Guide to the 2019 Leakage Monitor

Accurate as of September 2019 for revA, by Alex Hodges (resident breadboarder extraordinary)

Introduction

Version three of the CCM leakage monitor allows large improvements in precision and accuracy. Time resolution is improved to around 0.4ms, current resolution to 0.25pA, power supply is simplified to a single +/-15V Molex connector and a battery change every 2 years, and data acquisition is integrated, exposing a single DC voltage output proportional to leakage current.

Description and Explanation

The ammeter has two parts. The transmitter side floats at high voltage, measuring current then encoding it into a frequency modulated square wave and transmitting it down to the the lab, where the modulated square wave is reconstructed and demodulated.

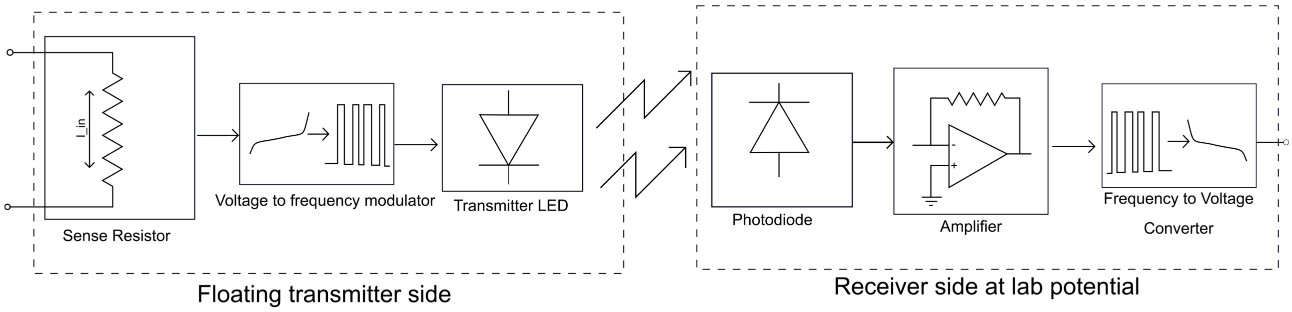


Figure 1: Block diagram of full ammeter showing data flow from measurement to outside world data connection.

