqrt{\pi}\,\langle I_{obj}\rangle^{3/2}\,\sigma_z}$$

$$p(\theta_K) = b\left[\cancel{\frac{1}{a}\sqrt{\infty}\,\exp(-\infty)}^{\,0} - \cancel{\frac{1}{a}\sqrt{0}\,\exp(0)}^{\,0} + \frac{j\sqrt{\pi}}{2a^{3/2}}\left[\mathrm{erf}\left(\sqrt{|a|\infty}\right)^{1} - \cancel{\mathrm{erf}(0)}^{\,0}\right]\right]$$

$$= b\left[\frac{j\sqrt{\pi}}{2a^{3/2}}\right] = \frac{b\cdot j\sqrt{\pi}}{2}\cdot\left(\frac{-1}{\langle I_{obj}\rangle}\left(1+\frac{\theta_K^2}{4\sigma_z^2}\right)\right)^{-3/2}$$

$$= \frac{b\cdot j\sqrt{\pi}}{2}\cdot(-1)^{-3/2}\cdot\left(\frac{1}{\langle I_{obj}\rangle}\left(1+\frac{\theta_K^2}{4\sigma_z^2}\right)\right)^{-3/2}$$

$$= \frac{b\sqrt{\pi}}{2}\left(\frac{1}{\langle I_{obj}\rangle}\left(1+\frac{\theta_K^2}{4\sigma_z^2}\right)\right)^{-3/2}$$

$$= \frac{\sqrt{\pi}}{2\cdot 2\cancel{\sqrt{\pi}}\,\cancel{\langle I_{obj}\rangle^{3/2}}\,\sigma_z}\cdot\frac{1}{\cancel{\langle I_{obj}\rangle^{-3/2}}}\cdot\left(1+\frac{\theta_K^2}{4\sigma_z^2}\right)^{-3/2}$$

$$= \frac{1}{4\sigma_z}\left(\frac{4\sigma_z^2+\theta_K^2}{4\sigma_z^2}\right)^{-3/2}$$

$$p(\theta_K) = \frac{1}{4\sigma_z}\cdot\left(\frac{4\sigma_z^2}{4\sigma_z^2+\theta_K^2}\right)^{3/2}$$

$p(\theta_K)$ is almost Lorentzian. Lorentzian distribution is known for having no higher moments.

Fixed.

~~Has to be wrong because of scaling with~~ $\cancel{\langle I_{obj}\rangle}$

Is $p(I_{obj},\theta_K)$ even normalized?

- Fix expression for $p(\theta_K)$: OK normalized
- Compute second moment $\langle\theta_x\rangle^2 = \int_{-\infty}^{\infty}\theta_x^2\,p(\theta_x)\,d\theta_x$

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$$p(\theta_K) = \frac{1}{4\sigma_z}\left(\frac{4\sigma_z^2}{4\sigma_z^2 + \theta_K^2}\right)^{3/2}$$

$p(x) = \frac{1}{\pi(1+x^2)}$ is Lorentzian

$$= \frac{1}{4\sigma_z}\left(1 + \frac{\theta_K^2}{4\sigma_z^2}\right)^{-3/2}$$

$$= \frac{1}{4\sigma_z}\left(\cancel{\frac{4\sigma_z^2}{4\sigma_z^2}} \cdot \frac{1}{1+\left(\frac{\theta_K}{2\sigma_z}\right)^2}\right)^{3/2}$$

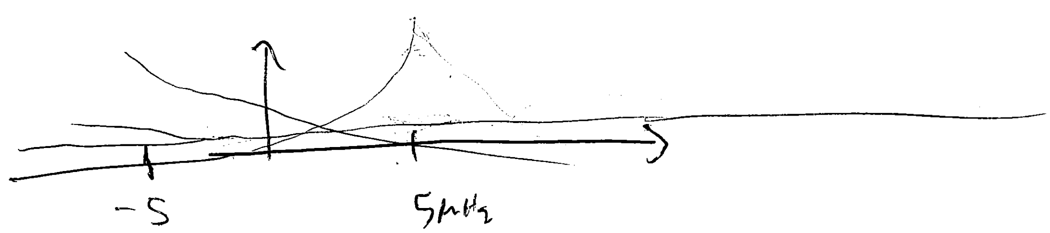
$$= \frac{1}{4\sigma_z}\left(\frac{1}{1+\left(\theta_K/2\sigma_z\right)^2}\right)^{3/2}$$

$$\langle \theta_K \rangle^2 = \int_{-\infty}^{\infty} p(\theta_K)\theta_K^2 \, d\theta_K = \int_{-\infty}^{\infty} \frac{1}{4\sigma_z}\left(\frac{1}{1+\left(\frac{\theta_K}{2\sigma_z}\right)^2}\right)^{3/2} \theta_K^2 \, d\theta_K$$

Could ~~do~~ do: ~~...~~ $2\int_0^\infty d\theta_K$

$$u = \left(\frac{\theta_K}{2\sigma_z}\right)^2 \qquad du = \frac{2\theta_K}{4\sigma_z^2} d\theta_K$$

does not converge . . .



Signal is third order polynomial
We do a first-order least-squares fit
→ Which is really just a linear combination of our points
How much does third order term affect the slope that we find?

But first: What is derivative of our third order polynomial

$R(z)$ = gaussian with std deviation = ΔR ~~with center~~
$h(t) = \sum_{k=1}^{N} a_k \delta(t - z_k)$
z_k are Gaussian with std dev = σ_z, a_k are? we don't really care

$I(z) = R(z) \exp\left(j 2\pi \frac{z}{\lambda}\right)$ IGM for a single scatterer at $z=0$.

$I_h(z) = I(z) * h(t) = \sum_{k=1}^{N} a_k R(z - z_k) \exp\left(\frac{j 2\pi (z - z_k)}{\lambda}\right)$

$I_h(z) = R(z) * \sum_{k=1}^{N} a_k \exp\left(\frac{j 2\pi (z - z_k)}{\lambda}\right) = R(z) * h_c(z)$

$R(z)$ is smooth compared to $h_c(z)$
$R(z)$ ~~$\approx 1 + R_1(z) \cdot z + R_2$~~

We look for the peak in the envelope of $I_h(z)$, or the envelope squared
$|I_h(z)|^2 = \left| \sum_{k=1}^{N} a_k R(z - z_k) \exp\left(j 2\pi \left(\frac{z - z_k}{\lambda}\right)\right)\right|^2$

$\frac{Hz}{V} \cdot \frac{\text{Counts}}{Hz} \cdot \frac{V}{\text{Counts}} = \frac{\text{counts}}{\text{counts}}$

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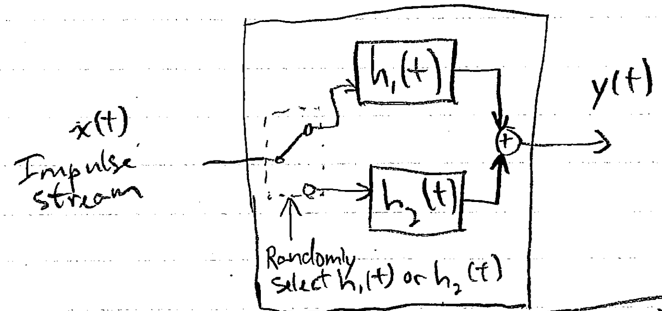
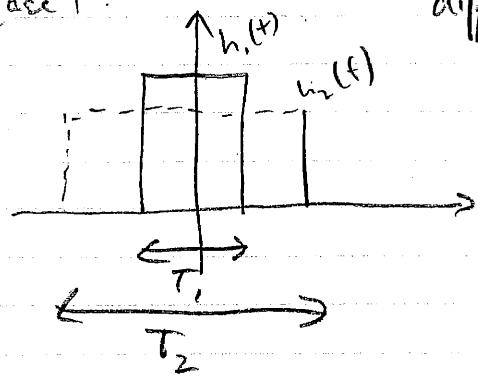
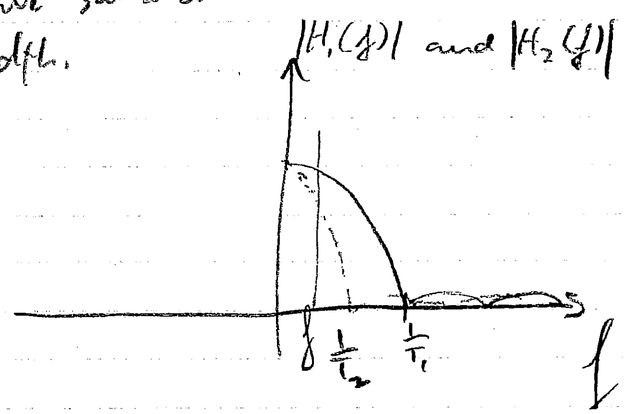


Simple model: send Dirac impulses to a stochastic system, which chooses either $h_1(t)$ or $h_2(t)$ as the impulse response for each.

Case 1: $h_1(t)$ and $h_2(t)$ have same center but different width.

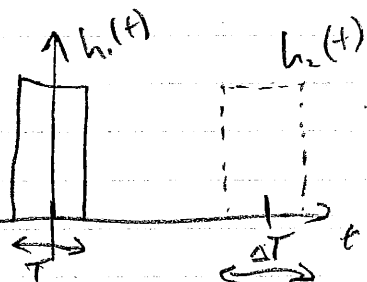


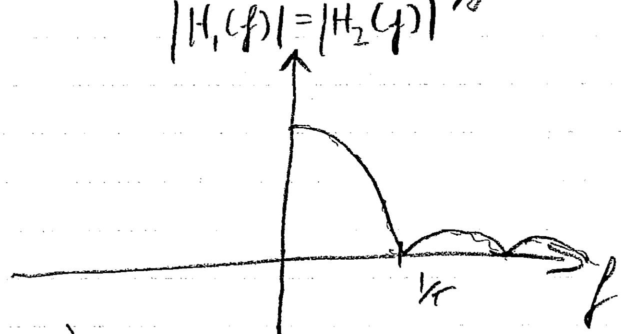
$\mathcal{F}\{\}$



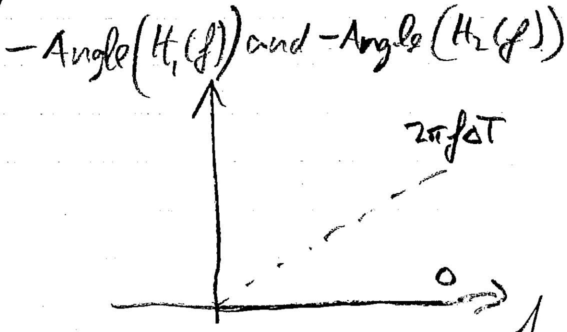
Results: - Gives a lot of excess amplitude noise at f.
- Gives no extra jitter (Center of mass of $h(t)$ doesn't change)
- Excess noise factor is high at f $F_{|H(f)|} > 0$.

