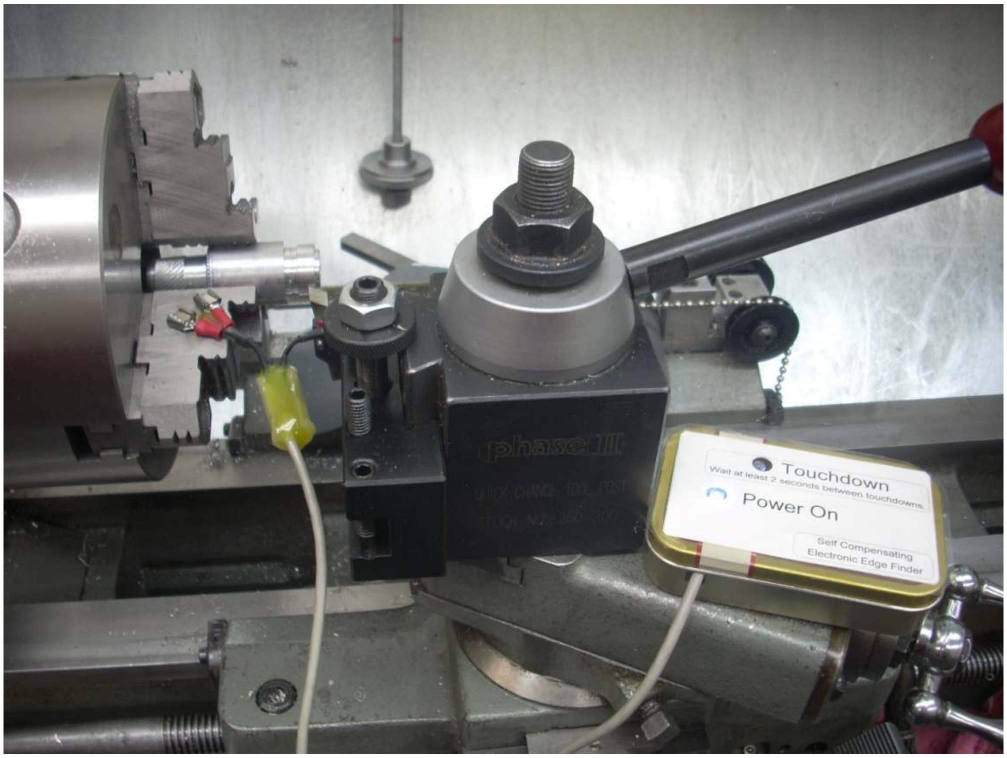
A Self Compensating Electronic Edge Finder, version 6.0

By R. G. Sparber

Copyleft protects this document.[1]

Purpose



This article describes an Electronic Edge Finder which connects directly to the cutter and spindle on a lathe without any modifications to the machine. It can detect when the cutter comes in contact with the work piece.

While the common EEF can tell the difference between an open circuit and a resistance less than around 500 ohms, this new EEF can tell when the resistance changes from 0.01 ohm down to 0.0098 ohms. In this way it deals with the low resistance of the machine rather than being insulated from it[2].

A test current of 20 milliamps, comparable to what is used with commonly available EEFs, is employed.

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[1] You are free to copy and distribute this document but not change it.

[2] If you own a lathe with a spindle to cutter resistance of greater than 2 ohms or a mill with a spindle to table resistance of greater than 2 ohms, then a simpler EEF circuit is available at http://rick.sparber.org/rctf.pdf.

The EEF can also be used on a mill. One probe is connected to the spindle and the other to the mill vise. When the edge of the end mill's flute comes in contact with the work piece, the EEF detects the change.

The EEF has been successfully tested on an Atlas/Craftsman 12" lathe, a RF30, a Shizuoka mill, a Monarch AA lathe, a Harding HLVH lathe, two different Bridgeport mills, and an Enco gear head lathe.

The circuit is reacting to a few millionths of a volt in order to detect touchdown. If a given shop has a high enough level of electrical noise, the circuit cannot function.

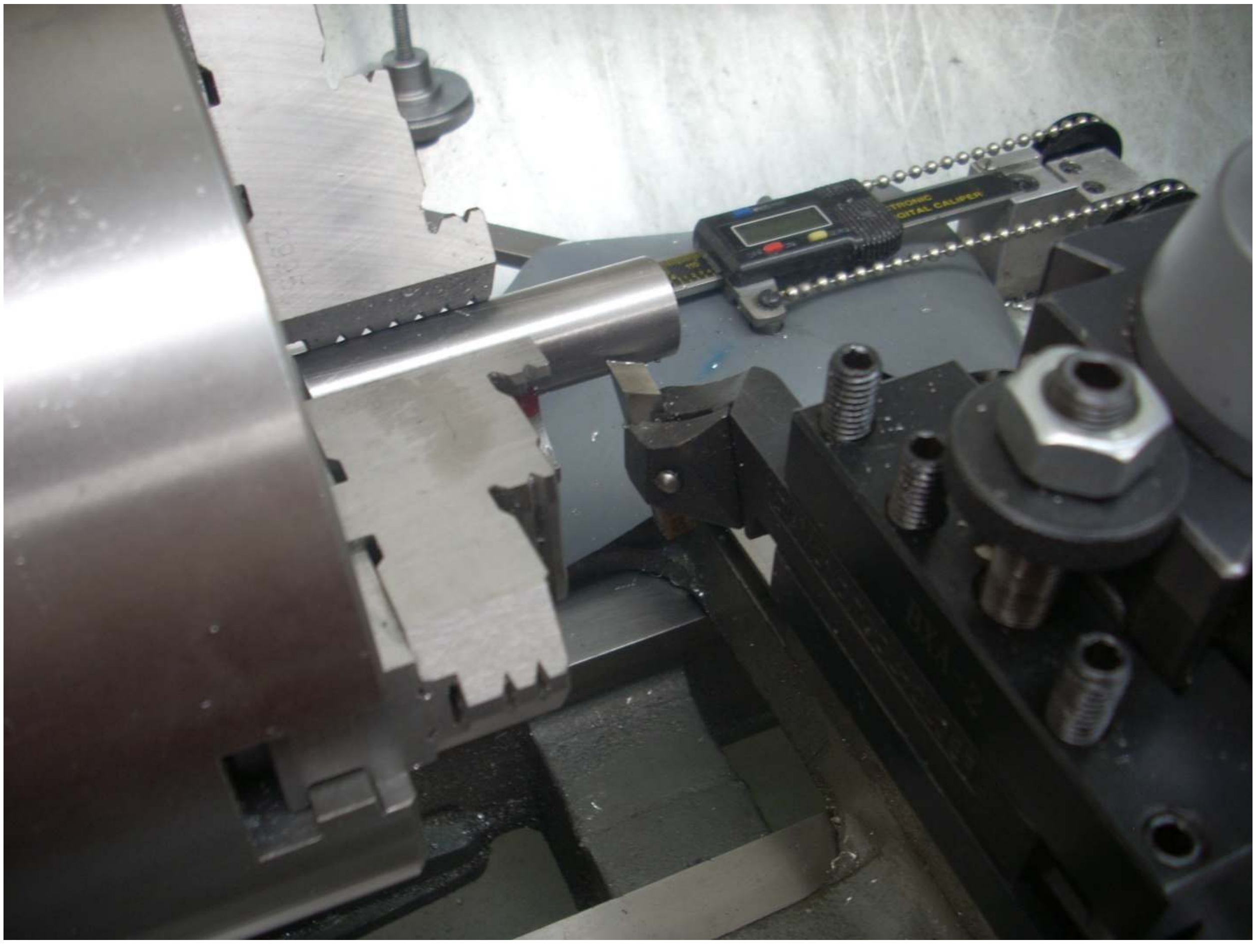
Contents

YouTube Video

You can see a YouTube video of this device being used on a lathe at

http://www.youtube.com/watch?v=XgFRIsJrA0s&list=UUQowQlSfFxybveyBDVOHXxw&index=1&feature=plcp

Background



You just finished turning the outside diameter of a work piece using a right hand cutter and now want to pick up the same surface with a left hand cutter. How would you do this?

There are many ways to set a new cutter down on a previously cut surface. The chosen method depend on the accuracy you want and your skill level. Some place a piece of thin paper between cutter and work piece and slide it up and down while feeding the cutter in. When the paper catches, you are the thickness of the paper away from touchdown. Now, if you are looking for high accuracy, then the amount that the paper has been compressed becomes an issue. The paper might be 0.003" thick and compress by 0.001" at the point of contact. That bothers me.

If I were to insulate the cutter from the rest of the machine, then a simple continuity checker can be used to detect touchdown[3].

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[3] See http://rick.sparber.org/LTEEF.pdf.

Adding insulation around a cutter is a hassle and for things like boring bars, hard to do.

Insulating the entire tool post was suggested by Scott G. Henion[4] as a universal solution. Since my goal was to not modify the machine in any way, I did not pursue this idea yet it is certainly elegantly simple.

The challenge for me was to build a circuit that is as easy to use as a simple continuity checker yet have it work without modifying the machine in any way. The problem at hand is not all that obvious so has been dealt with in another article[5].

One solution, proposed by Jerrold Tiers[6], is to place an AC powered magnetic field around the cutter. This induces a voltage in the cutter that will cause a current to flow at touchdown. Although I was unable to find the proper components to try out this idea, I have no doubt that it would work.

My solution is to focus on passing a DC current through the machine and look for the ***change*** in resistance at touchdown.

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[4] of the atlas_craftsman Yahoo group.
[5] See http://rick.sparber.org/ueef.pdf
[6] of the atlas_craftsman Yahoo group.

The controversy started when I discovered that the task was easy to do if I used 1 amp as my test current. Some felt that this could cause arcing inside the bearings and lead to pitting of the surface. Since it is impossible to prove that this is not the case even when the applied voltage is only 0.01V, I decided to develop another circuit that used only 0.15 amps. It works well but required the turning of a knob to compensate for variations in resistance. That bothers me.

So this time around I have developed a circuit that passes only 20 mA at no more than 0.7V through the machine and automatically compensates for resistance.

Since Electronic Edge Finders bought at Enco® put out about this much current and more voltage, I think I can safely say that any concerns about damaging the bearings can now be put to rest.

Some have suggested methods involving touchdown with the machine running. There is a "Catch-22[7]" here. The instant the cutter touches the reference surface, it will slice away the surface. So you will never touch down for more than an instant and then the reference surface has been damaged.

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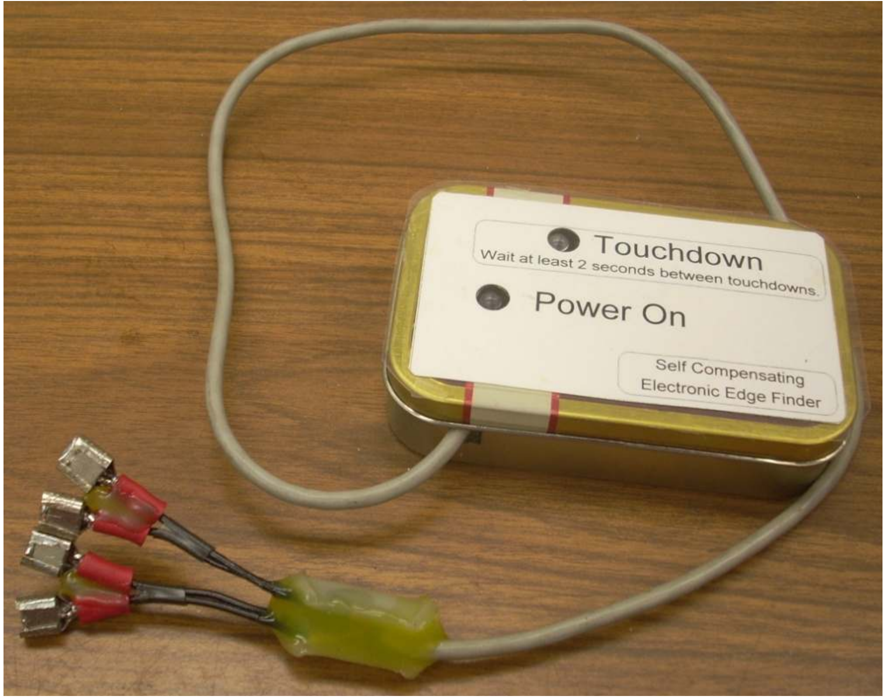
[7] See http://en.wikipedia.org/wiki/Catch-22_(logic)

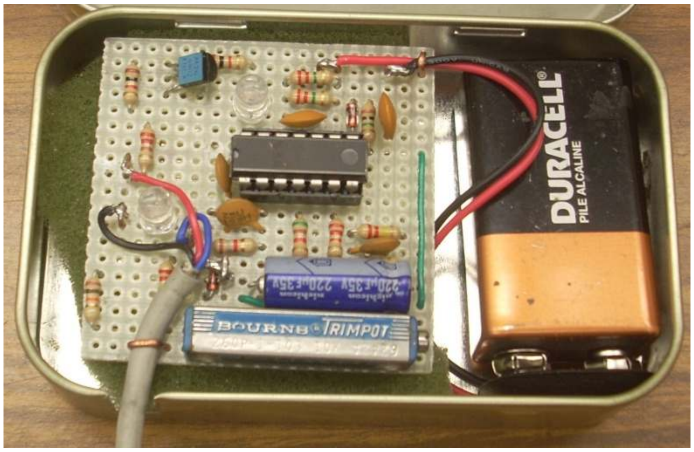
The Self Compensating EEF



When the probes are connected to the lathe or mill, the power-on light comes on. The circuit then measures the resistance it sees and stores the result. At touchdown, this resistance suddenly drops by as little as160 millionths of an ohm . This drop is detected by the circuit and the Touchdown light flashers.

Power for the EEF comes from a single 9V battery which supplies less than 25 mA only during use.

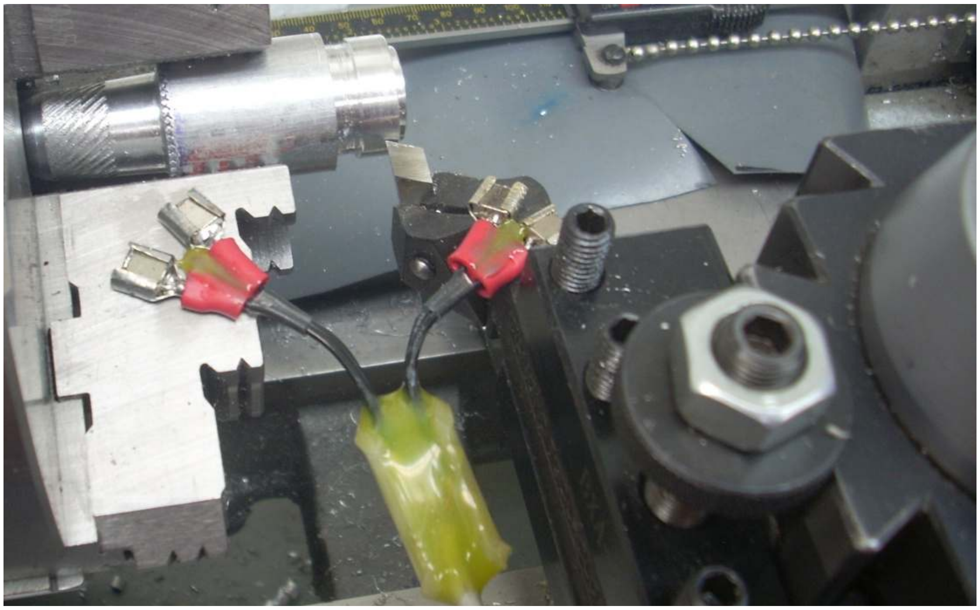
Magnets have been incorporated into the ends of the probes so an electrical connection is made by simply placing them on clean, steel surfaces.



The circuit uses one commonly available integrated circuit (LM324), one transistor, two diodes, two LEDs, 15 resistors including a trim pot, and 6 capacitors. It is powered by a single 9V battery.

There are two probes. Each probe has two magnetic clips. One probe contains the High Current 1clip along with the Signal Probe 1 clip. The other probe contains the High Current 2 clip along with the Signal Probe 2 clip.

EFF Operation on a Lathe



One of the probes connects to the jaws of the chuck. The other probe connects to the cutter holder.

Since both probes are now connected, the power-on light comes on.

When the cutter comes in contact with the work piece, the touchdown LED will flash for at least 0.1 seconds. If desired, you can back the cutter out and repeat the process to verify you have set the zero point correctly. Wait at least 2 seconds between touchdowns to give the circuit time to re-compensate.

When done, remove the clips from the machine and store them on the insulated top area on the EEF. This turns off power.

The circuit works normally for chuck to cutter resistances of at least 0.01 ohms and a touchdown resistance of less than 0.6 ohms.

Shop Experience

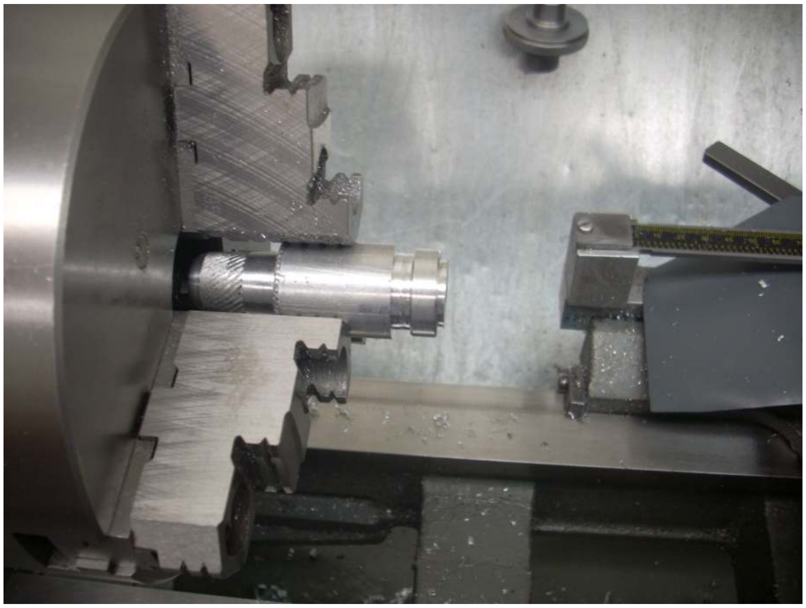
Normally I would take two or more cuts of equal in-feed, measure the result, take an average, and have a correction factor that will get me the desired Depth Of Cut. Then the next cut would use this correction factor at the same in-feed and the resulting finish cut would be as close as possible to idea. This procedure compensates for time invariant error but nothing can compensate for time variant error which is random.

I see the EEF being used in a different situation. To start, I would make one or more calibration cuts which begin at touchdown each time. These cuts would tell me the average DOC for a given in-feed.

Notice that in the first case my cut includes the initial and final spring of the cutter. In the second case, using the EEF, my cut only includes the spring at the end of the cut. Initially there is no spring because the cutter is just barely touching the surface.

I'm now ready to start machining. I touchdown using the EEF and set my zero. When I feed in by the calibrated in-feed, I should be able to predict the actual DOC to within the accuracy of the lathe.

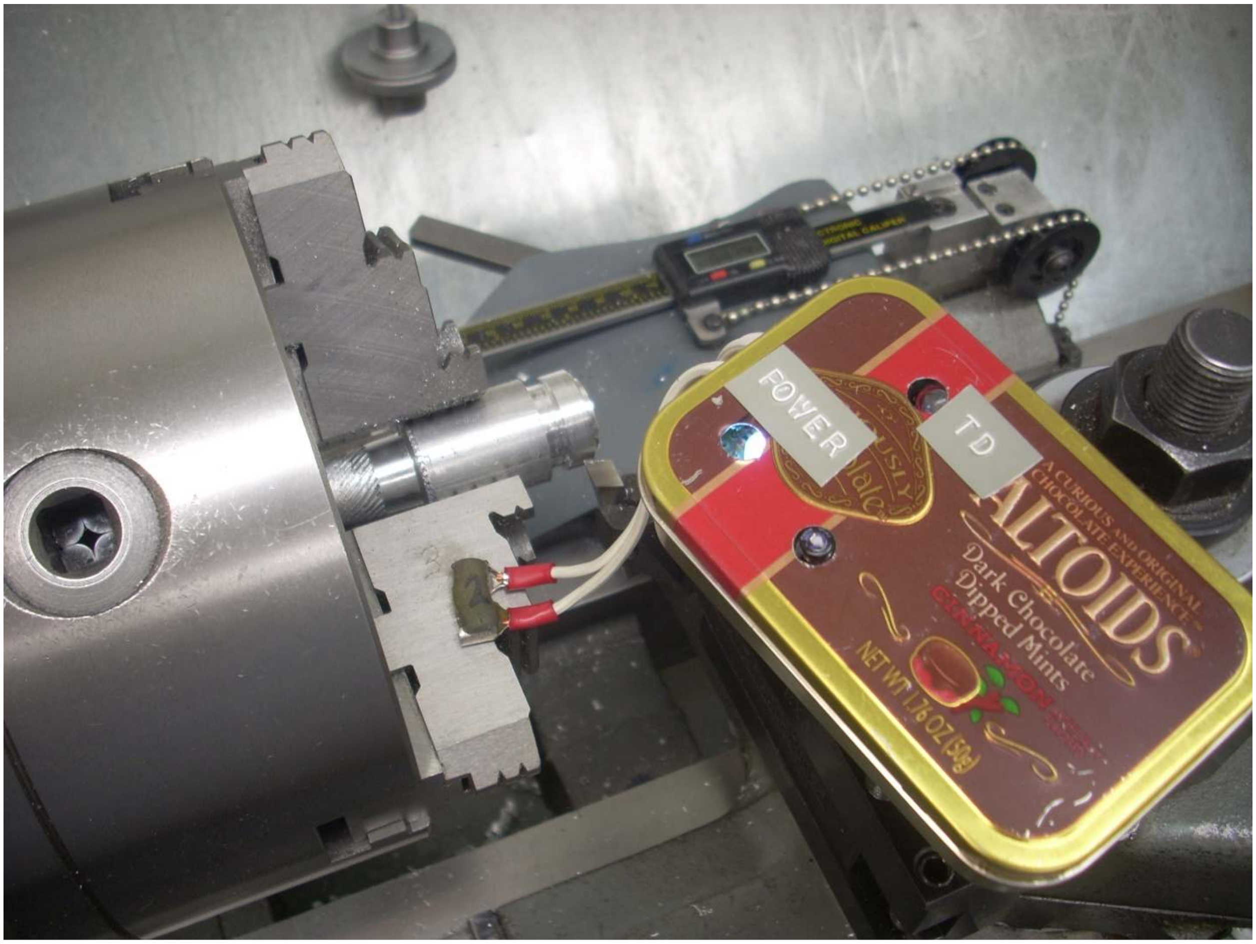
The calibration must be done under the same conditions as the cut just performed.



