Lathe Electronic Edge Finder, Model 2, version 1.0

By R. G. Sparber

Introduction

The Lathe Electronic Edge Finder (LEEF) is able to detect the cutter coming in contact with the workpiece. No modification to the lathe, cutter, or workpiece is needed[2]. One magnetic probe is placed onto the chuck and another one near the cutter. An LED lights when the tip of the cutter just touches the surface of the workpiece.

This task would be very simple if the cutter was insulated from the rest of the lathe. But in most cases, this is not an attractive solution.

The LEEF is intended to be used on a previously turned surface. For example, a work piece is chucked up and turned true. Then there is call for a change in cutters. By changing the cutter, the initial zero set is likely lost. By using the LEEF, you quickly reestablish zero with the new cutter.

Since there is no need for the user to see the point of contact between cutter and workpiece, the LEEF can also be used with a boring bar to touchdown at the bottom of a hole.

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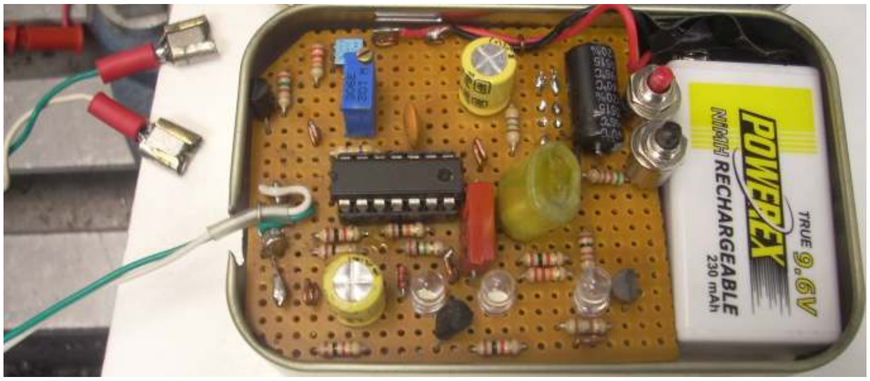
[1] You are free to copy and distribute this document but not change it. My intent is to do defensive publishing (http://en.wikipedia.org/wiki/Defensive_publication).

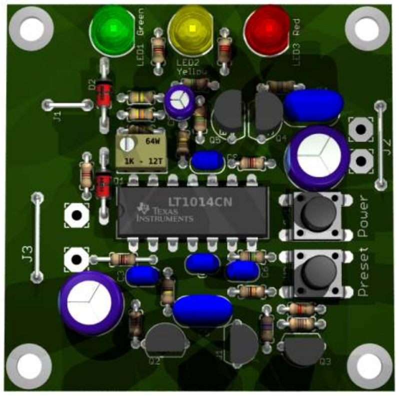
[2] Cutter and workpiece must be electrically conductive.

This article is part of a series[3] that explains the problem being solved, previous iterations of this circuit, plus shop performance. I will focus only on this new design. To see the LEEF in action, watch the video at http://www.youtube.com/watch?v=4ilY9Pse7TU&feature=youtu.be



This is my prototype.

The next iteration was a single sided circuit board:



This is a computer generated picture created by Mark Cason showing one version of the finished product.

The board is 50 mm on a side. Not shown is the top side ground plane.

This is the first hand etched and drilled prototype that used almost the same artwork except is single sided. I didn't have an amber LED so used a green one for the center LED.

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[3] See http://rick.sparber.org/ueef.pdf for a description of the problem being solved. See http://rick.sparber.org/ma.htm#6 for the complete set of articles.

The LEEF detects the change in resistance between cutter and workpiece caused by the cutter coming in contact with the workpiece. No modification to the lathe is needed. The resistance before touchdown can be as low as 0.01 ohms or as high as 1.4 ohm. This is essentially the resistance of the spindle bearings. The test current is 20 mA. Induced 60 Hz noise can be 40 times larger than the signal being detected.

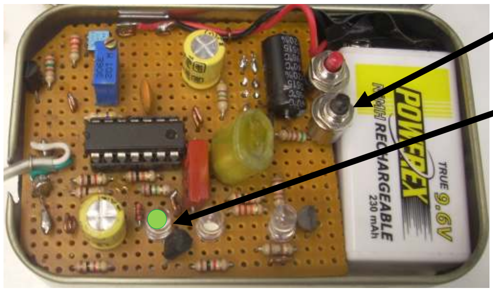
A LEEF, Model 1 also exists. It performs the same task as the design presented in this article but has a minimum resistance threshold of 2 ohms. If your machine has a spindle bearings resistance above 2 ohms, then the Model 1 is a better choice. It is easier to use and costs much less than the Model 2. For details see

http://rick.sparber.org/LEEF_Model_1.pdf

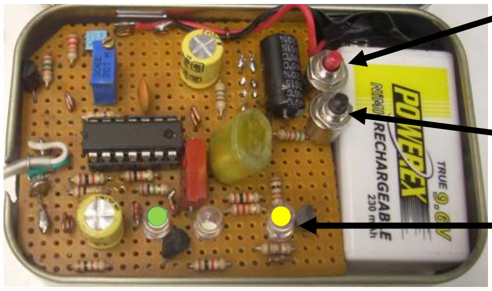
Operation

All contact surfaces must be free of oil. I use alcohol as a cleaner. Drop one probe on the chuck and the other near the cutter. These probes are magnetic. Be careful not to bump the probes as this will disrupt the reading.

Note: With the updated circuit, you only push the power on once and the reset once. A newer version eliminates the reset button and you only push the power on button.



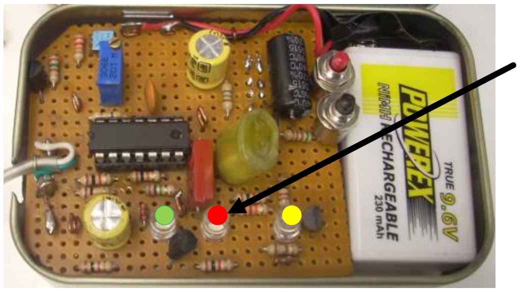
Press the ***Power On and Reset*** button. This brings up power and lights the ***Power On*** LED.



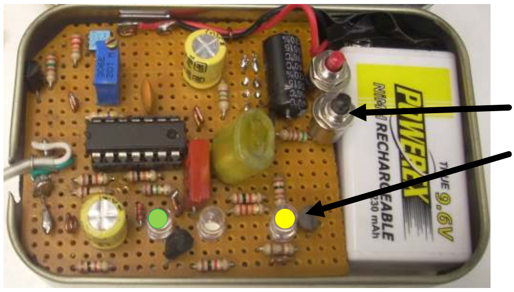
Then press the ***Preset*** button to stabilize the high gain amplifier.

A second push of the ***Power On and Reset*** button causes the circuit to reset.

Within a few seconds the ***Ready*** LED starts to flash.

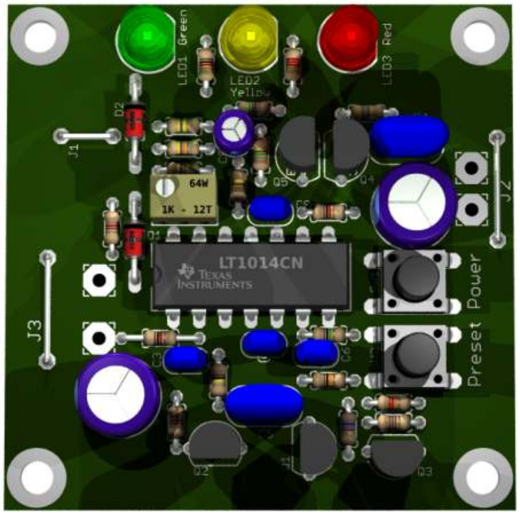


Feed in the cutter until the ***Touchdown*** LED flashes on. Zero your lathe's cross slide dial.



If you wish to repeat the touchdown event, back the cutter out and press the ***Power On and Reset*** button again. The ***Ready*** LED will start to flash within a few seconds to say the LEEF is ready to detect another touchdown event.

When done, remove the probes from the lathe which also turns off LEEF power.



The Ready and Touchdown LED tell you about your shop environment and machine. If the Ready LED (shown here in amber) is steady, it says you have no electrical noise induced in your machine or your spindle resistance is greater than 1.3 ohms. The more it flickers, the more noise you have. If the flicker rate is constant, then your noise is constant. If it is random, you have a lot of voltage spikes. Spikes are commonly generated by motors turning on and off quickly. Excessive noise will interfere with proper circuit operation.

The Touchdown LED (shown here in red) in conjunction with the Ready LED tells you about your machine. If you touchdown, see the touchdown LED flash, and back away from touchdown, the time it takes before the Ready LED starts to flash again will tell you about your spindle resistance. The longer this time delay, the larger the resistance. A time delay of 3 seconds correlates to the minimum spindle resistance and maximum contact resistance between cutter and workpiece. The longer the time delay, the larger the spindle resistance. Of course, normally you would press the reset button if the wait time became excessive.

If at touchdown both the Ready and Touchdown LEDs turn on, your spindle resistance is greater than 1.4 ohms. This just means that you probably could have used the LEEF, Model 1 which cost less and is easier to use.

The power on LED (shown in green) will come on even when the battery is too low for the circuit to work. So if the LEEF begins to behave abnormally, try replacing the battery. Given a 200 mA-H battery, you should get about 8 hours of run time from a battery.

Requirements and Goal

My design requirements were:

1. Be able to detect the touchdown down of the cutter on the workpiece when the spindle resistance is as small as 0.01 ohms and the contact resistance of the cutter on the workpiece is no greater than 0.6 ohms.
2. Maximum spindle resistance shall be 1 ohm.
3. Be able to detect touchdown even when 100 microvolts of 60 Hz is present across the unknown resistance.
4. Test current less than 25 mA.
5. Open circuit test voltage less than 9V.
6. Automatic power down.
7. Less than 10 seconds delay from power up until ready to detect.
8. Less than 10 seconds delay from previous touchdown event until ready to detect again.
9. Visual indication of the circuit's state
   a. Power
   b. Ready to detect
   c. Touchdown
10. Run on a single 9V battery.

My design goal were

1. minimize the complexity and cost of the circuit.
2. Detect touchdown for spindle resistances greater than 1 ohm and contact resistance of the cutter on the workpiece of no greater than 0.6 ohms.

With the latest updates, I can now detect resistances above 5 ohms and still see the READY LED light before touchdown.

Background

Many problems had to be solved in order to make a circuit that could detect touchdown when the spindle resistance is as low as 0.01 ohms and the contact resistance is no more than 0.6 ohms. This situation causes a change in resistance of as little as 163 micro ohms. Furthermore, it has to do it in an industrial environment which has a high level of 60 Hz noise. The journey of this development was very much like pealing an onion. I would solve one problem, cause or reveal a new one, and then start the cycle over. Eventually I did get to the point where all known problems were solved and no new ones became evident. That does not mean that new problems won't pop up in the future.

The circuit is reacting to an input as small as 3.2 micro volts. This demands special attention to board layout to minimize internal noise pick up and DC voltage drops across critical circuit paths. Yet I had to connect the high sensitivity parts of the circuit to the high noise generating parts without unwanted interactions. At one point in the circuit I have a voltage transition of 8V yet if it generated a 1.6 microvolt spike at the input, the LEEF would not operate correctly.

I was extremely fortunate to have Mark Cason designing the layout. The layout is the single most important analog component.

Contents

Limitations

All designs have limitations. In this case, I am trying to detect a very tiny drop in voltage while a large amount of AC noise is present. Any noise that drops in level over a brief amount of time may look identical to the signal I am trying to process. So given an environment with a lot of voltage spikes, the circuit might not work unless its sensitivity is reduced.

Such is the case around at least some CNC machines. My limited experience with these machines tells me I might be able to ignore their spikes but can't guarantee it. Filtering in my circuit that would block these spikes would also tend to block the signal I am trying to detect. An alternative is to set the threshold so it takes a bigger change in resistance to trigger a touchdown indication. So we are up against the physics of it[4] and any solution must first change the situation so different conditions exists. It is a classic case of "if you can't solve the problem, change the problem".

The circuit can function correctly until the spindle resistance gets above 1 ohm. Above 1 ohm, the circuit will still detect touchdown as long as the touchdown contact resistance is no more than 0.6 ohms. The touchdown indication will be different: the Ready LED will ***not*** flash to indicate the circuit is ready to detect touchdown. Then at touchdown, both the Touchdown and Ready LEDs light at the same time. Do understand that if you have a spindle resistance above 1 ohm, a much simpler circuit will do the job. See http://rick.sparber.org/rctf.pdf for details.

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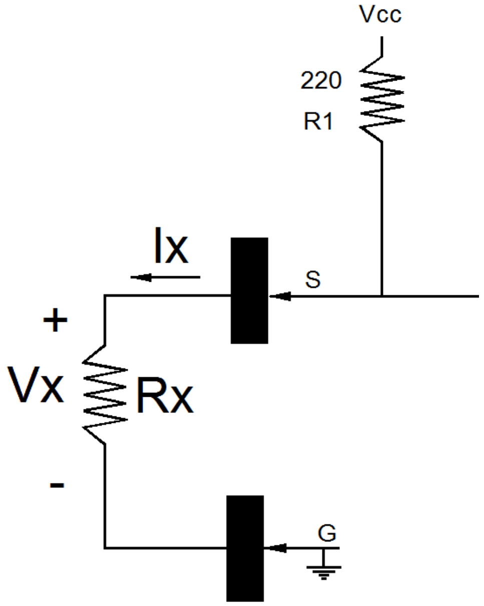
[4] By this I mean up against things that cannot be solved.

Theory and Practice

You can find the complete schematic on page 20. Leading up to this point, I will present the design evolution and theory.

Isn't It Just an Ohm Meter?



I am just trying to measure a resistance. To do this I have a DC voltage, V_{cc} which feeds through a resistor, R_1, and into my unknown resistance, R_x. A test current, I_x, flows. Between test probes "S" and G" I am able to measure the resulting voltage across my unknown resistance. It is called V_x.

$$V_x = R_x \times I_x \tag{1}$$

So if I simply generate I_x and measure V_x, I can calculate R_x.

$$R_x = \frac{V_x}{I_x}$$

R_x has two values. Before touchdown it can be as small as 0.01 ohms. After touchdown it can be as small as 0.009836 ohms.

If $I_x = 1$ amp, then

$$V_x = R_x \times I_x$$

$$V_x = 0.01\ ohms \times 1\ amp$$

$$V_x = 0.01\ volts$$

So before touchdown we measure 0.01 volts. After touchdown R_x equals 0.009836 ohms which generates 0.009836 volts when 1 amp flows through it. Although these are tiny voltages, it is possible to measure them with a good Digital Volt Meter.

We have solved the problem of detecting touchdown but created a new one. Passing 1 amp through bearings can cause damage due to arcing.

Drop The Test Current Down

OK, let's drop the test current down to a more reasonable level. How about 0.02 amps which can also be written as 20 milli amps or 20 mA? From equation (1) we can calculate V_x again.

$$V_x = R_x \times I_x \qquad (1)$$

$$V_x = 0.01\, ohms \times 20\, mA$$

$$V_x = 0.0002\, volts \;\; or\; 200\, micro\; volts$$

Damn! Now this is a very tiny voltage. Furthermore, after touchdown we will see

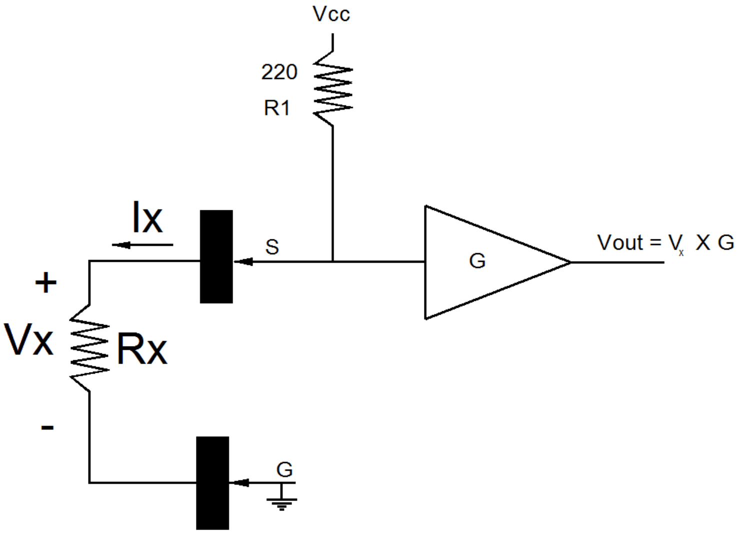
$$V_x = 0.009836\, ohms \times 20\, mA$$

$$V_x = 0.0001967\, volts \;\; or\; 196.7\, micro\; volts$$

a change of only 3.28 micro volts. To give you an idea how tiny this is, consider that the voltage picked up from a small radio antenna is just a few micro volts.

We have solved the problem of detecting touchdown without damaging the bearings but created a new one. How do we detect these two tiny voltages that are so close together?

Add A Voltage Amplifier



OK, let's use a voltage amplifier to raise the voltage back up to a reasonable value.

My voltage amplifier will take the voltage V_x and multiply it by a constant I call G, for gain (not to be confused with test probe "G", for ground). The resulting voltage, V_{out}, will follow the equation

$$V_{out} = V_x \times G$$

$$V_{out} = (R_x \times I_x) \times G \qquad (2)$$

If I pick a gain, G, of 10,000, then my pre-touchdown V_{out} will be

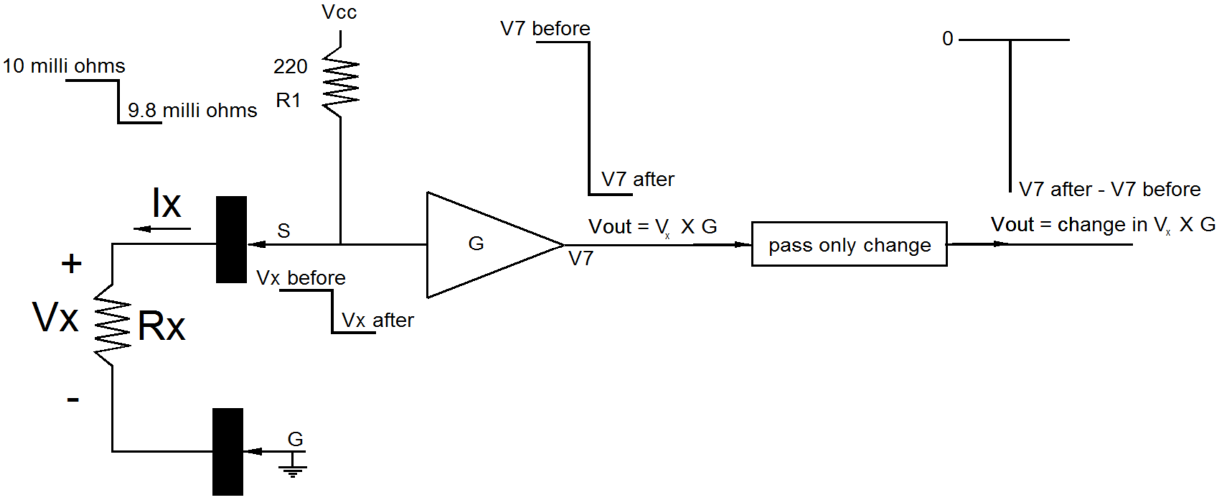
$$V_{out} = (R_x \times I_x) \times G$$
$$V_{out} = (0.01\, ohms \times 20\, mA) \times 10{,}000$$
$$V_{out} = (200\, micro\, volts \times 10{,}000)$$
$$V_{out} = 2\, volts$$

And after touchdown my V_{out} will be 1.97 volts.

We have solved the problem of measuring a tiny V_x but now have a new problem. What if my pre-touchdown resistance changes just a little. I will then get a different V_{out}. I would have to record this pre-touchdown voltage and use it to determine if my touchdown voltage was right. That sounds like an annoying bit of work.

Block What I Don't Want

Let's think about this a little. Do I really care what value I get for V_{out}? Don't I really care only that it changed? After all, my goal is to detect touchdown which is a drop in resistance. R_x equals 0.01 ohms or 10 milli ohms before touchdown and 0.009836 ohms or 9.836 milli ohms after touchdown. Before touchdown, V_x equals 10 milli ohms times my test current I_x. After touchdown V_x equals 9.836 milli ohms times my test current. In other words, V_x will drop in value during touchdown. This drop occurs extremely fast since the cutter is not touching the workpiece one instant and is touching in the next instant.



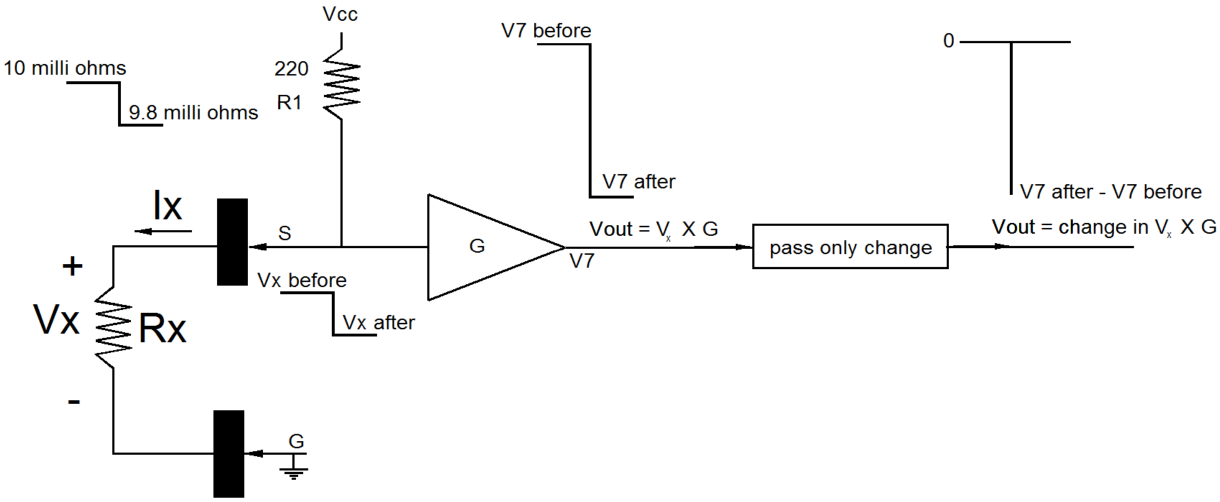
V_x passes through our voltage amplifier so we again get a nice big signal. The output of my amplifier is called V7.