Case 2: $h_1(t)$ and $h_2(t)$ have same width but different centers





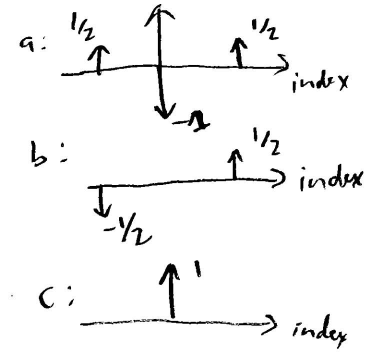
Results: - No excess amplitude noise (for $\Delta T/ \ll 1$)
- Excess timing jitter equal to $\frac{\Delta T}{2}$.
- Excess noise factor is $F_{|H(f)|} = 0$.



$$a = \frac{1}{2}\left[y(-1) - 2y(0) + y(1)\right] = \frac{1}{2}\left[y_{-1} - 2y_0 + y_1\right]$$

$$b = \frac{1}{2}\left[y(1) - y(-1)\right] = \frac{1}{2}\left[y_1 - y_{-1}\right]$$

$$c = y(0) = y_0$$



$$K_{peak} = \frac{-b}{2a} = \frac{1}{2}\,\frac{y_{-1} - y_1}{y_{-1} - 2y_0 + y_1}$$

$$Y_{peak} = aK_{peak}^2 + bK_{peak} + c$$

$$= \frac{b^2}{2a} + \frac{-b^2}{2a} + c$$

noiseless measurement vector

Assume three noise samples:

$$y_{-1} = \tilde{y}_{-1} + n_{-1}$$
$$y_0 = \tilde{y}_0 + n_0$$
$$y_1 = \tilde{y}_1 + n_1$$

Transformation from the measurement vector $[y_{-1}, y_0, y_1]$ to the peak position is: $K_{peak} = \frac{1}{2}\frac{y_{-1}-y_1}{y_{-1}-2y_0+y_1}$

ie it is non-linear, but we approximate it with the following:

$$K_{peak} = \frac{1}{2}\,\frac{\tilde{y}_{-1} - \tilde{y}_1 + n_{-1} - n_1}{\tilde{y}_{-1} + n_{-1} + -2\tilde{y}_0 - 2n_0 + \tilde{y}_1 + n_1} = \frac{1}{2}\,\frac{\tilde{y}_{-1} - \tilde{y}_1 + n_{-1} - n_1}{\alpha\left(1 + \frac{n_{-1} - 2n_0 + n_1}{\alpha}\right)} \approx \frac{1}{2}\,\frac{\beta}{\alpha}\,\frac{\left(1 + \frac{n_{-1}-n_1}{\beta}\right)}{\left(1 + \frac{n_{-1}-2n_0+n_1}{\alpha}\right)}$$

$$\simeq \frac{1}{2}\,\frac{\beta}{\alpha}\left(1 + \frac{n_{-1}-n_1}{\beta} - \frac{n_{-1}-2n_0+n_1}{\alpha}\right) = \tilde{K}_{peak}\left(1 + \frac{n_{-1}-n_1}{\beta} - \frac{n_{-1}-2n_0+n_1}{\alpha}\right)$$

with $\beta = \tilde{y}_{-1} - \tilde{y}_1$ $\quad \tilde{K}_{peak} = \frac{1}{2}\frac{\beta}{\alpha}$ (the true peak position)

$\alpha = \tilde{y}_{-1} - 2\tilde{y}_0 + \tilde{y}_1$

$$\varepsilon = K_{peak} - \tilde{K}_{peak} = \frac{1}{2}\,\frac{\beta}{\alpha}\left(\frac{n_{-1}-n_1}{\beta} - \frac{n_{-1}-2n_0+n_1}{\alpha}\right)$$

$$\text{Error} = \frac{1}{2}\left(\frac{n_{-1}-n_1}{\alpha} - \frac{n_{-1}-2n_0+n_1}{\alpha^2}\cdot\beta\right)$$

$$= \frac{1}{2}\left[n_{-1}\left(\frac{1}{\alpha} - \frac{\beta}{\alpha^2}\right) + n_0\left(\frac{-2\beta}{\alpha^2}\right) + n_1\left(\frac{-1}{\alpha} - \frac{\beta}{\alpha^2}\right)\right]$$

The coefficients in parenthesis should be $2\frac{\partial \text{Error}}{\partial n_{-1}}, 2\frac{\partial \text{error}}{\partial n_0}, 2\frac{\partial \text{Error}}{\partial n_1}$.

Potential complication: coefficients depend on the values of β and α, which depends on the subsample peak position.

1. Let's decompose the noise differently:

$$a = \frac{1}{2}\left[Y_{-1} - 2Y_0 + Y_1 \right]$$

$$b = \frac{1}{2}\left[Y_1 - Y_{-1} \right]$$

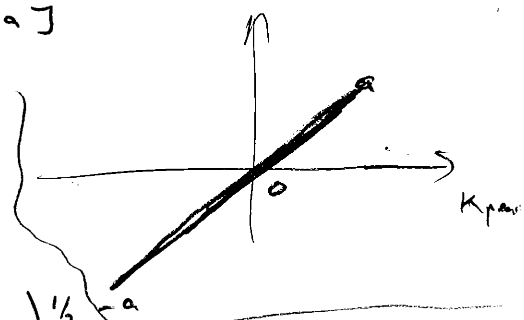
$$K_{peak} = \frac{-b}{2a}$$

$$\frac{\partial K_{peak}}{\partial b} = \frac{-1}{2a}$$

$$\frac{\partial K_{peak}}{\partial a} = \frac{+b}{2a^2}$$

4. a is pretty much constant with respect to K_{peak} but b varies linearly with K_{peak}, so the ~~b~~ noise gets converted to K_{peak} noise differently depending on the subsample position, while the b noise has pretty much always the same conversion coefficient, $\frac{-1}{2a}$.

5. Assume that we know a for out signal, and K_{peak} varies within $[-0{,}5,\ 0{,}5]$:
How does b scale? ~~so~~ $-2aK_{peak} = b \in [-a,\ a]$
What is the RMS conversion coefficients?



$$\sqrt{\left\langle \left(\frac{\partial K_{peak}}{\partial b} \right)^2 \right\rangle} = \sqrt{\frac{1}{4a^2}} = \frac{1}{2a}$$

RMS value of $\frac{\partial K_{peak}}{\partial a}$

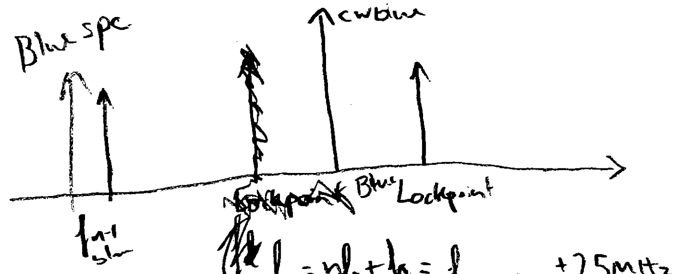
$$\sqrt{\left\langle \left(\frac{\partial K_{peak}}{\partial a} \right)^2 \right\rangle} = \left(\frac{1}{1} \int_{-0{,}5}^{0{,}5} (-2aK_{peak})^2 \, dK_{peak} \right)^{1/2} \cdot \frac{1}{2a^2} = \left(4a^2 \int_{-0{,}5}^{0{,}5} K_{peak}^2 \, dK_{peak} \right)^{1/2} \cdot \frac{1}{2a^2}$$

$$= \frac{2a}{2a^2} \left(\frac{1}{3} K_{peak}^3 \Big|_{-0{,}5}^{0{,}5} \right)^{1/2} = \frac{2|a|}{\sqrt{3} \cdot 2a^2} \left(0{,}5^3 + 0{,}5^3 \right) = \frac{2a}{\sqrt{3}} \cdot \frac{2}{2a^2} \cdot \frac{1}{2^3} = \frac{4\cancel{a}\cancel{1}}{\cancel{4}\sqrt{3}|a|} = \frac{1}{4\sqrt{3}|a|}$$

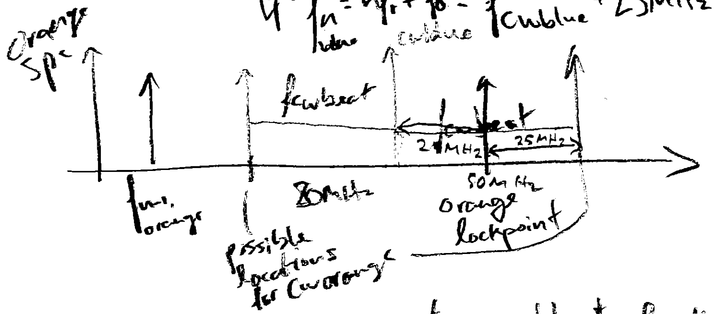
~~$std(K_{peak}) = \sqrt{\left\langle \left(\frac{\partial K_{peak}}{\partial b} \right)^2 \right\rangle} \cdot std(b) +$~~

3. $$Var(K_{peak}) = \left\langle \left(\frac{\partial K_{peak}}{\partial b} \right)^2 \right\rangle \cdot Var(b) + \left\langle \left(\frac{\partial K_{peak}}{\partial a} \right)^2 \right\rangle \cdot Var(a)$$

2. $$Var(K_{peak}) = Var\left(\frac{\partial K_{peak}}{\partial b} \cdot b_{noise} \right) + Var\left(\frac{\partial K_{peak}}{\partial a} \cdot a_{noise} \right)$$