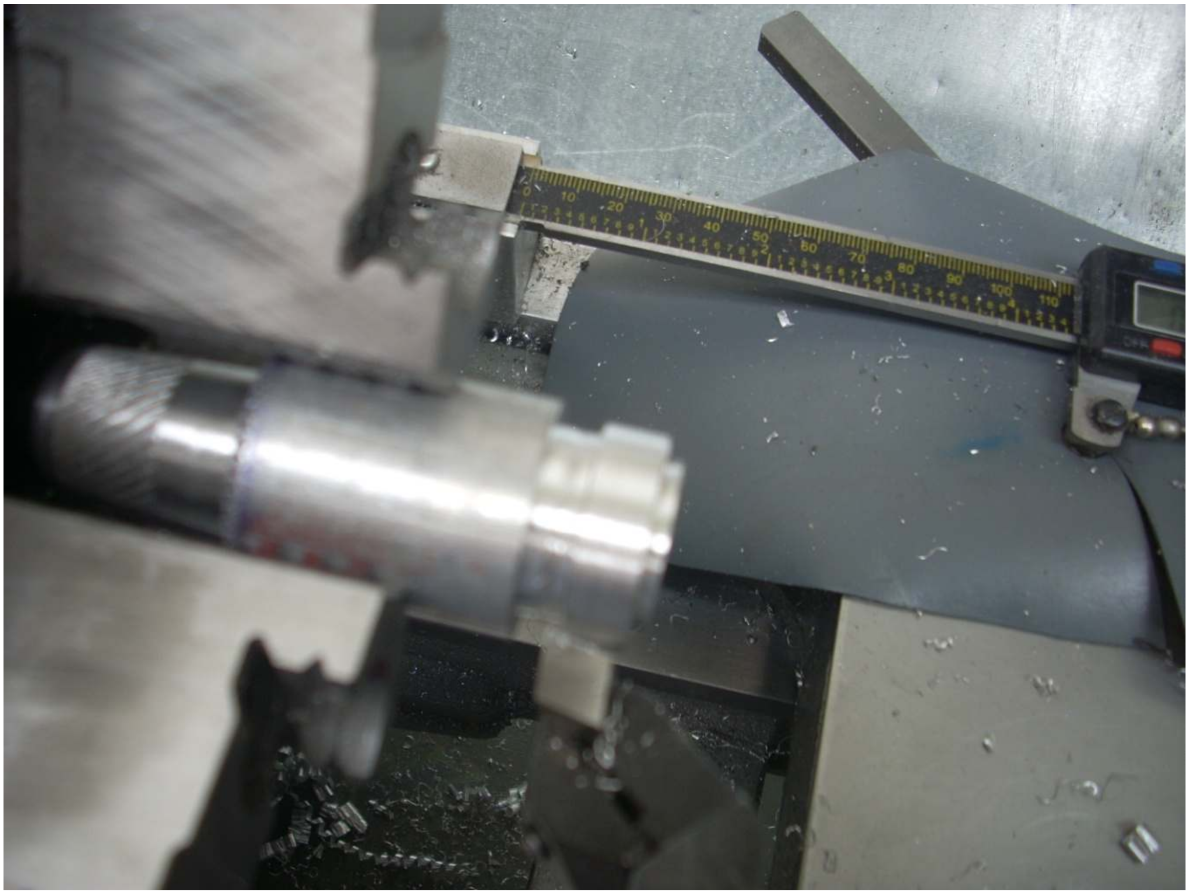
I started by turning a test piece out of 6061 aluminum. The lathe is a 12" Atlas/Craftsman in very good shape (IMHO).

The corners of the test piece have been deburred and then cleaned with alcohol.

The cutter was also cleaned with alcohol. All surfaces are then dried.



Then I used my EEF to set my in-feed dial to zero at touchdown. With the EEF a safe distance away, I fed in 0.005" on my cross feed dial.



I made one pass with the cutter using the power feed. Then the cutter was backed away, the test piece was deburred, and both test piece and cutter were cleaned with alcohol and dried.

I measured the Outside Diameter of the test piece using my Mitutoyo IP65 digital mic. The reading was taken after turning the clicker enough to make 3 clicks. This provides a consistent torque on the spindle so consistent force between mic and test piece.

I then repeated the process by first establishing a new zero on my in-feed dial. Then feeding in 0.005", taking a cut, and measuring the resulting diameter.

Here is the raw data plus reduction to average and deviation:

Pass	mic'd diameter, inches	change in radius, inches
1st ref	0.91940	n/a
1	0.91190	0.00375
2	0.90445	0.00373
3	0.89815	0.00315
4	0.89070	0.00372
5	0.88290	0.00390
6	0.87675	0.00307
	average =	*0.0036*
	deviation (+/-) =	0.0003

In each case I fed in 0.005" and got a Depth Of Cut of 0.0036" ± 0.0003.

Now, the challenge during machining is to get as close to a final DOC of 0.0036" as possible. Then the correction factor will be as accurate as possible.

Note that if you calibrate all of the cutters you plan to use, you can change cutters and maintain the same accuracy. It is important to understand that this correction factor depends on all parameters of the cut being the same. It also depends on the cutter not dulling "too much". At some point you will have to at least recalibrate and later sharpen the cutter.

The above procedure does not replace the standard method of taking repeated cuts to get to a given diameter. It is just another "tool in the box" to solve the sticky problem of starting with a finished reference surface.

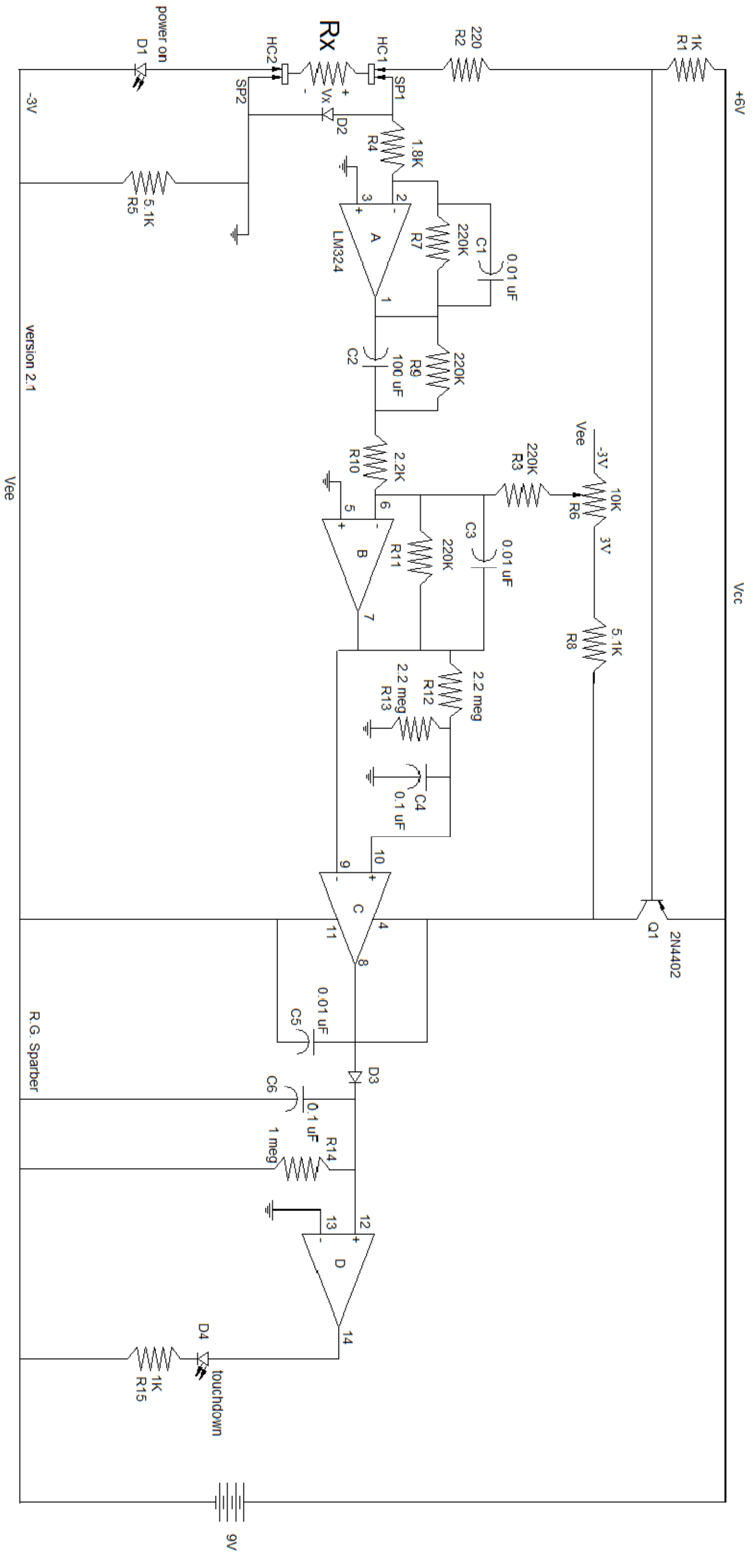
Schematic

Here is the full schematic. A magnified copy can be found starting on page 63.

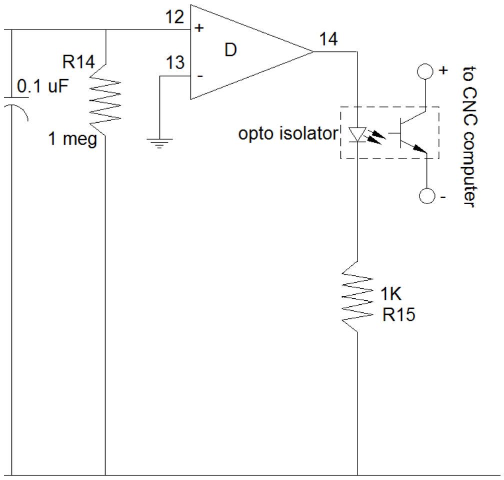
Some readers may want to just build the EEF and are not interested in how the circuit works. Others will want to know more details.

For the electronics geeks in our ranks, I will give details for each functional block. Then I will put these blocks together and deal with how it works.

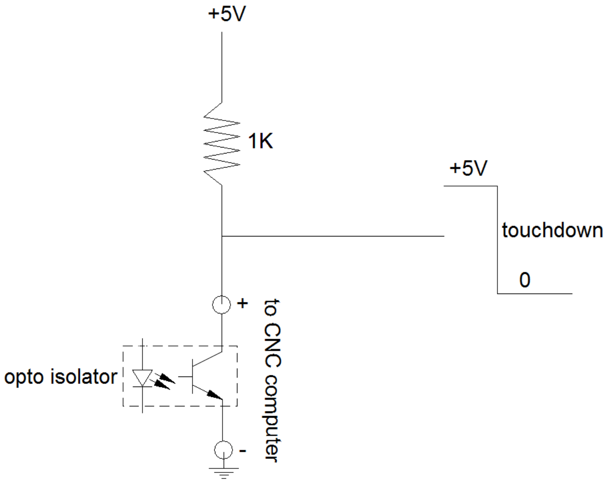
Note that you may need to view this schematic at a higher zoom level to see all lines.



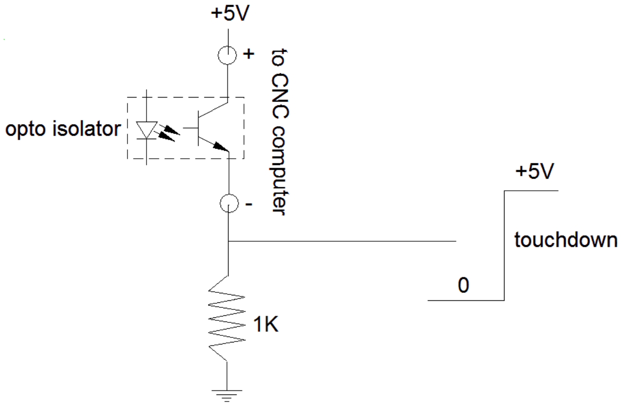
Interfacing with a CNC System



You can substitute an opto isolator for D4 so the EEF can safely interface with a CNC system. At touchdown, you would get conduction through the opto.



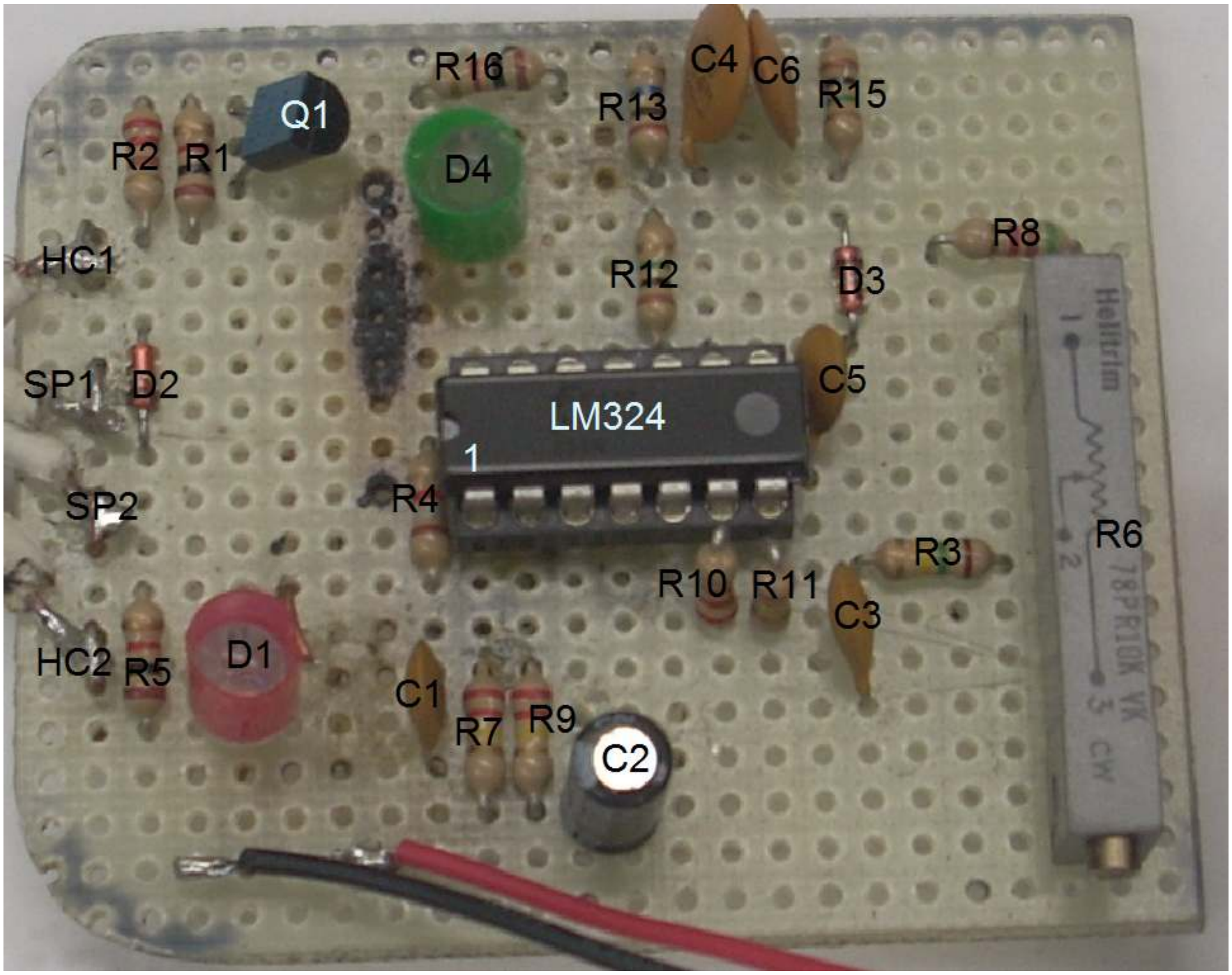
If your computer needs to see a high to low transition at touchdown, connect the - lead of the opto to the computer's ground. Connect the + lead to a 1K resistor. The other end of the resistor goes to +5 provided by the computer. Then connect the + lead to the input port so the software can see it.



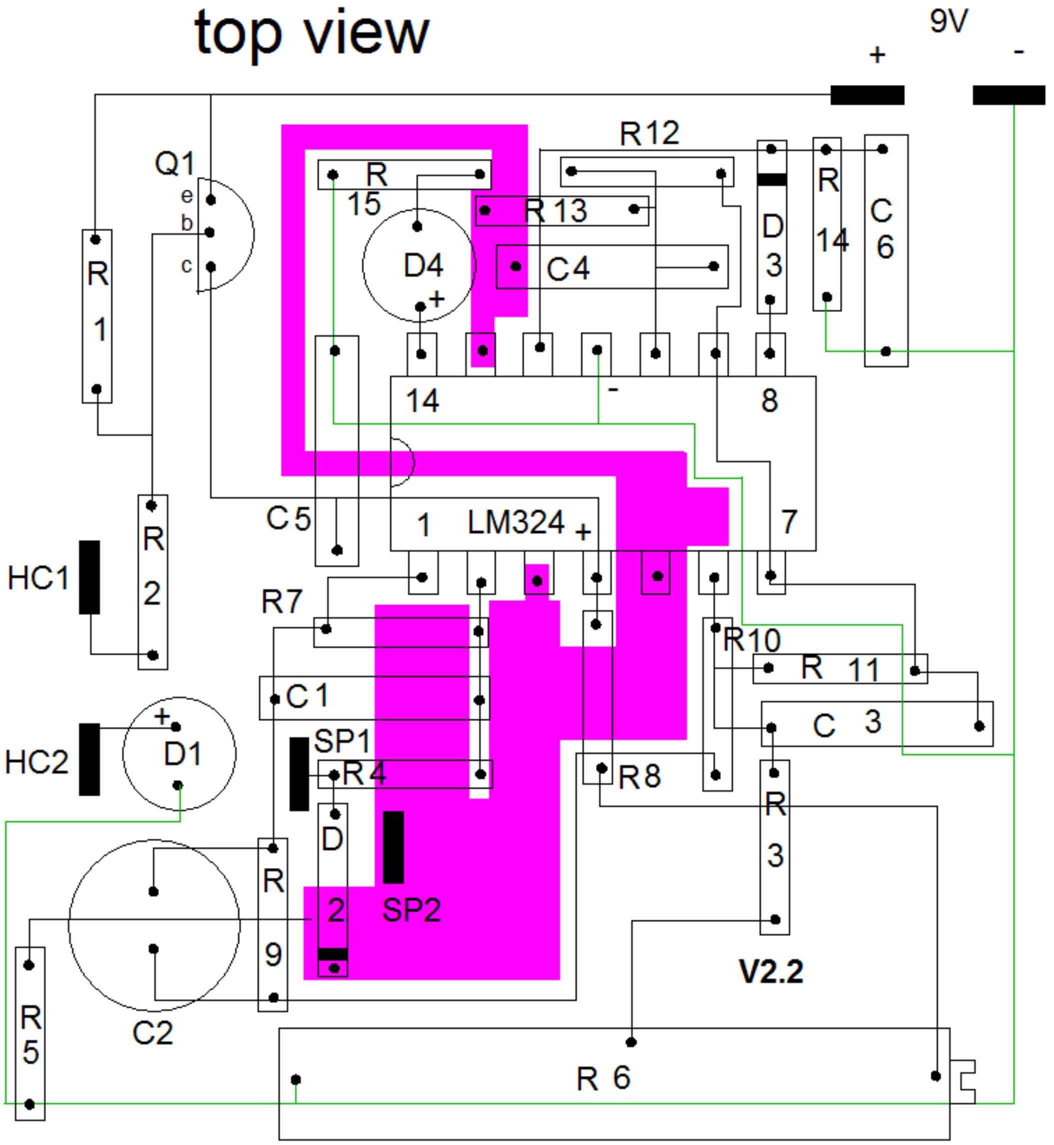
If your computer needs to see a low to high transition at touchdown, connect the + lead of the opto to the computer's +5V. connect the - lead to a 1K resistor. The other end of the resistor connects to ground. Connect the - lead to the input port so the software can see it. This may not work if interfacing to TTL but should be fine for CMOS. See your interface specs.

Parts Placement

This is not the ideal placement of parts but is how I did it in my first prototype. It shows you what the various parts look like.



My R6 is larger than the trim pots commonly found today. You will see that the circuit board artwork uses the more compact style.



This layout can be done as a circuit board with copper on only the bottom side. The purple areas are ground and are designed to quiet the most sensitive nodes. If you wish to go double sided copper, make the top layer all ground. Thermal breaks[8] around all connections to ground would help during the soldering process.

To avoid problems, I suggest you measure each resistor before soldering it in place.