I have added a new box which follows my voltage amplifier. For now, I will call it "pass only change". It blocks V7 before touchdown because it is constant. So instead of getting 2V at the output, I will get 0. But when V7 jumps down at touchdown, the "pass only change" block sends it right through. I can then write

$$change\ in\ V_{out} = (change\ in\ R_x \times I_x) \times G \qquad (3)$$

Recall from page 7 that R_x drops by as little as 164 micro ohms as we touchdown. Since this is a drop, the value is negative. I can then write

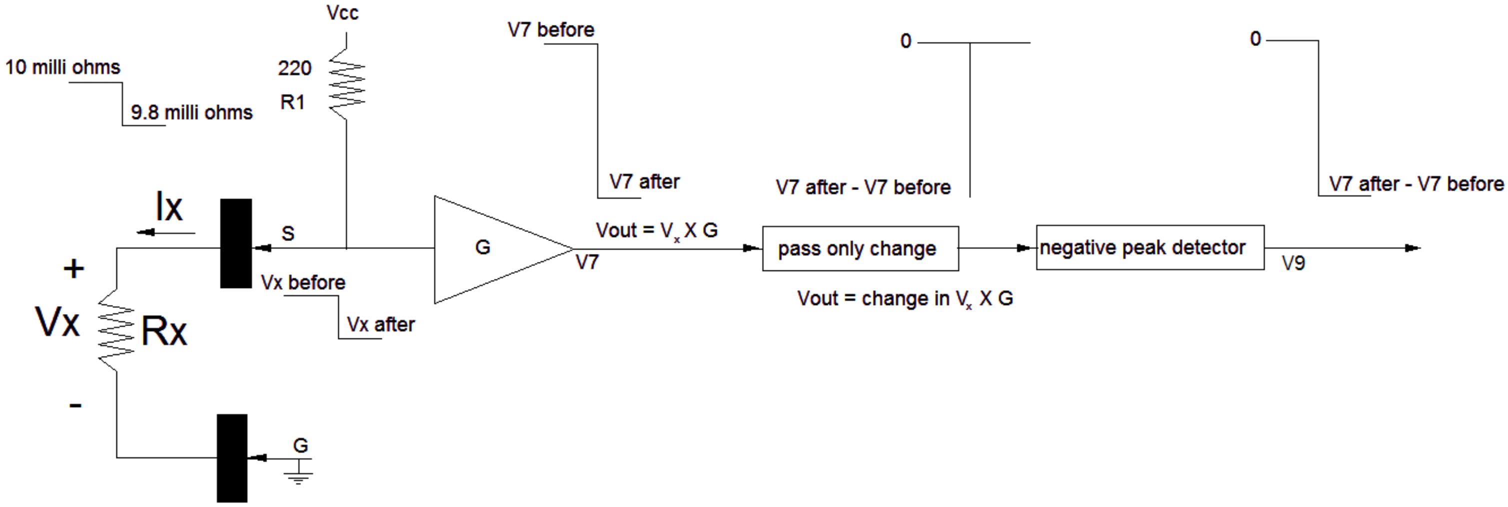
$$\begin{aligned} change\ in\ V_{out} &= (change\ in\ R_x \times I_x) \times G \\ change\ in\ V_{out} &= (-164\ micro\ ohms \times 20\ mA) \times 10{,}000 \\ change\ in\ V_{out} &= -32.8\ milli\ volts \end{aligned}$$



So as long as I see a drop in V_{out} of at least 32.8 milli volts, I will know we had touchdown. Now here is the cool part: I don't care what value I have for R_x. It will vary between machines. I only need to know if V_{out} drops by at least 32.8 milli volts and I will be sure we had touchdown.

Well, we almost got it. V_{out} is a very narrow spike with an amplitude of -32.8 milli volts. I would have trouble trying to see it if an LED was driven from this voltage. I need to make it wider and taller.

A Negative Peak Detector



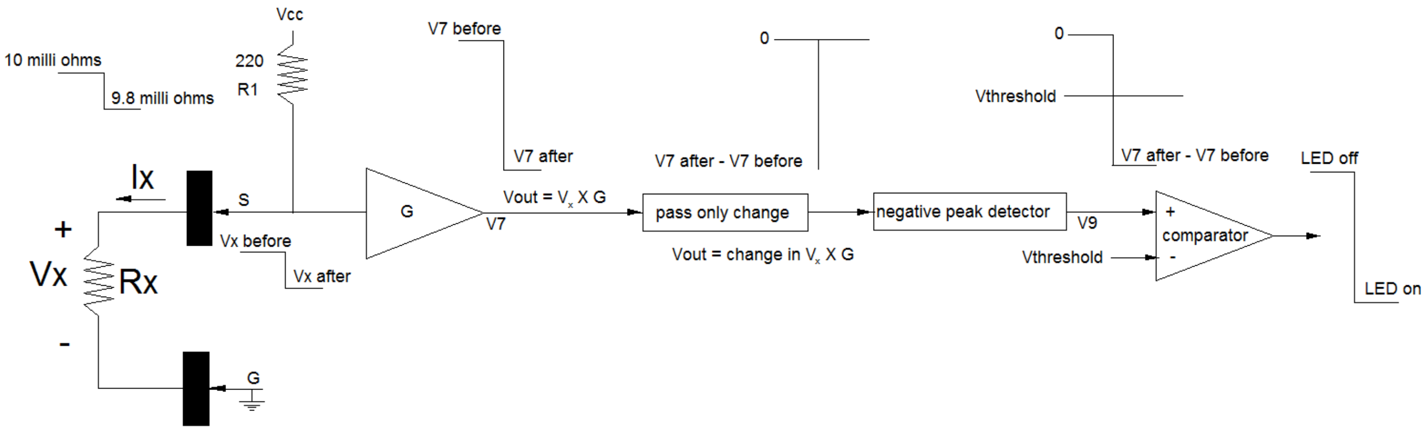
We will first work on making the signal wider. This new function always responds to the most negative peak voltage that comes along.

Before touchdown, the negative peak detector sees zero volts so that is what it outputs. But then the spike appears with its amplitude of -32.8 milli volts. This is now the most negative peak so becomes the new output. When the spike is over, the negative peak detector again sees zero volts. Since this new input is smaller than the spike, the output remains at the spike level.

In this way, we have greatly widened the spike into a constant value. It is still only a drop of 32.8 milli volts so we still need to make it bigger.

There are many ways to convert a spike into a level. You will later see why using a negative peak detector was a good choice.

Pulse Amplifier and LED Driver



The pulse is now wide enough, we just need to make it tall enough to drive an LED. This will be done with a comparator.

A comparator has a plus and a minus input.

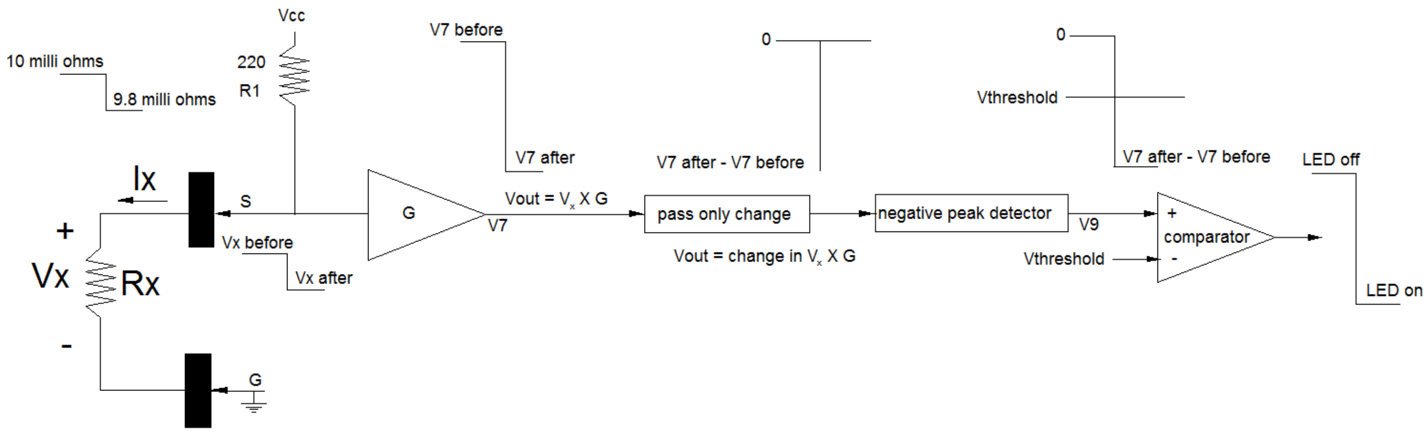
When the voltage applied to the + input is larger than the voltage applied to the - input, the output is at the most positive possible value.

When the voltage applied to the + input is smaller than the voltage applied to the - input, the output is at the most negative possible value. I will use this most negative possible value to drive my LED.

Our minimum drop in V9 due to touchdown is 32.8 mV. By setting our threshold at -16.4 mV, we split the difference between pre and post touchdown voltages.

Let's walk through the entire system to be sure we understand how this all fits together.

My probes, marked S and G, are connected across the spindle of the lathe. A resistance of 10 milli ohms is present before touchdown and a resistance of about 9.836 milli ohms is present after touchdown.



I pass a test current, I_x, out my "S" probe". It flows through the unknown resistance, R_x and returns via the "G" probe. The flow of current generates a voltage, V_x.

Before touchdown, V_x is at one value. After touchdown it is at a smaller value because R_x is smaller when the cutter contacts the work piece. V_x is fed into my voltage amplifier where it is multiplied by a value called G. The result is a much larger voltage at V7 that is one value before touchdown and a lower value after touchdown.

The voltage step down at V7 feeds into a "pass only change" function that blocks the constant voltage before and after touchdown. It only passes the change which will be a very narrow spike. Before and after this spike the output voltage is zero.

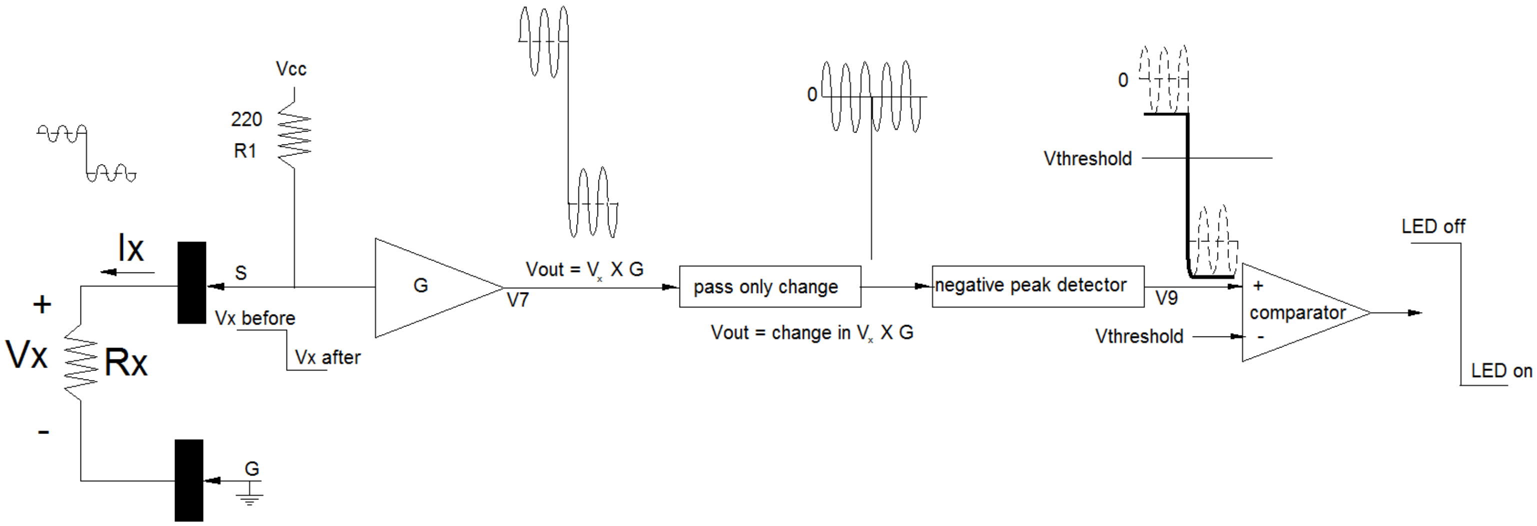
The amplitude of this spike equals the voltage at V7 after touchdown minus the voltage at V7 before touchdown. In the ideal case, this spike is zero seconds wide. Not too useful at this point.

The spike is fed into a negative peak detector which holds the most negative voltage it sees. So before touchdown it outputs zero at V9. In response to the spike, it outputs the peak of our spike. After touchdown it continues to output the peak of our spike.

V9 feeds into a comparator. When V9 is greater than a specified threshold, we turn an LED off. When V9 is less than this threshold, the LED turns on. In this way our touchdown event turns on an LED so the user knows to stop feeding the cutter into the workpiece.

It sounds like this may actually work. But alas, there are perils in a machine shop.

Machine Shops are Nasty Places



Consider what happens when there is electrical noise. If I have some AC voltage getting into V_x, it will be added to the drop in voltage caused by touchdown. Then the combination will be amplified by G. And since it is constantly changing, the amplified AC noise[5] will get through our "pass only change" box.

But then things start to get better. Our negative peak detector holds the most negative value of V9 before touchdown. So instead of AC, we get a constant value that sits on the "underside" of the noise. No matter how large the AC noise, we will always end up on this underside.

Then comes the spike. The negative peak detector grabs the tip of the spike and holds that value. But soon after the spike comes the next negative peak of the noise so our post-touchdown voltage just got a little larger.

From there, V9 feeds into the comparator . Oops, a problem. Due to the AC noise, my threshold voltage is no longer centered. In fact, given a large enough level of noise, I could exceed the threshold before touchdown. Then I would be unable to detect touchdown.

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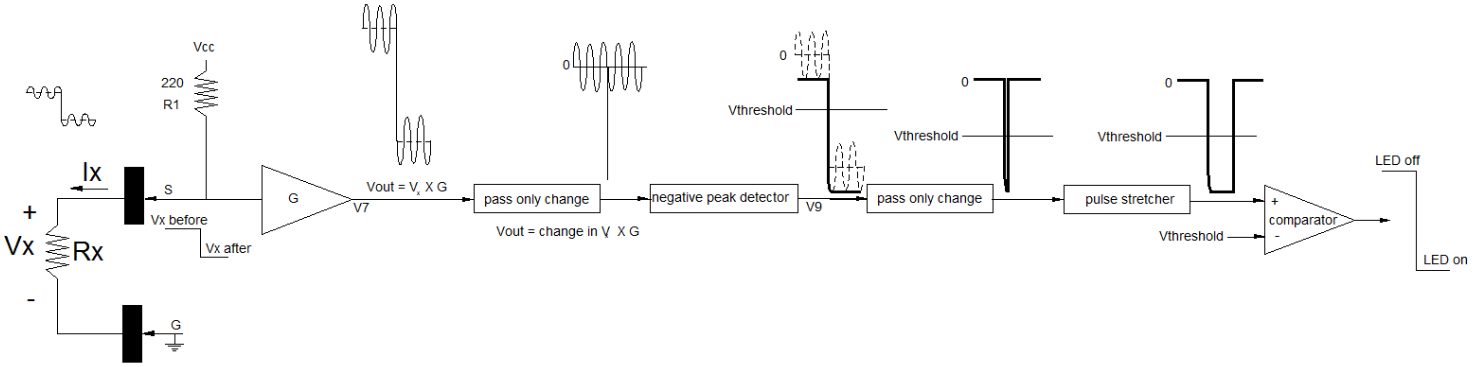
[5] The noise does not have to be a sine wave, it just has to be relatively constant. So a series of pulses would work here too as long as the amplitude did not change too much.

Using a Good Idea Twice

Thinking back to page 14, we talked about how the "pass only change" function was able to block the pre-touchdown voltage and substitute zero. It then passed only the change. Well, we need it again. V9 was shifted downward by the AC noise so our threshold is no longer centered.

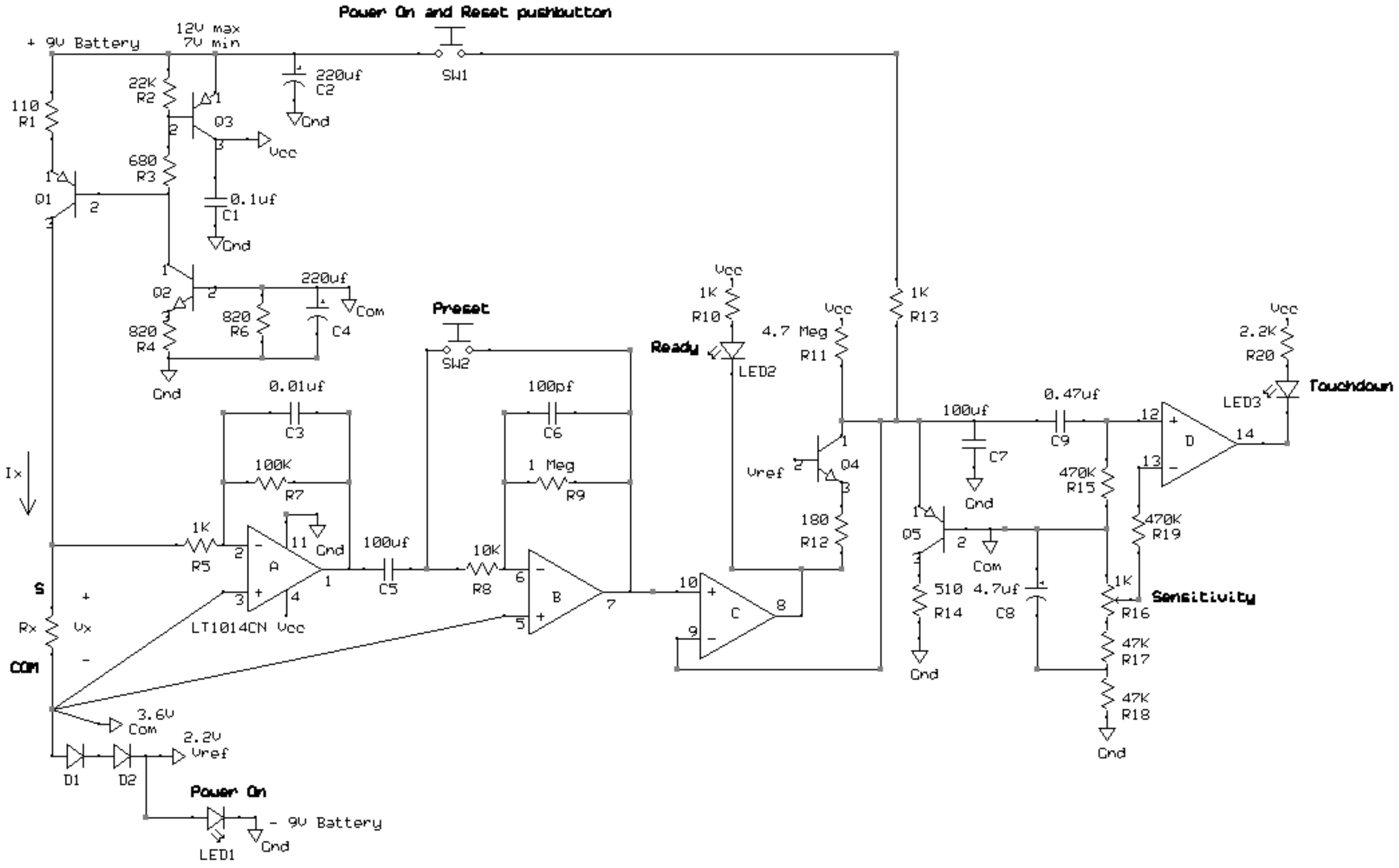


Going into our second "pass only change" block is V9. Coming out is a negative pulse that represents touchdown. And as before, it is just a skinny pulse. Too narrow to flash the LED. This time we don't need a negative peak detector. Instead, we will use a pulse stretcher. The pulse stretcher takes this narrow pulse and makes it slightly wider. This wider pulse feeds into our comparator with its threshold voltage. We get out a signal able to drive our LED that is easy to see.

We have just completed a tour of the main functions of the Lathe Electronic Edge Finder. I tried hard to present this evolution in a logical fashion assuming minimal background in electronics. To understand the finer points of the design, you must have some understanding of electronics. Of course, you can still build and enjoy this Lathe Electronic Edge Finder without fully understanding it.

Don't for a minute think that the actual circuit evolution was as straight forward as the story I have told in these preceding pages. Just like the making of sausage, you really do not want to know how it was done. I will admit that the total development time was around 6 months and I worked 5 to 6 days a week on it. The single most critical analog component was the board layout. It was masterfully done by Mark Cason.

The Schematic



Parts List By Name

These lists were developed by Mark Cason.