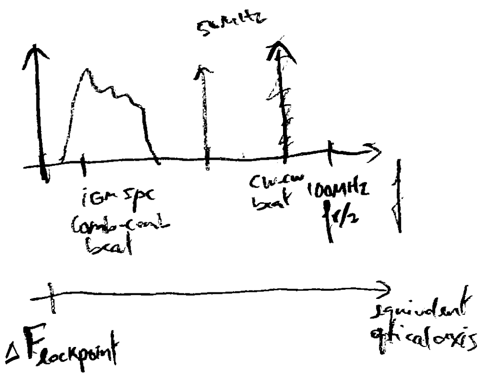


$f_n = n f_r + f_0 = f_{cwblue} + 25\,\mathrm{MHz}$



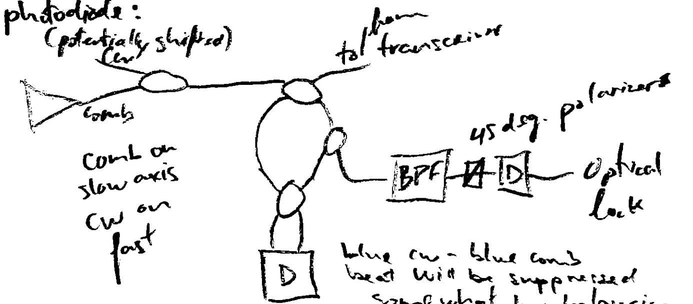
$f_{lockpoint\,orange} = f$

We want a beat spectrum that looks like:



$f_{beatcwcw} = f_{cwblue} - f_{cworange}$

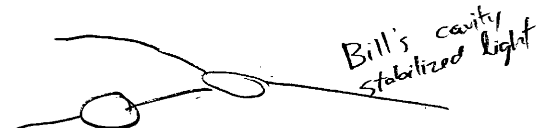
Then combine all these fields onto a photodiode:



Comb on slow axis
cw on fast

blue cw - blue comb beat will be suppressed somewhat by balancing.
+
combs fields are on one polarization while cw fields are on a different one.

↳ doesn't work because transceivers will let only 1 polarization through.



AOM
shift by ΔF_{cw}
Say 80 MHz

which is nice because we need an AOM anyway for Doppler cancellation (needed on both so we just use that frequency offset.

On the IGM detector, we want:

100 μW local cw
100 μW local comb in 10 THz BW

$\sqrt{2\,nW \cdot 100\,\mu W} = 50\,\mu W$

On the optical beat detector, we want ideally:

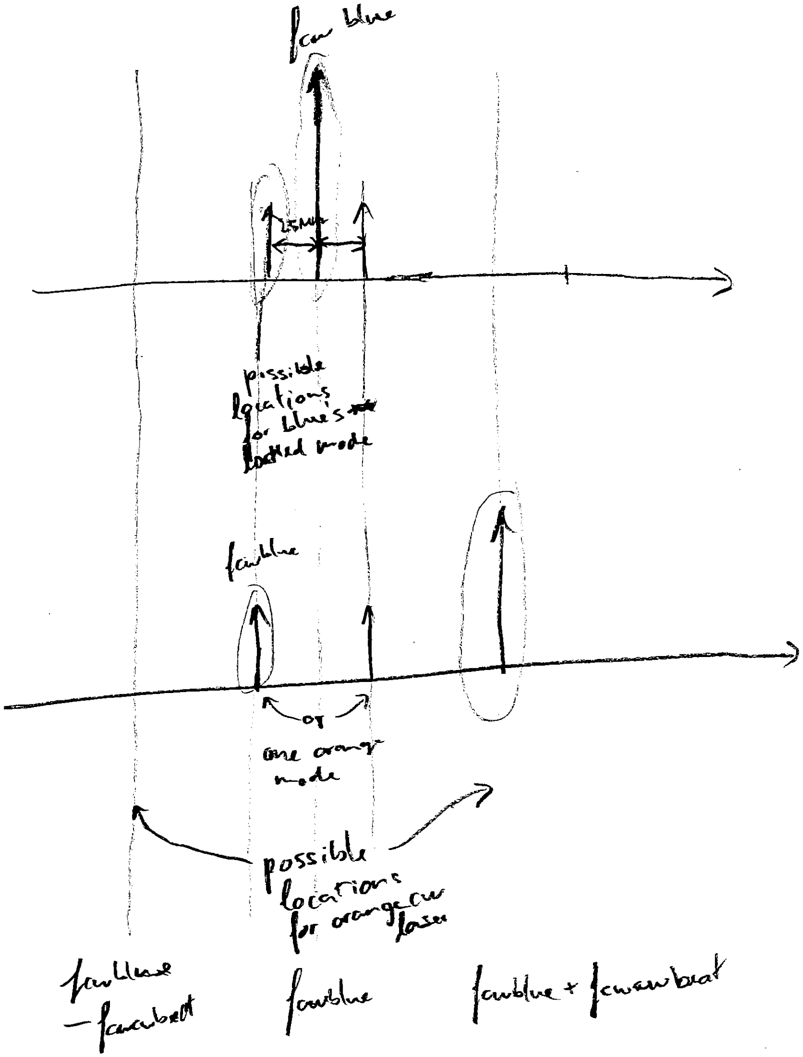
100 μW local cw
100 μW local comb in 100 GHz BW

But we'll have only $\sim 100\,\mu W \cdot \frac{100\,GHz}{10\,THz} \sim 1\,\mu W$ in 100 GHz

ie $100\,GHz = N \cdot 200\,MHz \Rightarrow N = 500$ modes

$\frac{1\,\mu W}{500\,modes} = \frac{2\,nW}{mode}$

Can use higher gain detector to have lower NEP or a GATOR. (GATOR paper used $20\frac{pW}{mode}$, 1 mW cw)



We would like $f_{cwcwbeat}$ to be well on one side of the be spectrum, for example $f_{cwcwbeat} = 75MHz$ or even $95MHz$

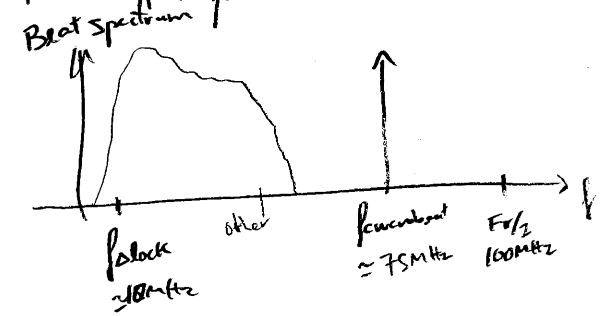
OK assume

$f_{cwcwbeat} = f_{cworange} - f_{cwblue}$ (orange cw laser is higher)

$f_{locked\ blue\ comb} = f_{cwblue} - 25MHz$ $\quad f_{lock} = 25MHz$

$= f_{cwblue} - f_{lock}$

Assume we want one mode from the orange laser to be ~10MHz offset from blue's laser



$f_{slock} = f_{lockedorange} - f_{lockedblue}$

$f_{locked\ orange} = f_{lockedblue} + f_{slock}$

$= f_{lockedblue} + 10\,MHz$

$= f_{cwblue} - 25MHz + 10MHz$

$f_{locked\ orange} = f_{cwblue} - f_{lock} + f_{slock}$

$f_{locked\ orange} = f_{cworange} \pm f_{lock}$

So we need $f_{cwblue} - f_{lock} + f_{slock} = f_{cworange} \pm f_{lock}$

We already have a constraint for $f_{cwblue} - f_{cworange}$:

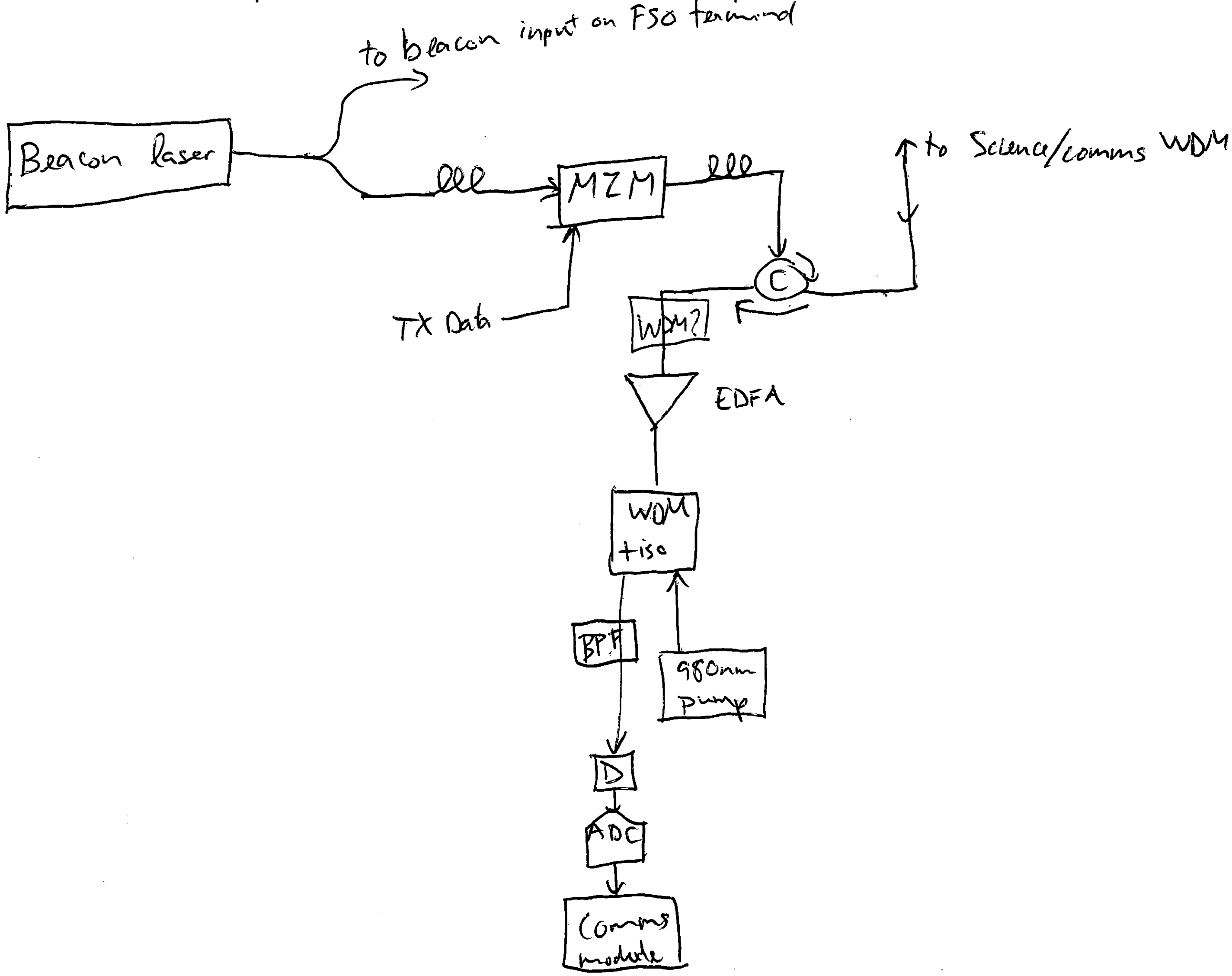
$f_{cwcwbeat} - f_{lock} \pm f_{lock} = -f_{slock}$