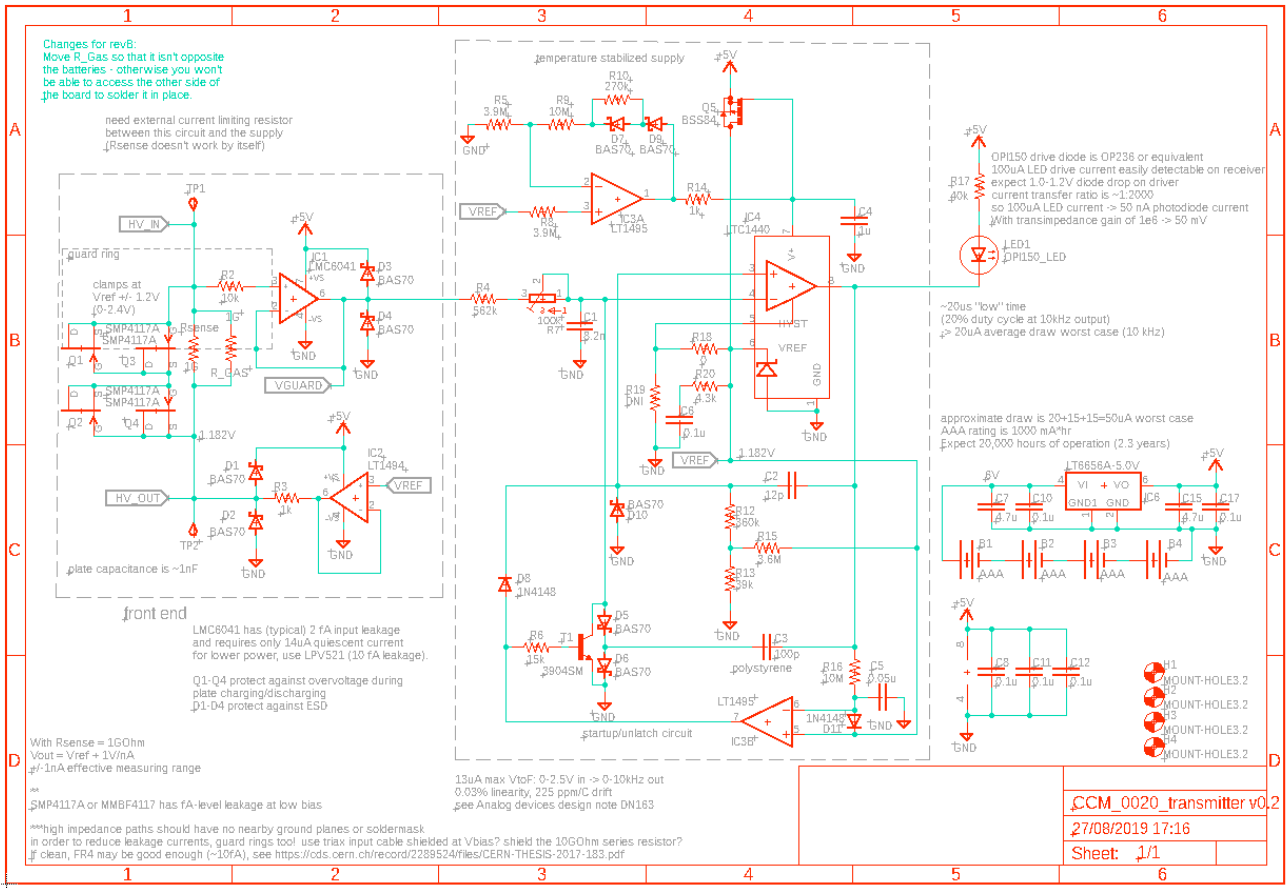


Figure 2: Transmitter side showing specific parts used, notes as well as bypass capacitor and power supply details.

Figure 2 shows the floating transmitter. The current being measured enters through the HV_in connector. It flows through R1, the sense resistor, before leaving via HV_OUT. Ideally no current flows through the diodes or R_GAS which are there as input protection. Since the currents being measured are from a high-impedance supply, we don't need to worry about R1 causing a fall in current since the voltage source has an equivalent resistance of more the 10 gigaohms.

The voltage at HV_IN is created at R1, and then buffered by IC1 configured as a simple voltage follower. IC2 is also configured as a buffer – it forces the HV_OUT voltage to be equal to a local voltage of 1.182V. This can also be expressed as saying that it forces the local ground to be 1.182V less than HV_OUT. Either way, this is what makes the ammeter floating, as all voltages are measured relative to this "ground", which itself moves relative to the lab depending on the voltage of the HV supply.

IC1's output is fed into the comparator inverting input at IC4. The non-inverting input is mostly high, and the output is mostly on. As the voltage to the left of C2 rises as it charges, D10 starts to conduct, discharging C2 and bringing the non-inverting input low, switching the output off. C2 stops discharging, D10 stops conducting, the non-inverting input becomes well above ground again, and the output quickly turns back on until C2 has sufficiently charged for the cycle to repeat. This leads to a high-duty cycle square wave.

IC3 is extra engineering to ensure linearity. IC3A takes the temperature-stabilized voltage source, V_ref, and converts it to a 4.3 V supply to power IC4. During startup, V_ref requires some current to begin working as it is powered internally, but the power supply for IC4 is controlled by V_ref. This circular dependency means V_ref can get stuck at 0.9V, where it does not have the power to startup fully. Q5 detects when IC4 is under-powered, and provides current to startup V_ref, switching off again when IC4 is powered up. IC4's output is also passed through an integrator (R15 and C5). This normally provides a high signal to IC3B's inverting input, higher than the 1.182V at the non-inverting, so the output is low, and T1 does not conduct. If IC4 stops oscillating, after a short time the inverting input will fall, and the output switches on. This discharges IC4's input, bringing it to ground by bypassing D5 and D6.