Part	Value	Device	Package
C1	0.1uF	TDK - FK18 Series	4.0mm x 5.5mm - 2.5mm Pitch
C2	220uF	Radial Electrolytic	8.0mm x 7.0mm - 3.5mm Pitch
C3	0.01uF	TDK - FK18 Series	4.0mm x 5.5mm - 2.5mm Pitch
C4	220uF	Radial Electrolytic	8.0mm x 7.0mm - 3.5mm Pitch
C5	100uF	TDK - FK22 Series	7.5mm x 8.0mm - 5.0mm Pitch
C6	100pF	TDK - FK18 Series	4.0mm x 5.5mm - 2.5mm Pitch
C7	100uF	TDK - FK22 Series	7.5mm x 8.0mm - 5.0mm Pitch
C8	4.7uF	Radial Electrolytic	3.0mm x 5.0mm - 1.0mm Pitch
C9	0.47uF	TDK - FK18 Series	4.0mm x 5.5mm - 2.5mm Pitch
D1	BAS33	Diode	DO35
D2	BAS33	Diode	DO35
D100	BAS33	Diode	DO35
IC1	LT1014CN	OP-Amp	DIL14
LED1	Green	LED - T1-3/4 (5mm)	V_f-2.2v I_f-25mA I_{fp}-140mA V_r-5v
LED2	Amber	LED - T1-3/4 (5mm)	V_f-2.1v I_f-50mA I_{fp}-200mA V_r-5v
LED3	Red	LED - T1-3/4 (5mm)	V_f-2.1v I_f-50mA I_{fp}-200mA V_r-5v
Q1	2N3906	PNP Transistor	TO92 Kink Lead
Q2	2N3904	NPN Transistor	TO92 Kink Lead
Q3	2N3906	PNP Transistor	TO92 Kink Lead
Q4	2N3904	NPN Transistor	TO92 Kink Lead
Q5	2N3906	PNP Transistor	TO92 Kink Lead

(continued on next page)

R1	110	Resistor	1/8 W Carbon film Resistor
R2	22K	Resistor	1/8 W Carbon film Resistor
R3	680	Resistor	1/8 W Carbon film Resistor
R4	820	Resistor	1/8 W Carbon film Resistor
R5	1K	Resistor	1/8 W Carbon film Resistor
R6	820	Resistor	1/8 W Carbon film Resistor
R7	100K	Resistor	1/8 W Carbon film Resistor
R8	10K	Resistor	1/8 W Carbon film Resistor
R9	1M	Resistor	1/8 W Carbon film Resistor
R10	1K	Resistor	1/8 W Carbon film Resistor
R11	~~4.7M~~ 68K	Resistor	1/8 W Carbon film Resistor
R12	180	Resistor	1/8 W Carbon film Resistor
R13	1K	Resistor	1/8 W Carbon film Resistor
R14	510	Resistor	1/8 W Carbon film Resistor
R15	470K	Resistor	1/8 W Carbon film Resistor
R16	1K - 12T	Trimmer Resistor	BI Technologies 64W Series
R17	47K	Resistor	1/8 W Carbon film Resistor
R18	47K	Resistor	1/8 W Carbon film Resistor
R19	470K	Resistor	1/8 W Carbon film Resistor
R20	2.2K	Resistor	1/8 W Carbon film Resistor

SW1	Power	Tactile Switch	6mm x 6mm x 9.5mm
SW2	Preset	Tactile Switch	6mm x 6mm x 9.5mm

===	9V	Battery Snap	Plastic molded "T" style - 6" lead

(end)

Parts List By Quantity and Names

Qty	Value	Parts	Device	Package
1	100pF	C6	TDK - FK18 Series	4.0mm x 5.5mm - 2.5mm Pitch
1	0.01uF	C3	TDK - FK18 Series	4.0mm x 5.5mm - 2.5mm Pitch
1	0.1uF	C1	TDK - FK18 Series	4.0mm x 5.5mm - 2.5mm Pitch
1	0.47uF	C9	TDK - FK18 Series	4.0mm x 5.5mm - 2.5mm Pitch
2	100uF	C5, C7	TDK - FK22 Series	7.5mm x 8.0mm - 5.0mm Pitch
1	4.7uF	C8	Radial Electrolytic	3.0mm x 5.0mm - 1.0mm Pitch
2	220uF	C2, C4	Radial Electrolytic	8.0mm x 7.0mm - 3.5mm Pitch
1	LT1014CN	IC1	OP-Amp	DIL14
1	110	R1	Resistor	1/8 W Carbon film Resistor
1	180	R12	Resistor	1/8 W Carbon film Resistor
1	510	R14	Resistor	1/8 W Carbon film Resistor
1	680	R3	Resistor	1/8 W Carbon film Resistor
2	820	R4, R6	Resistor	1/8 W Carbon film Resistor
3	1K	R5, R10, R13	Resistor	1/8 W Carbon film Resistor
1	2.2K	R20	Resistor	1/8 W Carbon film Resistor
1	10K	R8	Resistor	1/8 W Carbon film Resistor
1	22K	R2	Resistor	1/8 W Carbon film Resistor
2	47K	R17, R18	Resistor	1/8 W Carbon film Resistor
1	100K	R7	Resistor	1/8 W Carbon film Resistor
2	470K	R15, R19	Resistor	1/8 W Carbon film Resistor
1	1M	R9	Resistor	1/8 W Carbon film Resistor
1	4.7M	R11	Resistor	1/8 W Carbon film Resistor
1	1K - 12T	R16	Trimmer Resistor	BI Technologies 64W Series

(continued on next page)

2	2N3904	Q2, Q4	NPN Transistor	TO92 Kink Lead
3	2N3906	Q1, Q3, Q5	PNP Transistor	TO92 Kink Lead
2	BAS33	D1, D2	Diode	DO35
1	Green	LED1	LED - T1-3/4 (5mm)	Vf-2.2v If-25mA Ifp-140mA Vr-5v Ir-10uA
1	Amber	LED2	LED - T1-3/4 (5mm)	Vf-2.1v If-50mA Ifp-200mA Vr-5v Ir-100uA
1	Red	LED3	LED - T1-3/4 (5mm)	Vf-2.1v If-50mA Ifp-200mA Vr-5v Ir-100uA
2	Switch	SW1, SW2	Tactile Switch	6mm x 6mm x 9.5mm
1	9V	===	Battery Snap	Plastic molded "T" style - 6" lead

(end)

Mouser Bill Of Materials

http://www.mouser.com/Tools/part-list-import.aspx

Copy the following EXACTLY as shown, into the box at the website above:

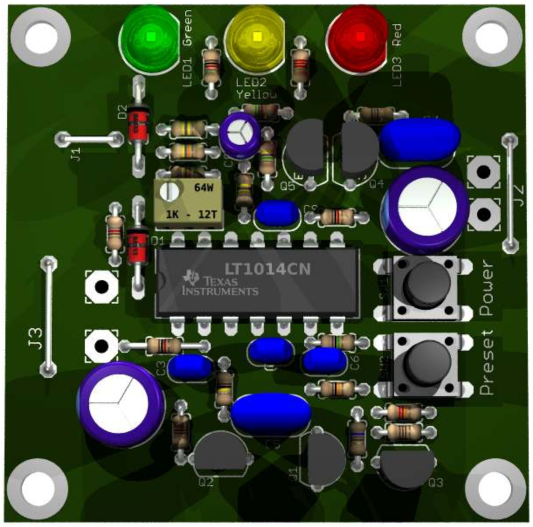
```
810-FK18C0G1H101J|1
810-FK18X7R1H103K|1
810-FK18X7R1E104K|1
810-FK18X7R1E474K|1
810-FK22Y5V1A107Z|2
647-UMA1C4R7MCD2|1
647-USR1C221MDD|2
595-LT1014CN|1
299-110-RC|1
299-180-RC|1
299-510-RC|1
299-680-RC|1
299-820-RC|2
299-1K-RC|3
299-2.2K-RC|1
299-10K-RC|1
299-22K-RC|1
299-47K-RC|2
299-100K-RC|1
299-470K-RC|2
299-1M-RC|1
299-4.7M-RC|1
858-64WR1KLF|1
512-2N3904TFR|2
512-2N3906TFR|3
78-BAS33-TAP|2
604-WP1503GT|1
941-C503BRCNCW0Z0AA1|1
941-C503BACSCY251030|1
611-PTS645SK95LFS|2
123-5116-GR|1
```

(end - do not copy this line)

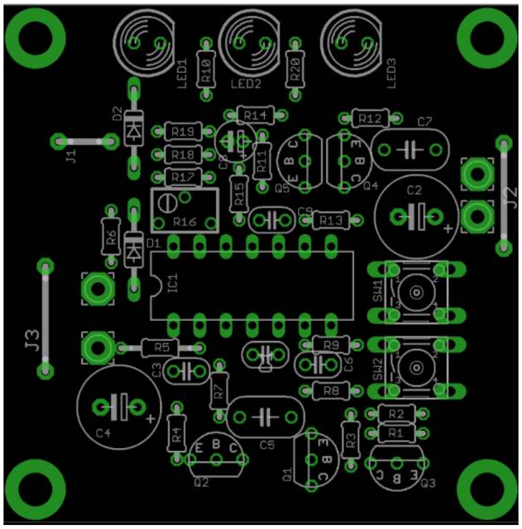
Single Sided Circuit Board Artwork

All artwork is from Mark Cason. If you want to etch your own board, it is easiest to make it single sided.

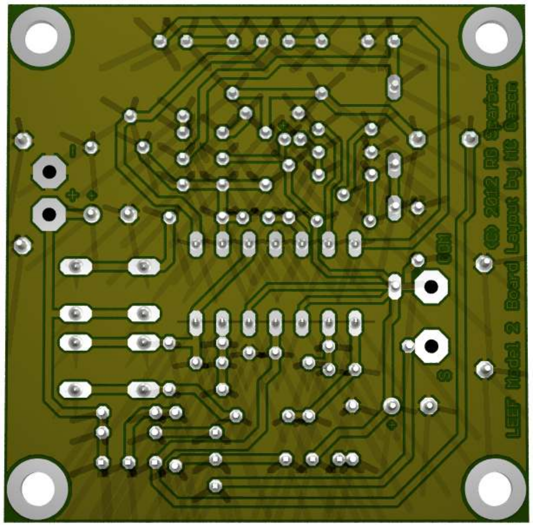
Populated Board



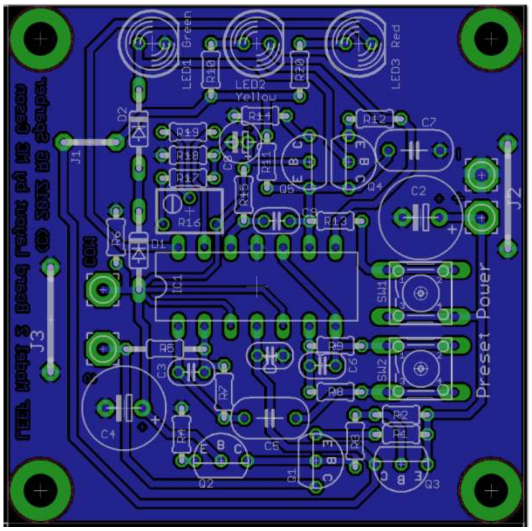
Parts Placement with Labels



Back Side



X-Ray View from Top

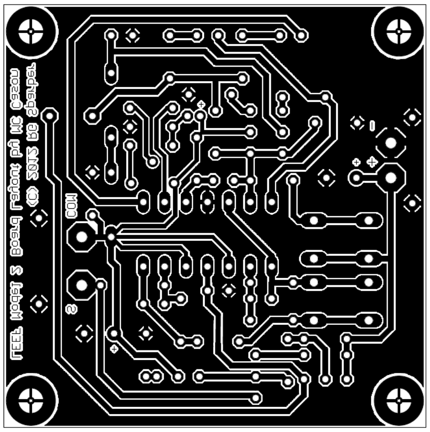


Note that there is one jumper, J1, used to complete the circuit. Jumpers 2 and 3 are optional cable reliefs.

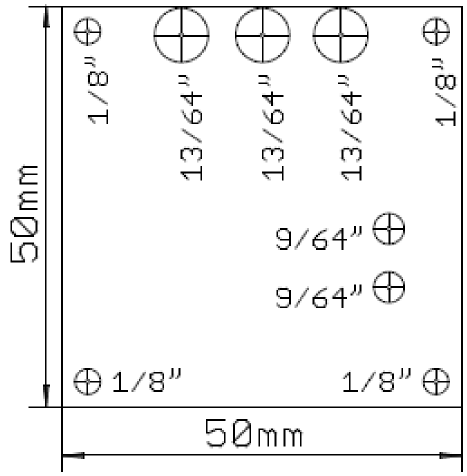
Here is the etching mask.

Scale this artwork so the board is 50mm by 50 mm. I do this by first printing the artwork at 100%. Then use a digital caliper set to metric to measure the width. Divide 50 mm by this measurement and multiply by 100%. Then round to the nearest integer to get the needed zoom percentage. For example, if you measured 47.4 mm, then change the zoom to $\frac{50mm}{47.4mm} \times 100\%$ = 105.49%. My zoom should be set to 105%. Print and recheck.

Laser print it onto clear plastic. Then use this mask for contact printing to a pretreated circuit board with ***text right reading***.



When drilling the board, choose drills that match the copper free diameters in the pads. Going larger will tear up the copper and make soldering much harder.



This is a template that can be used to drill the holes in the case.

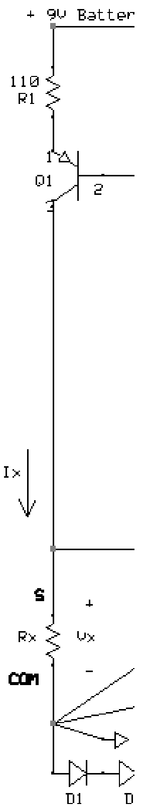
At some time in the near future, Mark Cason will be selling a double sided version that has been drilled along with a kit of parts. He plans to also sell a finished/tested version.

Detailed View

Many of the key design points of this circuit can be addressed as isolated concepts. They will be discussed first. Then we will get into elements of the design that cover many components. Only then can we consider the overall circuit behavior.

I have included measured voltages from my first circuit board based prototype next to some of the calculated values. They are shown with [square brackets around them in red].

Touchdown Sensing Physics



Our test current, I_x, flows out of probe "S" and into our unknown resistance, R_x and back into the circuit via probe COM. R_x has one value when the cutter is not in contact with the workpiece and a lower value when touching down.

I_x generates the unknown voltage V_x which is sensed between probe S and COM.

Not shown in the figure is that there is also an unknown contact resistance associated with each probe . This is the same as adding an unknown resistance to R_x. It is therefore the same as adding an unknown DC component to V_x.

The traditional way of dealing with this problem is to use a Kelvin Connection[6]. It has one set of probes that carry current and a second set of probes that just sense the voltage. In this way the sensed voltage is not tainted by the test current.

We don't need the Kelvin Connection because the circuit is only sensitive to changes in V_x. As long as the voltage drop in the probe is constant, it will be ignored. But if you wiggle the probes while the circuit is trying to detect touchdown, all bets are off. And if the circuit sees too much disruption of the test current, it will power down.

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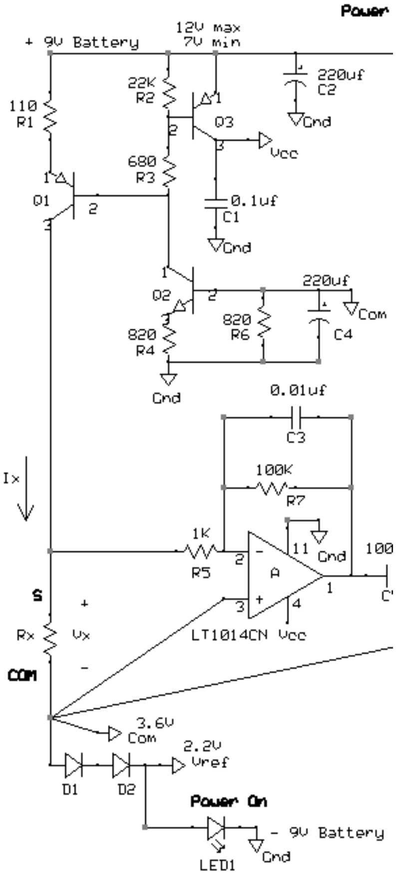
[6] See http://en.wikipedia.org/wiki/Four-terminal_sensing.

Power Control



At least in my shop, any battery powered box with a power switch will contain a dead battery. I will remember to turn the power switch to the on position so I can use the instrument. But after I have used the device, I put it aside and get on to making chips. It sits there until the battery goes dead.

So rule number one for my designs is ***there must be an automatic power off function!***

I satisfied this rule by having Q1 and Q3 be power switches and Q2 forms a path to keep power on.

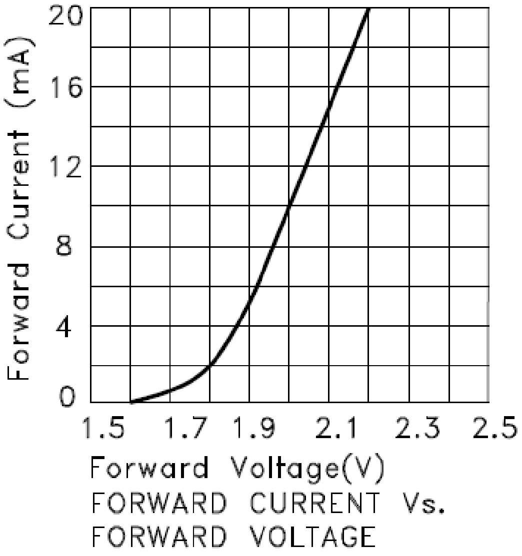
I will spare you the algebra and just say that it can be shown that

$$I_x = k_1 \times (V_{com} - V_{be2}) \qquad (4)$$

Where

$$k_1 = \frac{\left\{\frac{R_3}{\left(1+\frac{1}{\beta_1}\right)}\right\}}{R_4\left[R_1 + \frac{R_3}{(1+\beta_1)}\right]}$$

This set of equations tells us what current, I_x comes out of Q1 when a voltage, V_{com} is put in. That is half of the puzzle. We must next figure out how V_{com} is controlled by I_x. That involves D1, D2, and LED1 along with R6.



LED1 is a WP1503GT (green) which can be modeled by a voltage source and resistor. Extending the almost linear part of the curve to where it intersects the X axis gives us the voltage source. We find the resistor by taking data at two points along the almost linear part of the curve. I see 2.2V at 20 mA and 2.0V at 10 mA. I can then say that

$$R_{LED} = \frac{change\ in\ voltage}{change\ in\ current}$$

$$R_{LED} = \frac{2.2V - 2.0V}{20\ mA - 10\ mA} \qquad (5)$$

$$R_{LED} = 20\ ohms$$

To find the theoretical intercept voltage we note that

$V_{forwardLED} = V_{kneeLED} + R_{LED} \times I_{forward}$ (6)
Where $V_{kneeLED}$ is the voltage drop across the LED when almost no current flows. R_{LED} is the bulk resistance of the LED.

Filling in what we know using the lower data point gives

$2.0V = V_{kneeLED} + 20\ ohms \times 10\ mA$

Which can be solved for $V_{kneeLED}$

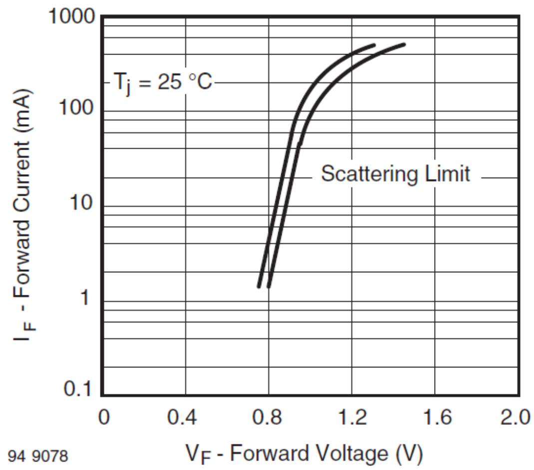
$V_{kneeLED} = 1.8V$

I can then write

$V_{forwardLED} = 1.8V + 20\ ohms \times I_{forward}$ (7)

As a check, I will plug in the second data point

$V_{forwardLED} = 1.8V + 20\ ohms \times 20\ mA = 2.2V\ \ OK$



I can model the BAS33 diodes the same way. They will have a knee voltage and a bulk resistance.

$$V_{forwardDiode} = V_{kneeDiode} + R_{Diode} \times I_{forward} \qquad (8)$$

I again look at the slope to get the resistance for my model.

$$R_{Diode} = \frac{change\ in\ voltage}{change\ in\ current}$$

There are two curves representing limits. I will shoot for the center to get typical.

$$R_{Diode} = \frac{0.95V - 0.55V}{100\ mA - 0.1\ mA}$$

$$R_{Diode} = 4.0\ ohms$$

The general case is

$$V_{forwardDiode} = V_{kneeDiode} + R_{Diode} \times I_{forward} \quad (9)$$

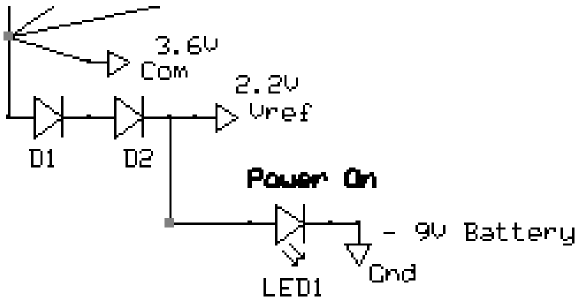
So I can find $V_{kneeDiode}$ by again using one of the data points

$$0.95V = V_{kneeDiode} + 4.0\ ohms\ \times 100mA \qquad (7)$$

Which gives me $V_{kneeDiode} = 0.55V$

So I can characterize the diode with

$$V_{forwardDiode} = 0.55V + 4.0\ ohms \times I_{forward} \qquad (10)$$



I can now describe my diode string that sets V_{com}:

$$V_{com} = V_{forwardLED} + 2 \times V_{forwardDiode} \quad (11)$$

Plugging in (6)

$$V_{forwardLED} = V_{kneeLED} + R_{LED} \times I_{forward} \quad (6)$$

and (9)

$$V_{forwardDiode} = V_{kneeDiode} + R_{Diode} \times I_{forward} \quad (9)$$

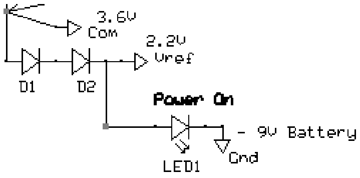
gives us

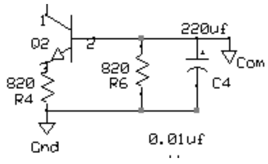
$$V_{com} = \{V_{kneeLED} + R_{LED} \times I_{forward}\} + 2 \times \{V_{kneeDiode} + R_{Diode} \times I_{forward}\}$$

$$V_{com} = \{V_{kneeTotal} + R_{Total} \times I_{forward}\}$$

Where $V_{kneeTotal} = V_{kneeLED} + 2 \times V_{kneeDiode}$

and $R_{Total} = R_{LED} + 2 \times R_{Diode}$ (12)





I_x has two major paths: the diode string and R6. The current through the rest of the COM node is very close to zero.

$$I_x = I_{forward} + \frac{V_{com}}{R_6}$$

or

$$I_{forward} = I_x - \frac{V_{com}}{R_6} \qquad (13)$$

Plug (13) into (12) and I get

$$V_{com} = V_{kneeTotal} + R_{Total} \times I_{forward} \qquad (12)$$

$$V_{com} = V_{kneeTotal} + R_{Total} \times [I_x - \frac{V_{com}}{R_6}]$$

or

$$V_{com} = \frac{V_{kneeTotal} + R_{Total} \times I_x}{1 + \frac{R_{Total}}{R_6}} \qquad (14)$$

This relates how I_x generates V_{com} as it passes through the diode string and R6.

From page 31 we have (4)

$$I_x = k_1 \times (V_{com} - V_{be2}) \qquad (4)$$

Taking a breath here, realize that (14) tells us how the diode string voltage reacts to I_x while (4) tells us how the feedback circuit current reacts to the diode string voltage. By combining these two equations we can see where they balance.

Plugging (4)

$$I_x = k_1 \times (V_{com} - V_{be2}) \qquad (4)$$

into (14)

$$V_{com} = \frac{V_{kneeTotal} + R_{Total} \times I_x}{1 + \frac{R_{Total}}{R_6}} \qquad (14)$$

I get

$$V_{com} = \frac{V_{kneeTotal} + R_{Total} \times [k_1 \times (V_{com} - V_{be2})]}{1 + \frac{R_{Total}}{R_6}}$$

or

$$V_{com} = \frac{(V_{kneeTotal} - R_{Total} \times \{k_1 \times V_{be2}\})}{1 + \frac{R_{LED}}{R_6} - R_{Total} \times k_1} \qquad (16)$$

where

$$k_1 = \frac{\left\{\frac{R_3}{\left(1 + \frac{1}{\beta_1}\right)}\right\}}{R_4\left[R_1 + \frac{R_3}{(1 + \beta_1)}\right]}$$

When I plug in typical numbers I get

$$k_1 = \frac{\left\{\frac{R_3}{\left(1+\frac{1}{\beta_1}\right)}\right\}}{R_4\left[R_1 + \frac{R_3}{(1+\ \beta_1)}\right]}$$

$$k_1 = \frac{\left\{\frac{680}{\left(1+\frac{1}{\beta_1}\right)}\right\}}{820\left[110+\frac{680}{(1+\beta_1)}\right]}$$

Consider two cases. In the first, $\beta_1$is "large". Then

$$k_1 = \frac{\{680\}}{820[110]}$$

$$k_1 = \frac{1}{132.7\ ohms} \quad (\beta_1\text{is maximum}) \quad (17)$$

In the second case, $\beta_1$is 80 which is the minimum at 10 mA.

$$k_1 = \frac{\{671.6\}}{820[118.4]}$$

$$k_1 = \frac{1}{144.6\ ohms} \quad (\beta_1\text{is minimum}) \quad (18)$$

So k_1 = varies from $\frac{1}{144.6\ ohms}$ to $\frac{1}{132.7\ ohms}$ as β_1 goes through its range of values. The average is $\frac{1}{139\ ohms}$.