Let's put numbers

$f_{cwcwbeat} - 25\,MHz \pm 25MHz = -10MHz$ or $50MHz$

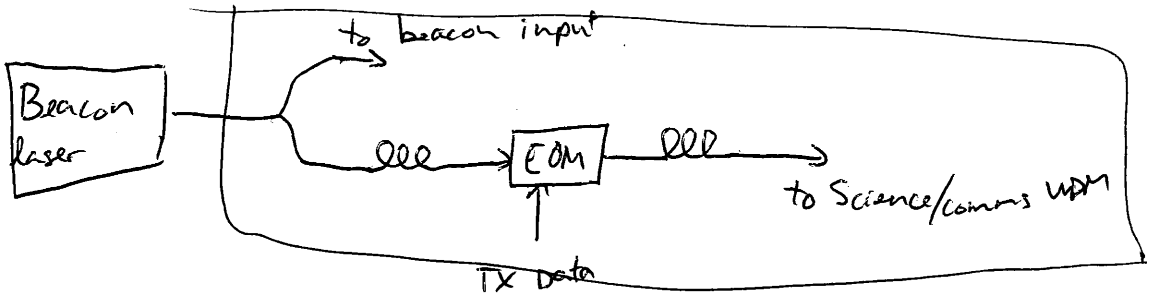
$75MHz - (50MHz \text{ or } 0) = -10MHz$

Optical comms package w/ optical pre-amp (incoherent detection), RX + TX

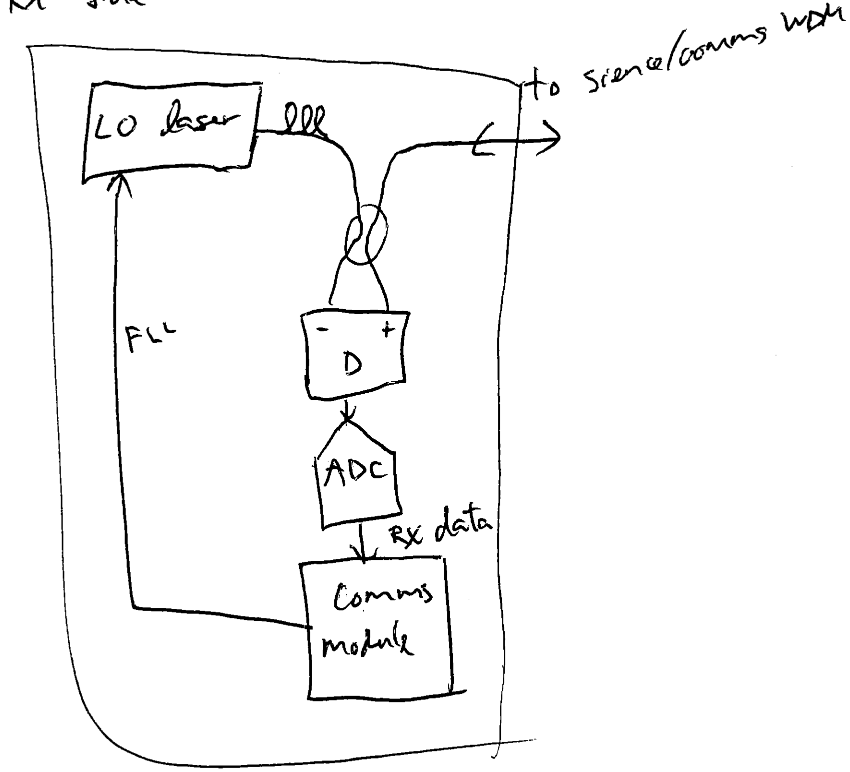


Coherent comms, separate RX and TX

TX side



Rx side



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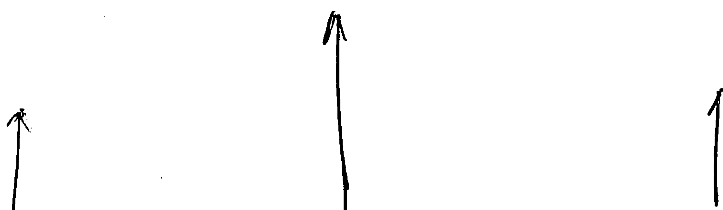
www.sovar.com

We want a linear combination of t_{-1}, t_0 and t_1 which gives 0 sensitivity to t_n and t_ϕ, and a gain of 1 through T



$t_{-1} = T - t_\phi - t_n \qquad t_0 = T + T_\phi \qquad t_1 = T - T_\phi + t_n$

$$\frac{1}{2}\left(t_{-1} + t_1\right) = T - t_\phi$$

$$t_0 + \frac{1}{2}\left(t_{-1} + t_1\right) = 2T$$

$$\frac{1}{2}\left(t_0 + \frac{1}{2}(t_{-1} + t_1)\right) = \frac{1}{2} t_0 + \frac{1}{4} t_{-1} + \frac{1}{4} t_1 = T$$

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Definitions used by Stable 32:

	Symbol	Units
Phase (really time)	$x[i]$	[Seconds]
(Fractional) Frequency $y[i] = \frac{\Delta f}{f}$	$y[i]$	[unitless]
Sampling period	tau, τ_0	
Real phase	$\phi(t)$	[radians]

to go from real phase to time: $\phi(t) = x(t) \cdot 2\pi\nu_0 = x(t) \cdot \frac{2\pi}{\text{Period}}$

$$x(t) = \phi(t) \cdot \frac{\text{Period}}{2\pi} = \frac{\phi(t)}{2\pi\nu_0}$$

To go from frequency to phase:

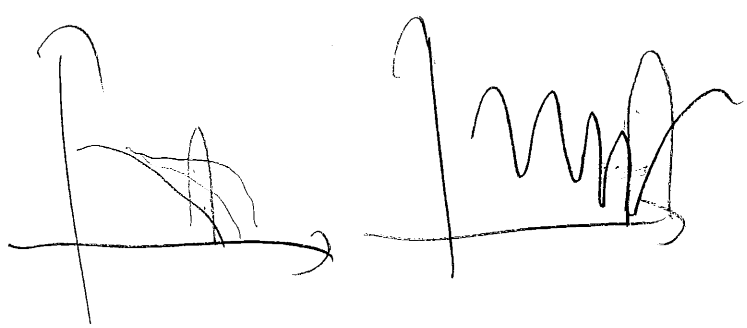
$$\phi(t) = 2\pi \int_{-\infty}^{t} \Delta f(u)\,du \qquad \Delta f(t) = \frac{1}{2\pi}\frac{d\phi(t)}{dt}$$

Between normalized frequency and phase and time:

$$y = \frac{\Delta f(t)}{\nu_0} = \frac{1}{2\pi\nu_0}\frac{d\phi(t)}{dt} = \frac{1}{2\pi\nu_0}\frac{d}{dt}\left(x(t)\cdot 2\pi\nu_0\right) = \frac{dx(t)}{dt}$$

This is why it doesn't need the carrier frequency if it has time.

$$y = \frac{d\,\text{delay}(t)}{dt} = \text{Delay slope in } \frac{\text{seconds}}{\text{seconds}}$$