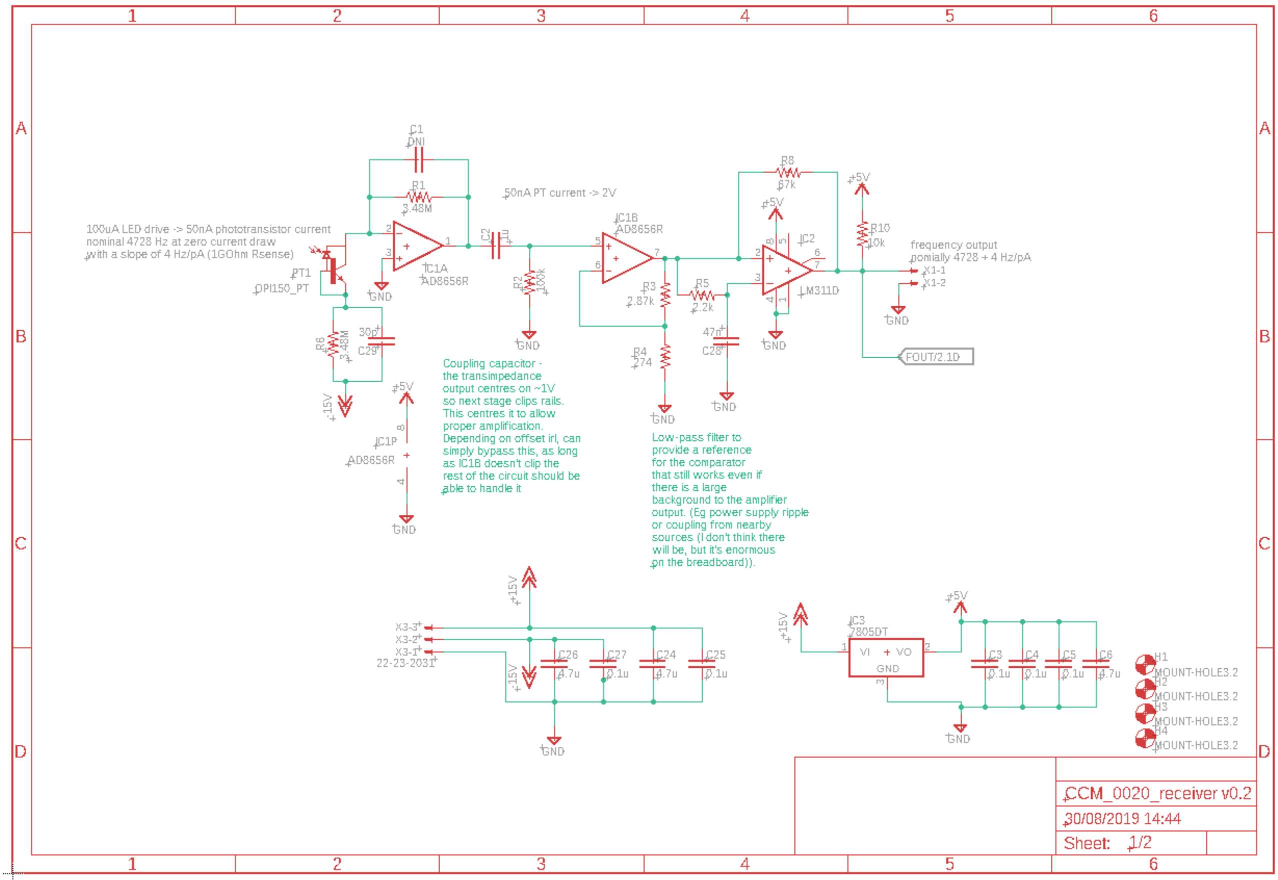


Figure 3: Signal reconstruction. These ICs reconstruct the transmitted signal, expose it via a Molex connector, and pass it on for demodulation.

Since IC4's inverting input is suddenly lower than the inverting, IC4's output switches on again and it starts oscillating like normal.

Data is transmitted out with the comparator output connected to the cathode of an LED. When the cathode is low, the LED feels a voltage across it and emits light. The high duty cycle means that the LED is only on a small fraction of the time, minimizing power consumption. The 40k resistor also limits current and therefore power.

The LT6656 stabilizes the voltage supply from the battery, while C4, C8, C11 and C12 are bypass capacitors for each IC.

On the lab side (Figure 3), the LED's light is detected by a phototransistor. The base and emitter are tied together, meaning it behaves like a photodiode. Tying these to -15V increases linearity while C29 decreases distortion.