In preparation for plugging in the rest of the values, we need to calculate two intermediate terms.

$V_{kneeTotal} = V_{kneeLED} + 2 \times V_{kneeDiode}$
$V_{kneeTotal} = 1.8V + 2 \times 0.55V$
$V_{kneeTotal} = 1.8V + 1.1V$
$V_{kneeTotal} = 2.9V$

$R_{Total} = R_{LED} + 2 \times R_{Diode}$
$R_{Total} = 20\ ohms + 2 \times 4.0\ ohms$
$R_{Total} = 28\ ohms$

I can now find V_{com} by using (16)

$$V_{com} = \frac{(\{V_{kneeTotal}\} - \{R_{Total}\} \times \{k_1 \times V_{be2}\})}{1 + \frac{R_{Total}}{R_6} - R_{Total} \times k_1} \quad (16)$$

Using the average value for k_1

$$V_{com} = \frac{\{2.9V\} - \{28\ ohms\} \times \{\frac{1}{139\ ohms} \times 0.65V\})}{1 + \frac{28\ ohms}{820} - 28\ ohms \times \frac{1}{139\ ohms}} \quad (16)$$

$V_{com} = 3.33V$ [3.83V]

Plug this into (4) which describes how the feedback circuit works and get

$$I_x = k_1 \times (V_{com} - V_{be2}) \quad (4)$$

$$I_x = \frac{1}{139\ ohms} \times (3.33V - 0.65V)$$

$$I_x = 19.3mA \quad [22.9\ mA]$$

This tells us that given typical values, I_x equals 19.3 mA.

As a final check, let's find the current through just the diode string.

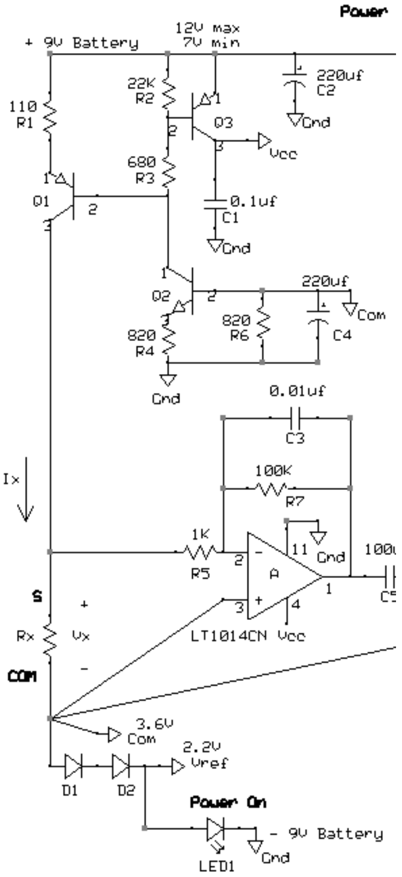
$$I_{forward} = I_x - \frac{V_{com}}{R_4} \quad (13)$$

$$I_{forward} = 19.3\ mA - \frac{3.33V}{820\ ohms}$$

$$I_{forward} = 15.2\ mA$$

Which is safely below the recommended maximum of 30 mA for the LED.

If this design was going into production, I would need to do a full variational analysis and probably adjust the circuit to better center I_x.



You may wonder why I have chosen such a complicated way to generate I_x. After all, wouldn't a simple resistor do the job?

Well, for a long time I did simply have a 220 ohm resistor tied between R_x and the +9V Battery rail. At some point I reduced the noise in the circuit to where it became clear that this resistor was taking the noise on the +9V Battery rail and passing it to V_x.

Say I have an R_x equal to 0.01 ohms which is the smallest value the circuit has to process. This gives me a voltage divider for the noise:

$$v_x = \frac{0.01\ ohms}{0.01\ ohms + 220\ ohms} v_+ \qquad (19)$$

$$v_x \cong \frac{0.01\ ohms}{220\ ohms} v_+ = \frac{1}{22{,}000} v_+.$$

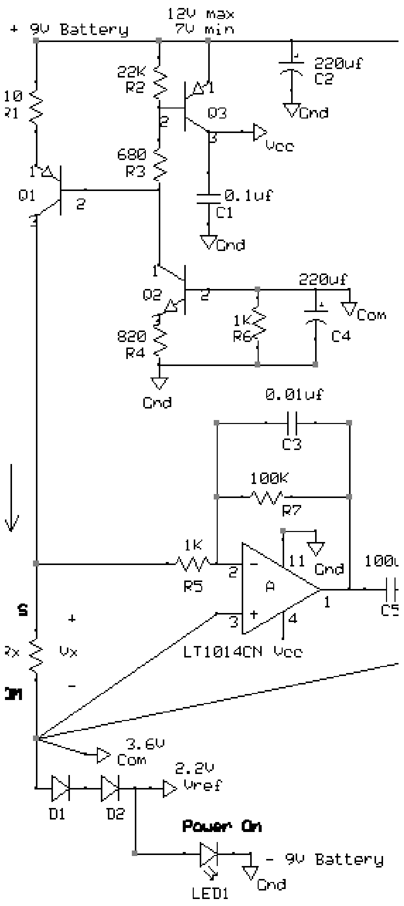
Where v_x is the AC noise in V_x and v_+ is the AC noise in the +9V Battery rail.

So if I got 1μV at v_x I would expect to see $1\mu V \times 22{,}000 = 22$ mV on v_+. Measurements showed noise of this magnitude. Now, given 1 μV at V_x, we get $1\ \mu V \times 10{,}000 = 10\ mV$ at V7. When you are trying to measure a 32 mV signal, 10 mV hurts a lot.

Finding the source of this noise was a challenge because it came from the negative peak detector reacting to the 10 mV spike that in turn was generated by the negative peak detector. In other words, the stimulus was aligned with the response so it was tough trying to sort them out by looking at the signals.

My first attempt at reducing the noise was to split the 220 ohm resistor into two 110 ohm resistors. At the junction I added a very large filter capacitor. It did reduce this noise by a factor of 2 but I really needed a lot more attenuation.

The current sources formed by Q1 and Q2 do that well. Noise on the +9V Battery rail is no longer detectable at V7. So it was well worth adding this much complexity to the circuit.

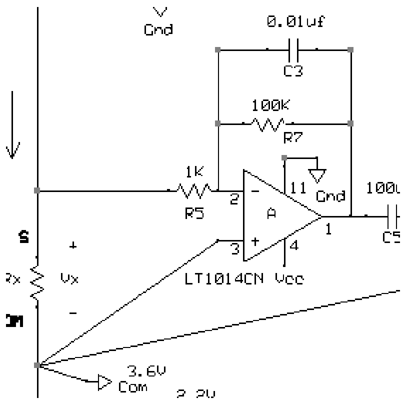


Capacitors C1 reduces the high frequency noise on V_{cc} due to current pulses generated by flashing LEDs. C2 does the same thing for low frequencies. Placing C2 at the battery terminal rather than V_{cc} permits a faster power down when the probes are disconnected.

C4 minimizes the change in voltage between the COM net and ground. This voltage movement has the potential of disabling the circuit as will be explained later.

The negative peak detector uses V_{ref} which is equal to two diode drops below V_{com} and is typically 2.2V with respect to ground.

The First Gain Stage



This is a simple inverting amplifier. The DC gain is set by the ratio of R7 to R5 so comes out to

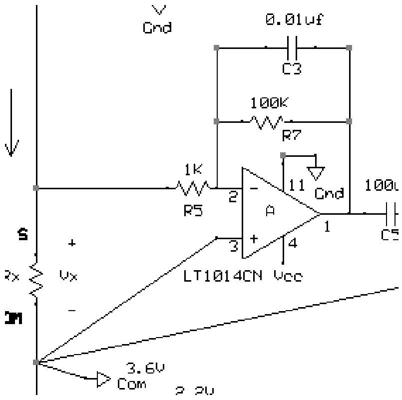
$$-\frac{R7}{R5} = -\frac{100K}{1K} = -100 \qquad (20)$$

Although R3 is rather small, it is much larger than any reasonable R_x. So as far as input loading is concerned, this gain stage has a very high input resistance.

The input bias current for this op amp is 30 nA maximum. This current splits between R5 and R7. $\frac{100}{101} \times 30\ nA$ pass through R5 and $\frac{1}{101} \times\ 30\ nA$ passes through R7. This means that pin 2 is raised by $1K\ \times 29.7\ nA = 29.7\ \mu V$. The current through R7 is 0.3 nA so the voltage across R7 equals $0.3\ nA\ \times 100K = 30\ \mu V$. With pin 2 at 29.7 μV and R7 dropping 30μV, the output is at -0.3 μV due to input bias current. It is therefore safe to ignore it.

Input offset voltage is another potential problem. It can be as high as ±0.3 mV. That drives the output to as much as $\pm 0.3\ mV\ \times (-101) =\ \pm 30.3\ mV$. Good, not a problem. [-108 mV with the probes tied together which implies a contact resistance of between $\frac{108 mV - 30\ mV}{22.9\ mA\ \times 100}$ = 34 milli ohms and $\frac{108 mV + 30\ mV}{22.9\ mA\ \times 100}$ = 60 milli ohms]

We cannot completely ignore the DC at the output of this gain stage. We must be able to pass AC without distortion. So let's look at the maximum AC. This gain stage multiplies the input by 100 and feeds it to the second gain stage which also multiplies by 100. If the output of the second gain stage is at maximum, it can output a signal centered around COM that is ± 1.3V. That means that the input to this gain stage cannot be larger than $\frac{\pm 1.3V}{100} = 13\ mV$ which equals the output of the first gain stage at pin 1 and is tiny.



We do have a case where this first gain stage does saturate. It occurs at large values of R_x. The maximum I_x is about 25 mA as found on page 6.

Note that the op amp is referenced to COM and not ground. When the op amp is at negative saturation, it is really hitting ground. So the maximum V_x ***with respect to COM*** equals -3.6V. This means that V_x

$$V_x = \frac{-V_{com}}{-100} \qquad (21)$$

$$V_x = \frac{-3.6V}{-100}$$

$$V_x = 36\,mV.$$

But

$$V_x = R_x \times I_x \qquad (1)$$

$$36\,mV = R_x \times 25\,mA$$

This gives us a maximum R_x of $\frac{36\,mV}{25\,mA} = 1.4\,ohms$. As long as the contact resistance between workpiece and cutter (in isolation) meets the spec of being no larger than 0.6 ohms, there is no problem here. But if the parallel combination of the spindle resistance and the contact resistance equaled more than 1.4 ohms, the circuit would not be able to detect touchdown. The first gain stage would stay in saturation. Of course, given resistances this big, a much simpler electronic edge finder can be used[7].

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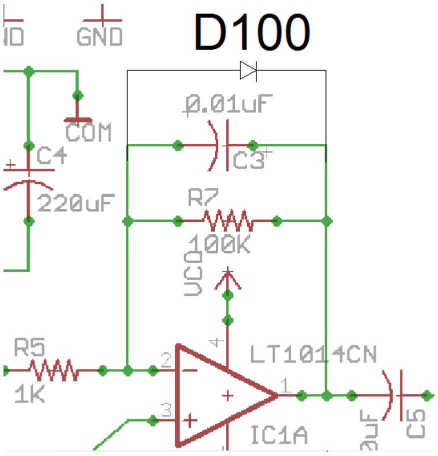
[7] See http://rick.sparber.org/rctf.pdf .

Update to First Stage



I have added a diode, D100 to the feedback loop. This prevents op amp A from going more than 0.6V below COM and prevents entering negative saturation. I have found that once this op amp goes into saturation, it disturbs node COM and upsets the rest of the circuit. It also prevents the READY LED from flashing. At touchdown, both the READY and Touchdown LEDs flash but there is no indication that the circuit is actually ready.

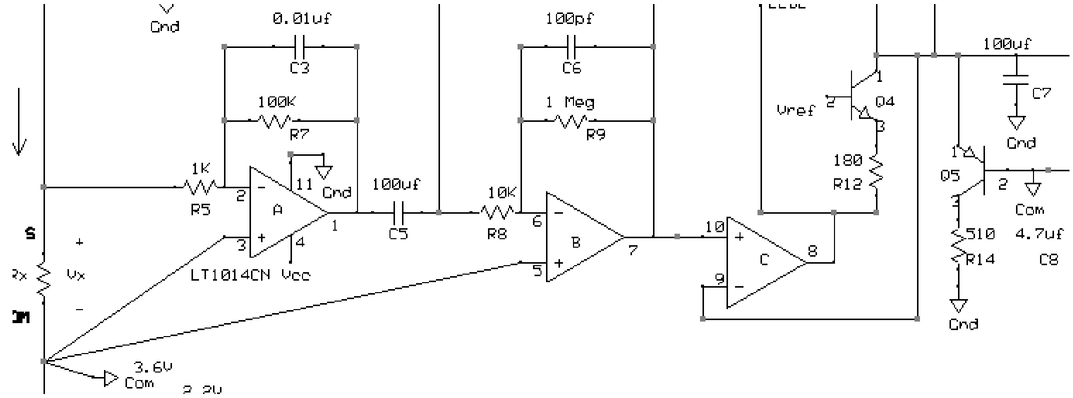
When V1 is at -0.6V, the voltage at the input is at $\frac{-0.6V}{-100} = 6\,mV$.

First consider the case of no AC noise. Given a test current of 25 mA, this translates into an R_x of $\frac{6\,mV}{25\,mA} = 0.24\,ohms$. So any R_x greater than 0.24 ohms looks to the circuit like it is 0.24 ohms as long as it is small enough to keep the circuit powered up. The circuit should power up when an R_x of up to 100 ohms is connected.

We will consider the case of AC noise later because the negative peak detector is the limiting factor here.

Note that if R_x is large enough, V1 will stay at about -0.6V even if some AC noise is present. V1 will have very little AC in it.

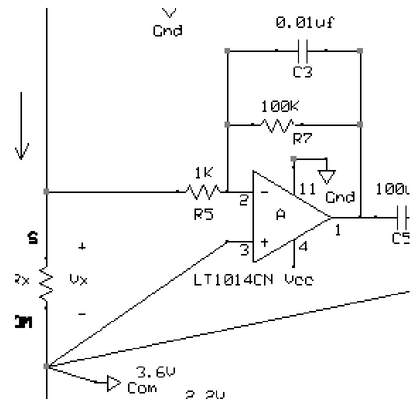
So the addition of this feedback diode enables the circuit to work correctly with spindle resistances between 0.01 ohms and 100 ohms.



When we have R_x large enough to cause the first gain stage to approach saturation, we do run the risk of distorting an AC noise. However, the voltage transition at touchdown is so large that the circuit can't miss it. So there really is no issue with AC in this case.

The flip side to causing AC distortion is that AC cannot pass through the amplifier if it is in saturation. With V1 sitting at a rail and dead quiet, the output of the second gain stage, V7, will also be dead quiet. That means that the negative peak detector will see a dead quiet signal.

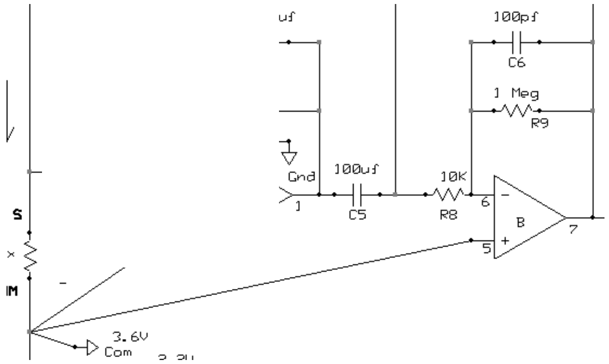
We will later show that when the negative peak detector is in linear mode the Ready LED does not flash. It will likely detect touchdown.



My first line of defense against noise spikes developed across R_x is C3. C3 in combination with R7 gives me a low pass filter with a corner at 159 Hz. I chose this corner frequency so it would attenuate spikes a bit yet not blunt the step generated by touchdown too much. As you can see, it has no effect on 60 Hz noise.

Note the odd way I drew the connection from non-inverting input of the op amp to my COM node. It is essential that this be a direct connection with no other circuits sharing it. Any noise injected into this input is multiplied by 10,000 by the time it gets to pin 7. If I had 0.3 micro volts of noise at pin 3 relative to COM it would become 3 millivolts of noise at pin 7. That is 10% of my valid minimum signal. Additionally, the area flanking this node should be ground to minimize capacitive coupling.

The Second Gain Stage



V_x is multiplied by -100 as it appears at the output of the first gain stage, pin 1. The second gain stage is going to multiply it by -100 again. This is only possible if I first block any DC. Otherwise, I could only tolerate about 30 mV at the output of the first gain stage before I hit the rails at the output of the second stage. My blocking capacitor, C6, is a 100 μf ceramic capacitor which has low leakage and no polarity.

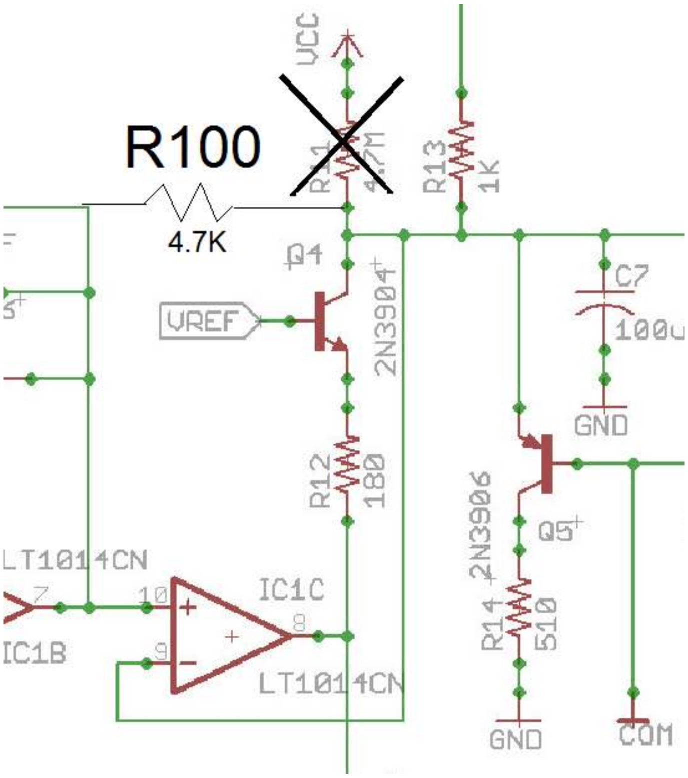
My hat off to Mark Cason for finding this device for me.

I need as large a time constant as possible here so have set R8 equal to 10K. This lets me pass the touchdown signal with minimal loss.

An R8 of 10K forces me to use a 1 Meg for R9 in order to get my gain of -100. I will deal with the effect of input offset current and R9 later.

Capacitor C6 was chosen to have a time constant 10 times faster than the one in the first gain stage. In this way it does not adversely affect the touchdown voltage transition. Its main purpose is to block noise spikes that get into C5 and/or R8. A 1 milli volt spike at V7 can be generated by 10 micro volts of noise at the junction of C5 and R8. I can't afford that.

The Negative Peak Detector with Update

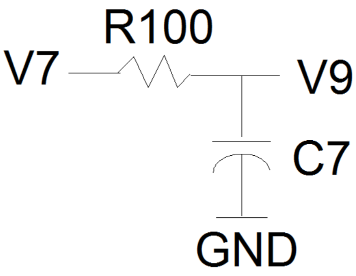


I'll admit it. This was the hardest part of the design. It is a mix of analog and digital which makes noise management difficult. I need to be able to move a central node in this circuit by many volts quickly yet almost park it at any point with fractions of a milli volt of drift.

The circuit has two states. It is either moving the voltage at pin 9, call it V9, up or it is moving it down.

V9 moves up slowly. This movement is defined by an exponential with a time constant equal to R100 times C7.

When V9 is moving up, we can use a simpler model to explain what is happening:



First, consider what $v_7(t)$ is doing. Assume it has a large 60 Hz component and no DC offset. Given R100 = 4.7K and C7 = 100 uf, it can be shown that

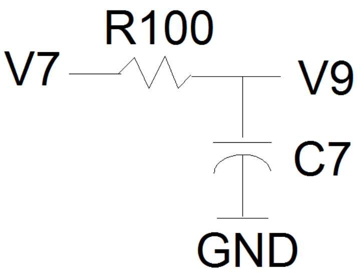
$$v_9(t) = \frac{-j\dfrac{1}{\omega C_7}}{R100 - j\,\dfrac{1}{\omega C_7}} \times v_7(0)cos(\omega t)$$

$$\omega = 2\pi f = 377\ radians/second$$

$$v_9(t) = \frac{-j\dfrac{1}{377\ x\ 100uf}}{4.7K - j\,\dfrac{1}{377\ x\ 100\ uf}} \times v_7(0)cos(\omega t)$$

Cosine is in radians.

$$v_9(t) = \frac{-j26.5}{4.7K - j26.8} \times v_7(0)cos(377t)$$



$$v_9(t) = \frac{-j26.5}{4.7K - j26.8} \times v_7(0)cos(377t)$$

$$v_9(t) = 5.64 \times 10^{-3} \times v_7(0)cos(377t - 1.57) \quad (21.1)$$

Cosine is in radians.

Next, assume *$v_7(t)$* is zero and look at how *$v_9(t)$* moves from *$v_9(0)$* towards zero. Recall that *$v_9(0) = v_7(0)$*. We have an RC circuit with the standard format of

$$v_9(t) = v_7(0)e^{\left(\frac{t}{-\tau}\right)} \qquad (21.2)$$

Where τ = R100 x C7.

This equation says that *$v_9(t)$* starts out at the negative peak of v_7 and discharges towards zero.

Using Superposition, we can add (21.1) to (21.2) to get the total behavior at *$v_9(t)$*:

$$v_9(t) = 5.64 \times 10^{-3} \times v_7(0)cos(377t - 1.57) + v_7(0)e^{\left(\frac{t}{-\tau}\right)} \qquad (21.3)$$

τ = R100 x C7 which nominally equals 0.47 seconds.