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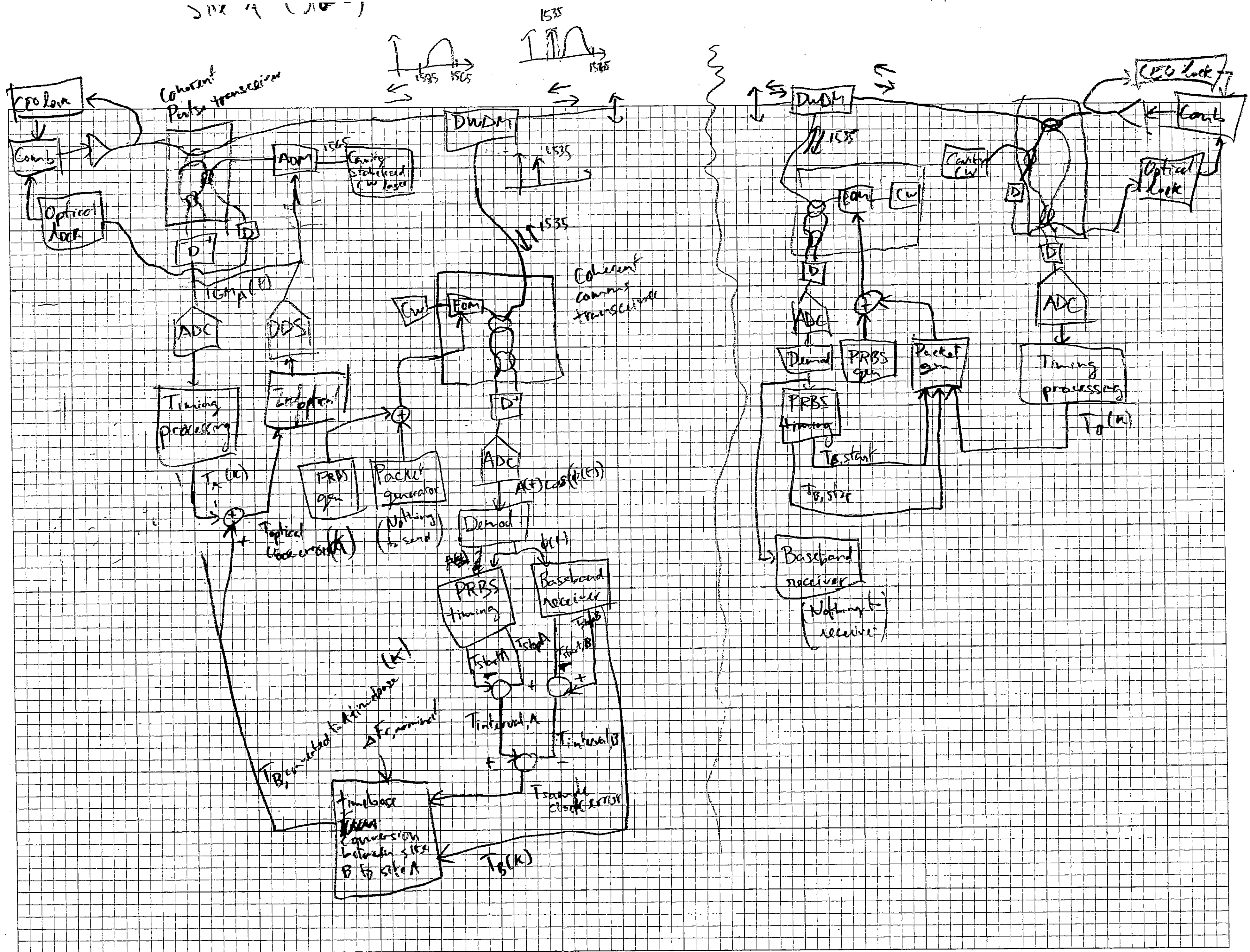
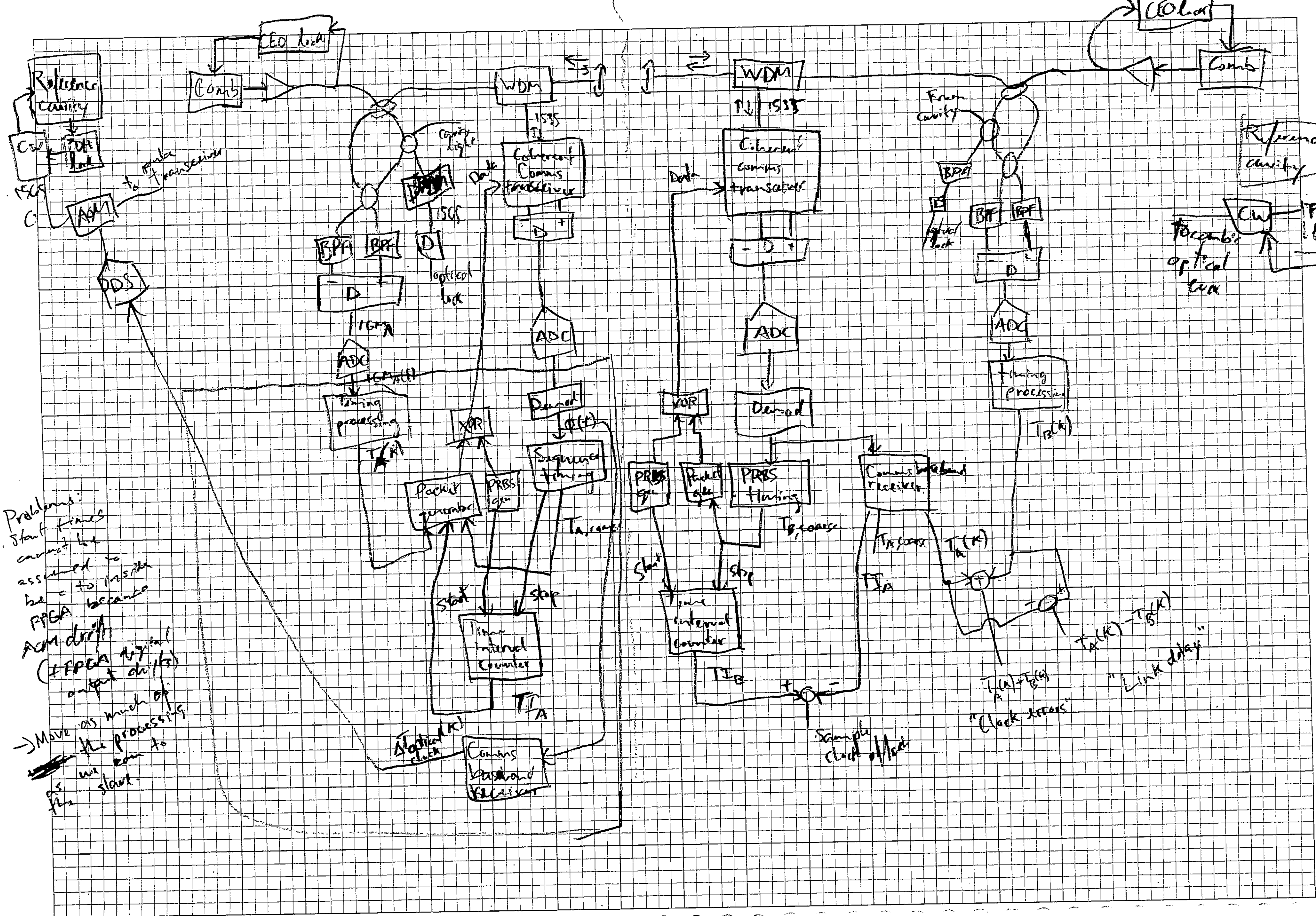

1535
1535 1565
1565
CEO lock
Comb
Optical lock
Coherent Pulse transceiver
AOM
1565
Cavity stabilized CW laser
DWDM
1535
D
ADC
DDS
Timing processing
Int optical
$T_{A}(k)$
$T_{optical\ lock\ error}(k)$
PRBS gen
Packet generator
(Nothing to send)
CW
EOM
Coherent comms transceiver
ADC
$A(t)\cos(\phi(t))$
Demod
$\phi(t)$
PRBS timing
Baseband receiver
$T_{start,A}$
$T_{stop,A}$
$T_{stop,B}$
$T_{start,B}$
$T_{interval,A}$
$T_{interval,B}$
$T_{sample\ clock\ error}$
$T_{B,\ converted\ to\ A\ timebase}(k)$
$\Delta F_{nominal}$
timebase conversion between site B to site A
$T_{B}(k)$
DWDM
1535
Cavity CW
Optical lock
CEO lock
Comb
EOM
CW
ADC
Demod
PRBS gen
Packet gen
PRBS timing
Timing processing
$T_{B}(n)$
$T_{B,start}$
$T_{B,stop}$
Baseband receiver
(Nothing to receive)

Site A (slave)

Site B (master)

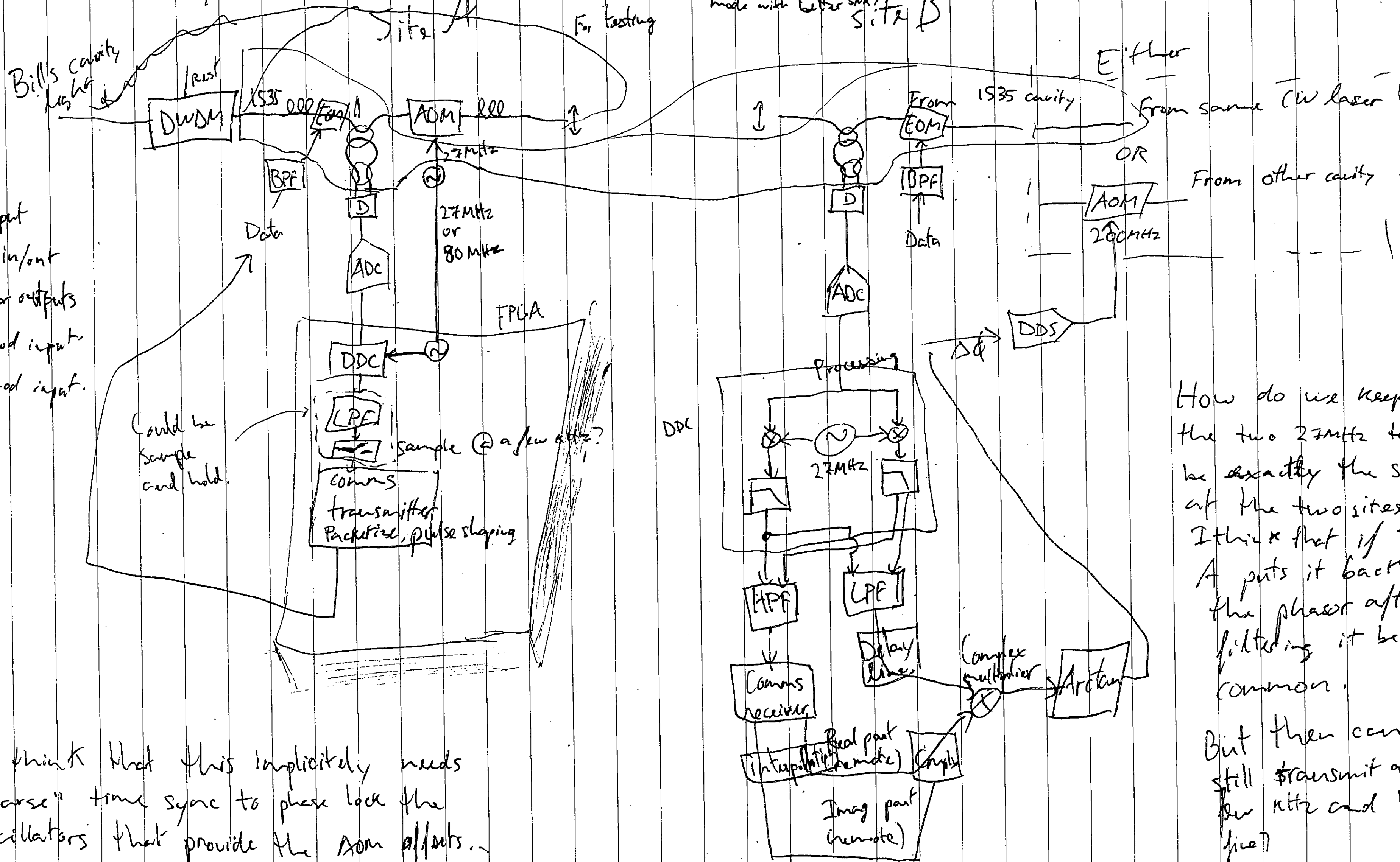


Problems: start times cannot be assumed to be = to inspite because FPGA AOM drift (+FPGA digital output drifts)

→ Move as much of the processing as we can to the slave.

Bill's 1535 lasers are offset by 204M between the original and the special cavities.

Two-way CW time transfer (really, syntonization only) + comms link + probably coarse time sync. (simply use preamble?) or have a dedicated mode with better SNR?