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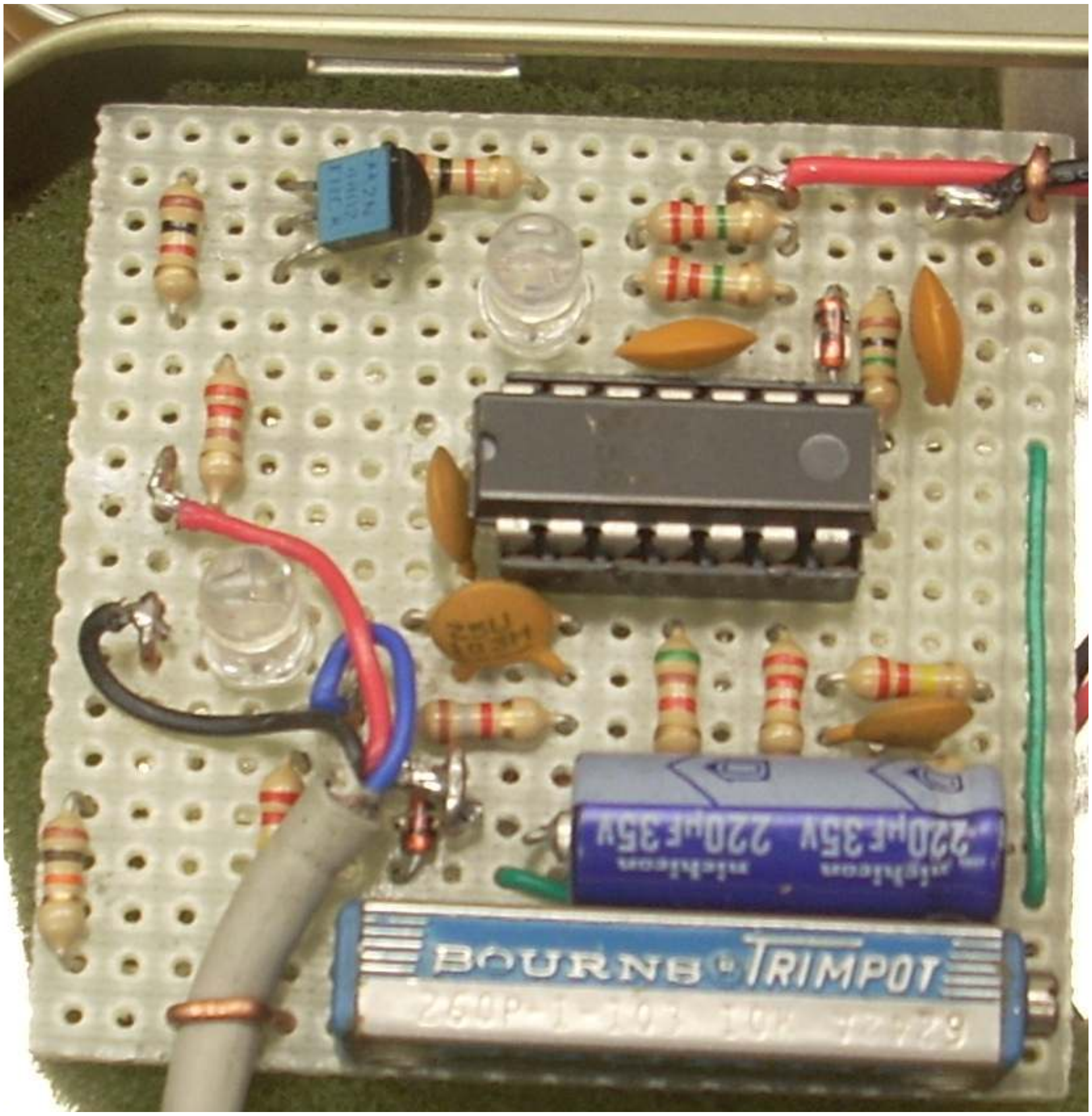
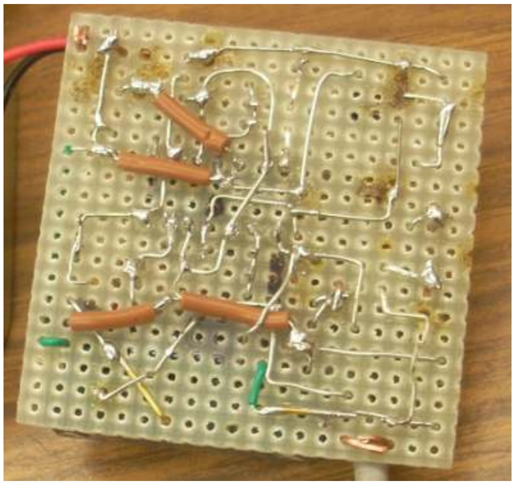
[8] A thermal break looks like a miniature 4 spoke wheel. It provides an electrical connection while reducing the amount of heat needed during soldering.



This is a hand wired version which uses my suggested parts placement except for C2 and R3.



During the prototype stage, I can't afford the time or expense of etching a circuit board. Instead I use a wiring method called "point to point". I have followed the layout shown on page 17 but use wire for my paths. I let some wires cross since it is more mechanically stable than routing long distances around other wires.

Parts List

Resistors (all 5% ¼ watt)

Value	*Quantity*	*Places used*
220	1	2
1K	2	1, 15
5.1K	2	5, 8
1.8K	1	4
2.2K	1	10
ten turn 10K trim pot	1	6
220K	4	3, 7, 9, 11
1 Meg	1	14
2.2 Meg	2	12, 13

Capacitors

Value	*Quantity*	*Places used*
0.01 uf ceramic	3	1, 3, 5
0.1 uf ceramic	2	4, 6
100 uf 16V non-polarized electrolytic	1	2

Diodes

Description	Quantity	Places used
Super Bright white LED	2	1, 4
general purpose signal diode	2	2, 3

Transistor: 2N4402 (or any other general purpose PNP)

Integrated Circuit: LM324 (a quad op amp)

Clips: see page 21.

9V battery connector

9V battery

Altoids® box (minus the candy…) or some other metal enclosure (not plastic)

Perforated circuit board material or etched circuit board.

Trim Pot Adjustment Procedure

In order to maximize sensitivity to touchdown while minimizing sensitivity to electrical noise, we need to connect the circuit up to the lathe or mill.

On a Lathe

Degrease a small patch of the ways. Then place all connectors on this patch. Adjust R6, the trim pot, until the touchdown LED just stops flickering.

Move the probes so one is on the cutter or tool post while the other is on the chuck. Put a metal bar in the chuck. Wait a few seconds and then do a trial touchdown. The touchdown LED should flash for at least 0.1 seconds. If the touchdown LED flickers, slightly adjust the trim pot until it remains dark.

On a Mill

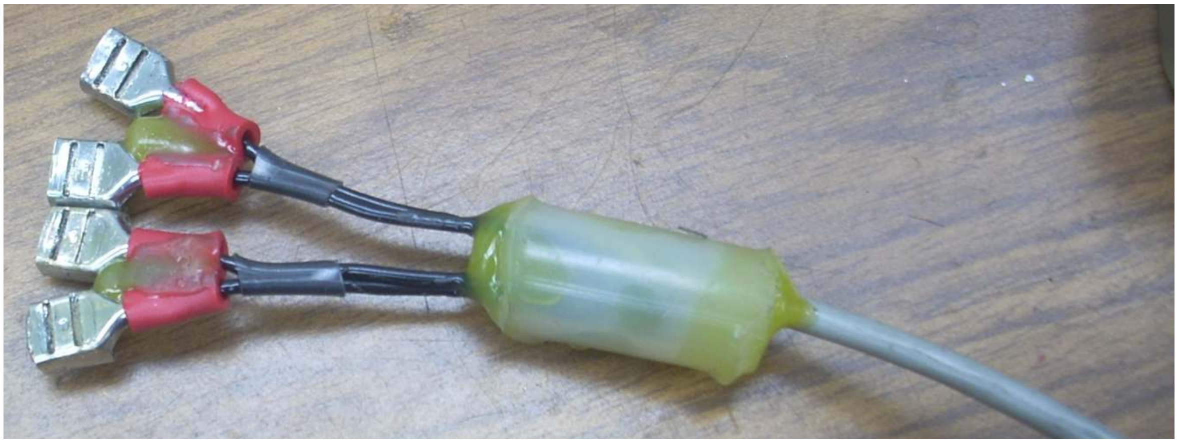
Degrease a small patch of the table. Then place all connectors on this patch. Adjust R6, the trim pot, until the touchdown LED just stops flickering.

Move the probes so one is on the spindle while the other is on the vise. Put a metal block in the vise. Wait a few seconds and then do a trial touchdown. If the touchdown LED flickers, slightly adjust the trim pot until it remains dark. The touchdown LED should flash for at least 0.1 seconds.

This adjustment should not be needed again unless the EEF is moved to a different shop that has significantly different electrical noise.

It is possible that the electrical noise level around the machine is so high that the circuit cannot operate correctly. Keeping the test leads short will minimize this effect. Some detective work may be needed to find the source of the noise. I welcome you to contact me at rgsparber@aol.com if you have problems.

The Magnetic Probes



This is a view of the contacting surfaces of the probes.

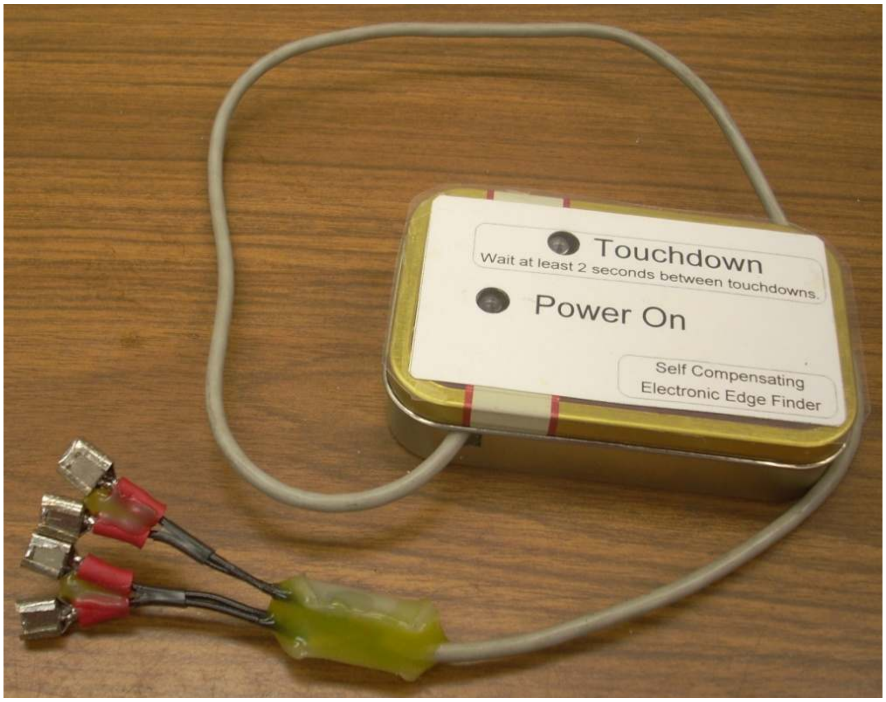
Making each clip is explained in

http://rick.sparber.org/electronics/mwc.pdf

Before crimping in each magnet, I soldered the wire to the clip. Then I crimped the wire to secure it. Place the magnets in the clips so that the two clips in a pair repel at their edges. This will keep them apart. After securing the first pair, place the second pair of magnets so they repel at the edge but also attract the other set of probes. You can see in the picture that the inner two clips are touching while the outer two clips are pushing away.

The bottom clip is HC1 and its mate is SP1. On the top probe, the bottom clip is SP2 and the top clip is HC2. This puts the high current clips on the outside of the connection and the sensing clips on the inside.

A top view of the probes.



Of critical importance is the cable that connects between the circuit and these probes. It is a coaxial cable with 3 conductors inside[9].

This cable insures that noise cannot enter the circuit via the cable. The ground shield of the cable connects to SP2.

Heat glue was used to pot the splice between coaxial cable and probe wires.

The coaxial cable is about 12" long and the probe wires are about 1" long. Minimize the area defined by the probe cables and clips since it is where much of the electrical noise comes from.

Except around swarf, I do like the magnetic clip approach. Given that swarf completely disrupts the finding of touchdown, it is reasonable to assume that the area is free of the stuff. Shop experience may change my mind.

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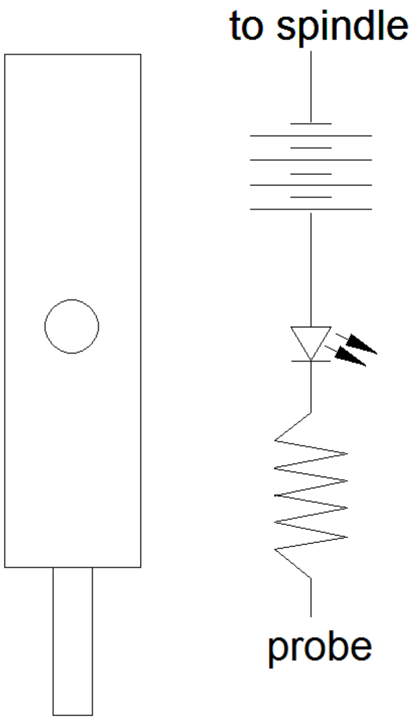
[9] I scrounged this cable from an old Apple Computer system.

Theory of Operation

A Common Electronic Edge Finder



Common EEFs are often used on manual[10] mills to find the exact point where the center of rotation of the spindle is a known distance from a reference surface. These EEFs consist of a conductive body that is held in the spindle. A conductive cylinder, the probe, extends out the bottom which is insulated from the body. A battery is connected to the conductive body. The other end of the battery connects to an LED and resistor in series. The LED lights when current flows out the probe, through the work piece, into the machine, and out the spindle. This works very well because there is a large change in resistance from essentially infinite to zero resistance. I am not aware of a similar device that is used on a manual lathe.

The New Electronic Edge Finder

The EEF presented here works in a different environment. No attachment similar to a common EEF is used.

In the case of a lathe, one wire is connected to the spindle and the other wire connects to the cutter. This resistance measures as a dead short using a commonly available Volt-Ohm meter. But if you have the right equipment, you will find that the resistance can be as small as 0.010 ohms. When the cutter comes in contact with the work piece held in the spindle, this resistance falls by as little as 0.16 milli ohms. The circuit is sensitive to this tiny drop in resistance. For more information, please see http://rick.sparber.org/ueef.pdf .

The circuit uses the pre-touchdown resistance to calculate a threshold and then monitors for a drop in this resistance that indicates touchdown has occurred.

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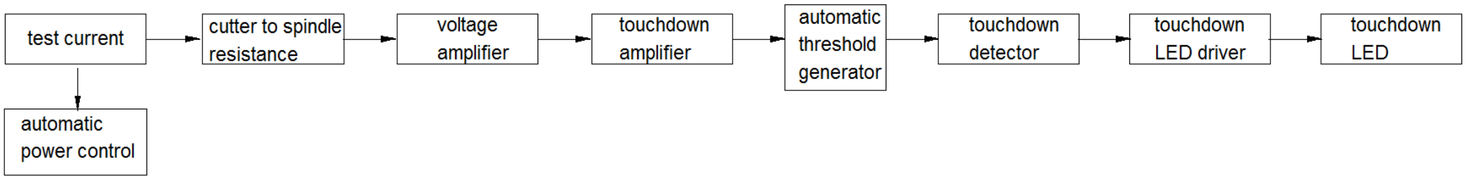
[10] Versus a CNC mill.

System Block Diagram



Starting on the left, we have a test current generator. It applies 20 mA to the resistance between cutter and spindle. It also signals the automatic power control circuit. As long as the test current flows, power is applied to the rest of the circuit.

The cutter to spindle resistance can be as small as 10 milli ohms so the voltage generated by the 20 mA test current is only 200 micro volts. The voltage picked up by a TV antenna is on the order of tens of micro volts. So our test voltage is very small.

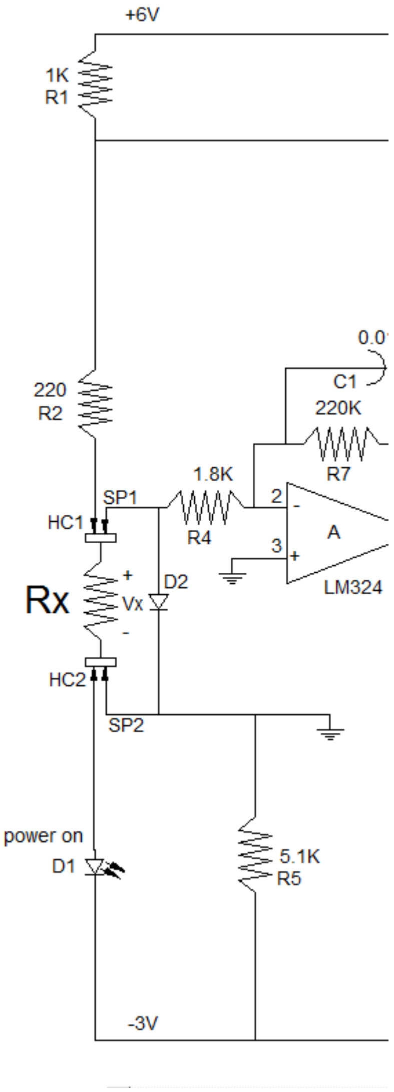
Our drop in resistance at touchdown is on the order of 0.16 milli ohms which generates a change in voltage of -3.2 micro volts. It is this tiny drop in voltage that the circuit is able to detect.

The circuit has two phases. In the first phase it is constantly calibrating to the non-touchdown resistance. In the second phase it is detecting the sudden drop in resistance.

During calibration, the voltage generated across the cutter to spindle resistance is amplified by a factor of -122 by the voltage amplifier. It then passes through the touchdown amplifier which multiplies it by -1. The voltage is then applied to the automatic threshold generator which computes and stores the proper threshold that must be crossed in order to signal touchdown.

During the second phase, the circuit reacts to only the sudden drop in resistance. It causes a tiny drop in the voltage generated across the cutter to spindle resistance. This voltage drop is amplified by the voltage amplifier just like during the first phase. But then it is further amplified by -100 in the touchdown amplifier. The signal is then sent to the touchdown detector where the threshold has already been set. The detector sends a signal to the touchdown LED driver which in turn lights the touchdown LED.

Test Current Generator



The test current, I_x, is set by R_2:

$$I_x = \frac{Vbattery - Veb1 - VD1}{R2} \qquad (1)$$

$$I_x = \frac{6V - 0.75V}{220\ ohms}$$

I_x = 23.9 mA

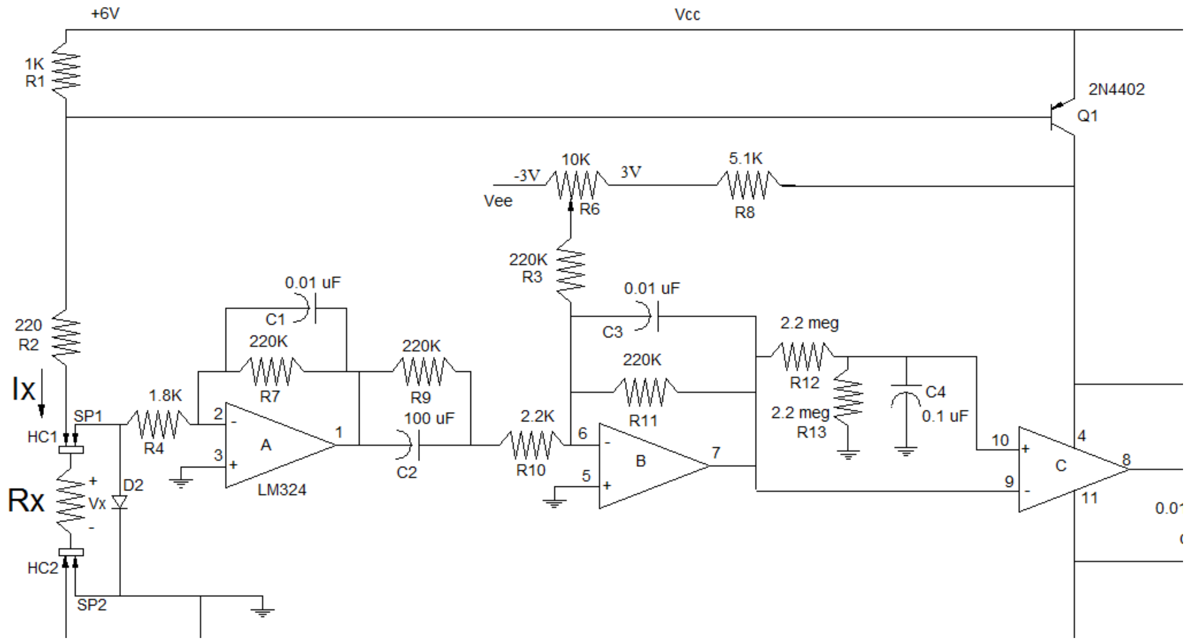
I saw a droop of about 0.8V in the battery voltage when under load and measured close to 20 mA in my prototype. The exact value is not important as will be shown later.

I_x flows through a pair of High Current probes marked HC1 and HC2. The resistance of these probe wires and the contact resistance of the clips causes a voltage drop. This drop is tiny compared to the battery voltage but could be huge compared to the voltage being sensed. To get around this problem, a second set of probes is used to only sense the voltage across R_x. Almost no current flows in these signal probe (SP) leads so no appreciable voltage is generated in their wires or probes. This arrangement is called a Kelvin connection and is commonly used when measuring very low resistances.

If R_x is above 35 ohms, diode D2 turns on to limit the voltage to less than 0.7 volts.

SP2 defines ground for the circuit. When this probe is disconnected, R5 proves a path to the negative battery terminal. That prevents the ground node from floating which can make it susceptible to electrostatic discharge damage.

I_x does a few other things besides generating the input voltage for the voltage amplifier. As it flows through R1, it turns on the power control circuit to be presented next. This current also flows through D1 which lights it to indicate power is on. The voltage drop across D1 is used to define the negative terminal of the battery to be at about 3 volts below ground.

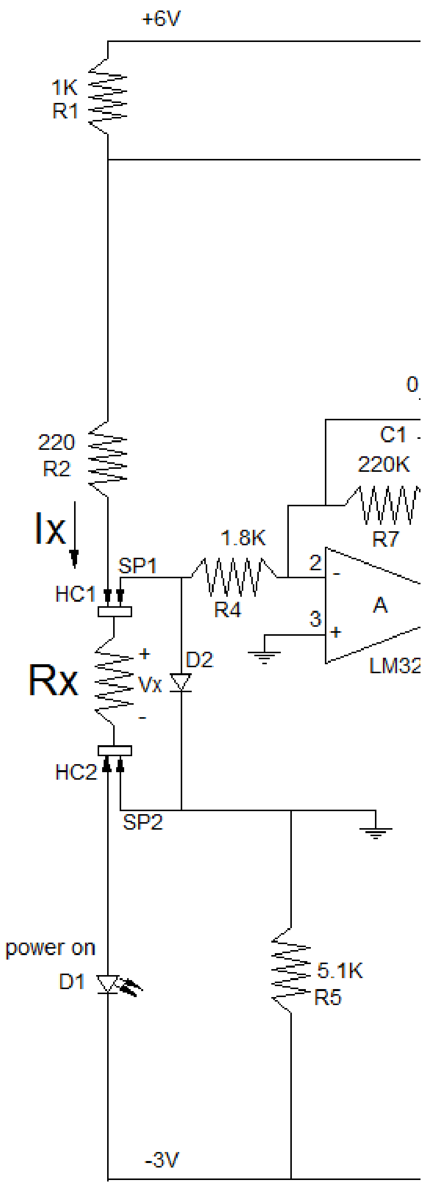


Power Control

The flow of I_x through R1 causes the emitter-base junction voltage of Q1 to rise. When this voltage rises until Q1 turns on. Then the voltage is almost constant at 0.75V. This means that $\frac{0.75V}{1K}$ = 0.75 mA flows through R1 and the rest flows out of the base of Q1. So my base drive is 20 mA - 0.75 mA = 19.3 mA. The collector current on Q1 is far less than 20 mA so Q1 is driven deep into saturation. The quad op amp integrated circuit is then essentially tied to Vcc, the positive terminal of the battery.