It will be later shown that we leave this simple model when t = 15.8 mS so

$$v_9(t) = 5.64 \times 10^{-3} \times v_7(0)cos(377\ x\ 15.8\ ms - 1.57) + v_7(0)e^{\left(\frac{15.8mS}{-0.47}\right)}$$

$$v_9(t) = v_7(0)(5.64 \times 10^{-3} \times cos(377\ x\ 15.8\ mS - 1.57) + e^{\left(\frac{15.8mS}{-0.47}\right)})$$

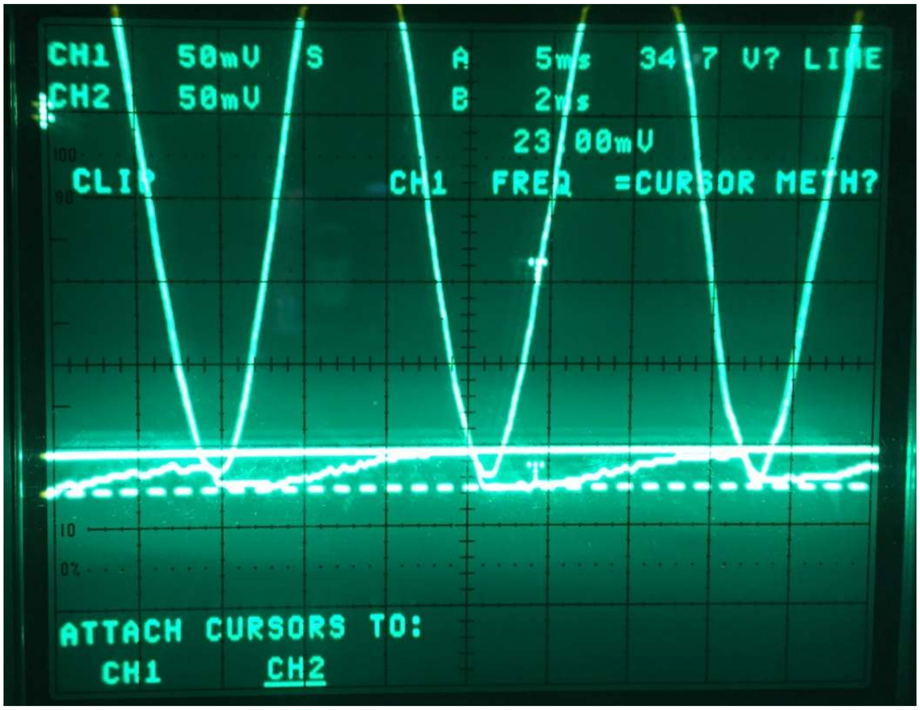
$$v_9(t) = v_7(0)(-0.00181 + 0.967)$$

In other words, the AC component can be ignored. This was actually the intent of connecting R100 to V7. I wanted to connect to COM but could not afford to pull this much current out of the COM node. By connecting R100 to V7, I knew the AC component would be blocked by C7 and I would be left with COM ± V_{os} of op amp C.

Of particular interest is how much v_9 rises during this 15.8 mS:

$$\text{rise in } v_9(t) = v_7(0)(1 - 0.967) = 0.033 \times v_7(0)$$

$$\text{rise in } v_9(t) = 0.033 \times v_7(0) \qquad (21.4)$$



OK, let's test out our new equation using real data. My bench model has a R100 = 4.64K and C7 = 69 uf. So RC = 0.32. This means that

$$\text{rise in } v_9(t) = 0.048 \times v_7(0)$$

$v_7(0)$ equals (-4.5 divisions x 50 mV/division) = -0.225V.

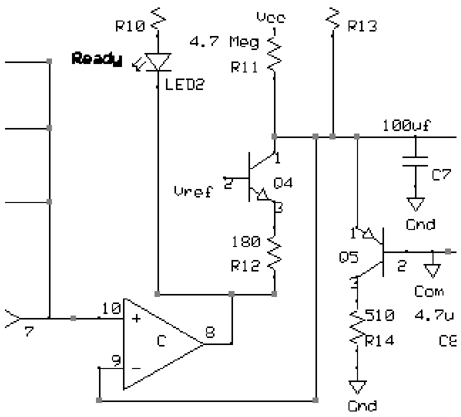
$$\text{rise in } v_9(t) = 0.048 \times v_7(0)$$

$$\text{rise in } v_9(t) = 0.048 \times 0.225V$$

$$\text{rise in } v_9(t) = 11\ mV$$

The picture shows a rise in $v_9(t)$ of 23 mV. So what is wrong here? If you look closely at $v_9(t)$, you will see that it changes on each cycle. This hints of noise. By setting the scope to record the envelope of this signal, we see that it changes by 9.5 $mV_{p\text{-}p}$. So a calculated rise of 11 mV might actually be as high as 20.5 mV. Recall that $v_7(t)$ equals 10,000 times the input voltage so 9.5 $mV_{p\text{-}p}$ would take 0.95µV of noise at the input. I realize this is not a very satisfying answer but after a few weeks work, that is the best I got.

Next we will focus on how V9 moves down.

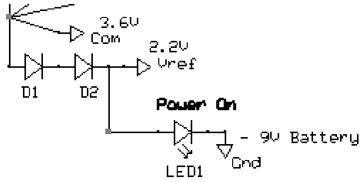


V9 can move down very quickly. This is accomplished by Q4.

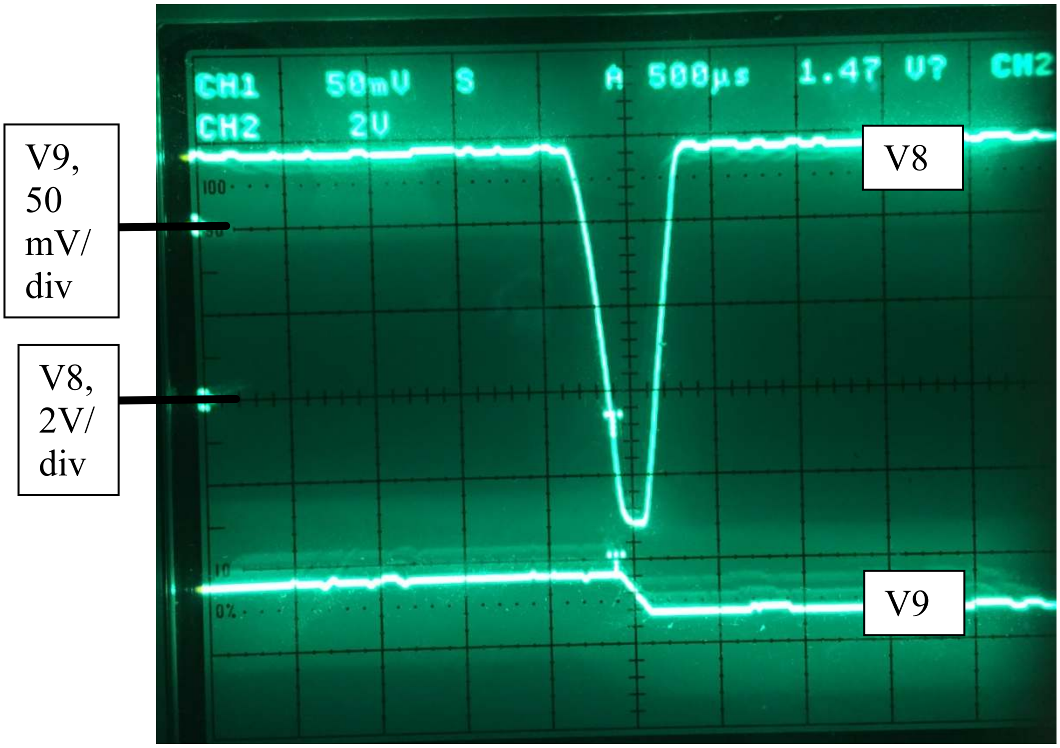
V8 can either be at ground which is negative saturation, or at a voltage more positive than V_{ref} which is about 2.2V as was shown back on page 32. While V8 is at ground, Q4 is on. While V8 is above V_{ref}, Q4 is off.

When V10 is less than V9, V8 is at ground. Q4's base is at about 2.2V so Q4 acts as a current source. The current equals

$$\frac{(\mathrm{Vref}-\mathrm{Vbe3})}{R12}=\frac{(2.2V-0.65V)}{180}=8.6\ \mathrm{mA}. \qquad (22)$$

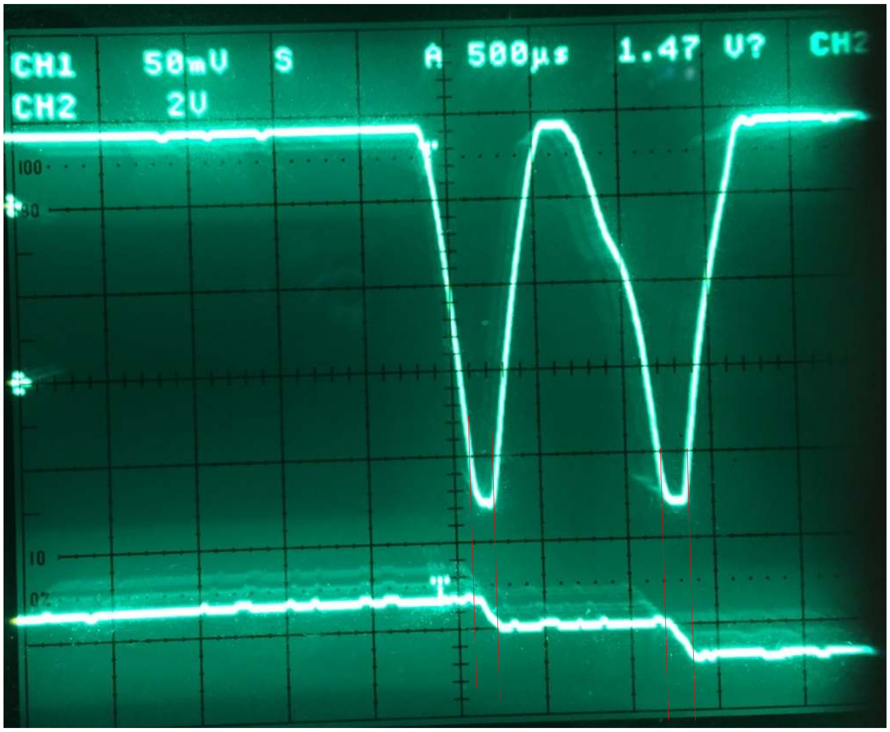


This current will flow as long as V9 is above V_{ref}. Below this voltage, Q4 is cut off. This also puts a limit on V7. The best way to see this is to look at voltages with respect to the COM node. V_{ref} is 1.3V below this reference. So as long as V7 is no more negative than -1.3V with respect to COM, Q4 will operate correctly.



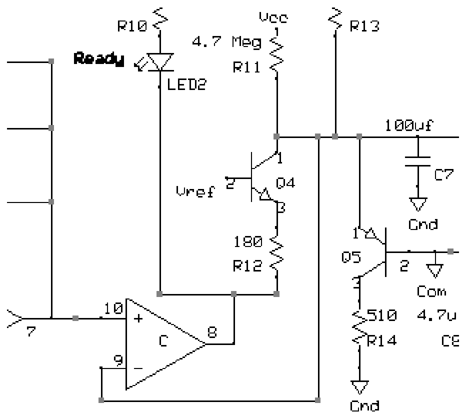
The top trace is V8. Most of the time it is at 2.8 divisions x 2V/div = 5.6V above COM. To turn on Q4, V8 dips below about - 2V.

The bottom trace is V9. It is at - 4.2 x 50mv = -219 mV before V8 dips and - 4.6 x 50 mV = - 230 mV after V8 dips low.



In this example, V8 dips twice and you can see how V9 steps down twice.

Note that in both of these examples, V8 did not reach down to ground which is about 3.6V below COM. The circuit was able to get sufficient current flow through Q4 with less drive voltage.



When V8 is at ground, Q4 is full on and its collector current is drawn from C7. We know that

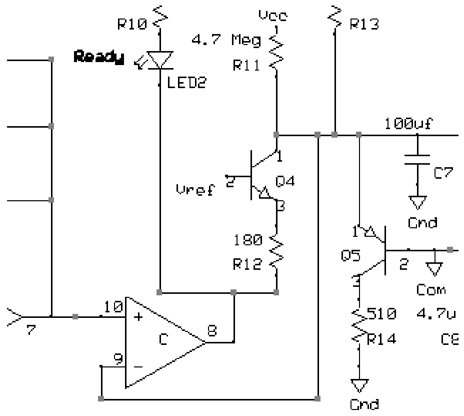
$$I_{C_4} = C_7 \frac{dv_9}{dt} \tag{23}$$

Plugging in what we know yields

$$8.6\,mA = 100\ \mu F\ \times \frac{dv_9}{dt}$$

So $\frac{dv_9}{dt}$ = 86 mV /ms.

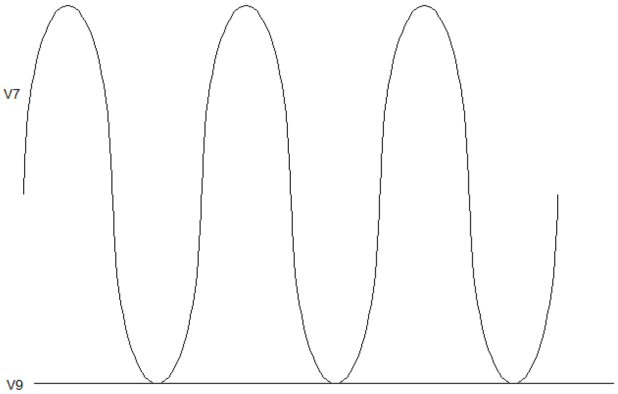
Why do we care how fast we can pull V9? The answer is rather subtle. Consider the case of detecting touchdown when a large AC voltage is present.



First consider the steady state case of holding the negative peak when a large AC signal is present.

Here we have V7 being applied to the negative peak detector and V9 parked at its negative peak. Since this is a 60 Hz signal, the period is 16.7 ms. This means that every 16.7 ms, V9 is pulled down to match the negative peak of V7. It is then free to slowly rise for about 16.7 ms until it hits V7 coming down again. During that 16.7 ms, the current flowing in R11 feeds into C7 and causes it to rise. Ignoring op amp C's bias current plus any leakage currents in Q4 and Q5, this current is around $\frac{9V-V_{com}}{R_{11}} = \frac{9V-3.6V}{68K} = 79\ \mu A$. Since V9 is going to be changing only a tiny amount compared to the voltage across R_{11}, we can treat this as a constant current and say that

$$I_{R_{11}} = C_7 \frac{dv_9}{dt} \tag{24}$$



Plugging in what we know yields

$$79\mu A = 100\ \mu F\ \times \frac{dv_9}{16.7\ ms}$$

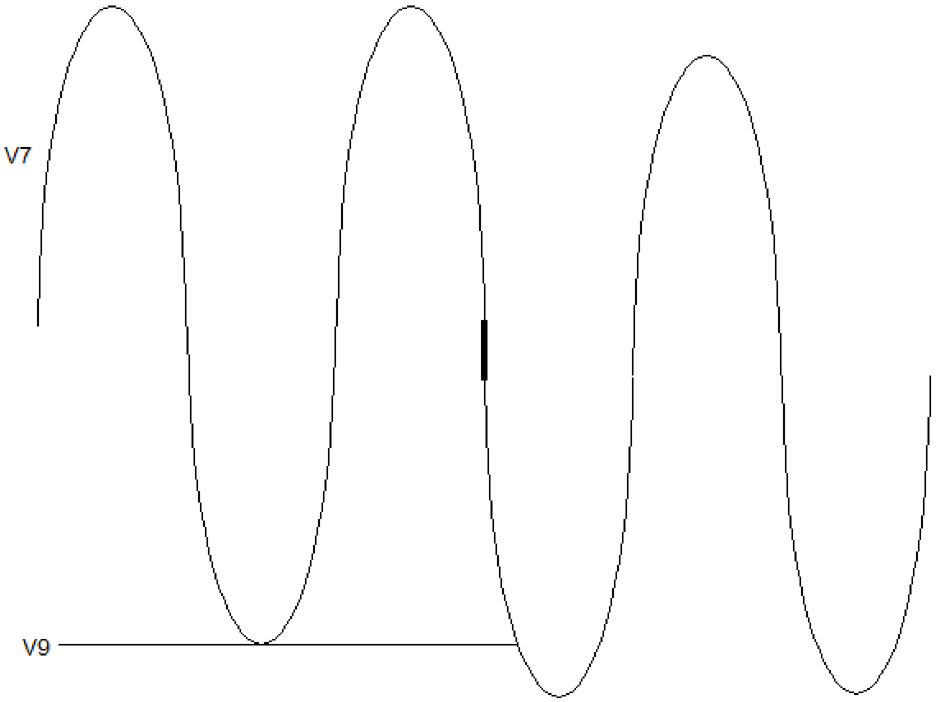
$$\text{So}\ \ dv_9 = \frac{79\ \mu A \times 16.7\ ms}{100 \mu F} = 13\ mV.$$

V9 will have drifted up 13 mV. This is a rate of

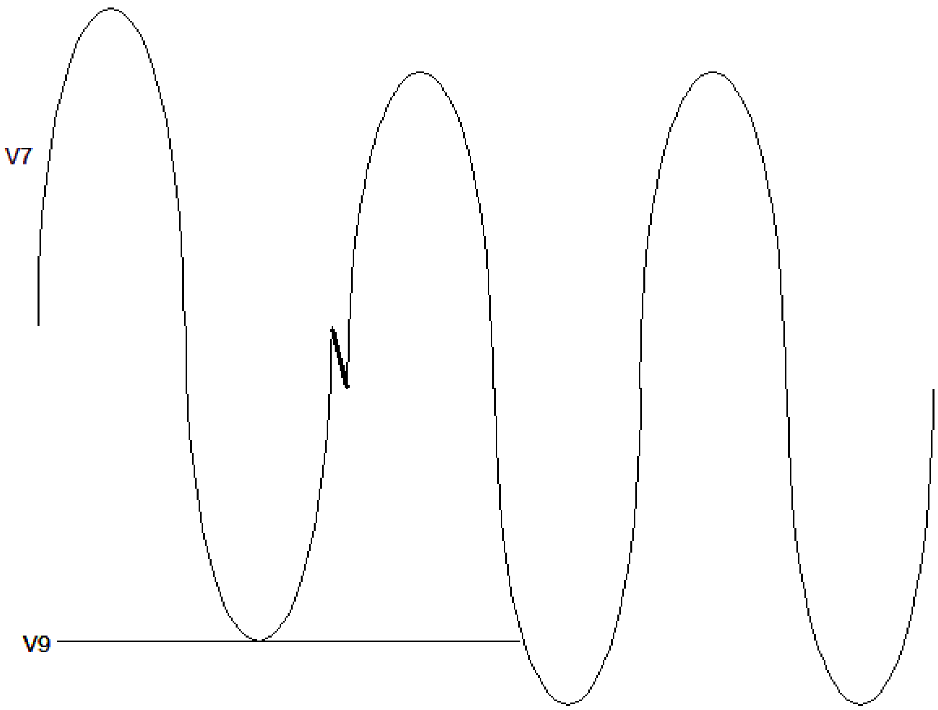
$$\frac{dv_9}{dt} = \frac{I_{R11}}{C_7} = \frac{79\ \mu A}{100\ \mu F} = \ 0.79\ mV/ms \tag{25}$$

To recap, I can move V9 down at a rate of 86 mV/ms and it will drift up at a rate over 100 times slower. This is the analog equivalent of digital memory.

Now consider the case where we have touchdown which causes a sudden drop in V7.

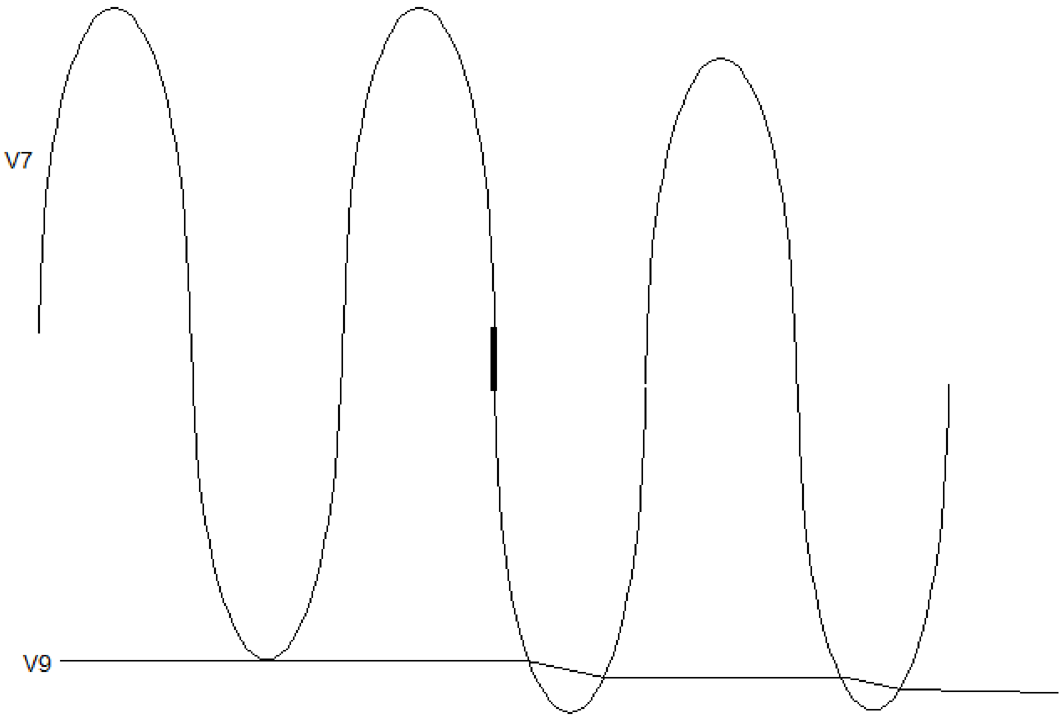


I have highlighted this event with the thick vertical line. Touchdown causes a sudden drop in V7. The AC noise is added to this drop so you can see that the negative peak has shifted by the same amount as the length of the vertical line. This drop in V7 can occur anywhere along the sine wave but will only be detected by the negative peak detector when V7 drops below the previously stable V9 value.



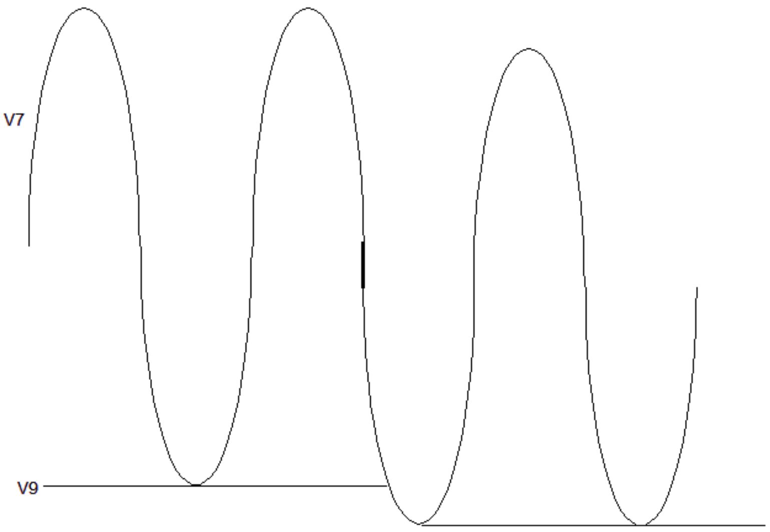
If touchdown occurred while V7 was rising, you would see something like this. I have slowed down the rate of change of the touchdown signal to make it easier to see. The drop in V7 occurred while V7 was rising, but in the end, the negative peak detector sees it near the next negative peak.

Now consider two cases.