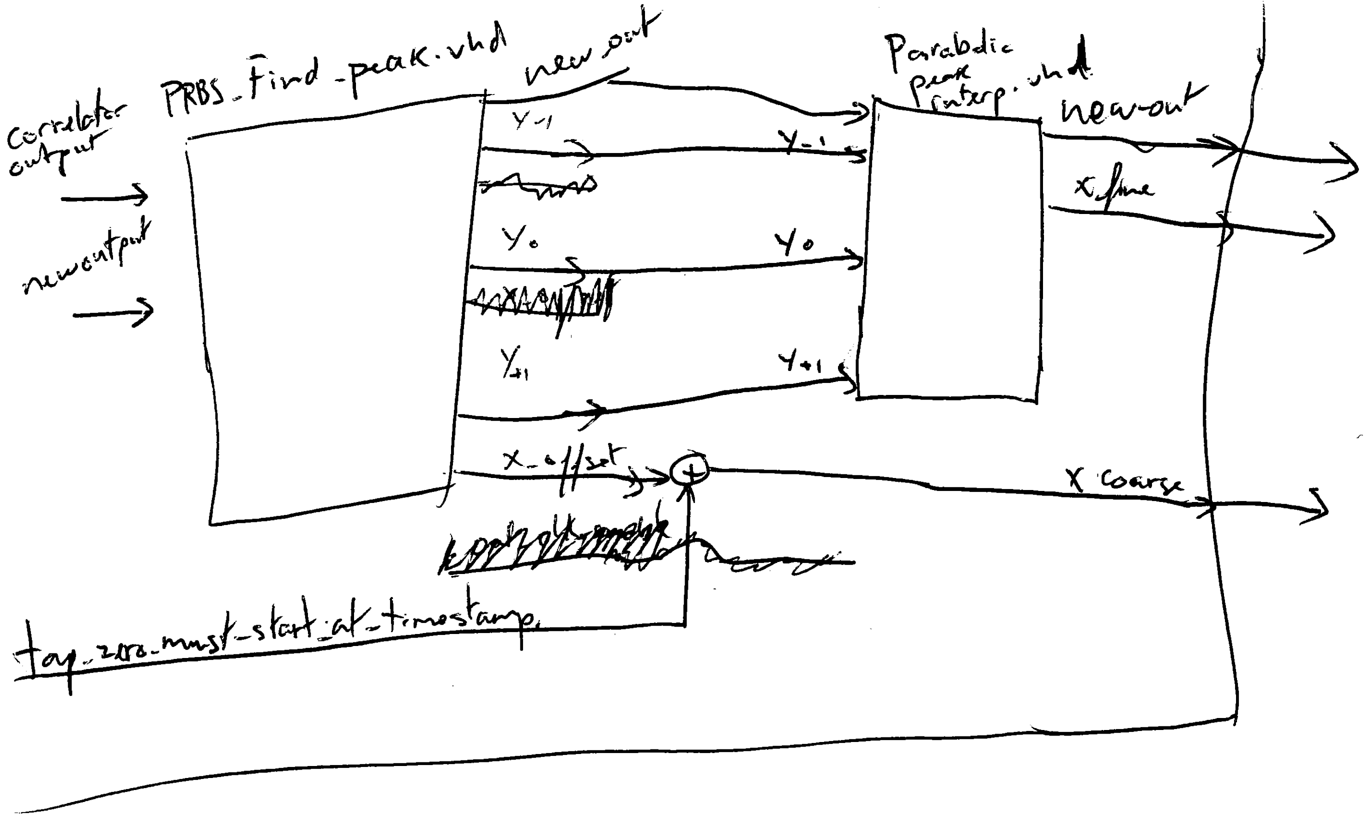
1 CW input
2 Link in/out
4 Detector outputs
2 EOM mod inputs
1 AOM mod input.

How do we keep the two 27MHz to be exactly the same at the two sites? I think that if site A puts it back on the phasor after filtering it becomes common.

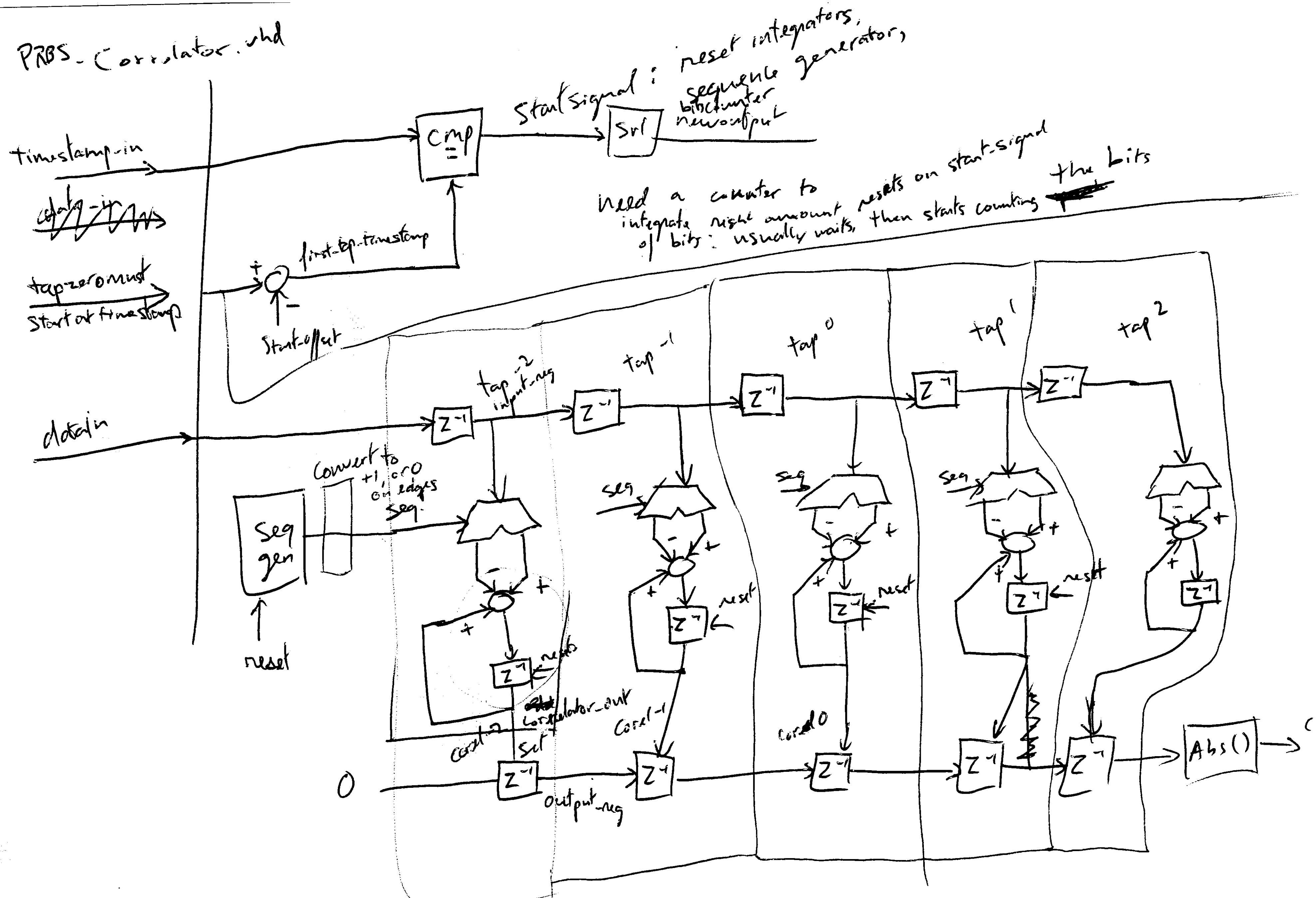
But then can we still transmit at a few kHz and be fine?

I think that this implicitly needs "coarse" time sync to phase lock the oscillators that provide the AOM offsets.

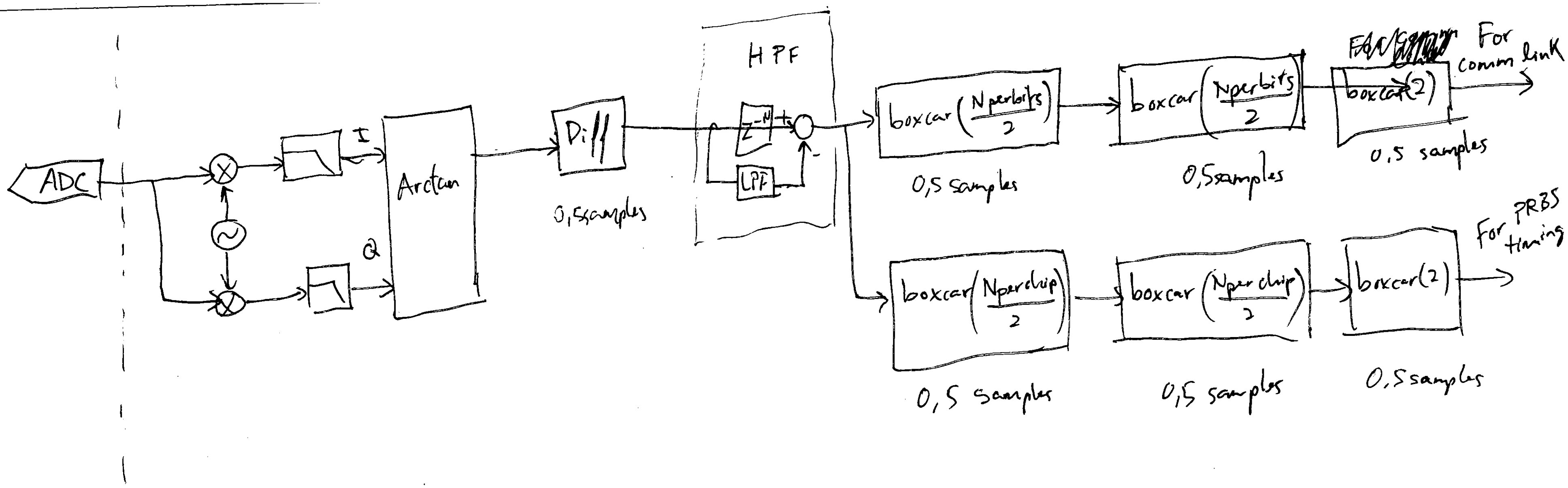
If we move the AOM to before the split, does this become in-loop?