IC1A is configured as a transimpedance amplifier – pin 2 is held as a virtual ground, allowing the photodiode to sink current easily and not be distorted by having to drive current. This sunk current is provided by the output, which flows back along R1, meaning the output voltage is equal to the input current * R1. (Photodiodes can be considered as nearly ideal current sinks). Therefore pin 1 is a voltage signal that ideally recreates the square wave input to the LED.

In practice there is significant ringing, which C1 cancels, large distortion, and a large DC background. Coupling capacitor C2 removes this, and has been chosen so that it will filter out a 50HZ background signal. IC1B then amplifies this signal to a level of around 100mV. This signal is low-pass filtered to generate a reference voltage at the mean of IC1B's output as there is typically still some offset. IC2 is a comparator which then reconstructs the input, which passes to the next stage for demodulation.

Note that this is how the breadboard works – the PCB designs used a fixed reference as they have a much nicer signal to play with.

IC3 generates a 5V signal from the 15V power supply, and C3 and C6 smooth it, while C4 and C5 are bypass capacitors.

Most of the rest of the PCB is used up by a phase locked loop (PLL), which acts as a frequency to voltage converter, and which we use to demodulate the square wave.

The two flip-flops, associated logic gates, T1, T2, and C7 act to control the VCO.

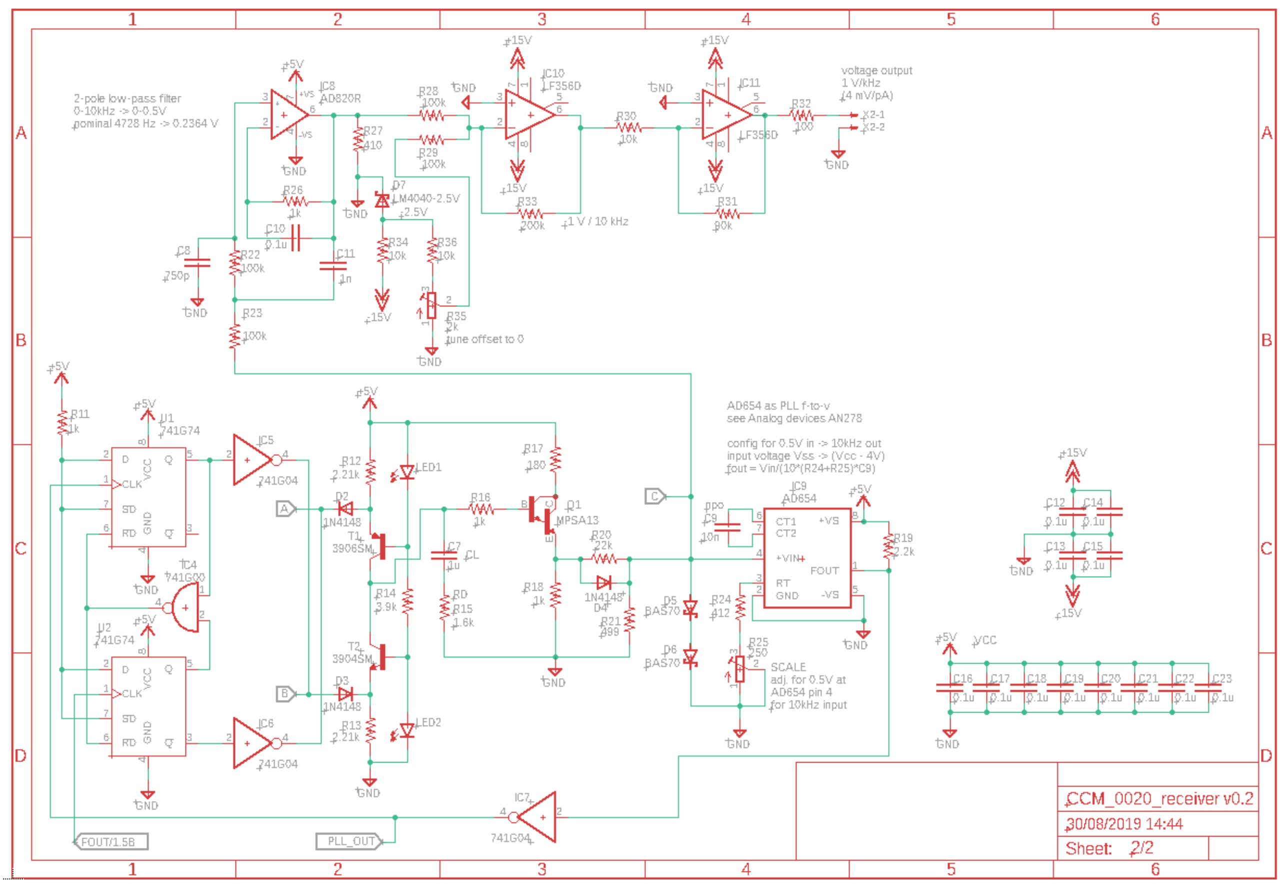


Figure 4: Part 2 of the receiver side. This acts as a frequency to voltage converter and then scales the output for maximum dynamic range.

The charge stored in C7 creates a voltage on the VCO's control pin which determines its frequency, while the flip-flops detect if the VCO output is drifting in phase from the input, and either charge or discharge C7 appropriately.

My mental image of how it works is as follows. Assume that PLL_OUT leads F_OUT. U1 accepts PLL_OUT and generates signal B. B goes high when PLL_OUT rises. When F_OUT later goes high, U2's Q pin goes high, causing the AND gate to trigger the reset pin and drive Q low. So signal B is a pulse train, where the pulse width equals the phase difference of the two waves, at the frequency of PLL_OUT.