In the first case, V9 moves down slowly relative to the rate of change of the AC noise. It can only fall while V7 is less than V9. But before V9 can catch up, V7 has turned around and headed back up again. On the next cycle, V9 again tries to reach the negative peak but runs out of time. Eventually it will get there. But in the meantime, the drop in V7 caused by touchdown is being divided into two or more pieces. If it was small, then V9 might not drop enough at one time to trigger a touchdown indication.

If the voltage drop due to touchdown is large, it might take a few cycles for V9 to catch up to the negative peak of V7. This is harmless as long as the first drop is large enough to detect.



In the second case, V9 keeps up with V7.

From page 16 we know that the minimum drop in V7 that we must detect is 32.8 mV. For V9 not to miss any of it, we must be able to slew 32.8 mV.

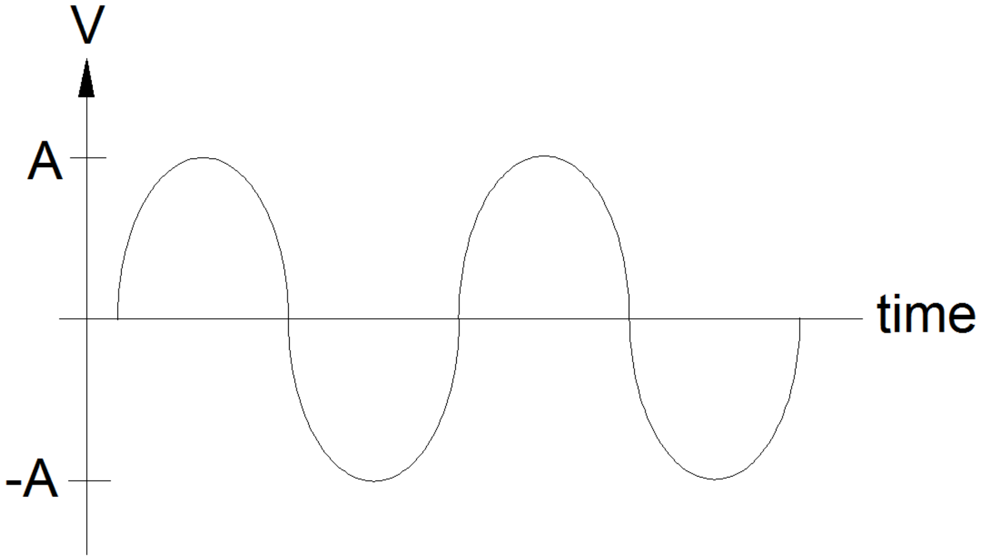
From page 52 we found that the slew rate for V9 is 86 mV /ms. So to move V9 32.8 mV

$$\frac{94\ mV}{1\ ms} = \frac{32.8\ mV}{X\ ms} \tag{26}$$

Solving for X we get

$$X\ ms = \frac{32.8\ mV}{86\ mV} \times 1\text{ ms} = 0.381\text{ ms} \tag{27}$$

So it will take Q4 0.381 ms to move V9 32.8 mV. This time interval is over at the negative peak of V7.



The 60 Hz noise signal can be described as

$$V7(t) = A \sin(\omega t) \quad \text{where } \omega = 2\pi f \quad \text{and the angle is in radians.} \tag{28}$$

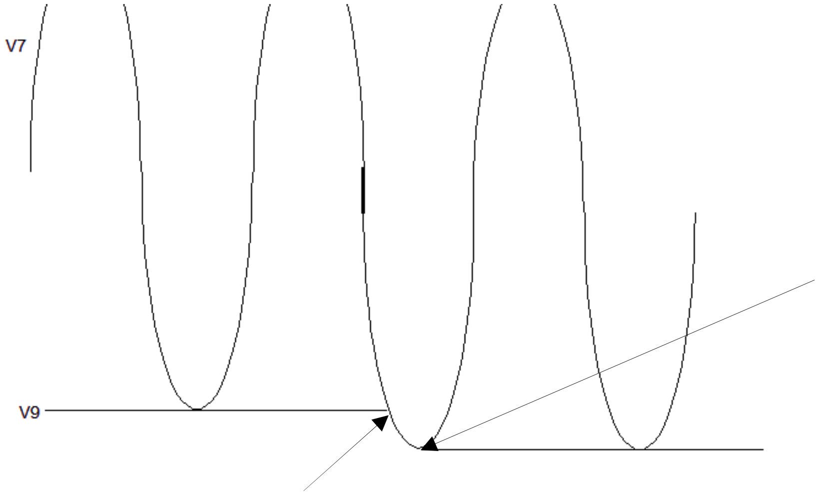
The frequency is 60 Hz so 2π X 60 Hz = 377:

$$V7(t) = A \sin(377t) \tag{29}$$

The maximum value for A can be found from knowing that V7 cannot be lower than V_{ref} (see page 50). V7 swings symmetrically around the COM node which is two diode drops above V_{ref}. So our sine wave will swing down to V_{ref} for its maximum negative value. Therefore, we can set $A = 2 \times V_{diode} = 2 \times 0.65V = 1.3V$.This gives us

$$V7(t) = 1.3 \sin(377t) \tag{30}$$

as our model for the maximum AC noise.



Given the frequency is 60 Hz, we know that the period is $\frac{1}{f}$ = 16.667 ms. The negative peak is ¾ the way through the cycle so is at ¾ X 16.667 ms or 12.5 ms.

If we back up by 0.381 ms, we will be at the point of contact where V9 hits V7. So that is at 12.5 ms - 0.381 ms = 12.12 ms. Plug that into the equation for V7(t) and we get

V7(12.12 ms) = 1.3sin(377 x 12.12 ms) = -1.29V. Don't forget that this is in radians and not degrees.

So V9 hits V7 when V7 is at -1.29V. In 0.381 ms, V7 is going to be at -1.3V. That is a movement of -1.3V - {-1.29V} = -10mV in 0.381 ms or -26.2 mV/ms.

We have already figured out from page 52 that V9 can be made to fall at a rate of 86 mV/ms. Therefore, V9 can reach the negative peak in time. This does not necessarily mean that V9 exactly follows V7.

Just for grins, let's see if V9 is fast enough to match the speed of V7.

We know from equation 30 that in the maximum case

$$V7(t) = 1.3\sin(377t) \qquad (30)$$

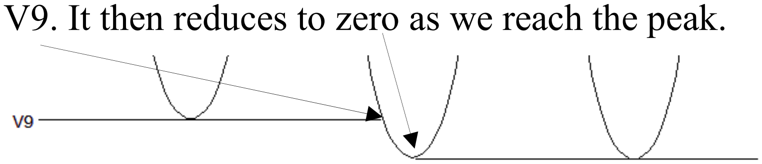
The rate of change of V7(t) can be found by taking the derivative:

$$\frac{dV7}{dt} = 1.3 \times 377 \cos(377t) \qquad (31)$$

$$\frac{dV7}{dt} = 490 \cos(377t) \qquad (32)$$

I have taken the derivative of the sin function and then used the chain rule on the angle.

From this figure we can see that the rate of change of V7 is highest at the point V7 equals V9. It then reduces to zero as we reach the peak.



V7 equals V9 at 12.12 ms as shown on page 59. So the maximum rate of change of V7 is at 12.12 ms:

$$\frac{dV7(12.12\ ms)}{dt} = 490\cos(377\ X\ 12.12\ ms) \qquad (33)$$

$$\frac{dV7(12.12\ ms)}{dt} = -70\ mV/ms \qquad (34)$$

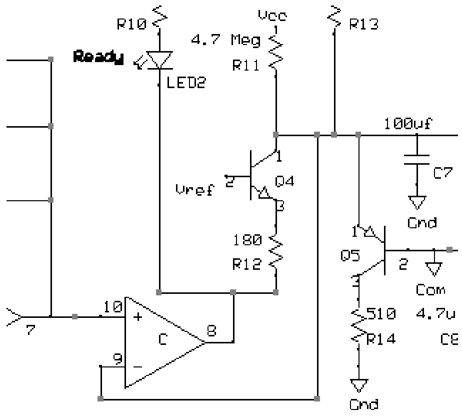
At -86 mV/ms, V9 is faster than the maximum rate of change of the AC noise component of V7 (-70 mV/ms).

Note that the circuit can be scaled. If we raise the battery voltage and also adjust V_{com} and V_{ref}, we can process a larger V7(ac).

To recap, the negative peak detector is able to pull V9 down fast enough to follow the maximum AC noise. This enables the detection of the smallest touchdown event with the maximum AC noise.

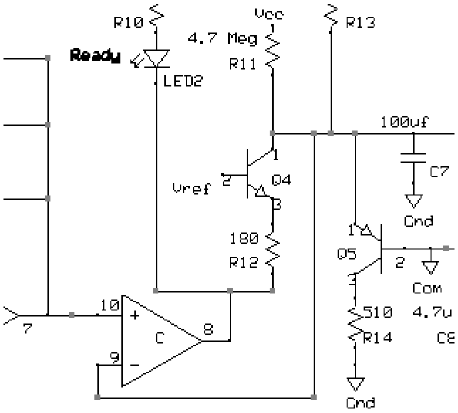
It took me a long time to reason this out so don't feel bad if this made your brain hurt.

Just in case you are totally lost, we just finished a discussion how the negative peak detector works long after power up when op amp C is working as a comparator. We will revisit this circuit later when we talk about power up.



Consider the current path that quickly pulls V9 down. We know it flows through the collector of Q4. From there, most of it flows out the emitter of Q4, through R12, and into pin 8 of op amp C. The output stage of the op amp conducts this current into ground. The current is pulled out of ground as it passes up into C7. Out the top of C7, we finish the path as we reach the collector of Q4. In this way the current path is fully defined, minimized, and controlled.

This current is the single largest source of noise in the circuit. It is absolutely essential that this current path avoid passing through any sensitive nodes. The most sensitive node is COM. Noise on COM directly couples into the amplifiers. So even though it would be nice to connect C9 to COM rather than ground, it injects way too much noise. I will talk more about the COM node later.

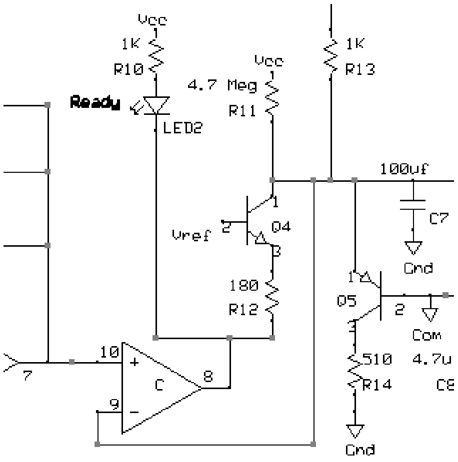


Now let's consider the special case where op amp C acts in a linear fashion. V8 will sit at a voltage close to V_{ref} - V_{be4}. The current flowing in Q4's collector will be near the current flowing in R11. A perfect balance will exist where V9 will be constant and equal to V10 plus the input offset voltage of op amp C.

This case is only possible if V7 is dead quiet. Any tiny drop in voltage will be quickly followed by Q4 turning on. Any tiny rise will require time for R11 to pull V9 up. Only when V7 is dead quiet can you sit in this linear mode.

As mentioned on page 43, if D100 is conducting, there is very little AC noise present. The second voltage gain stage will then be very quiet too and V7 will be dead quiet.

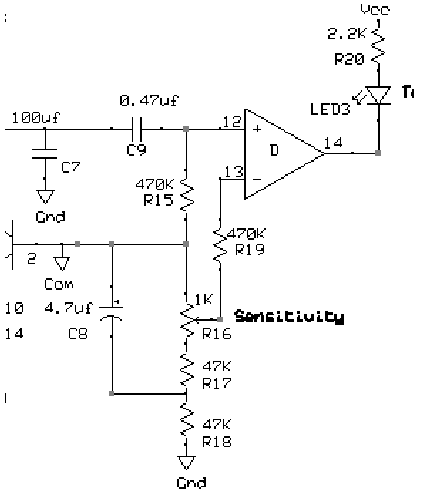
The Ready LED



Each time Q4 turns on, we also turn on LED2. Random and short flashes of light from LED2 tell us that the negative peak detector circuit is regulating V9 and is therefore stable. That means the circuit is ready to detect touchdown.

One thing that may surprise you is that V8 typically pulses low for just a few milliseconds. The resulting flashes of light from the LED are very brief yet easily seen.

Blocking Capacitor, Comparator, and Touchdown LED



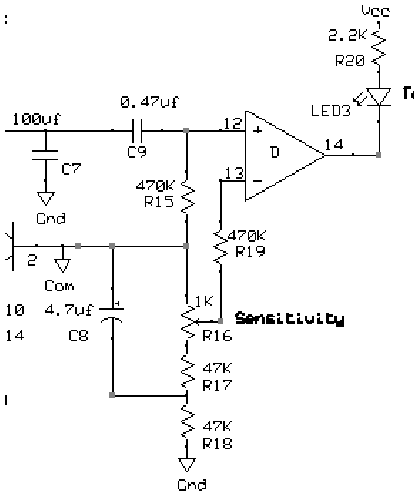
The voltage across C7 contains our touchdown event but the pre-touchdown DC voltage is a function of AC noise. We use C9 and R15 to remove this unknown DC voltage. The time constant here is slow enough to almost completely pass the touchdown edge without attenuation. The voltage at pin 12 will be near 0 with respect to the COM node before touchdown and will drop by at least 32.8 mV at touchdown. The larger the spindle resistance, the larger the voltage drop at pin 12.

Notice that the change in voltage at pin 12 is coupled through C9 and equals the change in voltage at the top of C7. The bottom of C7 is connected to ground. So the change in voltage at pin 12 is with respect to ground.

Now look at the voltage at pin 13. It comes from a voltage divider referenced to COM. Any noise between ground and COM directly appears at the input of the comparator. This may appear to be a design flaw but is actually a design compromise.

As mentioned on page 61, I cannot allow current from Q4 to pass through the COM node. So C7 must connect to ground. Yet I need V13 to be more negative than V12 and there is no voltage more negative than ground. Furthermore, I must permit V12 to dip below V13 as the touchdown occurs.

If I grounded V13 and put a positive DC bias in series with R19, I would get my initial bias of 16.4 mV on V12 relative to V13. But when V9 dips low at touchdown, V12 would also dip low and try to go below ground. This violates the input requirements of the op amp. By referencing the COM node which is more than 3V above ground, I have plenty of room for V12 to dip. By minimizing all change in voltage between COM and ground, I was able to make this arrangement work.



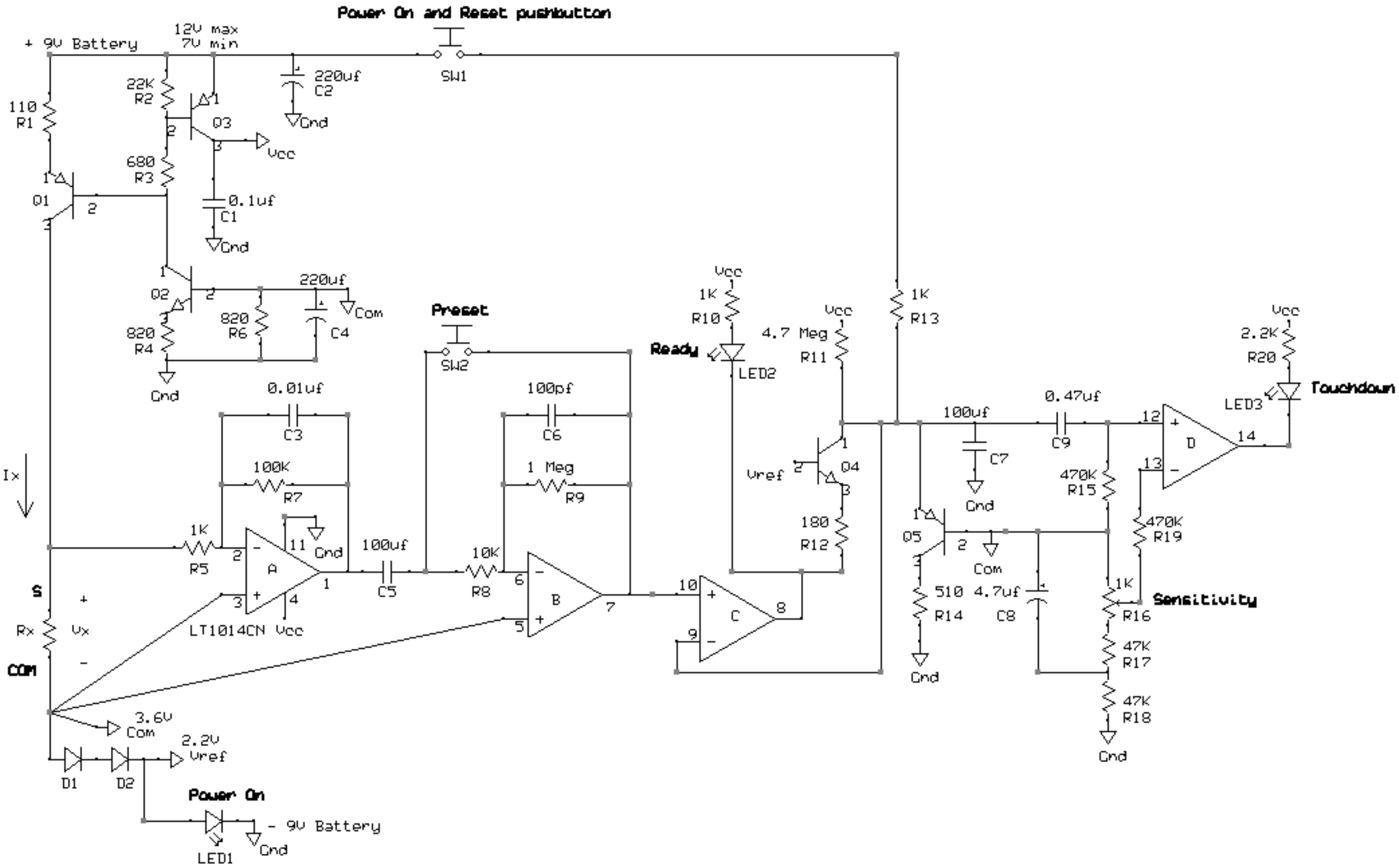
If the input offset voltage is zero and there is no spike noise, V13 is set to -16.4 mV. In practice, V13 is set at a value that accommodates input offset voltage and random voltage spikes in the shop. This is done by watching the Touchdown LED and adjusting the sensitivity until it does not flash in the idle state. Then the cutter is fed in and the Touchdown LED should flash at the moment of contact.

This is the moment of truth. If the random spike noise in your shop is greater than the touchdown signal, it is impossible for the circuit to tell them apart.

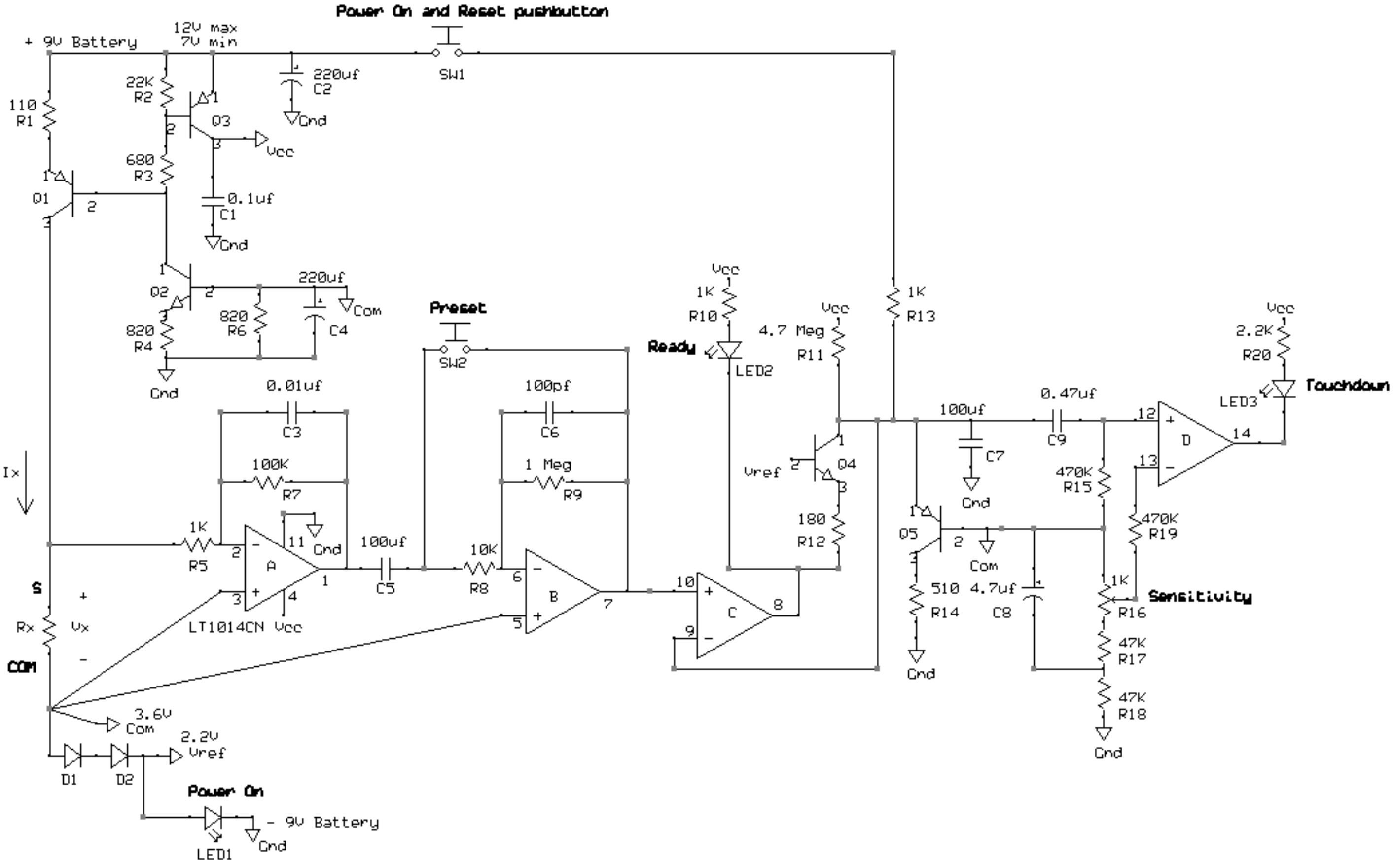
As long as V12 is lower than V13, the output, V14, will be near ground and LED3 will be lit. We will later talk about the time aspects of the circuit and you will see that LED3 stays on for plenty of time to see it.

The input bias current is a maximum of 30 nA. This current flowing in R15 generates a maximum of $I_{bias} \times R_{15} = \ 30\,nA\ \times 470K = 14.1\,mV$ $[10\ mV]$. R19 exists to balance the bias current. In this way, the voltage at the wiper of R16 is very close to the voltage on pin 12 with respect to pin 13.