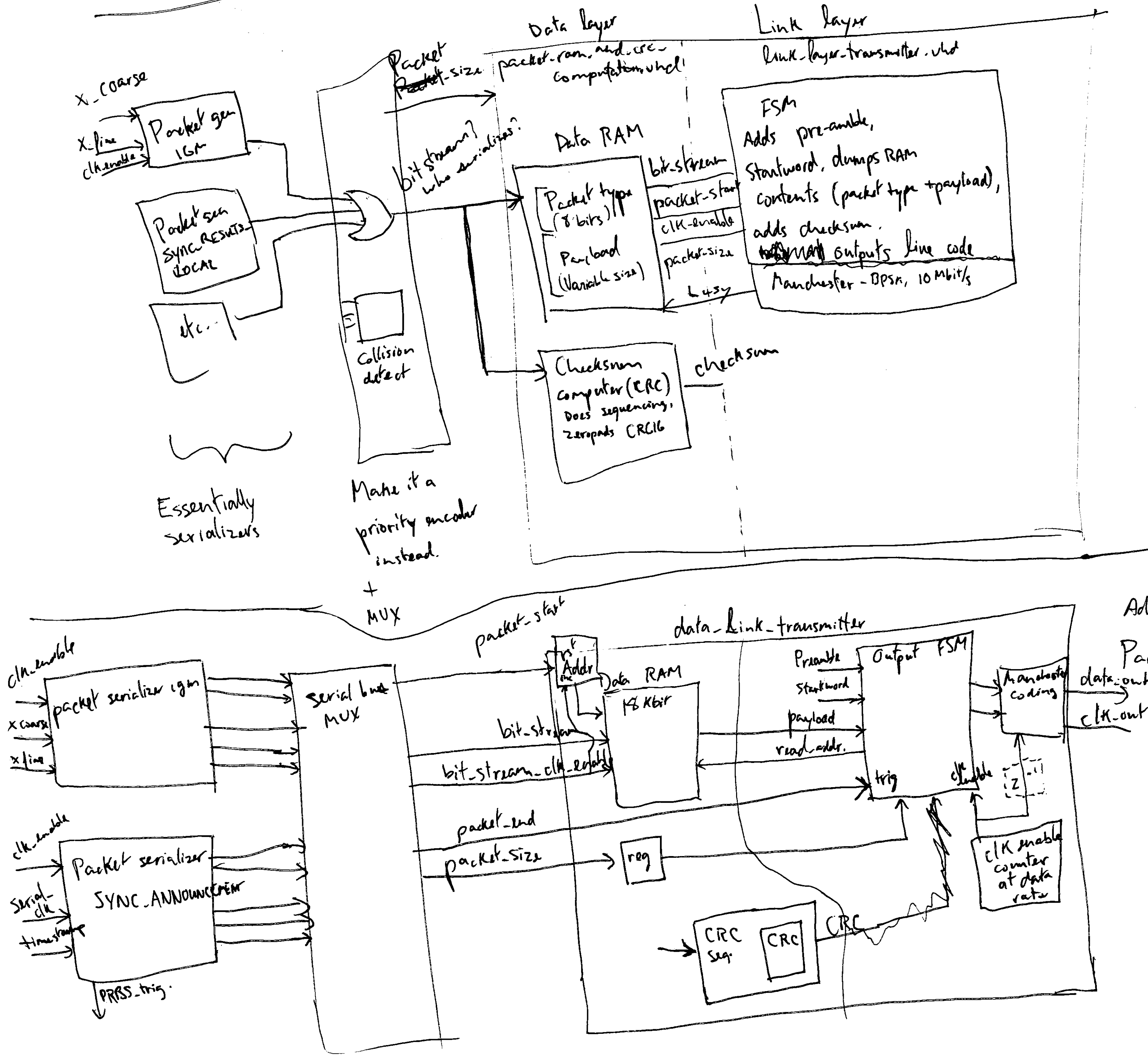
PRBS_Find_peak.vhd
Parabolic peak interp.vhd
correlator output
newoutput
new_out
Y-1
Y0
Y+1
x_offset
new_out
x_fine
x_coarse
tap_zero_must_start_at_timestamp



prbs_correlator_tap.vhd
integrator functions:
reset
hold
add
sub
needcarry chain: integrator
load ~~add~~ load carry

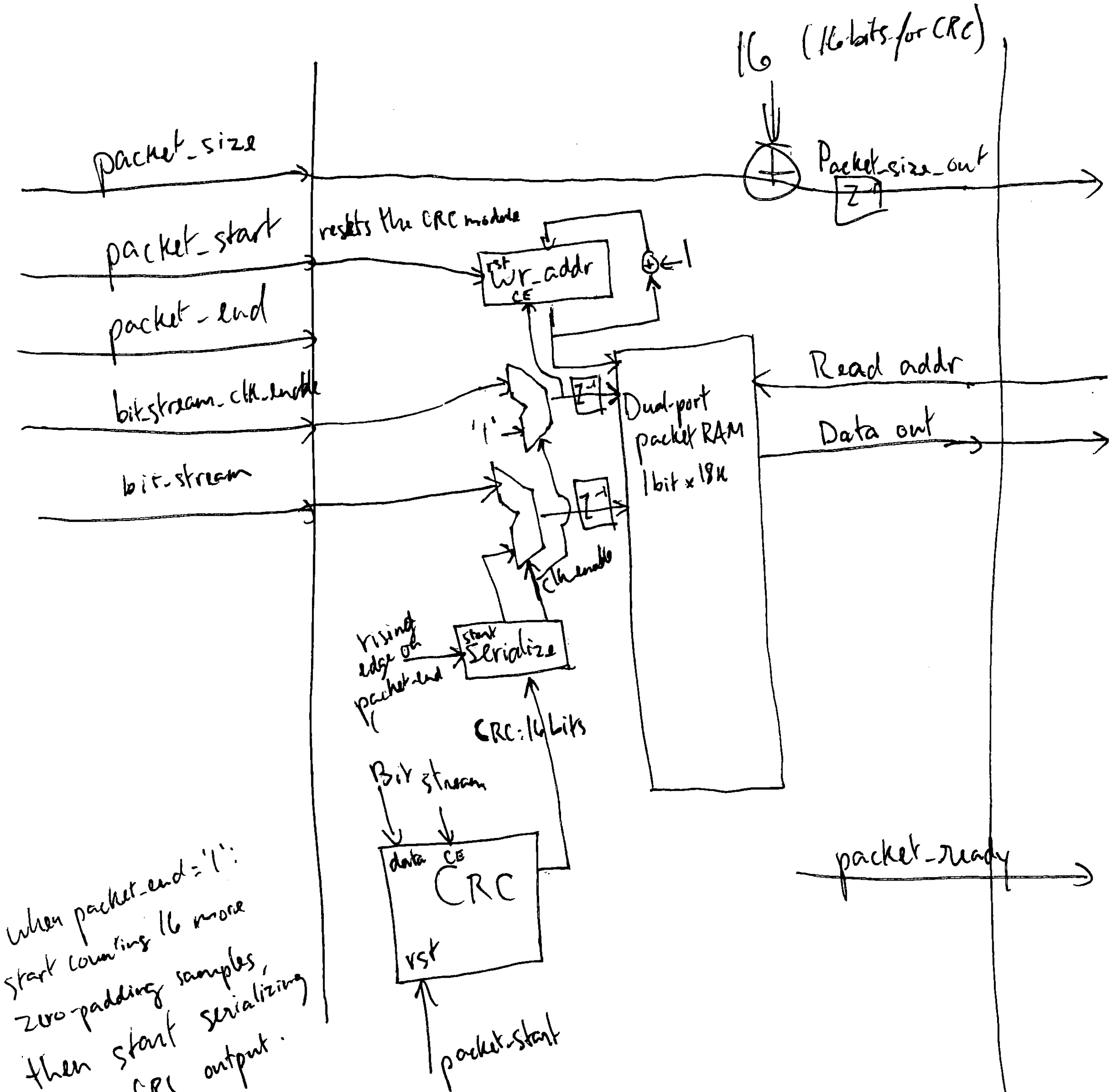

ADC
I
Q
Arctan
Diff
0,5 samples
HPF
LPF
boxcar(Nperbits/2)
0,5 samples
boxcar(Nperbits/2)
0,5 samples
boxcar(2)
0,5 samples
For comm link
boxcar(Nperchip/2)
0,5 samples
boxcar(Nperchip/2)
0,5 samples
boxcar(2)
0,5 samples
For PRBS timing

Transmitter design



Addr <= 0 when packet_start = '1'

Packet_end triggers the output F

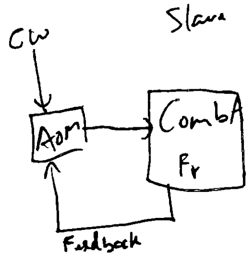


when packet_end='1': start counting 16 more zero-padding samples, then start serializing the CRC output.

Test system:
Looks at packet ready flag, then reads out the RAM at 10 Mbit/s from addr 0 to packet_size-1

→ Test a bit error too!

Slave



Needs to be able to translate timestamps from the master site to our own.