When HC1 is disconnected from the lathe or mill, I_x drops to zero, Q1 turns off, and power is removed from the circuit.

Cutter to Spindle Resistance



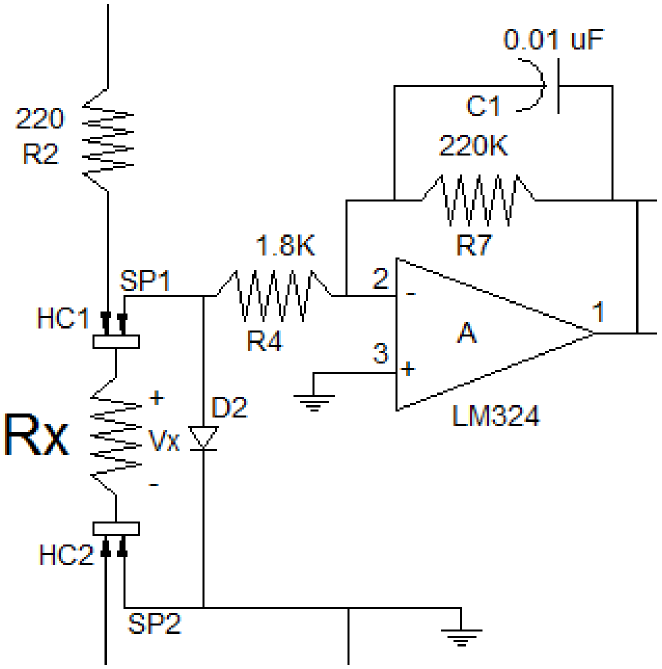
The cutter to Spindle Resistance is represented by R_x. It can be as small as 10 milli ohms.

$$V_x = R_x \times I_x \qquad (2)$$

This 10 milli ohms is shorted by the touchdown resistance which can be as large as 0.6 ohms. This causes the 10 milli ohms to drop to $\frac{Rx \times Rtd}{Rx+Rtd} = 9.84$ milli ohms at touchdown. That is a drop of 164 micro ohms. I call this change ΔR_x. Equation (2) still holds for changes in resistance.

$$\Delta V_x = \Delta R_x \times I_x \qquad (\Delta 2)$$

Voltage Amplifier



The voltage generated across R_x is carried by probes SP1 and SP2 to the voltage amplifier. The output voltage is at pin 1 and is called V1:

$$V1 = G_A \times V_x$$

$$\text{Where } G_A = -\frac{R7}{R4} \qquad (3)$$

$$V1 = -\frac{220K}{01.8K} \times V_x = -122\ V_x$$

assuming there is no input offset voltage. I will deal with this subject later.

Capacitor C1 in conjunction with R7 causes the gain of this amplifier to drop as the signal frequency rises. We want to pass some high frequencies in order to see touchdown but at the same time we do not want to see electrical noise present in the room. Given that we are trying to process micro volts of signal, it doesn't take much noise to cause trouble.

The gain is cut to $-\frac{122}{\sqrt{2}} = -86$ when the frequency equals $\frac{1}{2 \times \pi \times R7 \times C1} = 72$ Hz and continues to fall as the noise's frequency rises.

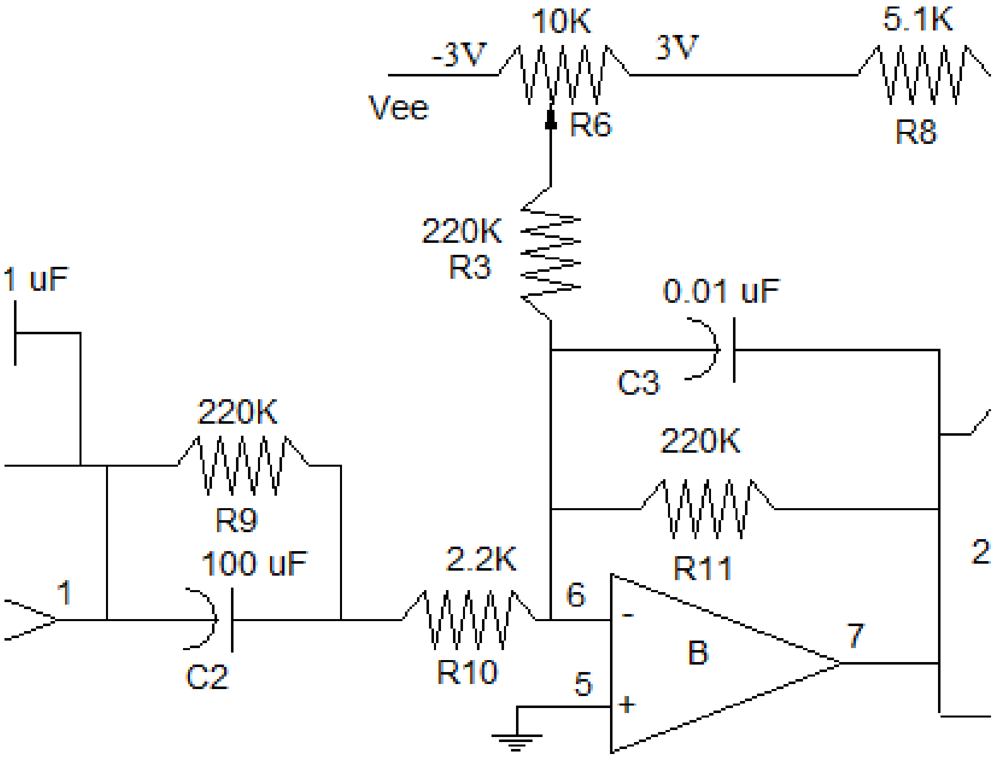
In the time domain, the amplifier has a time constant of $R7 \times C1 = 2.2$ milliseconds. This is the same time constant as the touchdown amplifier but 50 times smaller the next largest time constant. This fact will come in handy when analyzing other functional blocks.

The amplifier will also operate on changes in voltage which I will designate as Δ:

$$\Delta V1 = G_A \times \Delta V_x$$

$$\text{Where } G_A = -\frac{R7}{R4} \qquad (\Delta 3)$$

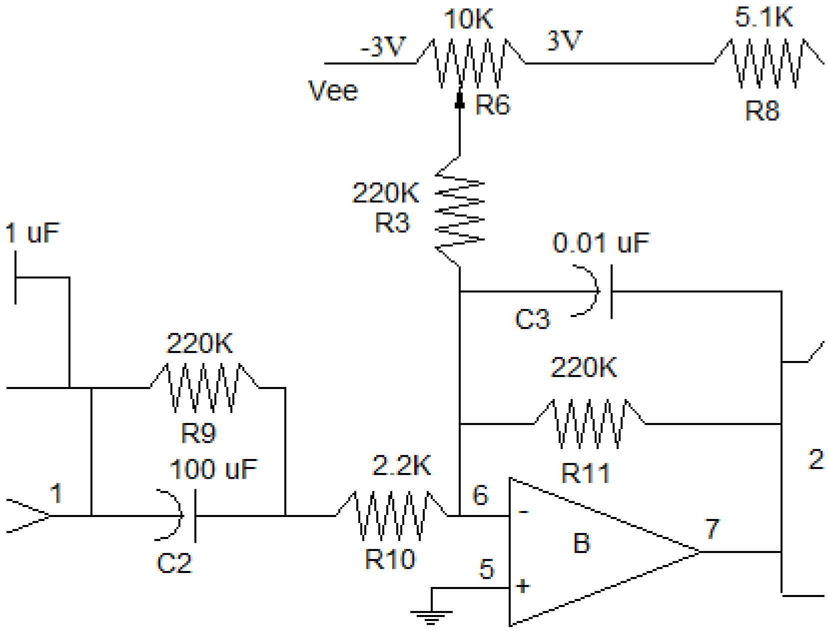
Touchdown Amplifier



This amplifier has two different gains. During the compensation phase, C2 charges up to V1 and then becomes an open circuit. C3 charges up to V7 and also becomes an open circuit. We are left with R9, R10, and R11:

Compensation phase: $V7 = -\frac{R11}{R9+R10} \times V1$

$$V7 = -\frac{220K}{220K+2.2K} \times V1 \cong -V1 \qquad (4)$$



During the touchdown phase V1 will rise with a time constant of 2.2 milli seconds as calculated during the analysis of the voltage amplifier. The time constant related to C2 equals $R10 \times C2$ = 220 milliseconds. This means that as far as C2 and R10 are concerned, V1 rises instantly and we can treat C2 as a short during this transition.

The rest of the amplifier is similar to the previous gain stage in that it has a time constant of 2.2 milliseconds. So I now have two circuits in series each with a time constant of 2.2 milliseconds, The drop in V7 therefore contains the slowing effect of these two time constants. The actual value is not that important because this voltage will be encountering circuits with time constants much slower than this. So we can still think of the drop in V7 as instantaneous.

Touchdown phase: $V7 = G_B \times V1$

$$\text{Where } G_B = -\frac{R11}{R10} \qquad (5)$$

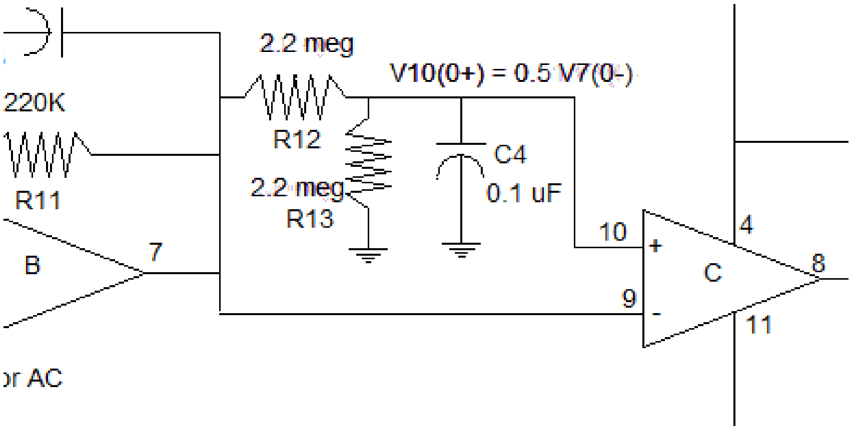
$$V7 = -\frac{220K}{2.2K} \times V1 = -100 \times V1$$

So to recap, the gain is -1 during compensation and -100 when passing the sudden drop in R_x.

This gain drops to to $-\frac{100}{\sqrt{2}} = -70.7$ at a frequency of 72 Hz due to C3 and R11.

Resistors R3, R6, and R8 compensate for offset voltage. I will explain this function later.

Automatic Threshold Generator



This functional block has the fanciest name and function yet is arguably the simplest circuit.

As with the touchdown amplifier, this circuit too has two behaviors depending on the phase we are in.

During the compensation phase, the capacitor, C4, charge up to a given DC voltage and can then be ignored. We will be in the compensation phase up to the moment of touchdown. In other words, up to t = 0^-. It is this moment in time that we will study.

The voltage V10(0^-) comes from a voltage divider made up of R12 and R13:

Compensation phase: $$V10(0^-) = \frac{R13}{R12+R13} \times V7(0^-)$$

$$V10(0^-) = \frac{2.2\ meg}{2.2\ meg+2.2\ meg} \times V7(0^-)$$

$$V10(0^-) = 0.5 \times V7(0^-) \qquad (6)$$

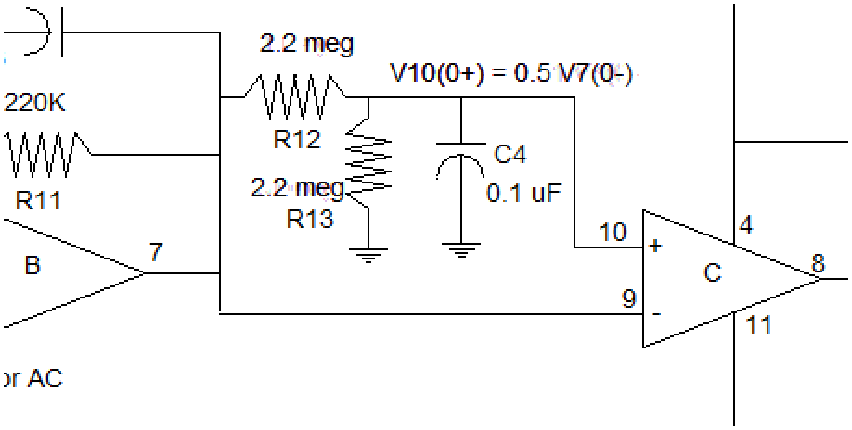
$$V9(0^-) = V7(0^-) \qquad (7)$$

Op amp "C" is used as a comparator. It responds to V10 - V9. Using (6) and (7):

$$V10(0^-) - V9(0^-) = \{0.5 \times V7(0^-)\} - V7(0^-)$$

$$V10(0^-) - V9(0^-) = - 0.5 \times V7(0^-) \qquad (8)$$

So as long as V7(0^-) is positive, V10(0^-)-V9(0^-) will be negative and V8(0^-) will sit near the op amp's negative rail, -3V.



During the edge phase, starting at $t = 0^{+}$, the voltage $V7(0^{+})$ and $V10(0^{+})$ both start to fall but at different rates. The time constant of V7(t) is on the order of a few milli seconds. The time constant of C4 with the parallel combination of R12 and R13 is 110 milli seconds. So I will again ignore the effects of the capacitors in the voltage and touchdown amplifiers which slow down V7(t). V10 immediately after touchdown essentially does not move. I can say

$$V10(0^{+}) \cong V10(0^{-}) \qquad (9)$$

$$V9(0^{+}) = V7(0^{+})$$

So at $t = 0^{+}$,

$$V10(0^{+}) - V9(0^{+}) = V10(0^{-}) - V9(0^{+})$$

Using (6) and (9) this becomes

$$V10(0^{-}) - V9(0^{+}) = \{0.5 \times V7(0^{-})\} - V7(0^{+})$$

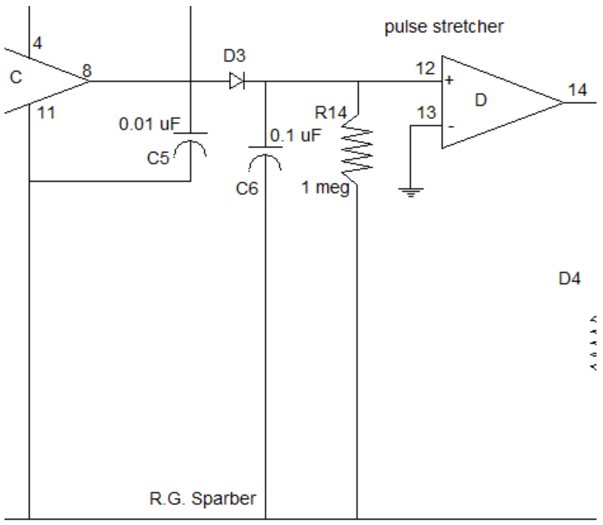
But $\Delta V7 = V7(0^{+}) - V7(0^{-})$ so $V7(0^{+}) = \Delta V7 + V7(0^{-})$

So $V10(0^{-}) - V9(0^{+}) = \{0.5 \times V7(0^{-})\} - \{\Delta V7 + V7(0^{-})\}$

$$V10(0^{-}) - V9(0^{+}) = \{-0.5 \times V7(0^{-})\} - \Delta V7 \qquad (10)$$

This equation is a bit strange in that I am using a voltage, V7(0-), which occurred before touchdown, with a voltage, ΔV7, that occurred after touchdown. This is possible because C4 acts as a memory device and stores the pre-touchdown information.

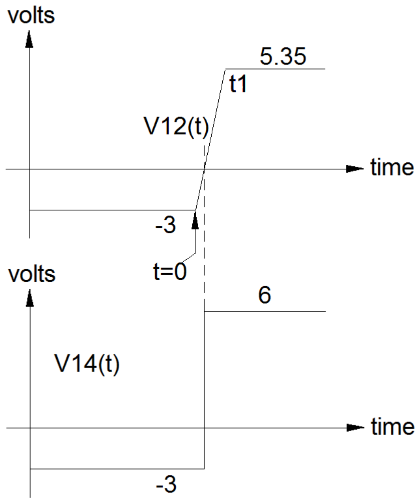
Touchdown Detector



Right after touchdown, we will have V8 high at about $t = 0^+$. It will go back low after various capacitors charge to their new values. All we care about is that V8 stays at +6V long enough for the touchdown detector to see it.

I define t_1 as the minimum time that V8 drops back to -3V.

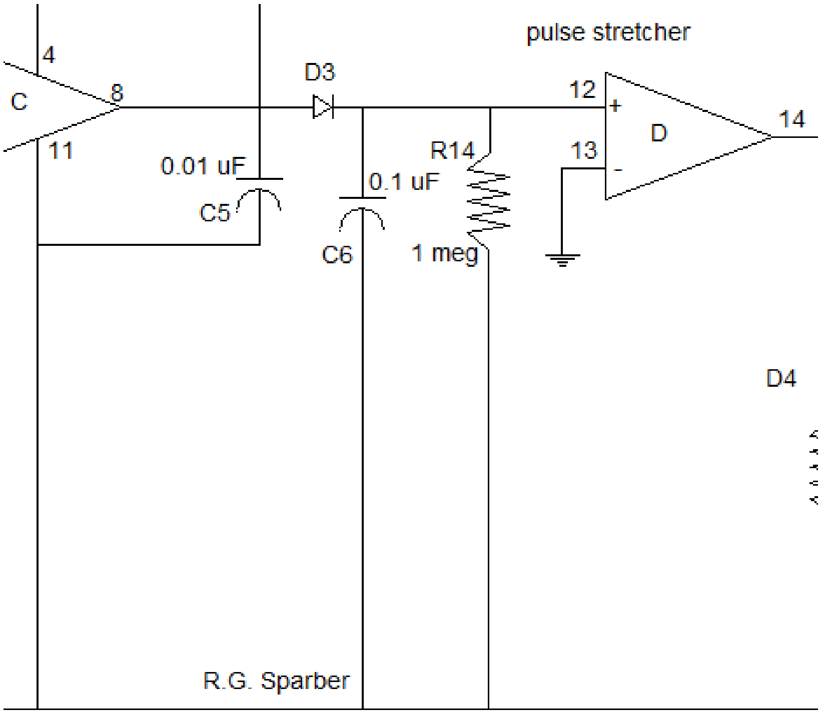
Op amp C has a guaranteed minimum output current of 20 mA. When V8 jumps from -3V up towards +6V, it will turn on D3 and start to charge C6. The voltage across C6 started at 0 because R14 discharged it. If there is time, it will charge up to about 6V - 0.65V = 5.35V. Once it crosses 0V with respect to ground, op amp D which acts as a comparator, will change state. V14 will swing from -3V to +6V.

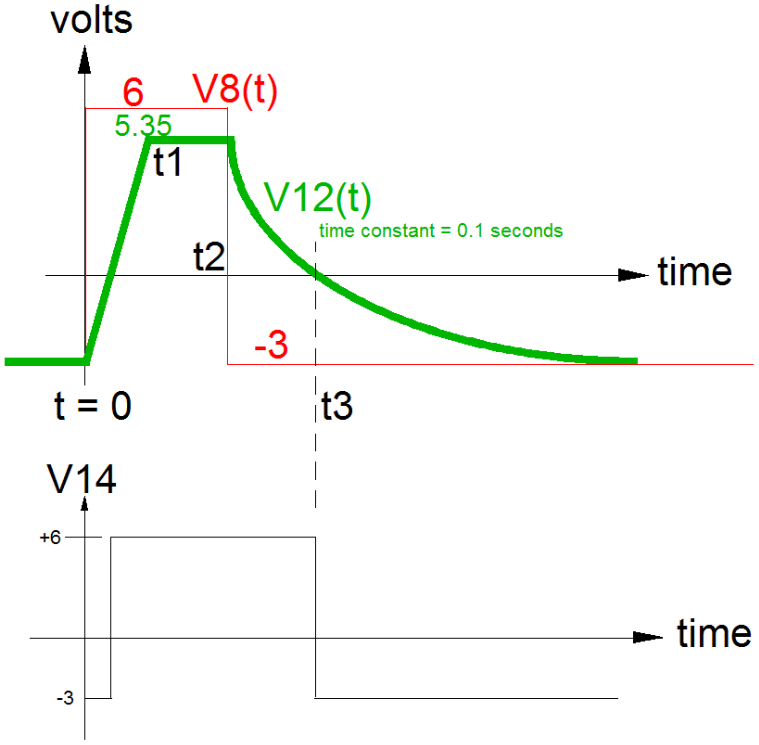


To be fully charged up to +5.35V, I need

$$5.35 = -3 + \frac{20\,mA \times (t1)}{0.1\,uF}$$

Which gives me a t1 = 42 micro seconds. So if V8 stays high for at least 42 microseconds, C5 will be fully charged. This 42 micro seconds is much smaller than the 110 milli second time constant affecting the input to op amp C. So we can be sure that there will be plenty of time for C6 to be fully charged.



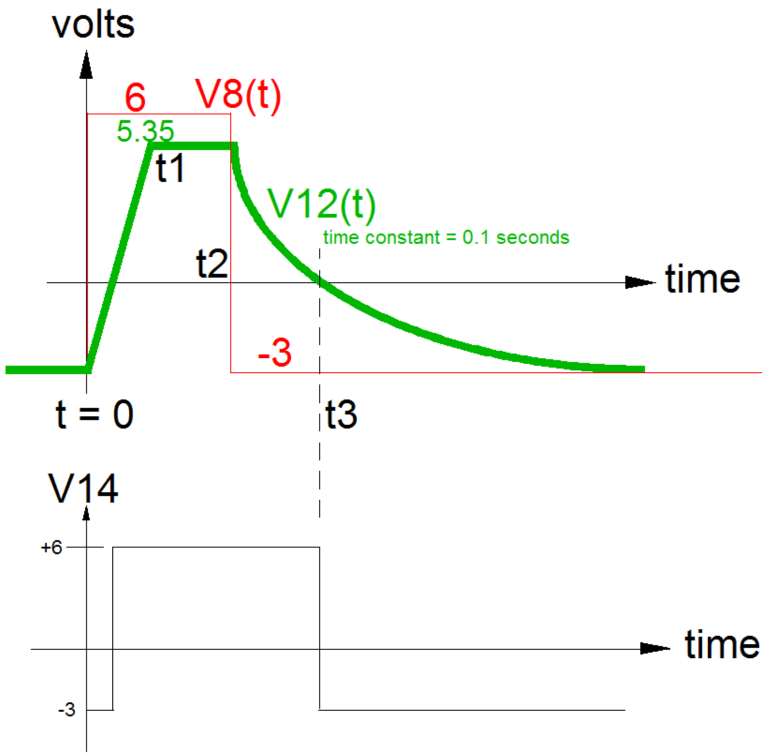


Recall that V14 swung up to +6V when C6 charged up past 0V. That was shortly after t = 0.