The COM Node



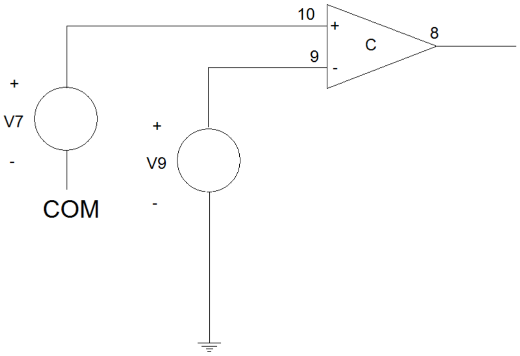
Let's follow where the COM node goes. Starting on the left, it is the reference node for V_x. A single point has been defined as my COM node. Leading directly from this point to pin 3 is my reference for the first gain stage. A second direct path exists from this point to pin 5. It is the reference for the second gain stage. COM is also connected to the touchdown LED's comparator. And lastly, COM connects to the base of Q5.



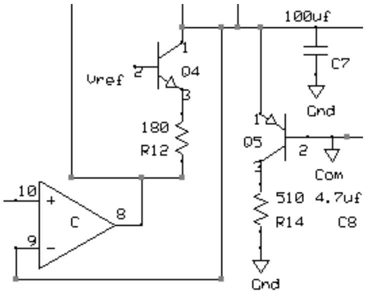
The voltage at pin 3 is extremely close to the voltage at the COM node. Any DC difference is due to the input bias current flowing out of pin 3 and the resistance of the path from pin 3 to the COM node. Any AC difference is due to noise pick up which has been minimized by keeping the path short and having a ground plane around it.

This enables the first gain stage to see almost entirely V_x with the least amount of noise from the rest of the circuit. The output, V1, is with respect to pin 3 so essentially with respect to COM. V1 is fed into our second gain stage which is also directly connected to the COM node. This second gain stage sees V1 with respect to pin 5 which is extremely close to the voltage at the COM node. Its output, V7, is almost entirely a function of V_x and is referenced to pin 5 so essentially the voltage a the COM node.

The key thing to remember here is that V7 is very quiet when measured ***with respect to the COM node***. I will emphasis this fact by taking about V7c, V7 with respect to the COM node.

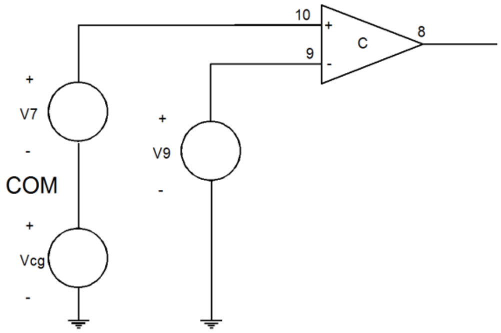


So far, I have talked about what I now call V7c, the voltage at pin 7 with respect to COM. Let's now start at C7.



The bottom end of C7 is connected to ground. This means the top end is at a voltage with respect to ground. But this node is also V9. I will emphasis this fact by taking about V9g, V9 with respect to ground.

Op amp C reacts to V10 - V9. But V10 equals V7 relative to COM and V9 is relative to ground. This picture is not complete. We need to know the voltage from COM to ground.



I will define Vcg as the voltage on COM with respect to ground. Vcg provides the missing piece. I can now talk about V10 - V9

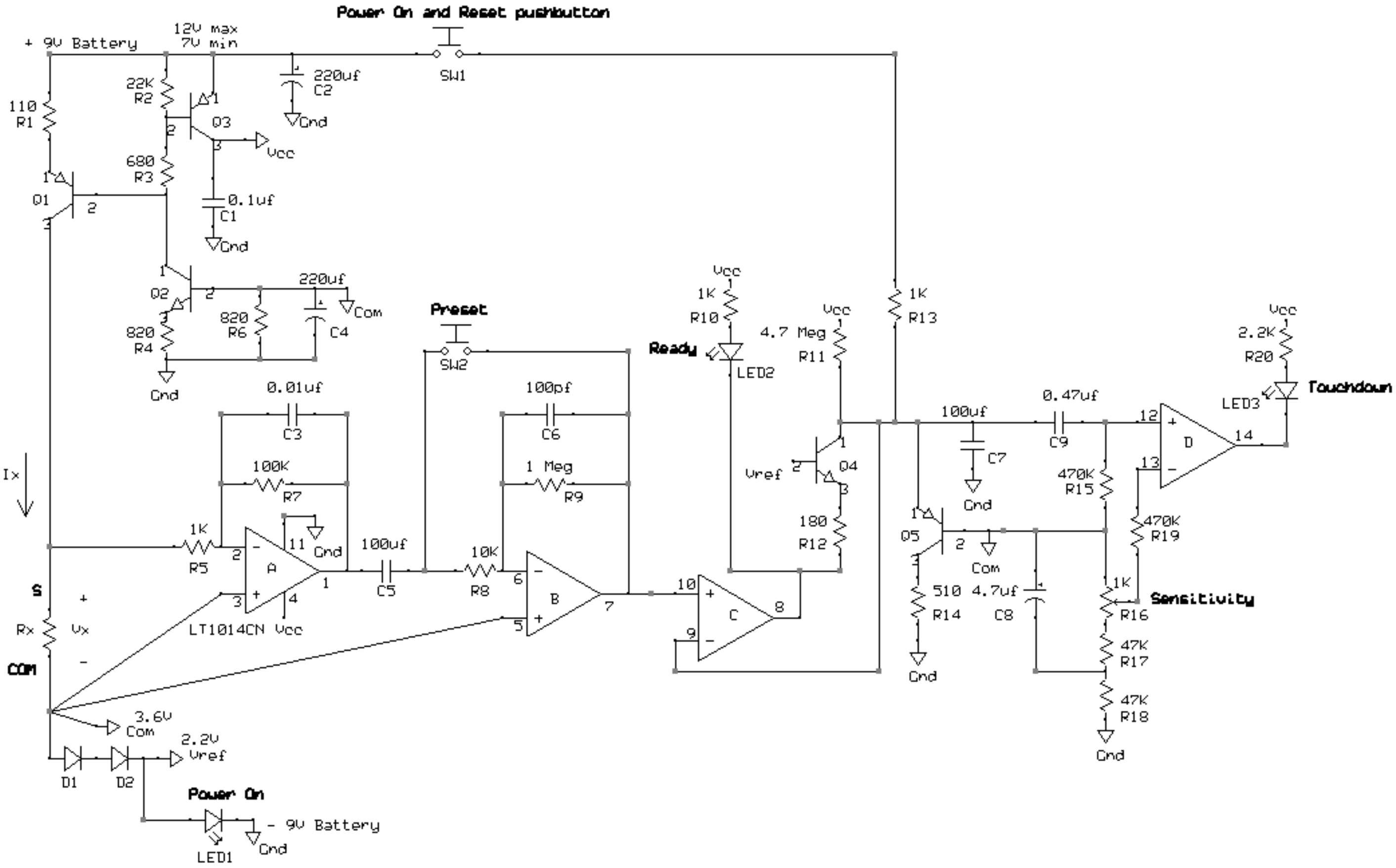
$$V_{10} - V_9 = \left(V_7 + V_{cg}\right) - \left(V_9\right)$$

The negative peak detector is going to drive V9 to be equal to the most negative peak of $V_7 + V_{cg}$.

What I wanted to drive V_9 to equals the most negative peak of V_7. But now you see that this nasty little V_{cg} term has crept in. If V_{cg} is pure DC, then V_9 will have a DC offset which will be blocked by C9. But if V_{cg} contains any noise, it will be combined with V7 and look like a valid signal to the op amp.

By limiting what connects to the COM node I have been able to make it almost noise free with respect to ground. As an added precaution, I have C4 tied between COM and ground.

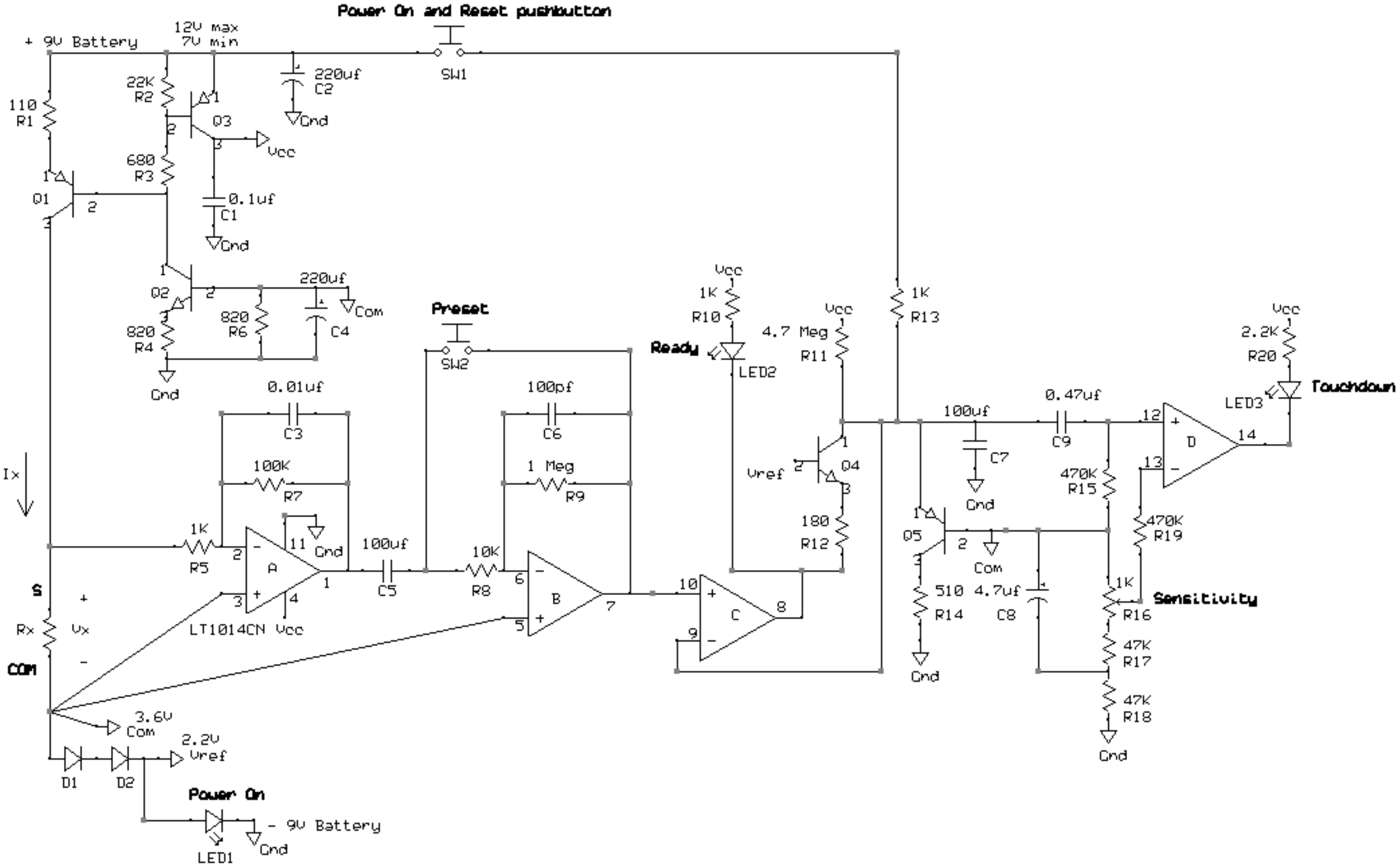
Preset and Reset



When the circuit is first powered up, we do not know the voltages on C_5 or C_7. In most cases they are not at their final value and it can take a very long time to get there. The solution is to preset C_5 and reset C_7.

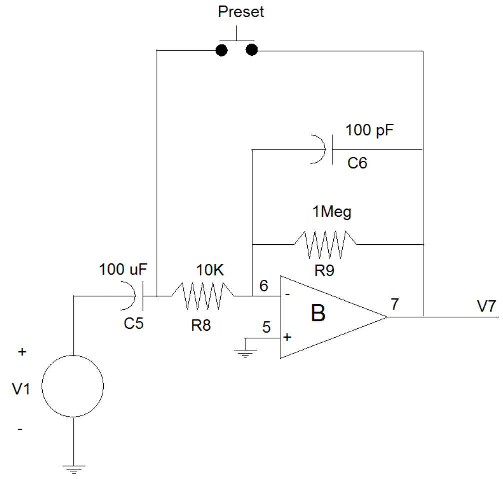
If C5 was fully discharged just before power up and R_x is small enough, op amp B will not go into saturation. This enables C5 to charge via just R8. The time constant equals $C_5 \times R_8 = 1\ second$. So in about 5 seconds V7 is stable. Given that the user is preparing to feed the cutter into the workpiece, they might not notice this delay.

If R_x is not small or if the voltage across C5 is large, op amp B can be driven into saturation. This causes a loss of feedback and C5 is charged by $R_8 + R_9$. The result is a 100 fold increase in the time constant and a very long wait. Worse yet, if V7 goes to negative saturation, it will pull V9 down to its minimum. Then we are stuck waiting for V9 to slowly rise up to a usable value.

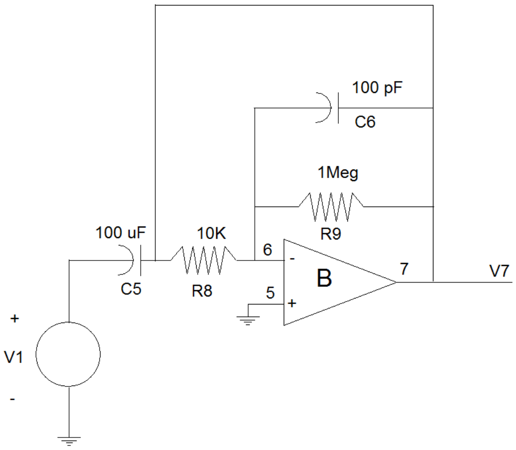


The solution is to add a Preset function. The user pushes the Preset button after power first comes on. It quickly charges C5 to near the correct value and brings V7 to stability in a fraction of a second.

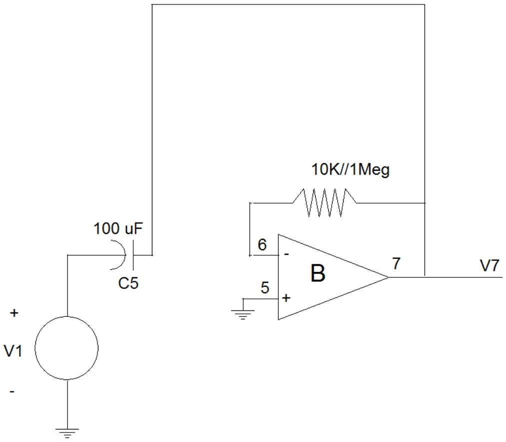
On the first pass of this explanation, I will assume an ideal op amp. Then I will add in the critical parasitics.



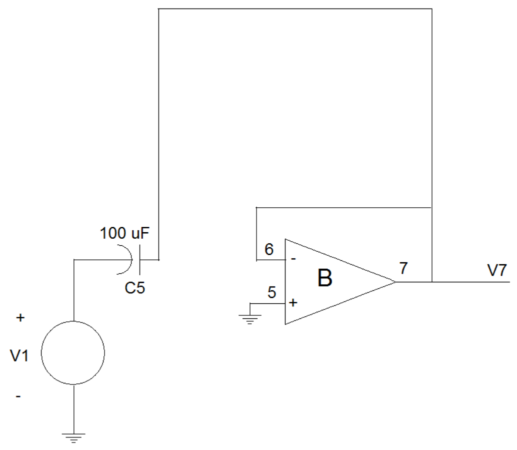
Here is the second gain stage. For now, assume that the output of the first gain stage has no AC component. I will represent it as a DC voltage source, V1. I don't know what the voltage is across C5 so don't know the voltage V7. The goal is to move V7 to zero and keep it there when the Preset pushbutton is released.



I push the Preset button. This connects the junction of C5 and R8 to V7.



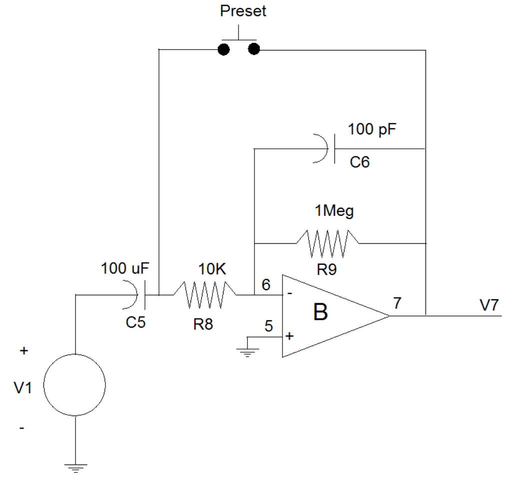
Let me redraw the circuit. I end up with R8 in parallel with R9. C5 is tied to the output of the op amp. I have not drawn C6 because it has no effect here.



Since no current can flow into the inputs of this ideal op amp, I can simplify the circuit further by replacing R8 and R9 with a short.

V7 of my ideal op amp is at zero volts because that is the voltage at the positive input.

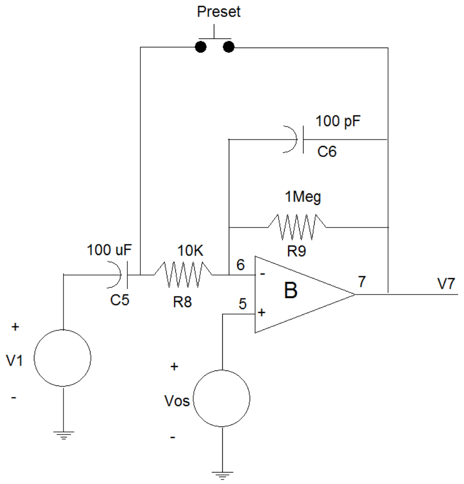
So the left end of C5 is at our unknown voltage, V1, and the right end is at ground.



When the Preset button is released, we are back to our origninal circuit but now the right end of C5 has been charged to zero volts. V6 is also at zero volts due to feedback. This puts zero volts across R8 so no current can flow in it. This means that no current flows in R9 or C6 either. Therefore, V7 is now at zero volts too.

We started with C5 as some unknown voltage. By pressing and releasing the Start button, V7 is quickly driven to zero volts.

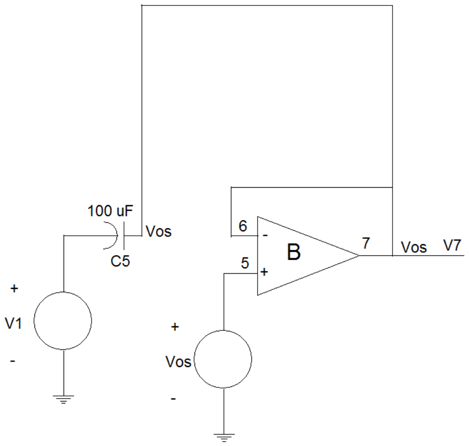
The Preset works great given an ideal op amp. Too bad you can't buy an ideal op amp.



The op amp is not ideal. It has an input offset voltage and an input bias current[8]. Let's consider the input offset voltage first.

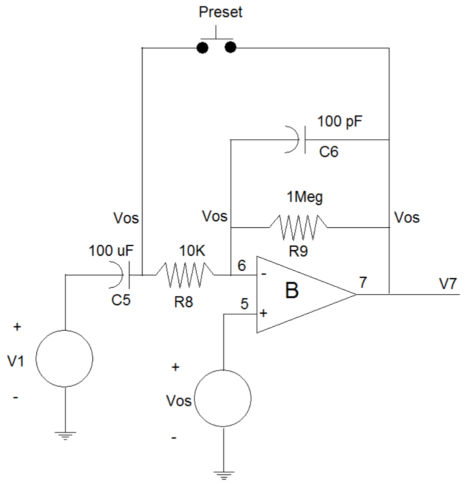
I will pull the input offset voltage, V_{os}, out of the op amp so it is easier to see. V5 is now at a voltage set by V_{os}. Due to the feedback, V6 is very close toV_{os}.

For the LT1014CN quad op amp, V_{os} worst case is somewhere between -0.3 mV and +0.3 mV.



While the Preset button is pushed, we have this circuit. V7 sits at V_{os} which is also the voltage at the right end of C5.

I need to rely on one non-ideal feature of the op amp for a moment. V7 acts as an ideal voltage source until the output current reaches a limit. Then it safely prevents more current to flow. I rely on that feature to charge up C5 safely and quickly.



When we release the Preset, the circuit reverts back to its original configuration.

Notice that R8 has V_{os} on both ends. This means that there is no voltage across it and therefore no current through it. No current flowing in R8 means no current flowing through R9. So V7 is equal to V_{os}. V7 was equal to V_{os} when the button was being pushed, and it is still equal to V_{os} now that it has been released.

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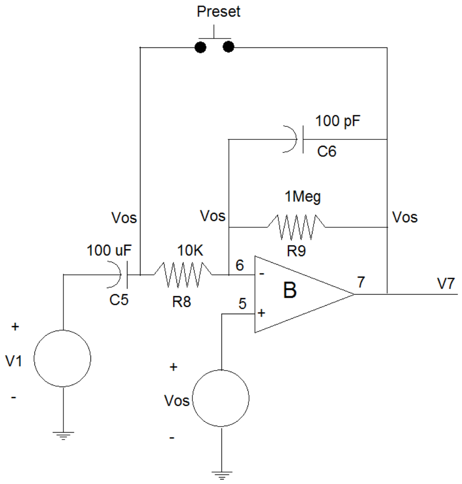
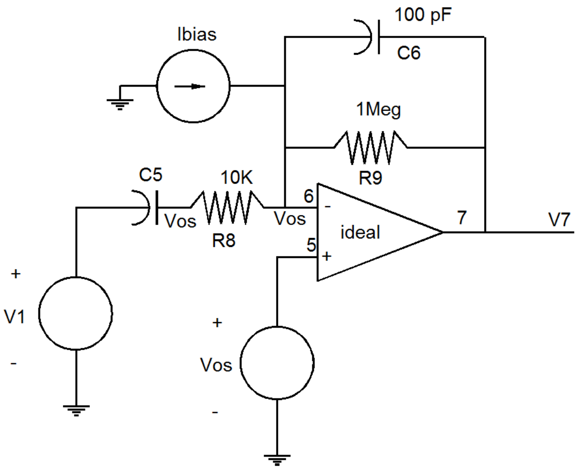
[8] It also has a gain less than infinite but that has little effect on the result in this case.



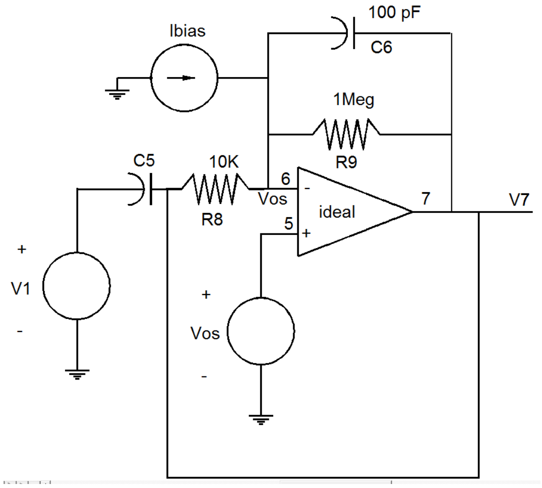
To recap: at power up, V7 was at some unknown value. We push the Preset button and V7 quickly moves to a voltage between -0.3 mV and +0.3 mV. We release the button and V7 does not move at all.