U2 accepts FOUT and generates signal A. When its Q pin goes high, U1's Q pin is already high, (since FOUT lags PLL_OUT), and the reset pin triggers within the gate delay of the AND gate. This causes signal A to be a pulse train with a pulse width of 50ns, at the frequency of FOUT. If PLL_OUT instead lags FOUT, the roles reverse. Either way, exactly one signal has a time spent high proportional to the phase difference.

The two transistors T1 and T2 are biased permanently on by the LEDs. Signal A is connected to T1's emitter. If A is low, current flows through T1. If B is low, current flows through T2. So if both currents flow at once, no net current goes into C7, but if only T1 has current flowing, it must flow into C7 and charge it, and if only T2 has current flowing it must source that current from C7, discharging it.

How does this lead to a stable equilibrium? When the PLL is in lock, it has a phase difference of zero, so signals A and B are identical, and C7 stays at a constant charge. If the outputs become difference, they start to drift in phase, and C7 responds until they are back in sync.

C7's voltage is buffered by the Darlington pair, which then controls the VCO and is our demodulated signal.

There is some noise on the PLL, so IC8 filters our demodulated signal with a 2-pole filter. The scaling of the AD654 VCO doesn't use our voltage range efficiently and will reduce the precision of the ADC that will ultimately be used, and so IC10 and IC11 scale the output voltage to be centered on 0V for 0 input current, and to use more of the available +/-15V range. This output is finally exposed via a Molex connector, and is a voltage proportional to the current on the transmitter side.

Board Layout Features

To minimize leakage there are a couple different physical design features on the transmitter PCB. Consider why these are needed. On a PCB there is typically a 100MΩ impedance between traces. If these traces have 1V between them, this allows 10pA of current to flow, which is enough to completely swamp any signal that we could try to measure. We need to stop this. There are three main techniques.

Most importantly is the guard ring, the wide copper band around the HV_IN connector. This is at the same voltage but it buffered (we take the IC1 output). So, the voltage between any traces in this ring is exactly zero – no leakage current should flow.

Next, the via fence in the guard ring has a double benefit. It stops leakage through the FR4 of the PCB itself, acting as a 3D guard barrier, as well as shielding against electromagnetic interference.

Lastly, the lack of silkscreen or soldermask within the ring helps reduce the risk of charge accumulation on the surface.