Master

Transmits the IGM timestamps + (eventually phase offset with Comb C)

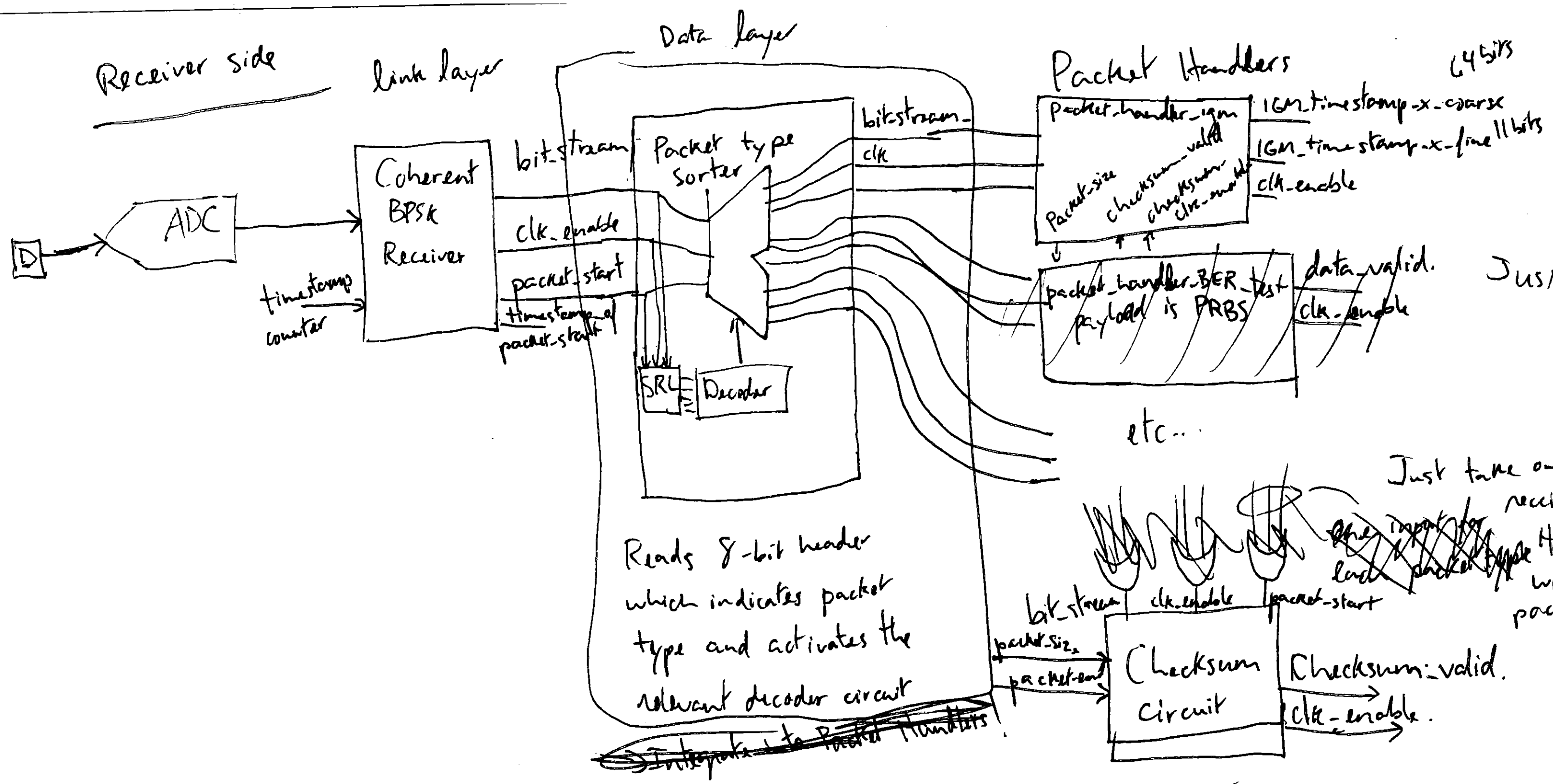
Protocol: Master is "active": initiates all comms.

1. Master detects an IGM, sends results packet.
2. Master checks if we need to run sync protocol (how long has it been since la

Sync protocol:

1. Master sends SYNC-ANNOUNCEMENT packet, indicating that PRBS will follow after N cycles after the start of the announcement.
2. Master sends the PRBS at pre-determined time.
3. Master sends local arrival timestamp of the PRBS.
4. Once slave receives master's timestamp, he sends his own SYNC-ANNOUNCEMENT.
5. Slave sends PRBS as announced.
6. Master receives the slave's PRBS, sends result packet.
7. Slave finally has all the required information to translate master's timestamps to its own clock.

Receiver side



Che

Just use checksum circuit...

Just take output of receiver....

~~One input for each packet type~~

How to do with variable packet lenght?

Checksum computed over the whole packet, starting at Startword or packet type, use CRC, bit-serial algorithm.

total packet size for IGM: 181 bits (1,8μs @10MS/s)

41% payload

Packet format

Pre-amble	Startword	Packet type	Payload	Checksum
1 0 1 0 1 0 (...) 50 bits	32 bits	8 bits	Variable size	16 bits
		IGM	IGM: 75 bits	
		~~[illegible]~~		
		SYNC_ANNOUNCEMENT	: 64 bits	
		SYNC_RESULTS_LOCAL		
		SYNC_RESULTS_REMOTE		

CCITT-CRC:

16 bits, polynomial:

$$x^{16} + x^{12} + x^5 + 1$$

In Hex: 0x1021 = ~~1 0000 0001 0000 0010 0001~~

$$\underbrace{1}_{\text{Implicit}}\ 0001\ 0000\ 0010\ 0001$$

$$x^{16} + x^{12} \qquad + x^5 + 1$$

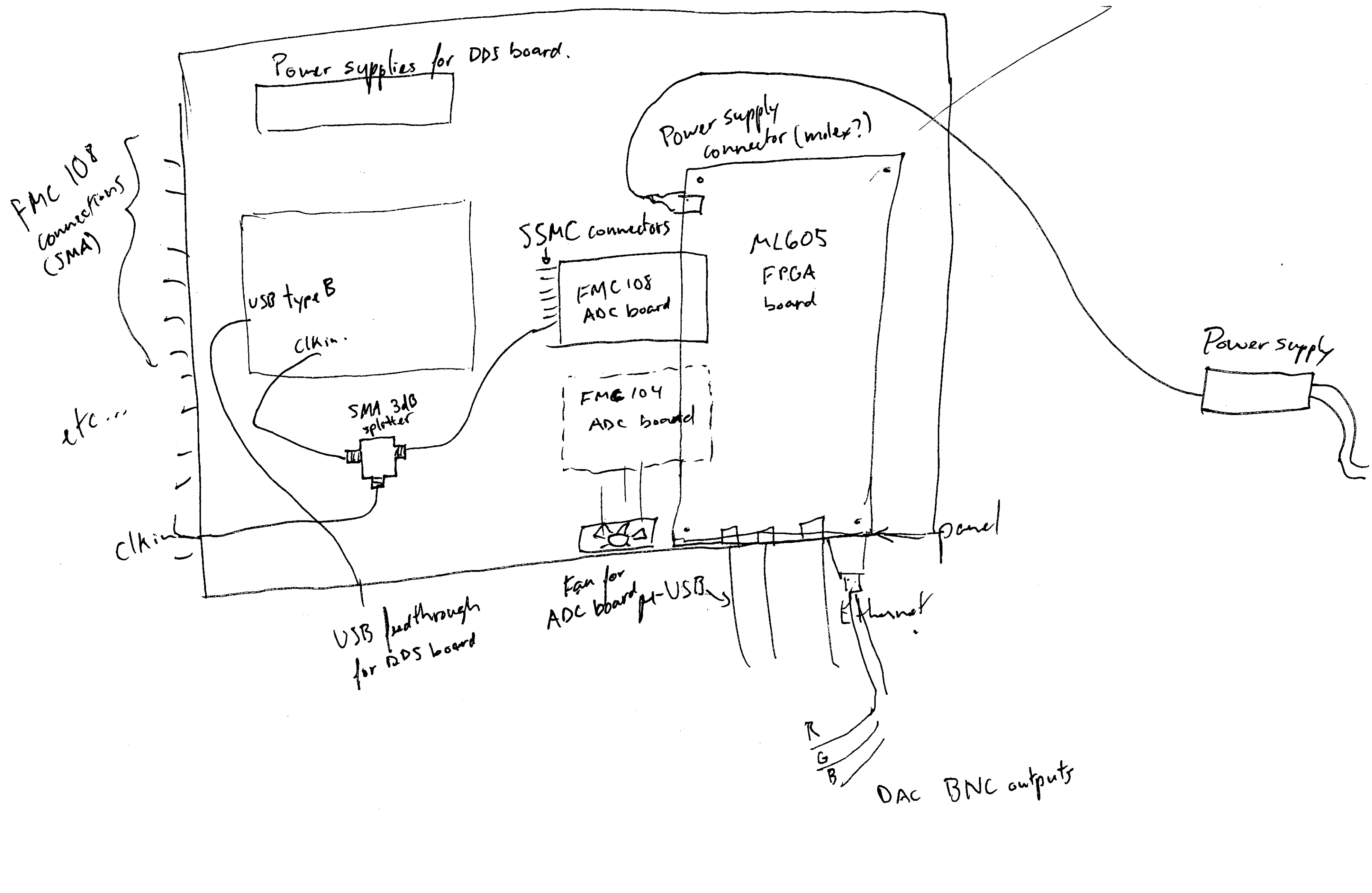
USB 2.0 - CRC16:

16 bits, polynomial:

$$x^{16} + x^{15} + x^2 + 1$$

Normal Hex: 0x8005 = $\underbrace{1}_{\text{Implicit}}$ 1000 0000 0000 0101

$$x^{16} + x^{15} \qquad + \qquad x^2 + 1$$



Power supplies for DDS board.
FMC 108 connections (SMA)
etc...
clkin
USB type B
clkin.
SMA 3dB splitter
SSMC connectors
FMC 108 ADC board
FMC 104 ADC board
Power supply connector (molex?)
ML605 FPGA board
Power supply
panel
USB feedthrough for DDS board
Fan for ADC board
μ-USB
Ethernet
R
G
B
DAC BNC outputs

- ML605 I/Os: (Virtex 6 FPGA eval board)

  - FMC108 ADC board: - Trig in / - Clock in / - 8 data channels } Total: 10 SMAs

  - 4 SMA GPIOs (could be used by DDS board internally)
  - 3 BNC outputs for VGA DAC (needs adapter DVI-VGA-BNC)
  - Gigabit Ethernet port (mount flush to panel)
  - LEDs on the panel could be nice (27 ohms series current-limiting resistor, no GND connector on header)
  - 2x USB 2.0 (mounted flush to panel) (micro-USB)

- DDS board IOs (AD9912 eval board)
  - 4 SMA outputs
  - Clock input shared with FMC108 through an internal splitter
  - USB

- Need a fan for the FMC108/FMC104 board.

LDO: Low dropout regulator.

4 BNCs { • Power supplies for DDS board: Need 4 LDOs, maybe fed by 5V?
  - 2x 1,8V rails → DVDD: 1,8V 250mA; → AVDD: 1,8V 150mA
  - 1x 3,3V — VDD_DAC3: 3,3V, 31mA
  - 1x Switchable 3,3V (manual switch ok) → VDDX_DRV: CMOS output driver spec says 10mA with no load. Can it drive 50Ω? If so, need $\frac{3,3V}{50\Omega} = 66mA$ }