V8 must stay at +6V for at least 42 micro seconds but could stay up longer. Say V8 drops back down to -3V at t2.

With V8 no longer holding up V12, C6 starts to discharge.

As long as V12 is greater than 0V, V14 will stay near 6V. At t3 V12 drops below 0V so V14 will drop to -3V.



C6 is discharged by R14 so we can write:

$$\mathrm{V12(t)} = \{\Delta \mathrm{V12}\}\{1 - e^{\frac{-(t-t2)}{\tau 6}}\} + \mathrm{V12}(0^-)$$

where $\tau 6 = R14 \times C6 = 0.1$ seconds (11)

I want to calculate t3 - t2, the time it takes for V12 to go from 5.35V down to 0:

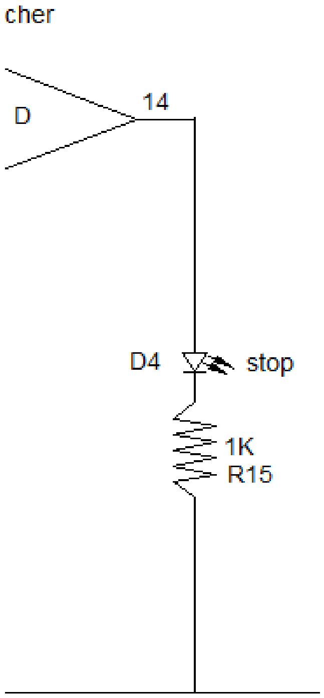
$$\mathrm{V12(t3 - t2)} = 0 = \{0 - 8.35\}\{1 - e^{\frac{-(t3-t2)}{\tau 6}}\} + 5.35$$

which gives me a t3 - t2 = $-\tau 6 \ln\left(1 - \frac{5.35}{8.35}\right) = -\tau 6(-1.02)$

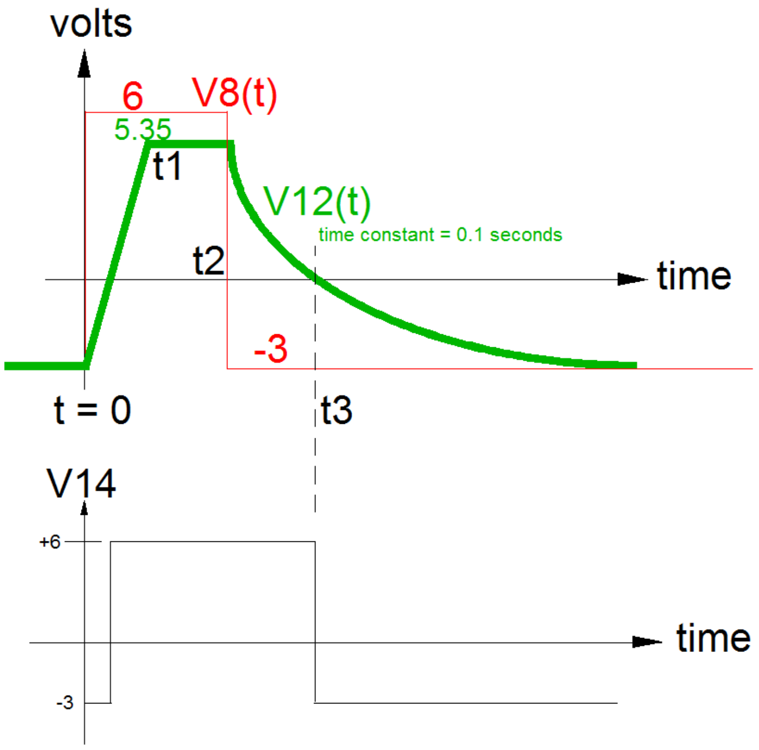
t3 - t2 = (- 0.1 seconds)(-1.02) = 0.1 seconds

To recap, if we assume that we get a pulse at V8 of at least 42 microseconds, the pulse stretcher will generate a pulse of 0.1 seconds. That is long enough to see as it flashers our touchdown LED.

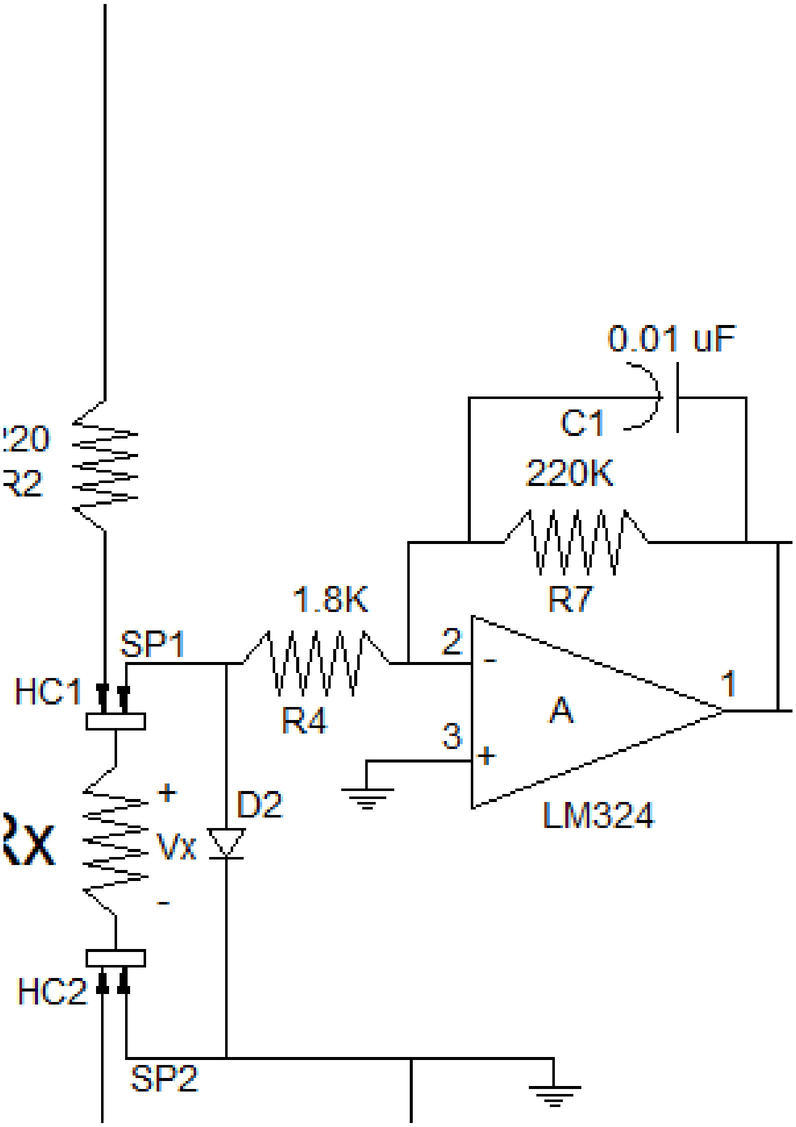
Touchdown LED



When the output of op amp D is at -3V, there is no voltage across the touchdown LED or its current limiting resistor, R15. But during the time from t2 to t3, V14 is at 6V so the voltage across R15 is at 6V - V_{D4} - (-3). D4 drops approximately 3V so R15 has about 6V across it. Since it is 1K, this means that the current through it is about 6 mA. This current also flows through D4 which is a Super Bright LED. The 0.1 second flash of light from D4 is easy to see.



Offset Adjustment



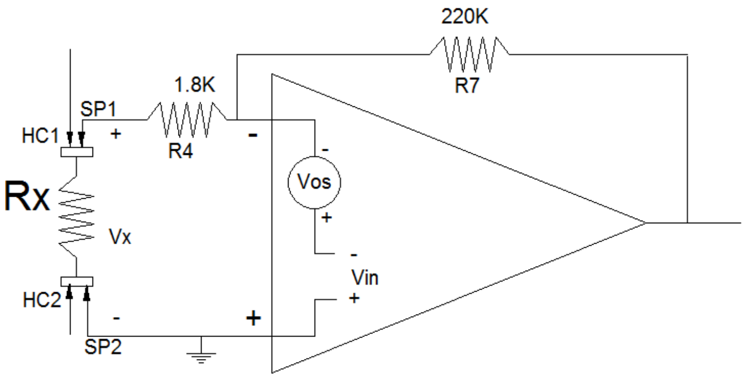
Way back in the discussion about the voltage amplifier, I skipped over the input voltage issue. We now know enough to see how it effects the overall circuit.

First consider the voltage. The voltage on SP1 with respect to ground is amplified by op amp A by an amount equal to

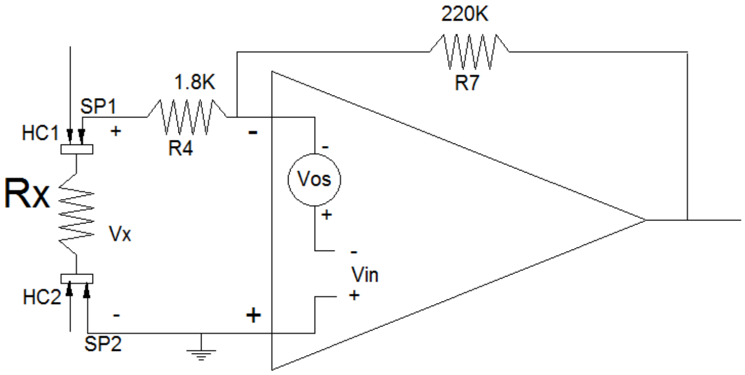
$$-\frac{R7}{R4} = \text{-}122.$$

Nice and simple.

But now let's include input offset voltage in our model.



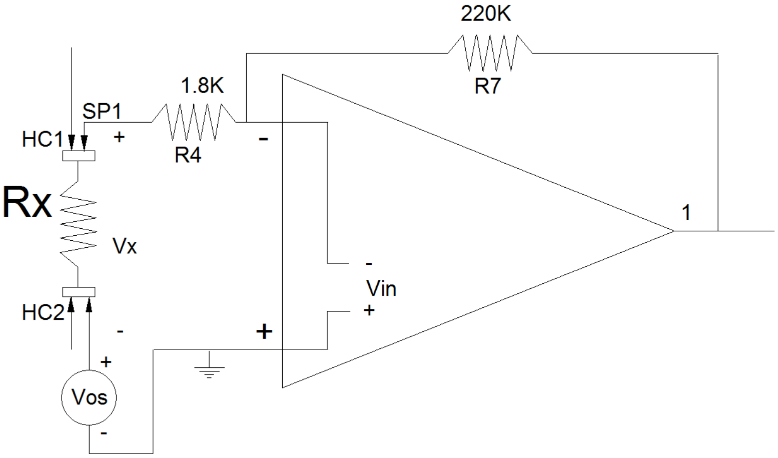
The input offset voltage is a tiny imperfection in all op amps which can have major consequences. When this voltage is zero, the voltage marked V_{in} is equal to the voltage applied to the + and - input terminals of the op amp. But when V_{os} is not zero, it directly effects the output



voltage. The polarity of V_{os} cannot be controlled by the manufacturer but they can specify a maximum magnitude. For the LM324, it is 9 millivolts.

The op amp is going to only react to the voltage V_{in}.

$V_{in} = -V_x - V_{os}$



An equivalent circuit that gives the same value to Vin is this one. It lets me treat the op amp as idea and lump the Vos problem in with Vx.

Equation (3) still applies but I must now include the effect of Vos inside V_x:

$$V1 = G_A \times V_x$$
$$\text{Where } G_A = -\frac{R7}{R4} \qquad (3)$$

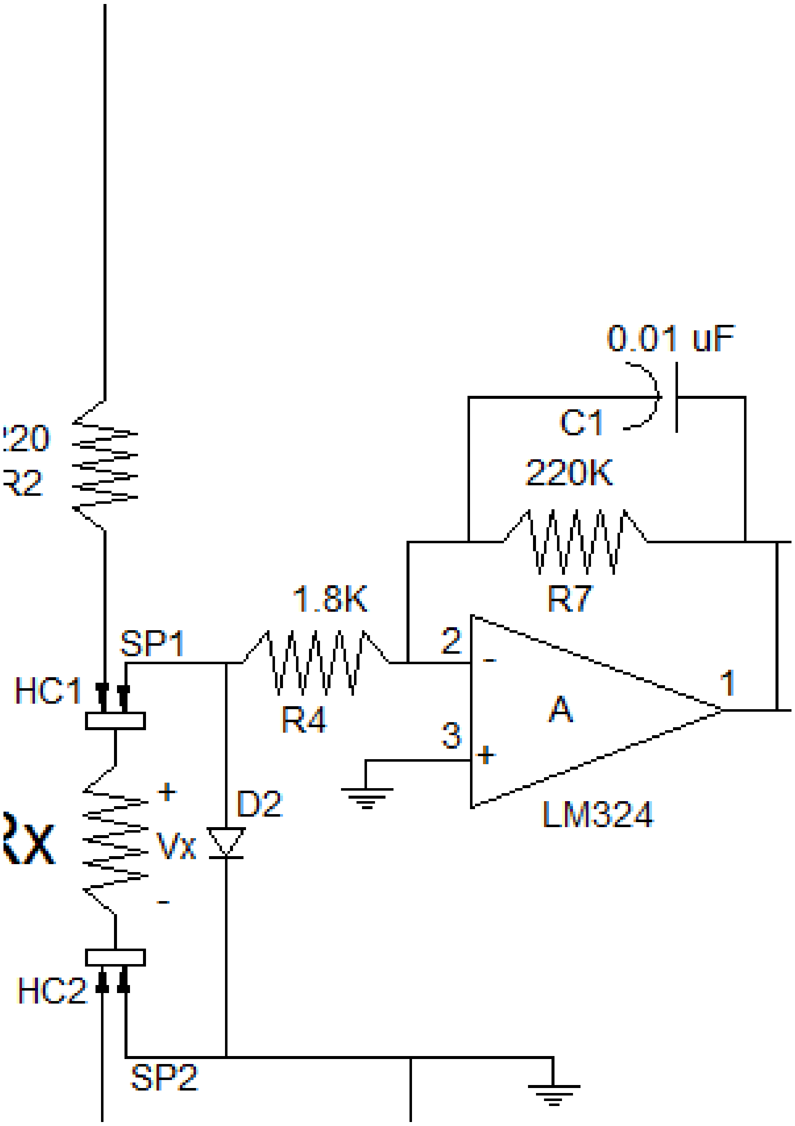
$$V1 = -\frac{220K}{1.8K} \times V_x = -122\ V_x$$

But now I have

$$V_x = I_x \times R_x + V_{os} \qquad (12)$$

and

$$V1 = -122 \times \{(I_x \times R_x) + V_{os}\}$$



So even when R_x is zero, I will get a V1 = -122 × V_{os}. Given that V_{os} can be as much as 9 mV, my V1 due to just this offset can be a large as 1.1V. Ouch!

The standard solution is to inject a compensating current into the circuit so most of this offset voltage is canceled. Originally[11] I did this but later discovered that the added lead length and components injected a sizable amount of noise right into the most sensitive node in the circuit, pin 2.

So in the end, I will accept an unknown voltage at pin 1 that can have a magnitude as large as 122 × 9 milli volts = 1.1 volts. This voltage can be positive or negative.

The output at pin 1 can swing down to about -3V. If my amplified V_{os} is at 9 milli volts, then V1 will be at -1.1V when $V_x = 0$. This means that I only have about 1.9 volts of travel on the output in the worst case. Reflected back to V_x this means I can handle up to $\frac{-1.9V}{-122}$ = 15.6 milli volts. Given an I_x of 20 milli amps, this means that my maximum R_x is $\frac{15.6\ milli\ volts}{20\ milli\ amps}$ = 780 milli ohms. Above this resistance, V1 will remain at about -3V. During touchdown, R_x will fall to at least 600 milli ohms because that is the maximum touchdown resistance. Since this value is less than our maximum, the rest of the circuit will see the change.

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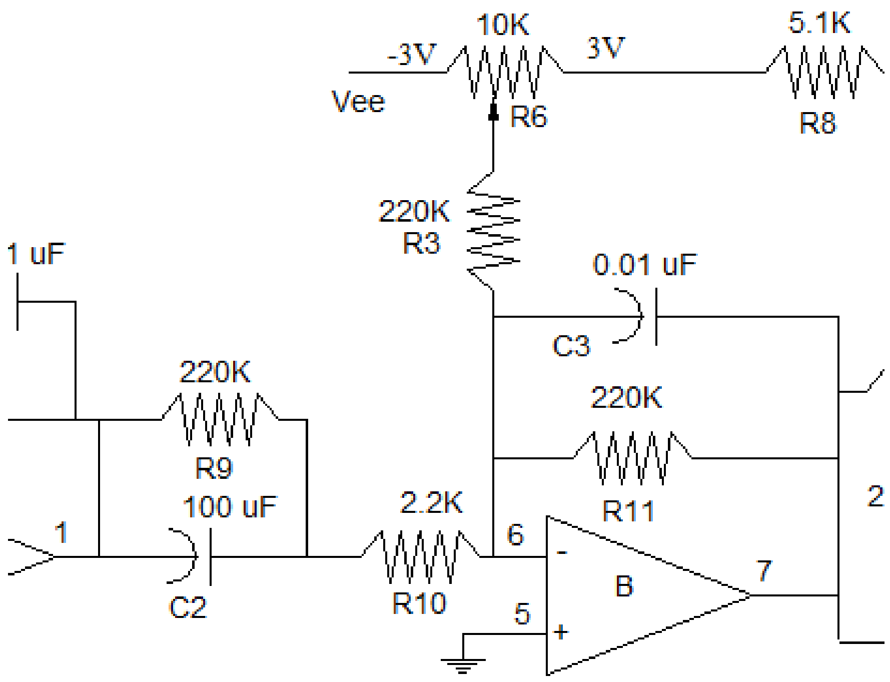
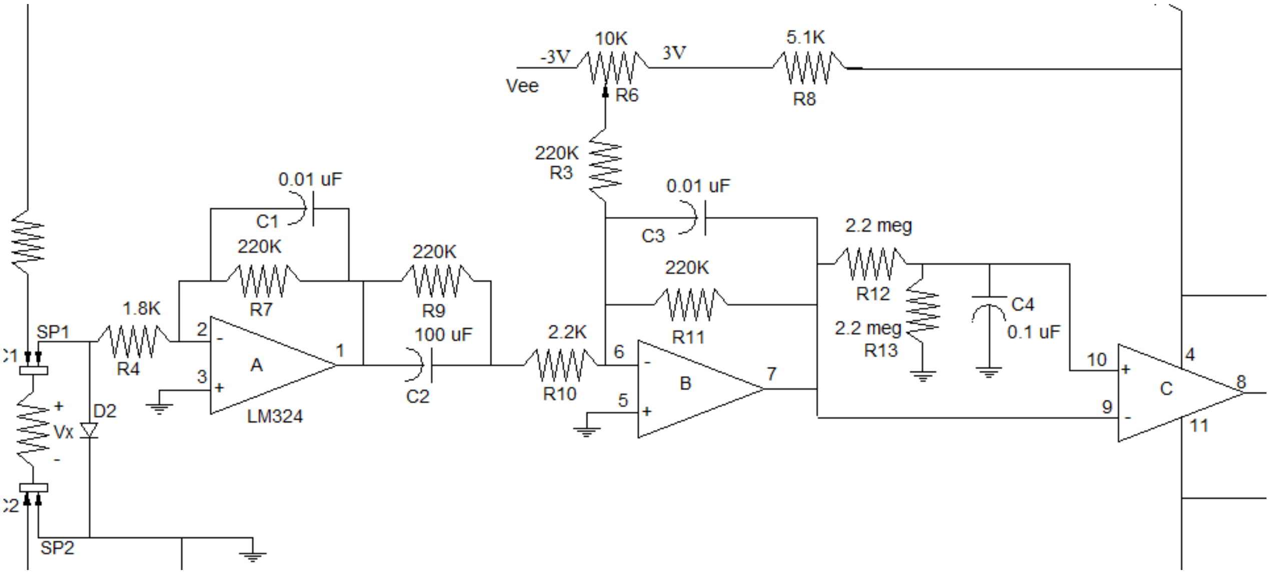
[11] Actually, the story is a bit crazier than that. I intended to inject an offset voltage compensating current into node 2 but made a mistake, and injected it into the touchdown amplifier. When I discovered my mistake, I "fixed" the circuit. Then the noise level was a lot higher and caused poor performance. So in the end, I put my mistake back in.



The touchdown amplifier including op amp B, has a gain of -1 during the compensation phase so it's V_{os} can only contribute from -9 mV to + 9 mV. But the amplified V_{os} from the voltage amplifier can add anywhere from -1.1V to +1.1V so it dominates.

The network made up of R3, R6, and R8 can be adjusted to move V7 from -3V to +3V . This voltage adds to the voltage generated by the amplified V_{os} from the voltage amplifier. So if V1 is at, say, -1.1V, we can adjust R6 to cancel it. In fact, there is plenty of adjust left over which will be used in the following stage.



Now we have to look at the entire circuit. The input offset voltage in op amp A can generate up to ± 1.1V at pin 1. If the wiper of R6 is set to zero volts, we would get ± 1.1V at pin 7. During the calibration phase, the voltage applied to op amp C equals -0.5 × $V7$ ± its own offset voltage. If this voltage is greater than zero, the touchdown LED will light. By adjusting R6 until this LED goes dark, we have canceled all offset voltages in op amp A, B, and C. This assumes there is no noise injected into the circuit with V_x.

Consider what happens if there is noise adding into V_x. When this noise voltage rises, the touchdown circuit ignores it. When the noise falls, the touchdown circuit might see it as a possible valid touchdown. This would only happen if the zero to peak noise voltage was large enough to make V10-V9 go positive.

But in our procedure to null out offset, we turned R6 until the touchdown LED goes dark. If there is noise present, we have raised the threshold so it does not turn on the LED.

The down side of this procedure is that if the zero to peak noise voltage is larger than ΔV_x, the circuit cannot detect touchdown. Capacitors C1 and C3 do attenuate this noise voltage as a function of frequency so they help some.