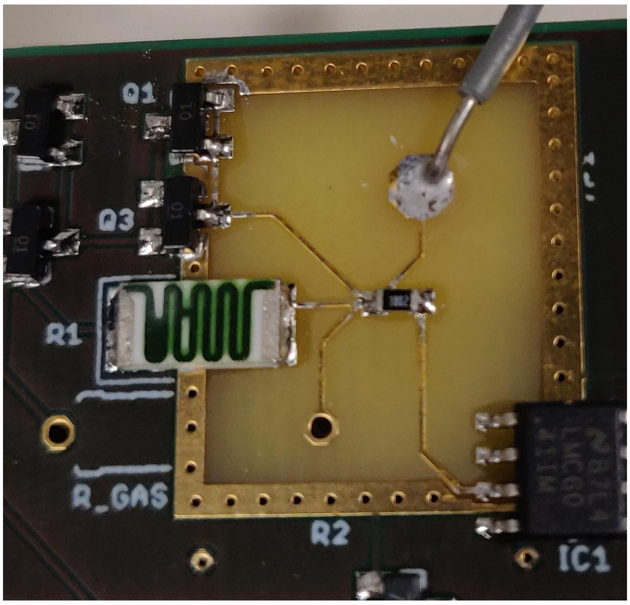


Figure 5: Detail of the HV_IN section of the transmitter PCB, showing anti-leakage design features.

Datasheet

Input range

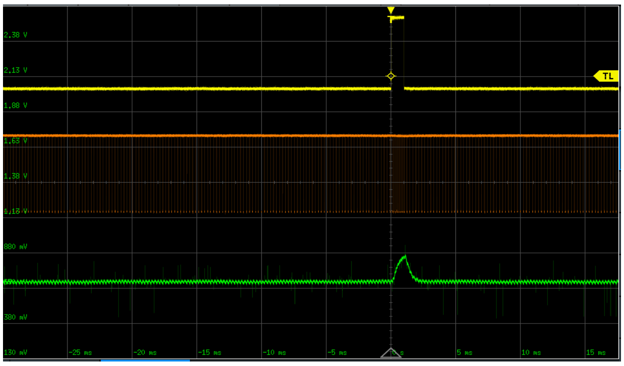
The current range is governed by 3 things: the sense resistor (R1), the range of the VCO and the range of the PLL. (The comparator should be able to handle everything fine).

The transmitter VCO is scaled for 0-10kHz from an input voltage of 0-2.5V. The zero current point creates a voltage of 1.182V; the sense resistor limits the scaling, while the VCO range determines the measurement range. In theory this allows you to measure current from -1.18nA to 1.32nA.

In practice, the PLL struggles to lock onto anything below roughly 3.5kHz. This limits us to a range of -430pA to 1.182nA.

Time resolution

The smallest visible feature is around 0.3ms, using the 3dB bandwidth. In practice, this allows you to confirm that a current spike happened, but the diminishing gain at those speeds means you'll have a hard time saying the magnitude of that spike. Staying within the linear region, you have a time resolution of 2ms where you don't have to worry about the gain-bandwidth product biting you.



Drawing 1: Step-response of ammeter. The yellow trace is the current in - a 0.4ms pulse. The orange is the comparator output (lines are thin, zoom in if it just looks flat). Green is the scaled output - a response is clearly visible.

Current Resolution

The counterpart to time resolution is current resolution. For a constant input current, how stable is the output voltage? The error on your measurement will obviously fall with measurement time as you collect more data points. Currently with a measurement time of ~40ms, you get can measure current with an accuracy of ±0.25pA. Much longer doesn't seem to add much benefit since that's the noise floor of the system. Shorter timescales add error – choose your bin size for the data collection with this in mind.

FAQ

Q: How often should I change the batteries?
A: About every 2 years according to the rated capacity of AAAs. My prototype is running off supposedly "dead" batteries since they were all I could find in the lab. So possibly significantly longer will work fine. We haven't included a power-good signal though, you'll just have to probe pin 6 of the voltage regulator and check whether it's getting enough power to regulate.

Q: What interface do I need?
A: You'll need a three-pin molex carrying GND/-15V/+15, and a pair of two-pin molexes to get the recreated wave and the DC voltage out.

Q: Why doesn't the transmitter side buffer (IC2) cause extra unmeasured current to flow, worsening the leakage?
A: Conservation of current means that the same amount of current flows into the transmitter side as leaves – there can't actually be any extra current added by IC2. This isn't to say that leakage isn't the problem – current could flow between HV_IN and HV_OUT without passing through R1 and being measured.