I will assume that all offset voltages have been canceled and my noise voltage is zero in the following analysis.

The Calibration Phase

We now have all of the equations necessary to define the calibration phase. This assumes you have nulled out all input offset voltages via R6 and the procedure defined in the last page plus that no electrical noise exists. Let me collect my equations first.

$$V_x = R_x \times I_x \qquad (2)$$

$$V1 = G_A \times V_x$$

$$\text{Where } G_A = -\frac{R7}{R4} \qquad (3)$$

$$V1 = -\frac{220K}{1.8K} \times V_x = -122\ V_x$$

$$V7 = -\frac{R11}{R9+R10} \times V1$$

$$V7 = -\frac{220K}{220K+2.2K} \times V1 \cong -V1 \qquad (4)$$

Note that V7 during the calibration phase equals $V7(0^-)$ to be used during the touchdown phase.

Next I will combine equations:

$V7(0^-) = V7$

$V7(0^-) = -V1$

$V7(0^-) = 122V_x$

$$V7(0^-) = 122 \times R_x \times I_x \qquad (13)$$

This equation tells us the voltage applied to the touchdown detector given R_x and I_x. The limiting case is for the smallest R_x which is 0.01 ohms. Assuming an I_x of 20 mA,

$V7(0^-) = 122 \times 0.01 \text{ ohms} \times 20 \text{ mA}$

$V7(0^-) = 24.4 \text{ milli volts}$

We have

$V10(0^-) - V9(0^-) = -0.5 \times V7(0^-)$ (8)

so

$V10(0^-) - V9(0^-) = -0.5 \times 24.4$ milli volts

$V10(0^-) - V9(0^-) = -12.2$ milli volts

This says that at our smallest R_x, op amp C's input is sitting at -12.2 milli volts so its output is at -3V.

For the selected resistor values,

$V10(0^-) - V9(0^-) = -61 \times R_x \times I_x$

$V10(0^-) - V9(0^-) = -61 \times R_x \times 20$ mA

$V10(0^-) - V9(0^-) = -1.22$ amps $\times R_x$ (8a)
where R_x is in ohms.

In the general case,

$V10(0^-) - V9(0^-) = -\left(\frac{R7}{2 \times R4}\right) \times (1 \text{ amp}) \times R_x$ (8b)
where R_x is in ohms.

The Touchdown Phase

I will again start by collecting equations.

$$\Delta V_x = \Delta R_x \times I_x \qquad (\Delta 2)$$

$$\Delta V1 = G_A \times \Delta V_x$$

Where $G_A = -\frac{R7}{R4}$

$$\Delta V1 = -\frac{220K}{1.8K} \times V_x = -122\ \Delta V_x \qquad (\Delta 3)$$

$$\Delta V7 = -\frac{R11}{R10} \times \Delta V1$$

$$\Delta V7 = -\frac{220K}{2.2K} \times V1 = -100\Delta V1 \qquad (4)$$

$$V7(0^-) = 122 \times R_x \times I_x \qquad (13)$$

$$V10(0^-) - V9(0^+) = \{-0.5 \times V7(0^-)\} - \Delta V7 \qquad (10)$$

Next I will combine equations:

$$V10(0^-) - V9(0^+) = \{-0.5 \times V7(0^-)\} - \Delta V7 \qquad (10)$$

$$V10(0^-) - V9(0^+) = \{-0.5 \times [122 \times R_x \times I_x]\} - \{-100\Delta V1\}$$

$$V10(0^-) - V9(0^+) = \{-0.5 \times [122 \times R_x \times I_x]\} - \{-100 \times (-122\Delta V_x)\}$$

$$V10(0^-) - V9(0^+) = \{-0.5 \times [122 \times R_x \times I_x]\} - \{-100 \times (-122 \times \Delta R_x \times I_x\}$$

For the selected resistor values,

$$V10(0^-) - V9(0^+) = I_x \times \{(-61 \times R_x) - (12{,}200 \times \Delta R_x)\} \qquad (14)$$

Where I_x is in amps and all resistance is in ohms.

In the general case,

$$V10(0^-) - V9(0^+) = -I_x \times \{[(\frac{R7}{2 \times R4}) \times R_x] + [(\frac{R7 \times R11}{R4 \times R10}) \times \Delta R_x]\} \qquad (14a)$$

Where I_x is in amps and all resistance is in ohms.

This equation tells us the voltage change at op amp C given R_x and ΔR_x . The limiting case is for the smallest R_x which is 0.01 ohms and the corresponding smallest ΔR_x which is -164 micro ohms.

$$V10(0^-) - V9(0^+) = I_x \times \{(-61 \times R_x) - (12{,}200 \times \Delta R_x)\} \qquad (14)$$

$$V10(0^-) - V9(0^+) = I_x \times \{(-61 \times 0.01 \text{ ohms}) - (12{,}200 \times [-164 \text{ micro ohms}])\}$$

$$V10(0^-) - V9(0^+) = I_x \times \{-0.610 \text{ ohms} + 2.00 \text{ ohms}\}$$

$$V10(0^-) - V9(0^+) = I_x \times \{1.39 \text{ ohms}\}$$

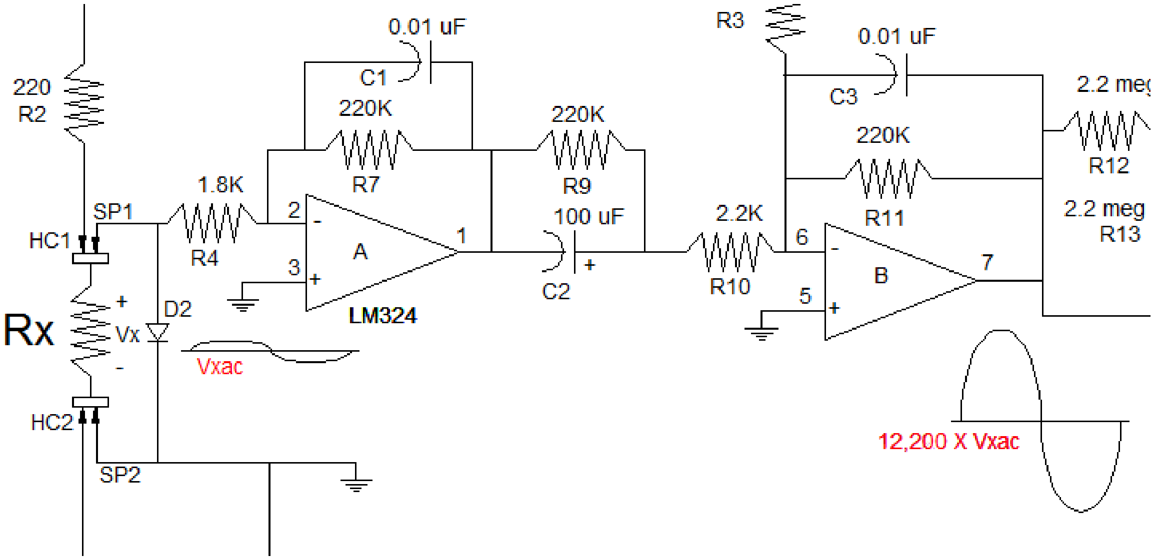
Given an I_x of 20 mA,

$$V10(0^-) - V9(0^+) = 27.8 \text{ milli volts}$$

Let's stop and reflect on all of this math. Given the smallest expected R_x of 0.01 ohms and a ΔR_x of -164 micro ohms, op amp C's input will be at -12.2 milli volts before touchdown and + 27.8 milli volts after touchdown. This transition will easily cause op amp C to change from outputting -3V to +6V.

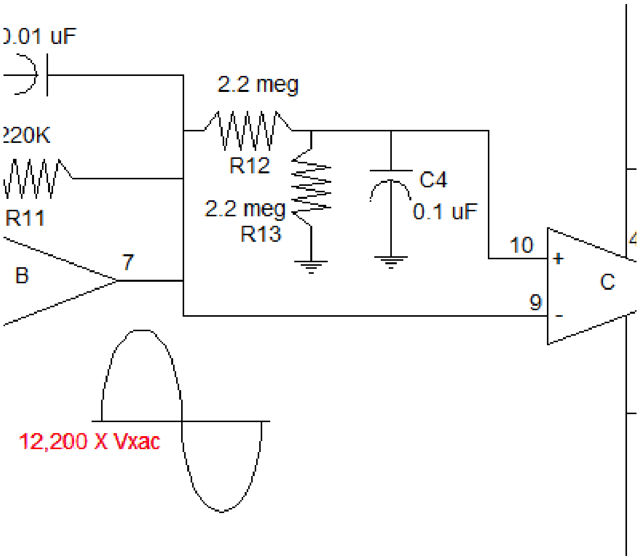
The Effect of Unwanted Electrical Noise

All of the calculations so far have assumed that there is no electrical noise mixing in with V_x. Although C1 and C3 tend to reduce this noise, some will still be there. This noise will erode our overdrive margins.

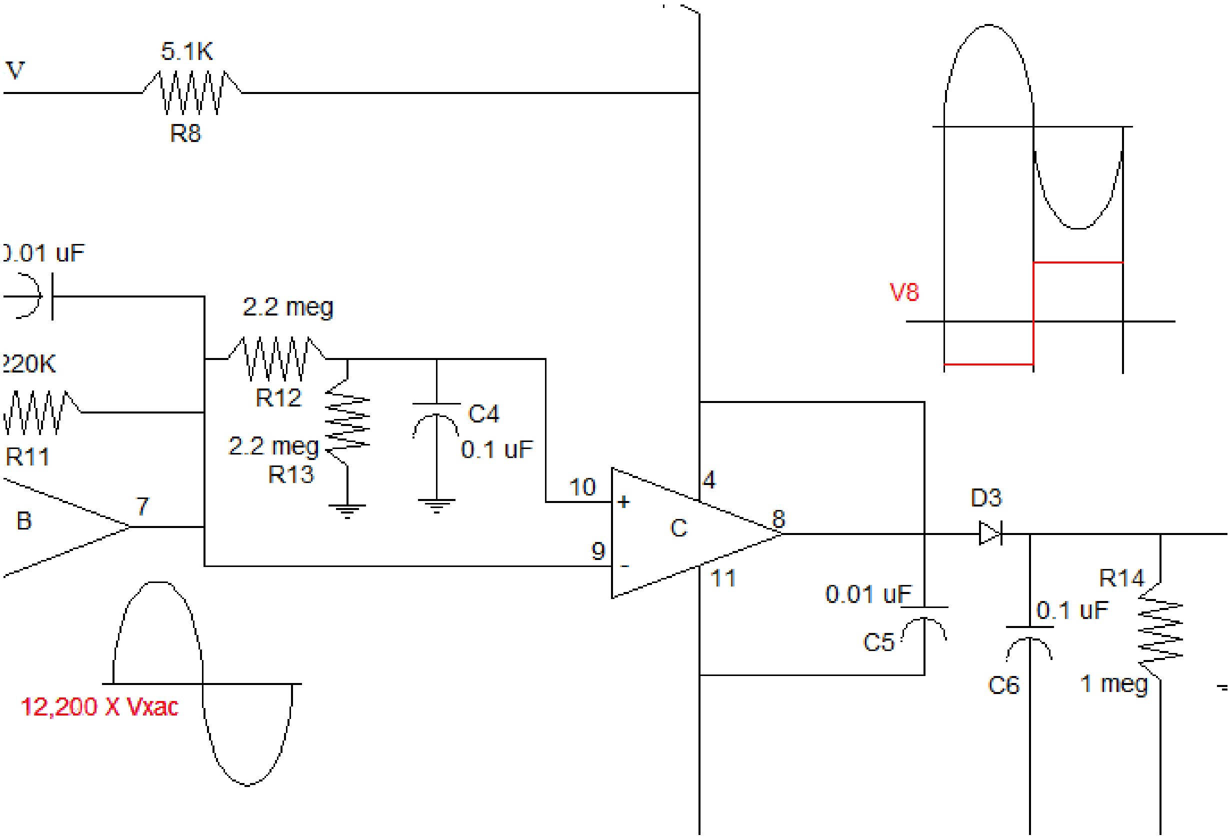


Starting at the left, I have V_{xac}. I have assumed it is sinusoidal and at a frequency that is far below the corner frequencies of the two voltage amplifiers. This is my worst case.

Since this noise is AC, it will be multiplied by 122 in the voltage amplifier and by 100 in the touchdown amplifier.



The amplified noise then goes straight into the inverting input of op amp C.



Assuming we are in the calibration phase, op amp C's output is at about -3V which causes the touchdown LED to be off. When the amplified noise voltage causes V9 to rise relative to V10, it has no effect on the output. But when the amplified noise voltage causesV9 to fall relative to V10, op amp C's output will go to +6V and turn on the touchdown LED. Not good.

Given that all offset voltage has been nulled and our R_x is 10 milli ohms, we should see a V10 - V9 of -12.2 milli volts DC.

A zero to peak AC voltage of 12.2 milli volts will therefore get us right to the edge of toggling op amp C. A zero to peak voltage divided by 1.4 equals its RMS value assume a sinusoidal wave shape. So if , then when my Digital Volt Meters reads $\frac{12.2\ milli\ volts\ zero\ to\ peak}{1.4}$ = 9 milli volts RMS, my touchdown LED is about to start flickering. This initial bias of -12.2 milli volts DC is my "noise margin".

Working backwards through the two amplifiers, this 9 milli volts RMS is equal to an AC noise signal at V_x of $\frac{9\text{ milli volts RMS}}{122 \times 100}$ = 0.7 micro volts RMS assuming the frequency is low enough. That is not much of a noise voltage. As the frequency goes up, the AC noise at V_x can be larger and still give the same voltage at V7.

If this noise voltage was caused by a noise current flowing in R_x, it would take $\frac{0.7\text{ micro volts RMS}}{10\ milli\ ohms}$ = 70 micro amps of noise current. Given that R_x is the cutter to spindle path through a lathe, it is hard to see how this current would flow since I don't see a closed path.

It is more likely that this 0.7 micro volts RMS noise voltage was picked up by induction[12]. The larger the area defined between the probes, the more induced voltage you would get. In order to minimize this noise pick up, I have run HC1, HC2, and SP1 inside a piece of coaxial cable. SP2 connects to the shield. This lets me run very short leads from the end of the coax to the clips and minimize noise pick up. I still have the area formed by the lathe but there isn't much I can do about that.

Recall that with an R_x of 0.01 ohms and a ΔR_x of -164 micro ohms, op amp C's input will be at -12.2 milli volts before touchdown and + 27.8 milli volts after touchdown. This is a change of voltage of 27.8 - (-12.2) = 40 milli volts. You could adjust the trim pot so the input to op amp C is at -39 milli volts during the calibration mode. Then at touchdown, this voltage would drop to +1 milli volts. That is enough drive to toggle the output. By doing this, you have raised the noise margin by a factor of $\frac{39}{12.2}$ = 3.2. The simple fact is that you will turn the trim pot until the touchdown LED stops flickering.

The bottom line here is that if the touchdown LED flickers when attached to a given machine, turning the offset pot until it stops flickering has a solid basis in the math. Just be careful to turn the trim pot too much or you will not be able to detect touchdown.

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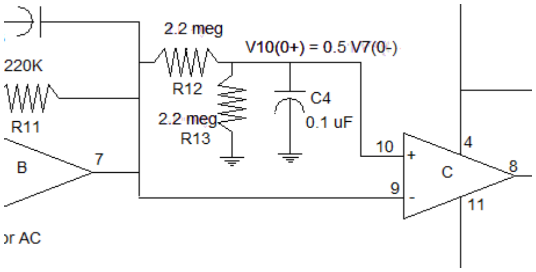
[12] The dominant frequency of noise measured in my shop was at 120 Hz.

Circuit Behavior As R_x Varies

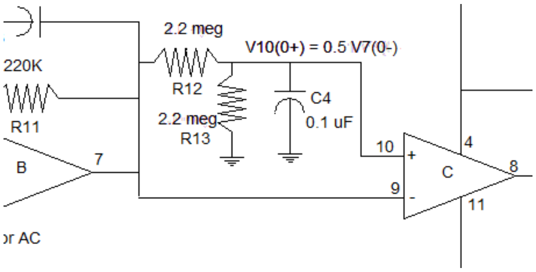


We took a close look at circuit behavior for R_x = 10 milli ohms but let's now look over a wider range.

I_x = 20 milli amps and $R_{touchdown}$ = 0.6 ohms.

		calibration phase:	**touchdown phase:**
Rx, milli ohms	delta Rx, milli ohms	V10(0-) - V9(0-) = -1.22 amps × Rx, mV	V10(0-) - V9(0+) = Ix × {(- 61 × Rx) - (12,200 × ΔRx)}, mV
5	-0.041	-6	4
10	-0.164	-12	28
15	-0.366	-18	71
20	-0.645	-24	133
25	-1.000	-31	214
30	-1.429	-37	312
35	-1.929	-43	428
40	-2.500	-49	561
45	-3.140	-55	711
50	-3.846	-61	877
55	-4.618	-67	1060
60	-5.455	-73	1258
65	-6.353	-79	1471
70	-7.313	-85	1699
75	-8.333	-92	1942
80	-9.412	-98	2199
85	-10.547	-104	2470
90	-11.739	-110	2755
95	-12.986	-116	3053
100	-14.286	-122	3364
200	-50.000	-244	11956
300	-100.000	-366	24034
400	-160.000	-488	38552
500	-227.273	-610	54845
600	-300.000	-732	hits neg sat

I advertise that the circuit works down to 10 milli ohms but you can see that it might work at 5. Before touchdown I have my comparator seeing -6 milli volts and after touchdown it sees +4 milli volts. As long as all offset voltages can been nulled and the AC noise is very close to zero, the circuit should work.



I_x = 20 milli amps and $R_{touchdown}$ = **0.6** ohms.

		calibration phase:	**touchdown phase:**
Rx, milli ohms	delta Rx, milli ohms	V10(0-) - V9(0-) = -1.22 amps × Rx, mV	V10(0-) - V9(0+) = Ix × {(- 61 × Rx) - (12,200 × ΔRx)}, mV
5	-0.041	**-6**	**4**

This case has the smallest drive to the comparator[13] but was not the point where the circuit was designed. That was done at 10 milli ohms. At larger values of R_x, all voltages are larger so the absolute values of the voltages is not as important.

I thought it was interesting to see that if I change $R_{touchdown}$ from 0.6 ohms down to 0.5 ohms, then the comparator sees -6 mV before touchdown and + 6 mV after touchdown. Nice and symmetric but due to a very tiny change.

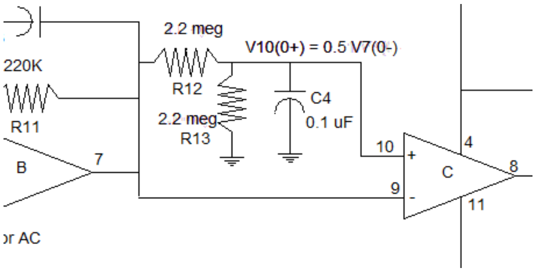
This tells me that it is not worth playing with the circuit just to get this symmetry.

I_x = 20 milli amps and $R_{touchdown}$ = **0.5** ohms.

		calibration phase:	**touchdown phase:**
Rx, milli ohms	delta Rx, milli ohms	V10(0-) - V9(0-) = -1.22 amps × Rx, mV	V10(0-) - V9(0+) = Ix × {(- 61 × Rx) - (12,200 × ΔRx)}, mV
5	-0.050	**-6**	**6**

---

[13] The op amp swings from about -3 to +6V. Typical gain is 100 dB which equals a voltage gain of 10^5. This means that 90 micro volts at its input is enough to move the output from -3V to +6V. So even 4 milli volts is plenty of drive.



I_x = 20 milli amps and $R_{touchdown}$ = 0.6 ohms.

Rx, milli ohms	delta Rx, milli ohms	**calibration phase:** V10(0-) - V9(0-) = -1.22 amps × Rx, mV	**touchdown phase:** V10(0-) - V9(0+) = Ix × {(- 61 × Rx) - (12,200 × ΔRx)}, mV
5	-0.041	-6	4
10	-0.164	-12	28
15	-0.366	-18	71
20	-0.645	-24	133
25	-1.000	-31	214
30	-1.429	-37	312
35	-1.929	-43	428

Note that the voltage applied to the comparator during calibration is related to R_x by a multiplier of -122. This means that a percentage change in R_x causes the same percentage change in the calibration phase voltage.

But the touchdown phase voltage is mostly a function of the change in R_x so goes up much faster than R_x. At an R_x of 5 milli ohms, my touchdown voltage is a miserly 4 milli volts. But when I double it to 10 milli ohms, my touchdown voltage is at a decent 28 milli volts. Going up to an R_x of 15 milli ohms gives us 71 milli volts at touchdown.

A Peak Into the Design Process

So far, you have seen just the answer. For some, this can be rather frustrating since there is no hint as to how we got here. Why this topology? Why these values?

In this section, I hope to share some of the messy process. I will spare you most of the false starts.

The brute force technique of applying a BIG current to a small resistance to generate a reasonably large voltage generated a lot of concern over damaging bearings inside the machine. Applying a small current to this small resistance generated a voltage that was difficult to see given the unknown offset voltage of each op amp. This offset voltage multiplied by the needed gain produced an output voltage large enough to saturate the final gain stage. So that was a bust.

But then it occurred to me that I really only cared about the change in R_x and not its absolute value. The change was abrupt so contained a lot of AC components. Why not have a low gain for R_x and a high gain for the change in R_x?

I wanted to have a test current, I_x, of 10 mA because it seemed very safe for bearings plus a minimum R_x of 5 milli ohms. I advertise a minimum of 10 milli ohms but wanted so margin.

Given a touchdown resistance of 0.6 ohms, this gave me a ΔR_x of -41.3 micro ohms. I assumed my comparator would need -10 milli volts of bias to stay in the low state. I would then give it a 20 milli volts step to get it to a bias of +10 milli volts. That would change its output to the high state and indicate touchdown.

So now I had all of the pieces: ΔR_x is -41.3 micro ohms, I_x = 10 milli amps, and ΔV at the comparator of 10 milli volts:

$$\Delta V = G_{ac} \times \Delta Rx \times Ix$$

So $G_{ac} = \frac{20\ milli\ volts}{(41.3 \text{ micro ohms} \times 10 \text{ milli amps})} = 48{,}430$

Assuming two gain stages, this meant that each stage had to have a gain of 220. That seemed kind of high. I doubled my test current to 20 milli amps and lowered my comparator bias to 5 milli volts:

So $G_{ac} = \frac{10\ milli\ volts}{(41.3 \text{ micro ohms} \times 20 \text{ milli amps})} = 12{,}110$

This can be done by two stages, each with a gain of 110.

Assuming a simple inverting amplifier, I liked having a feedback resistor of 220K which meant that the input resistor would need to be 2K. My junk box didn't have any of those. I had 1.8K and 2.2K. Using 1.8K, my gain was -122. Using 2.2K my gain was -100. Hmm. The overall gain came out to 12,200. So that worked OK.

So now I had the gain correct for amplifying the change in voltage related to a change in R_x. At first I used a potentiometer to adjust the threshold of the comparator but that was a real pain. Could I do this automatically?

After a bit of thought, I realized that R_x and ΔR_x are related. Could I use R_x to set the threshold? This would be most critical at minimum R_x so that is where I focused my design. After a few false starts, I came to the circuit I call my automatic threshold generator (see page31). It takes the amplified voltage generated across R_x before touchdown and stores it on a capacitor. Then when touchdown comes, it lets the change in voltage pass through to the inverting input of the comparator. I will not repeat the math here, but the result is that my threshold is proportional to R_x and my transition is proportional to the change in R_x. The calibration voltage out of my second gain stage had to be twice the needed bias due to the voltage divider formed by R12 and R13.

By this time I had also decided to try and use a simple inverting voltage amplifier for the first stage and a high pass amplifier for the second state. This second stage would have a gain of -1 at DC and a higher gain at AC when the change in input voltage came through at touchdown.

I still needed to decide how much of this gain to use to amplify the voltage across R_x during calibration.

With an R_x of 5 milli ohms and a test current of 20 milli amps, I get a voltage of 100 micro volts. I have already decided that I wanted an initial bias on the comparator of 5 milli volts which translates to a voltage out of the second gain stage of 10 milli volts. Call this voltage V7. This told me that my DC gain had to be

$$V7 = G_{dc} \times Vx$$

$$V7 = G_{dc} \times Rx \times Ix$$

$$10\ milli\ volts = G_{dc} \times 5\ milli\ ohms \times 20\ milli\ amps$$

So $G_{dc} = \frac{10\ milli\ volts}{5\ milli\ ohms \times 20\ milli\ amps} = 100$. This is the gain of my first voltage amplifier.

G_{ac} = 12,110 which equals the gain of my first amplifier times the gain of my second amplifier. But the first amplifier's gain is just G_{dc} so my second amplifier must be $\frac{12{,}110}{100} = 121$.

You may notice that I chose to have the first amplifier set to 122 and the second amplifier set to 100. My G_{ac} is about right but my G_{dc} is a bit high. I did this for two reasons. I wanted the time constant formed from C2 and R10 to be as large as possible. Using an R10 of 2.2K was therefore a little better than using the value of 1.8K. The second reason had to do with noise margin. The slightly higher DC gain set my calibration phase bias of op amp C a little higher than if G_{dc} was 100. That gives me a 22% increase in noise margin at my 5 milli ohm level.

After some testing in a shop with commercial grade machines, I decided to add feedback capacitors to reduce the sensitivity to noise.

The pulse stretcher was added because I wanted to insure a minimum time the touchdown LED would be on. I also had a spare op amp just screaming to be used.

I then turned to human factors concerns. In my shop, I really hate having to remember to turn battery powered devices off. So the circuit had to automatically power up when used and power down when put away. Well worth the single resistor and single transistor to do this function. I also wanted a power on indicator but didn't want to waste any current on it. By putting it in series with R_x, I got to use an existing current. This LED also gave me a voltage, Vee, that was about 3V below ground. That enabled the first gain stage output to swing below ground to almost -3V.

Going with the Kelvin connection solved two problems. It eliminated uncontrollable voltage drops in the probes which could obscure the voltage I was trying to measure. It also let me completely disconnect HC1 from the rest of the circuit to insure no leakage currents could flow. This current could partially turn on power and run down the battery.

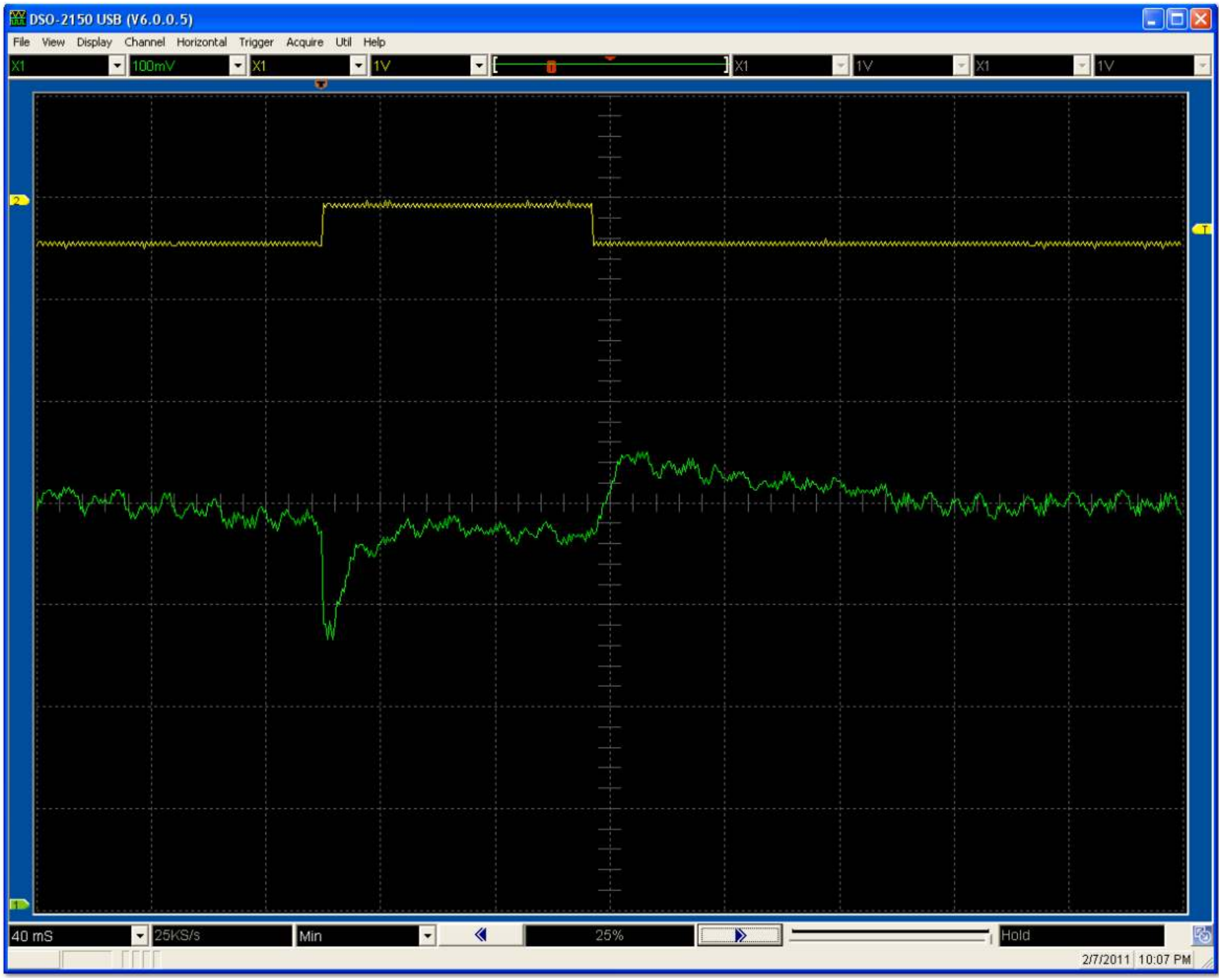
I have not fully explained why or how I chose this topology. It was not my first choice but evolved as I saw trouble with the previous design. Parts, like the feedback capacitors, were added based on field experience. But most came from just thinking about how it should work.

One word about component values. I tried my best to minimize the number of different values. This should make buying parts a little easier.

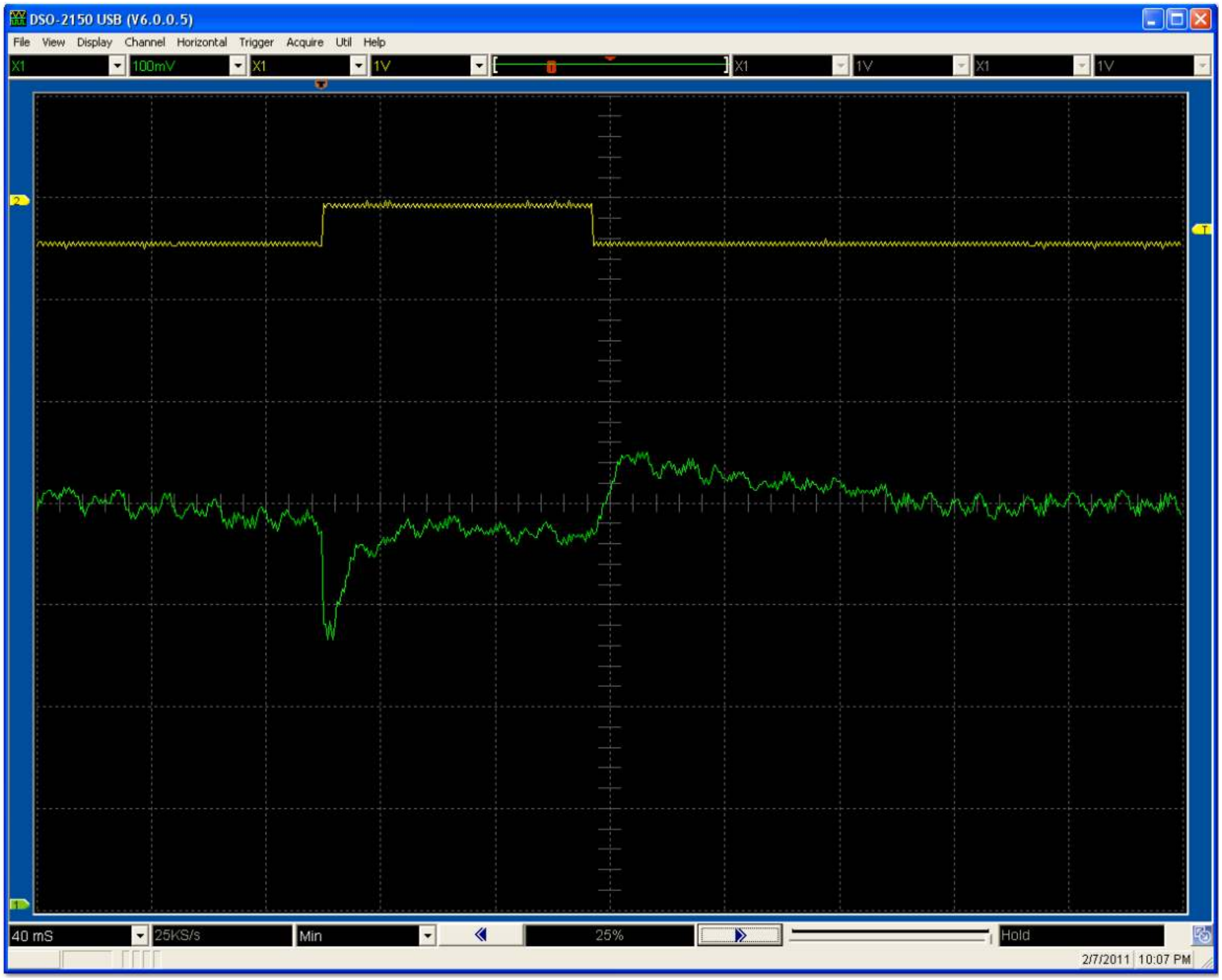
Scope Pictures

The top trace represents the voltage at the top of R15. When this voltage is low, the touchdown LED is off. When high, the LED is on.

The bottom trace is the voltage out of the touchdown amplifier.



First of all, look at the peak to peak noise voltage. It is about 0.2 divisions which at 100 milli volts per division equals 20 milli volts. Reflected back to V_x this is $\frac{20\ milli\ volts}{12{,}100} = 1.65$ micro volts. This is comparable to the voltage picked up by a radio antenna.

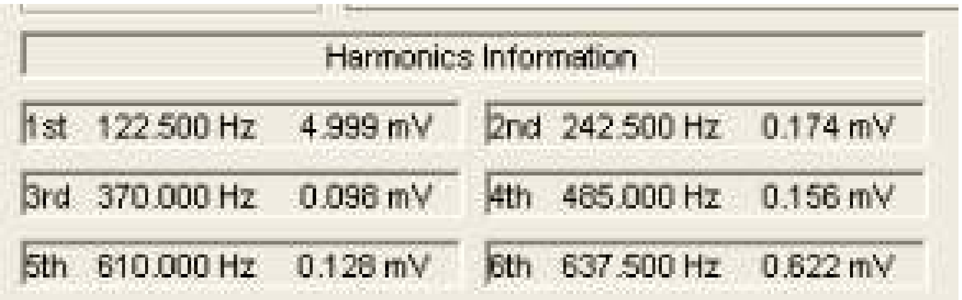


Next, look at ΔV7. It is about 1 division so 200 milli volts. Reflecting this back to V_x it is $\frac{200\ milli\ volts}{12{,}100}$ = 16.5 micro volts. This means our ΔR_x = $\frac{16.5\ \ micro\ volts}{20\ milli\ amps}$ = 0.8 milli ohms. This is not the minimum value the circuit can detect yet is certainly tiny.

So our signal is 16.5 micro volts and our noise is 1.65 micro volts. This is a signal to noise ratio of 10 to 1. That is a good thing.

And finally, look at how long the touchdown LED is lit. I read 2.5 divisions at 40 milli seconds per division = 0.1 seconds, as expected.

V7 jumps high when the LED turns off. I suspect this is crosstalk within the circuit. It is not due to breaking touchdown. Going to a copper circuit board with lots of ground should help reduce this cross talk although it is harmless.

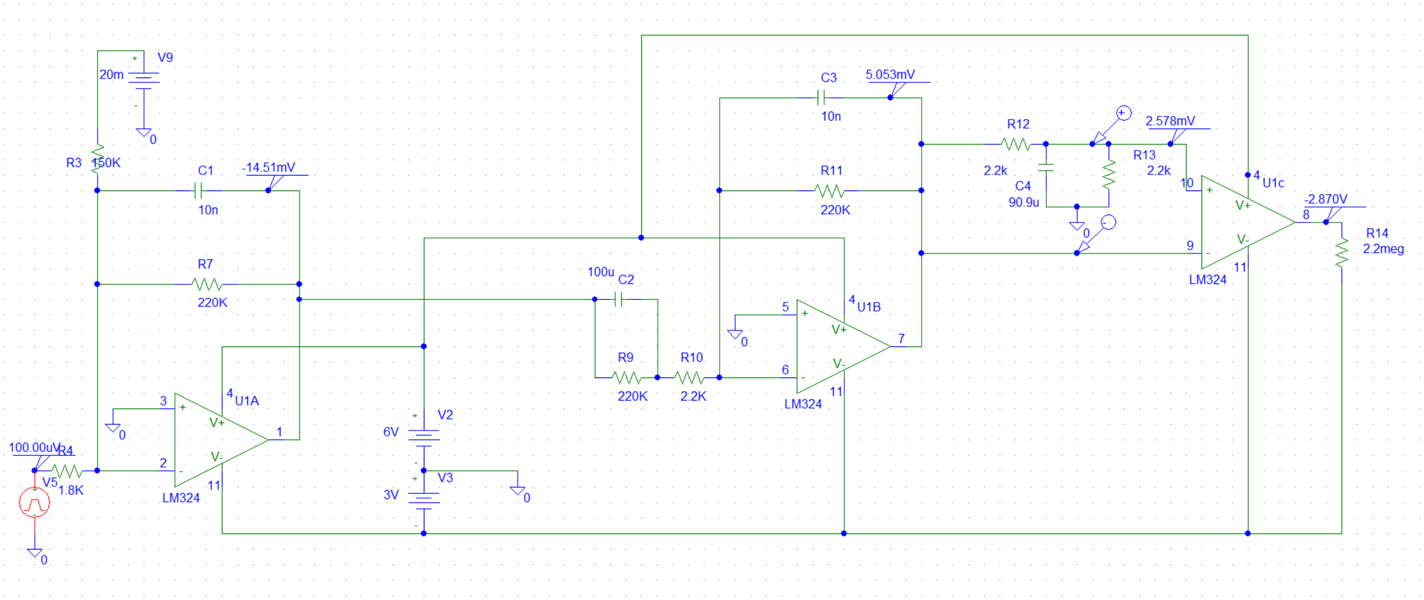


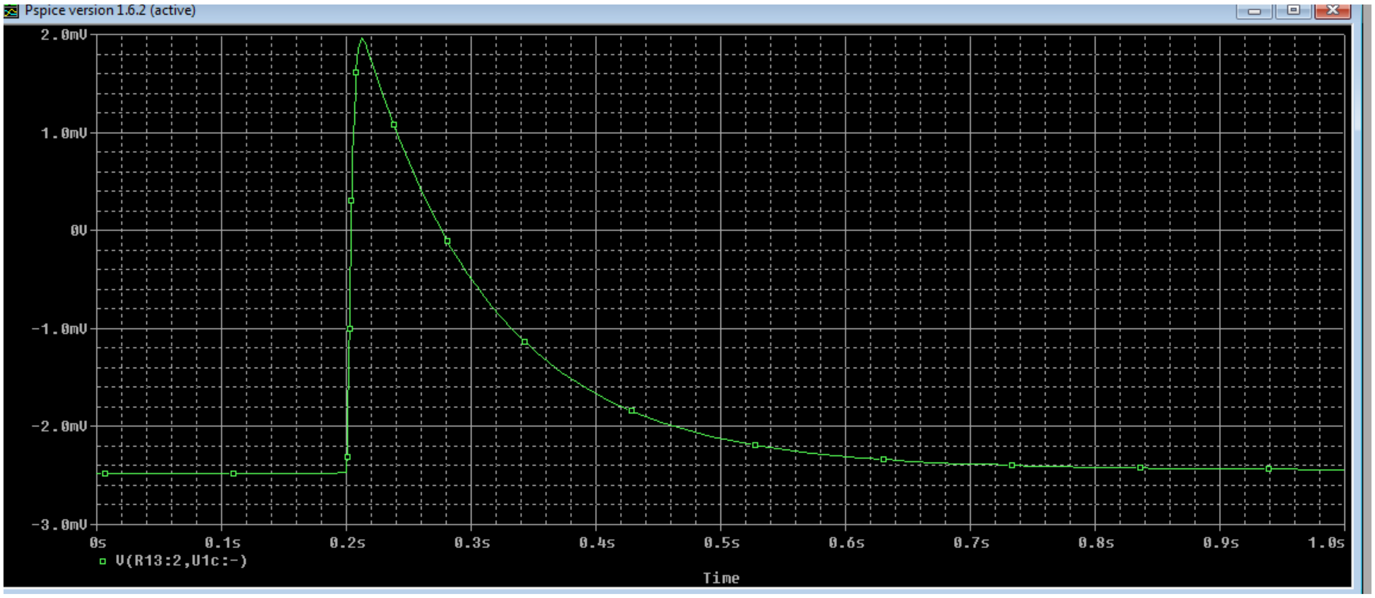
Harmonics Information

1st 122.500 Hz 4.999 mV
2nd 242.500 Hz 0.174 mV
3rd 370.000 Hz 0.098 mV
4th 485.000 Hz 0.156 mV
5th 610.000 Hz 0.128 mV
6th 637.500 Hz 0.622 mV

My scope has the ability to do a Fast Fourier Transform of the signal. It shows that most of the noise is the second harmonic of 60 Hz. Recall that both of the voltage amplifiers have a corner frequency of 72 Hz so this 122.5 Hz has already been attenuated by more than half.

PSpice 9.1 Simulation



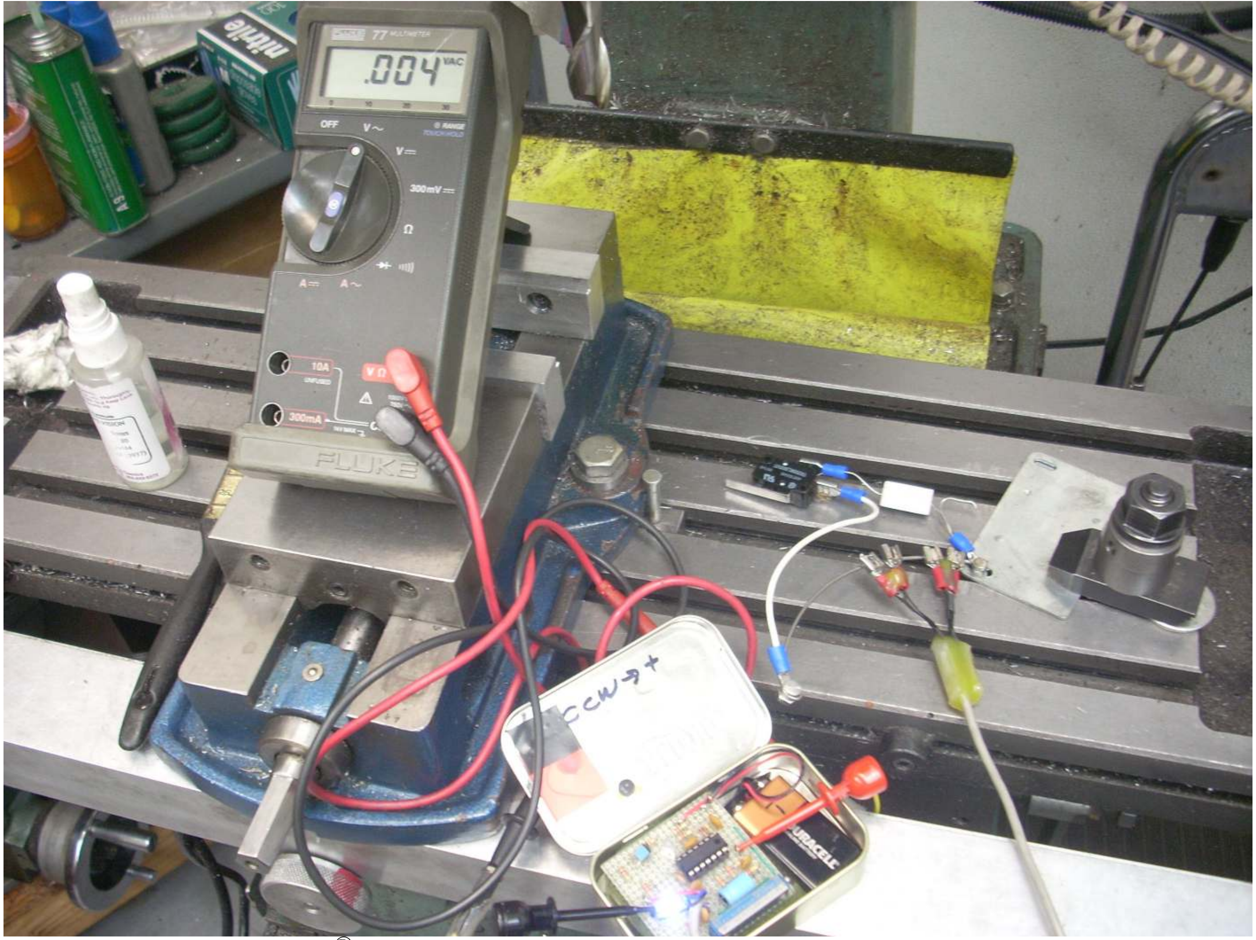


With the PSpice version 9.1 simulator I was able to see the minimum width of the pulse out of the touchdown detector. I used the minimum V_x and ΔV_x. The time that V10 - V9 is greater than 0V is more than 0.5 seconds. So we have no problem meeting our need of having it high for more than 42 micro seconds as was assumed on page 59.