Q: What do you need the op-amp in the transimpedance amplifier for? Couldn't you just use a resistor given that current will flow anyway, then V=IR will create the voltage for you?
A: Resistors are the most primitive I to V converters, and that's how we convert our pA current into a value to be buffered on the transmitter. However, for a low-impedance source they can distort the signal being measured, as the current source can't drive that resistance.

Q: The low-pass filter in the receiver introduces a phase delay – doesn't this mean your comparator reference won't trigger properly for a fast background signal?
A: You're absolutely right – it's tuned to *just* work for 50Hz at the moment, if you're American or have other high frequency ambient noise, you might have problems. They would manifest themselves as the comparator generating a wave at the right frequency, but only about half the time. Check the troubleshooting guide.

Q: This is all black magic and I'm amazed any of it works.
A: Me too. Now imagine persuading it all to work on a *breadboard*. There were lots of completely inexplicable problems that needed to be fixed.

Modifications

Gas Discharge Tube

The board provides two holes to mount a GDT to help stop the board getting damaged by current transients. However, this seems to leak about 5pA and adds noise (although possibly some of that could be removed by having it cleaner – we didn't check yet).

3D HV Shielding

It would help to have a guard-potential box surround the HV_in node to reduce leakage into the air – there are plans for such a box floating around, you'll need to cut and fold it to shape.

Cleaning

Dirt and oil can provide a low-impedance path that will leak pA – you need to keep the board very clean. Someone needs to make a proper cleaning schedule, and if you package this thing up there needs to be a desiccant pack in there (if it's not in a dry environment already).

Troubleshooting Guide

Problem	Solution
The output square wave is a constant ~40Hz independent of input current.	You have exceeded the maximum input voltage to the transmitter side VCO – reduce input current or decrease R1 (the sense resistor). (Less likely: VCO has lost power. Try replacing the batteries)
There is no output wave on the receiver board.	Most likely the low-pass filter leading to the comparator has too low a bandwidth, and the phase delay between its signal and the background of the transimpedance output is too long. Decrease the capacitor or resistor value.
The comparator output oscillates in phase at the right frequency but skips waves.	Most likely your hysteresis resistor is too small, increase it. Typically 67k works well. Unlikely, but possibly as in previous problem the phase between your -in pin and the +in background is too large. Decrease the filter capacitor.
The comparator output oscillates in phase but has occasional extra waves.	Almost certainly your hysteresis is too large – decrease it.
The comparator output is a nice square wave but the DC output does not change with input.	Your PLL is not locking on properly – possibly you are trying to measure a current that is too small and falling below the ~3kHz locking range. Either recalibrate the transmitter side with a smaller sense resistor, or change C9 for a larger value to expand the locking range. (20nF works well with slightly slower reaction times).
The transimpedance amplifier output is very distorted and is affecting the later stages.	Probably the capacitance values on the transimpedance amplifier were well suited to the breadboard it was designed on, but aren't so great for the PCB. Roughly speaking, the capacitor across the transimpedance feedback controls oscillation, while the capacitor from the OPI150 emitter to -15V controls distortion. The former is likely too small and the

	latter too big if you're seeing significant distortion.
The gain stage before the comparator holds a constant +5V output rather than oscillating.	This usually happens when the transimpedance amplifier's output has some large DC offset which is causing the gain stage to clip the rails. The coupling capacitor is meant to prevent this, but a high frequency DC offset might not get filtered out. Try reducing the value of R2, the resistor to ground. If that doesn't work, turn down the gain on the IC1B – the comparator should still be able to handle it, there's some margin built in.
You're trying to listen to the transmitter output directly from the transmitter board but the waveform is very distorted OR is much too fast (~15kHz) and does not respond to input current changes.	Attaching an oscilloscope probe to the OPI150's LED's cathode is tempting and a good way to check whether it matches the comparator output. However, the transmitter side ground is not equal to the receiver or oscilloscope ground. The middle pin of the trimpot does allow you to connect a wire to transmitter ground, but doing so alters the VCO frequency quite severely and makes it less sensitive. The oscilloscope will still trigger off a distorted VCO input, it's best to just not connect a transmitter ground and live with it. Or do the mad scientist approach – float your oscilloscope at transmitter ground as well.