Due to excessive input bias currents in the simulated op amps, I had to drop my R12 and R13 values by a factor of 1000. I compensated by raising C4 by the same factor.

The Test Bed

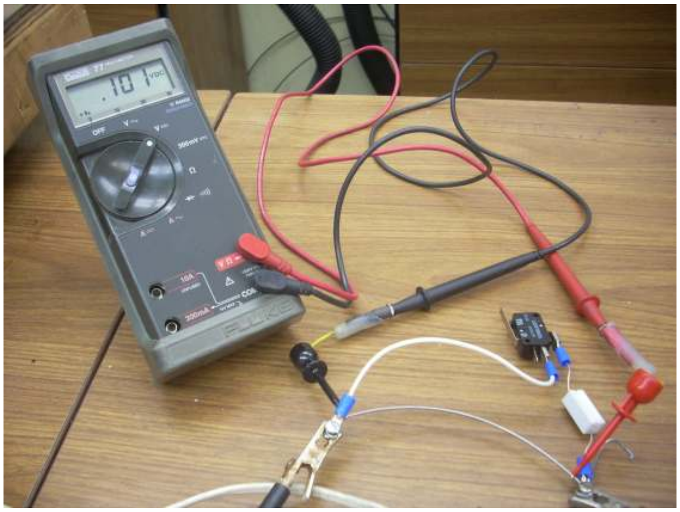
Nothing will drive you crazy faster than inconsistent results. I learned early on to set up a test bed for assessing the circuit's behavior that had minimum noise. I chose the table of my mill.



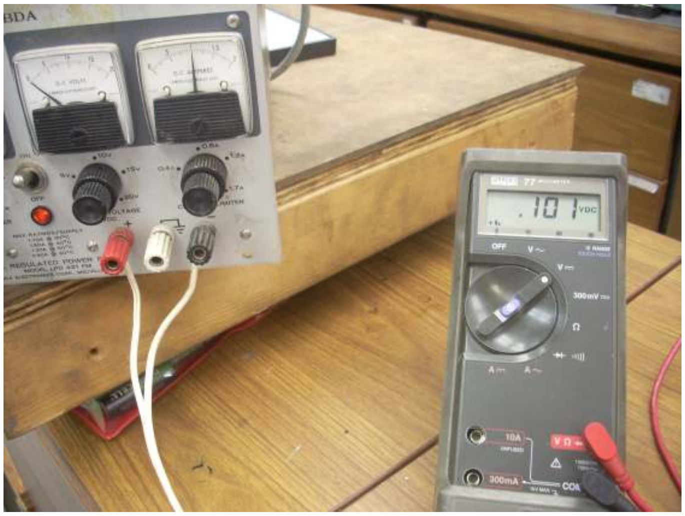
I use a piece of Kathal® wire as my variable R_x. One end is bolted to a steel plate which is clamped to the table. This provides a minimal and consistent noise level. I can't read the noise on the wire directly but by monitoring V7 with my meter set to AC RMS, I do get a relative sense. You see here that the meter reads 4 milli volts AC RMS. If I touch the grounded end of the wire, the reading does not change. If I touch the far end of the Kathal wire, I see a jump of about 50 milli volts RMS and the touchdown LED flashes.

If I take the meter reading as accurate, then this translates to a voltage at the probes of $\frac{4\ milli\ volts\ AC\ RMS}{12{,}200}$ = 0.3 micro volts. That 50 milli volts RMS reading translates to a probe voltage of 4 micro volts. So it should not be a surprise that the circuit has trouble in a noisy environment.

I keep the probes close to the ground clip and keep my switch in series with my touchdown resistor laying on the mill table. This gives me a consistent result free from random noise.



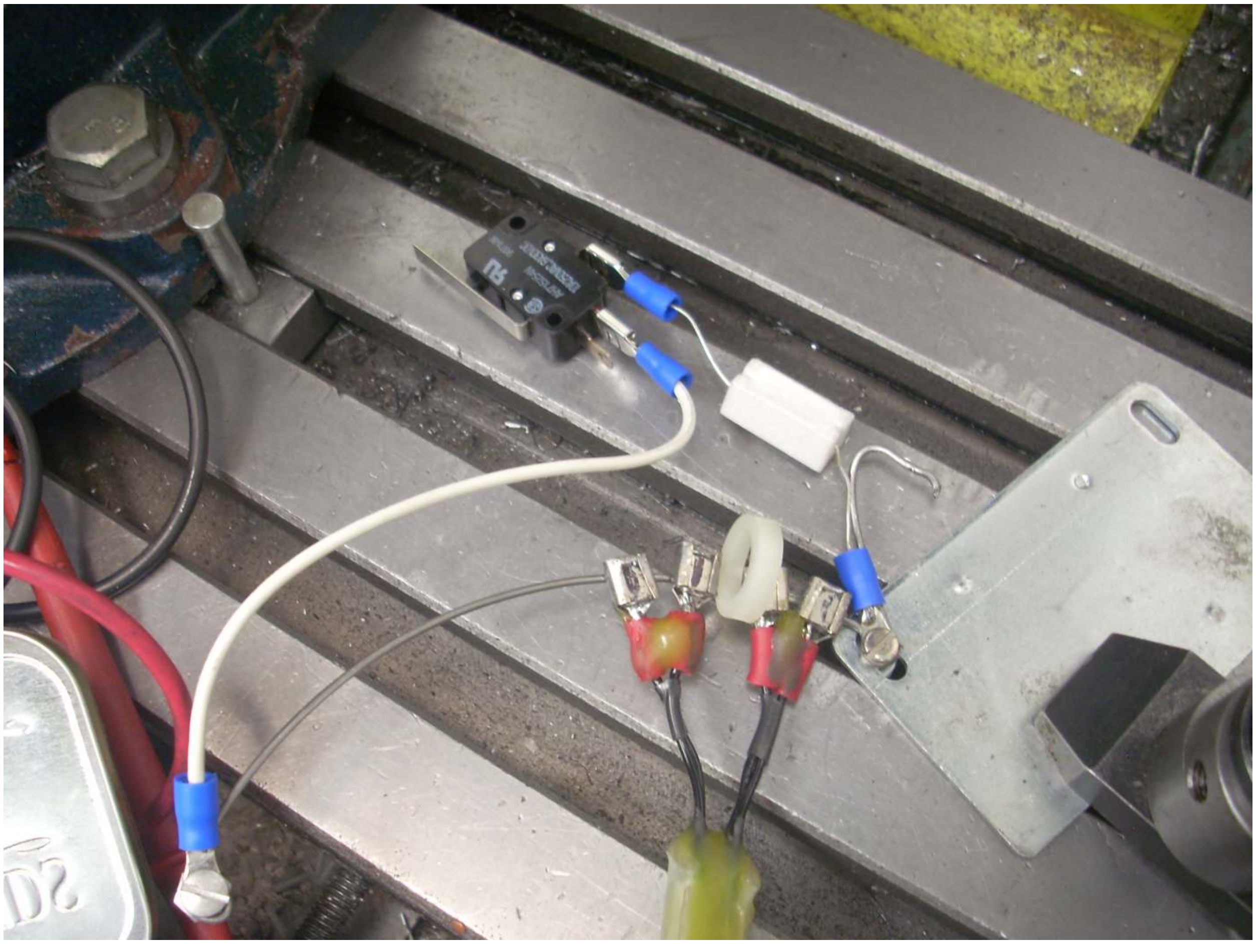
I measured the Kathal wire by applying 1 amp through it and measuring the voltage drop. Here you see the DVM reading 101 milli volts.



My power supply is putting out 1 amp. Note that I set the current by adjusting the voltage, not be adjusting the current limit. My current limit is set for 2 amps.

I measured the distance between the volt meter clips and found it to be 118 mm. This boils down to 0.88 milli ohms per mm.

Since the inner two clips on my probes are SP1 and SP2, I can measure the distance between them and calculate the R_x that they see.



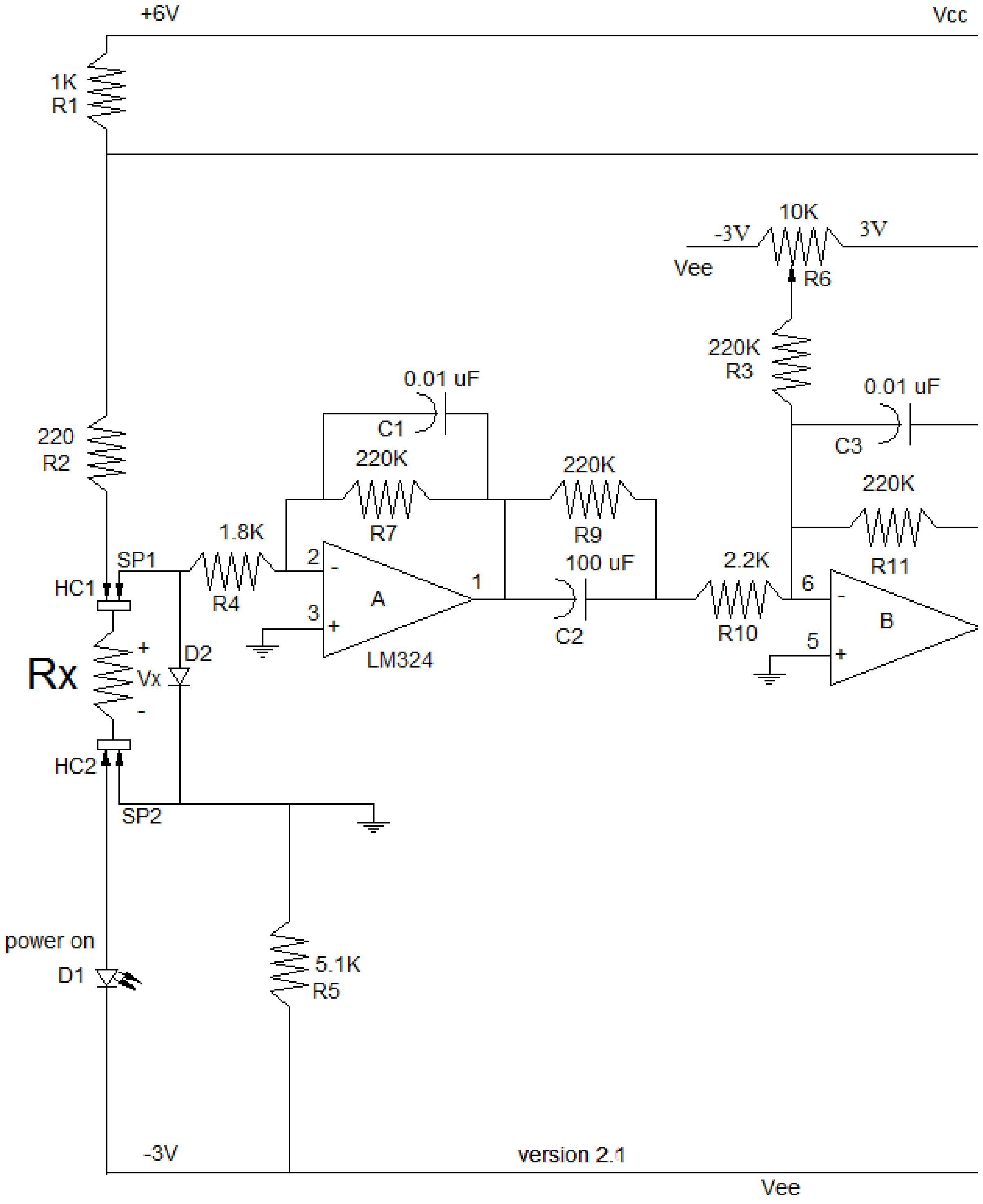
In this test I am using a nylon washer that is 4.4 mm thick. This translates to an R_x of 4.4 mm X 0.88 milli ohms/mm = 3.9 milli ohms. I was able to close the switch that simulates touchdown 50 times and saw the touchdown LED flash without fail.

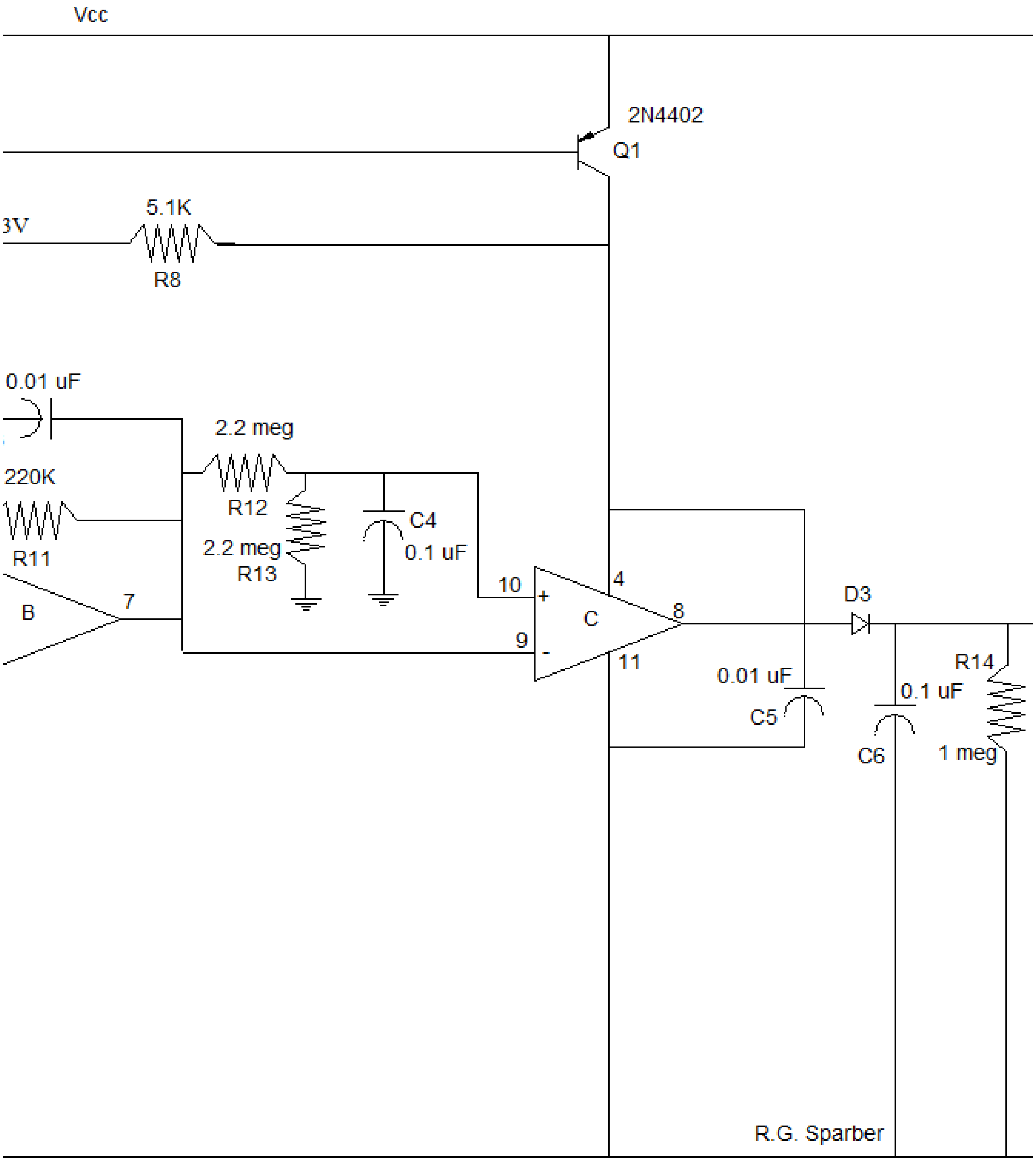
Given that R_x = 3.9 milli ohms and my touchdown resistor equals 0.47 ohms, I figure ΔR_x at -32 micro ohms. This means a ΔV_x = -0.6 micro volts.

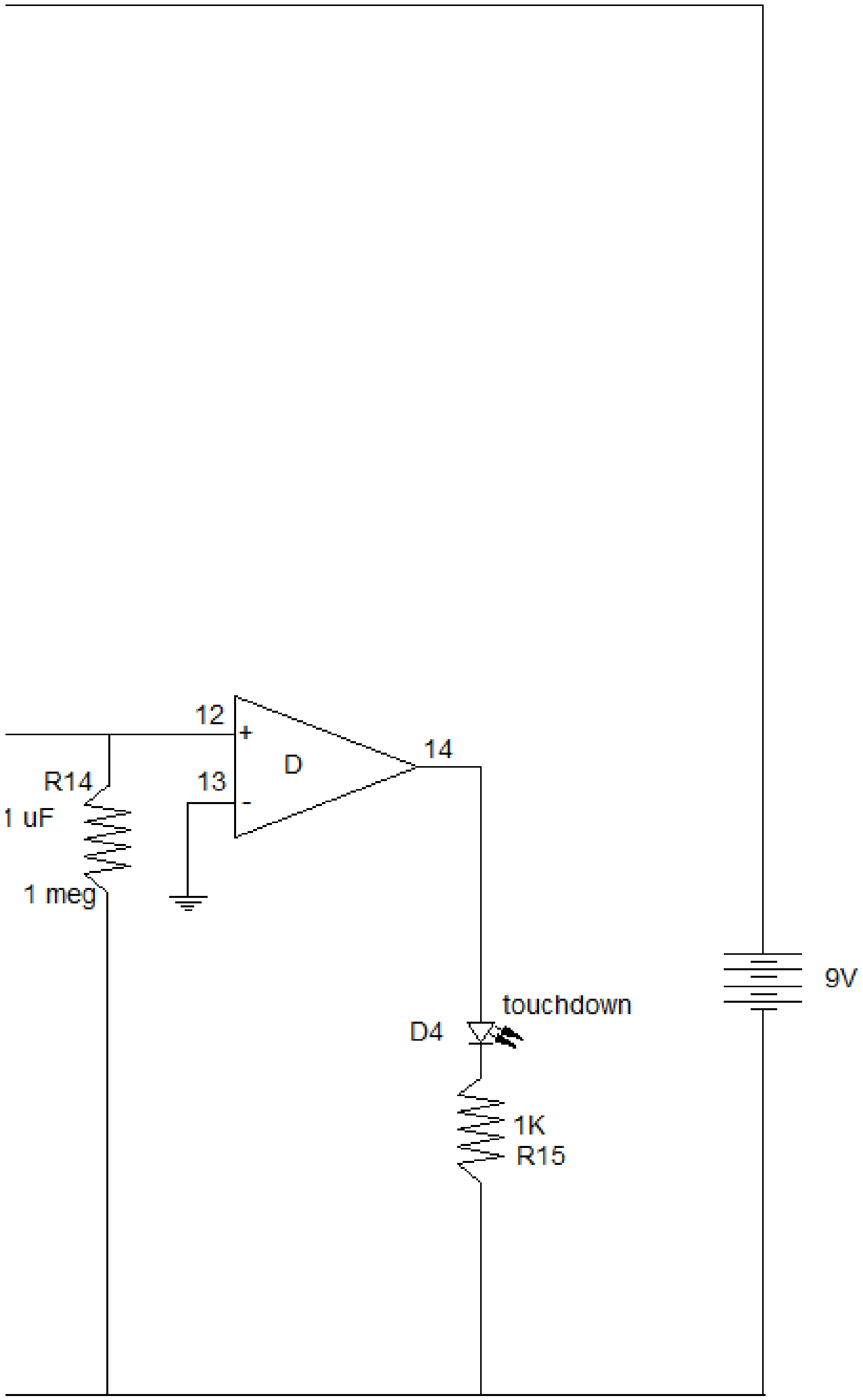
When I measured V7 I saw 103 to 113 milli volts DC. This means a V10 - V9 voltage of about -50 milli volts. Assuming that all offset voltages have been nulled and there being no noise, I expect a ΔV_x of $\frac{-50\ milli\ volts}{12{,}200}$ = 4 micro volts would be needed to cause the touchdown LED to flash.

Comparing my calculated ΔV_x from resistance with what I expect from circuit behavior, there is a discrepancy of about 3.5 micro volts. So if there was $\frac{3.5\ micro\ volts\ AC\ zero\ to\ peak}{1.4}$ = 2.5 micro volts RMS induced into my test bed at the probes, it would explain this discrepancy. Seems reasonable to me.

Magnified Schematic




Vcc
2N4402
Q1
3V
5.1K
R8
0.01 uF
2.2 meg
R12
220K
R11
2.2 meg
R13
C4
0.1 uF
B
7
10
+
9
-
C
4
8
11
D3
0.01 uF
C5
R14
0.1 uF
C6
1 meg
R.G. Sparber


12
+
D
14
R14
1 uF
13
-
1 meg
9V
touchdown
D4
1K
R15

Acknowledgements

Thanks to Dan Benoit for both his encouragement and use of his shop to test out this design. Thanks to Gregg Kricorissian who critiqued the circuit. Thanks to Andy Wander for suggesting the coax cable between box and probes. Although I took it out, it may still go back in at a later date. Thanks to Mark Cason for the through hole artwork. And finally, thanks to all those on the Yahoo groups valleymetal, mill_drill and atlas_craftsman for countless comments and suggestions.

These generous people again demonstrate that “all of us are smarter than any one of us”.

I welcome your comments and questions.

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