Next we will study the effects of input bias current.

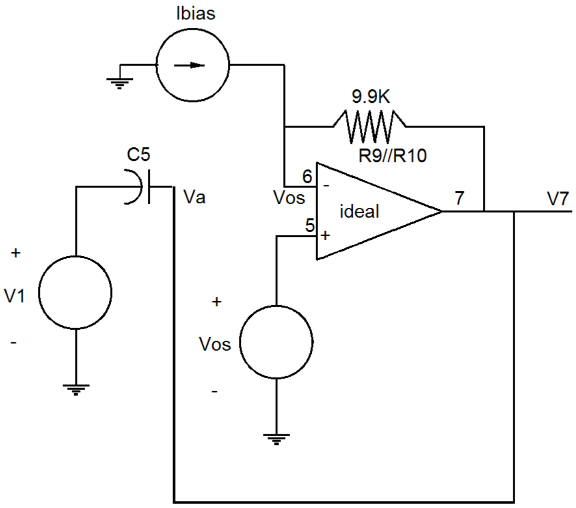


Input bias current flows out of both op amp inputs because PNP transistors are employed. The current flowing out of the non-inverting input feeds into zero resistance so does not change V5.

I have drawn the input bias current as a current source that feeds into the inverting input node. The effect is the same as having it come out of the inverting input but the drawing is less crowded.



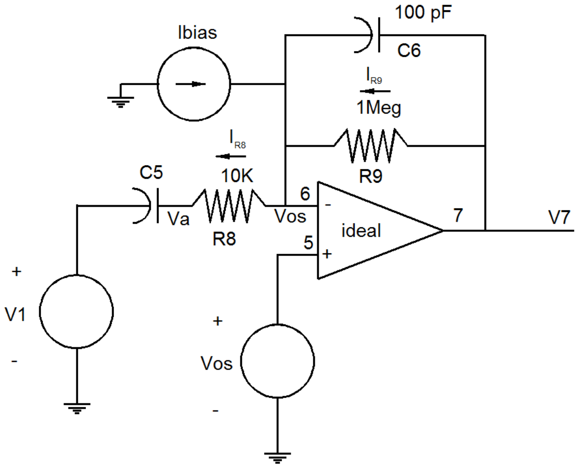
When the button is pushed, the junction of C5 and R8 is again connected to V7. Since we are still assuming the op amp has infinite gain, the inverting input is at the same voltage as the non-inverting input, V_{OS}.



Redrawing the circuit, we can see that the input bias current all flows through R9 in parallel with R10 which equals 9.9K. The voltage across the equivalent resistance of 9.9K equals $9.9K \times I_{bias}$. The left end is at V_{os} so

$$V_7 = V_{os} - (9.9K \times I_{bias}) \qquad (35)$$

This is also the voltage on the right end of C_5. Let me call it "V_a".



Now the button is released. R8 has V_{os} on the right end and at the instant the button is released, V_a on the left end still equals $V_{os} - (9.9K \times I_{bias})$. The initially current flowing through R8 towards C5 equals

$$I_{R8} = \frac{V_{os} - (V_a)}{R8}$$

$$I_{R8} = \frac{V_{os} - \left(V_{os} - (9.9K \times I_{bias})\right)}{R8}$$

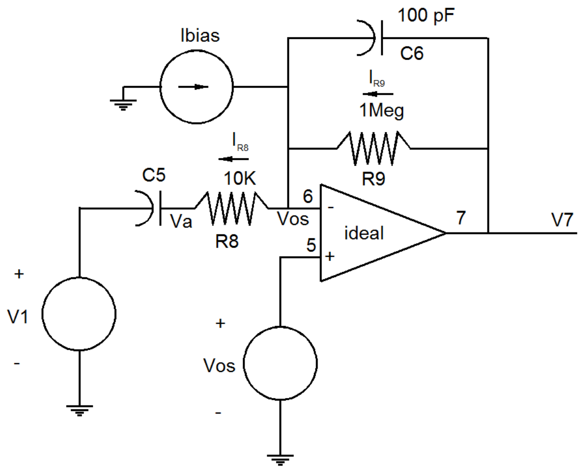
$$I_{R8} = \frac{\left((9.9K \times I_{bias})\right)}{R8}$$

This current leaves the inverting input node. Entering this node is our input bias current plus the current flowing through R9. So I can write

$$I_{R8} = I_{bias} + I_{R9}$$

I can rearrange the terms and get

$$I_{R9} = I_{R8} - I_{bias}$$



$$I_{R9} = \frac{((9.9K \times I_{bias}))}{R8} - I_{bias}$$

$$I_{R9} = \frac{((9.9K \times I_{bias}))}{10K} - I_{bias}$$

$$I_{R9} = (0.99 \times I_{bias}) - I_{bias}$$

$$I_{R9} = -0.01 \times I_{bias}$$

The minus sign says that the direction we assumed for this current is backwards. The actual current flows towards V7. We know the current through R9 and that the left end is tied to V_{os}. This lets us calculate V7

$$V_7 = V_{os} + (R_9 \times [-I_{R9}])$$

$$V_7 = V_{os} + (R_9 \times [0.01 \times I_{bias}])$$

Since this is at the instance the button is released, I will call this time t=0 and rename V_7 to $V_7(t)$. Then I can say

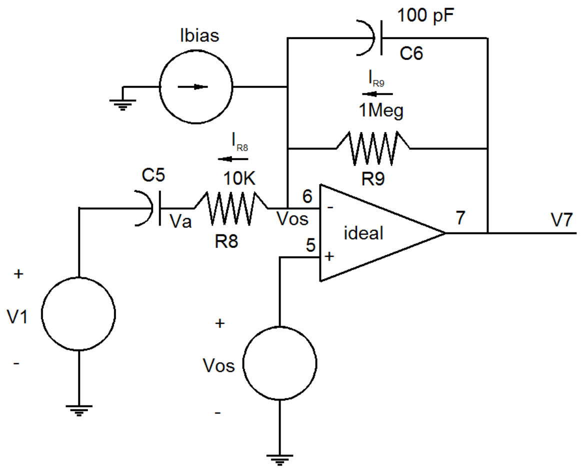
$$V_7(0) = V_{os} + (R_9 \times [0.01 \times I_{bias}])$$

The maximum I_{bias} is 30 nA for the LT1014CN. R9 = 1Meg. V_{os} can be as large as 0.3 mV. The most positive $V_7(0)$ is therefore

$$V_7(0) = 0.3mV + (1Meg \times [0.01 \times 30\,nA])$$

$$V_7(0) = 0.3mV + (0.3\,mV\ )$$

$$V_7(0) = 0.6\,mV$$



The minimum I_{bias} is zero and V_{os} can be -0.3 mV. Therefore, the most negative $V_7(0)$ is

$$V_7(0) = -0.3mV + (1Meg \times [0.01 \times 0])$$

$$V_7(0) = -0.3mV$$

We can therefore say that $V_7(0)$ is bounded by -0.3 mV and 0.6 mV.

Eventually C5 charges up and I_{R8} exponentially decays to zero. So after a "long" time, I am left with

$$I_{bias} = -I_{R9}$$

I can then write

$$V_7(\infty) = V_{os} - (R_9 \times I_{bias})$$

V7 starts at the voltage $V_7(0)$ and when it stops moving we are at $V_7(\infty)$. The change in voltage equals

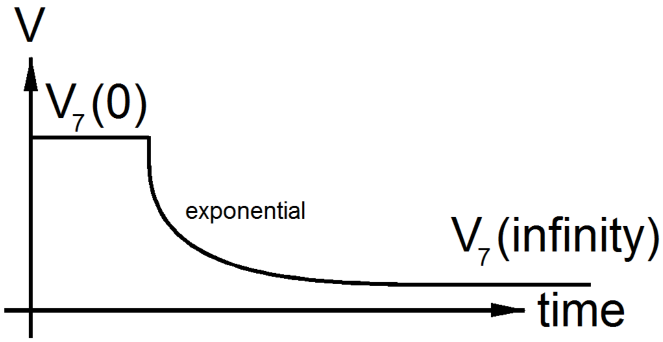
$$V_7(\infty) - V_7(0) = \{V_{os} - (R_9 \times I_{bias})\} - \{(V_{os} + (R_9 \times [0.01 \times I_{bias}]))\}$$

$$V_7(\infty) - V_7(0) = -1.01 \times (R_9 \times I_{bias})$$

Assuming the maximum I_{bias} of 30 nA and that R_9 is 1 Meg, we get

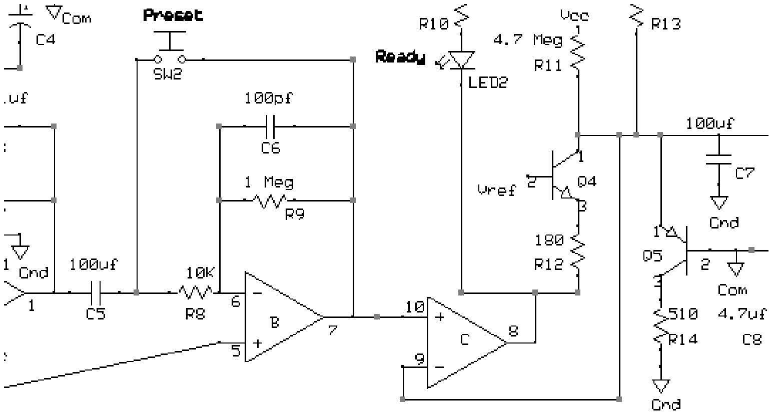
$$V_7(\infty) - V_7(0) = -1.01 \times (1\,Meg \times 30\,nA)$$

$$V_7(\infty) - V_7(0) = -30.3\,mV$$



So V_7 can fall by as much as 30.3 mV when the Preset button is released. Note that this drop is almost entirely a function of input bias current and the 1 Meg resistor. The input offset voltage canceled out of the equation.

Since the op amp stays out of saturation, the time constant is just $C_5 \times R_8$ which is $100\mu F \times 10K = 1$ second.



The negative peak detector will follow V7 as long as it goes to a lower value. If we get a touchdown event while V7 decays, V9 will follow the event. The slow decay of V7 has no effect[9].

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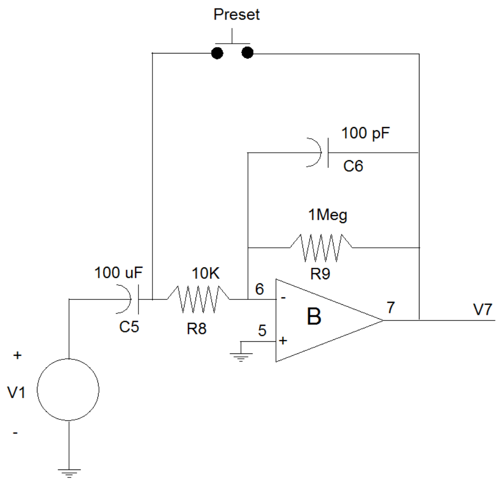
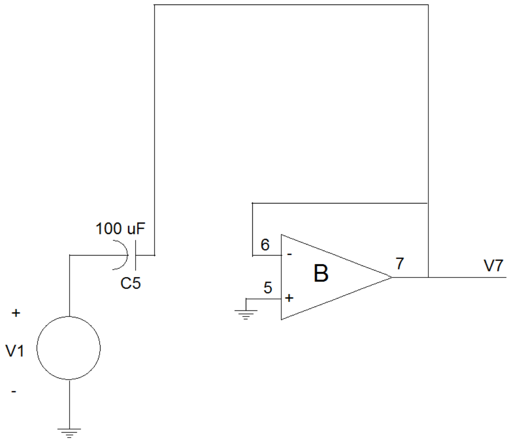
[9] If $V_7(t)$ rose in voltage, we would be in trouble since V9 would ignore the rise. A touchdown event would then have to cause V7 to drop below the previously lowest value which would be $V_7(0)$.



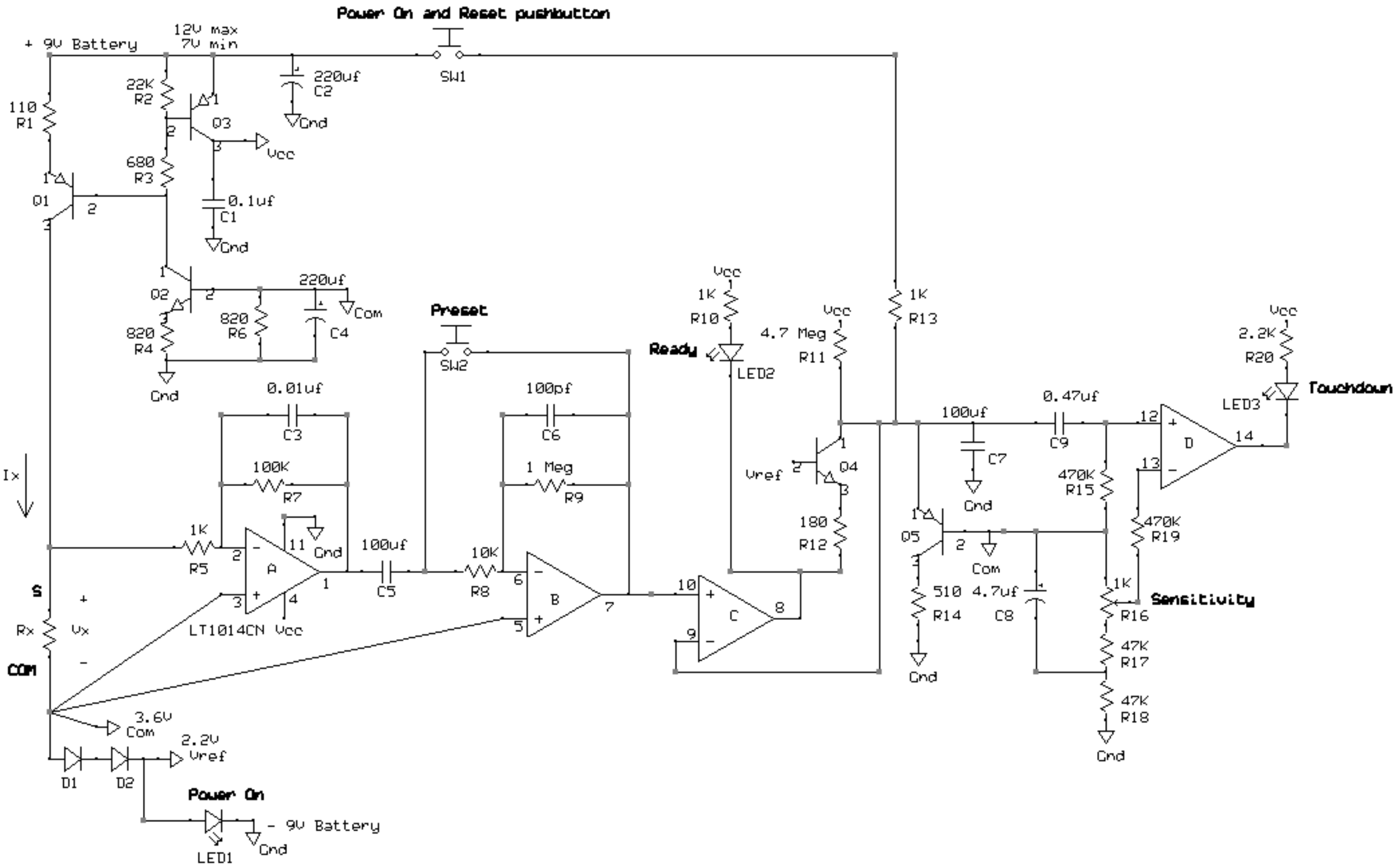
There is one case left to discuss. What if V1 is not just a DC value? What if there is AC mixed in?

First of all, consider the maximum AC value. We know from page 58 that V7 can handle up to $1.3V_{peak}$ of AC. To get $1.3V_{peak}$ out of this gain stage means that we put in $\frac{1.3\ \text{Vpeak}}{100} = 13\ mV_{peak}$. This would also be the voltage out of the first gain stage, V1.



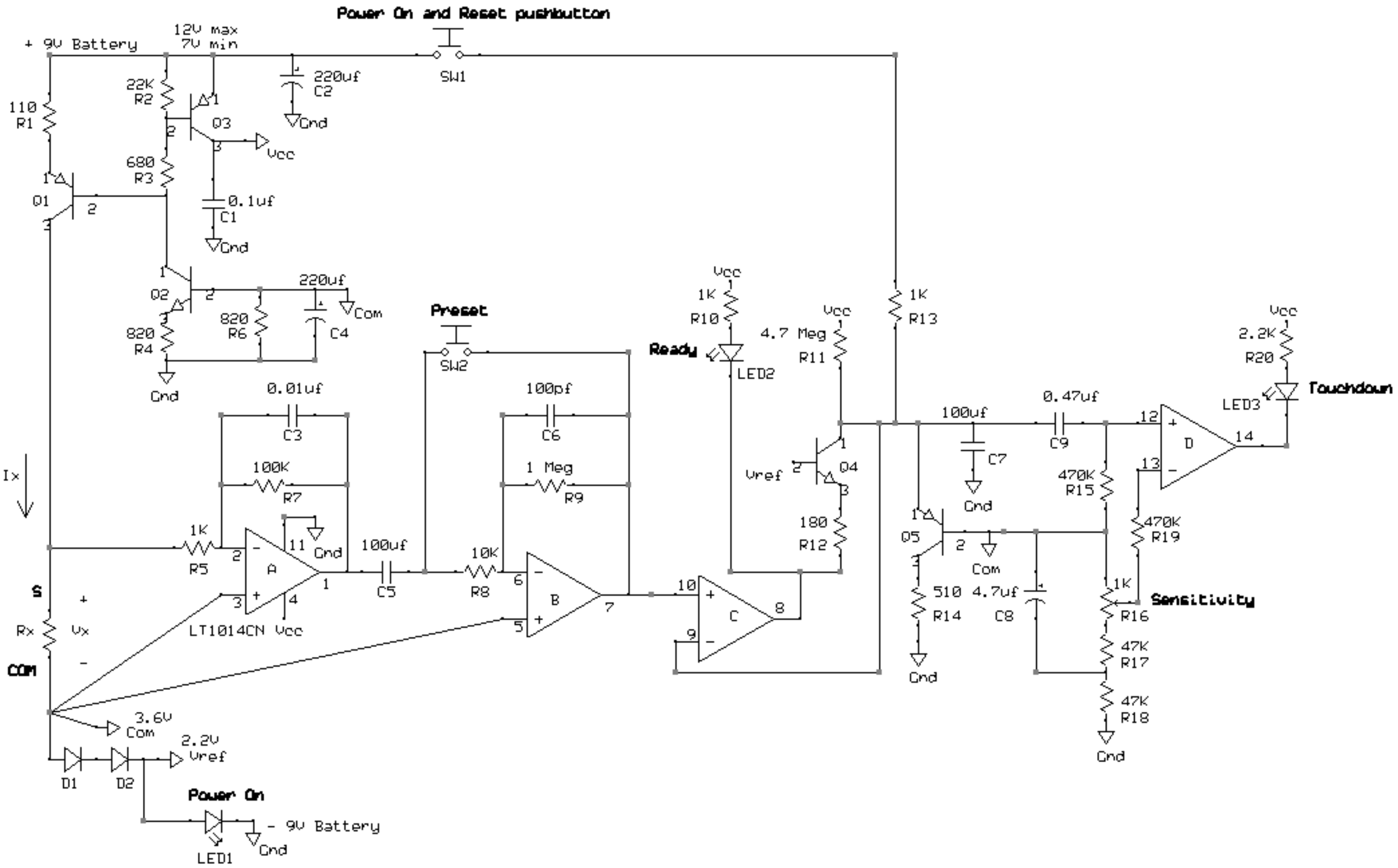
Now, think about what is going on when the Preset button is being pushed. We are forcing an AC voltage across C5. The impedance of C5 at 60 Hz is $-j\frac{1}{\omega C} = -j\frac{1}{377 \times 100\mu F} = -j26.5\ ohms$. In order to force 13 mV_{peak} across it takes $\frac{13\ mV}{26.5\ ohms} = 0.49\ mA$. This is well within the current sourcing and sinking capabilities of the op amps.

At the instant the Preset button is released, V7 will be at zero. Then as V1 continues to move, the amplifier will multiply it by -100 and drive V7. In this way, V7 will reflect just the AC component of V1 as it should.

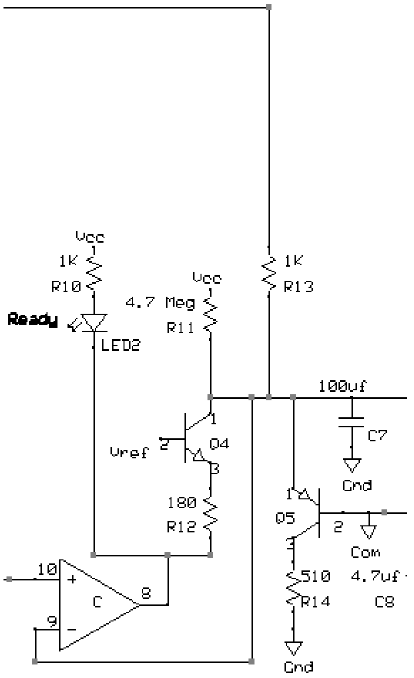


Now let's move on to the Reset for C7 which is part of the negative peak detector. This is done before each touchdown.

There are two cases. In the first case V10 is higher than V9. This causes V8 to be at positive saturation and Q4 to be off. R13 pulls up on V9 until Q5's emitter-base junction is forward biased. Then V9 is clamped to about 0.65V above the COM node. The current through R13 will be $\frac{9V-V_{EB5}-V_{com}}{R13} = \frac{9V-0.65V-3.6V}{R13} = 4.75\ mA$. With a minimum β_5 of 80, the maximum current dumped into the COM node is $\frac{4.75\ mA}{80} = 59\ \mu A$. This does disturb the voltage on COM relative to ground but no touchdown event is being detected at this time so it is harmless.



In the second case V10 is lower than V9. This causes op amp C to adjust V8 such that V9 falls to match the negative peak of V10. The worst case is when V7 is 1.3V below COM. This will cause op amp C to drive V8 to negative saturation which is near 0 volts. That turns Q4 on to the maximum value. We then get a collector current out of Q4 equal to $\frac{V_{ref}-V_{be2}}{R12} = \frac{2.2V-0.65V}{180\ ohms} = 8.6\ mA$. The maximum current flowing in R13 is when Q4 is just about to saturate. This means that V9 equals V_{ref}. $I_{R13} = \frac{9V-V_{ref}}{R13} = \frac{9V-2.2V}{1K} = 5.8\ mA$. This value is less than the maximum that can be sunk by Q4 so Q4 will always win. V9 will always be pulled down to match the negative peak of V7 when the Power On and Reset button is pushed.



In this worst case, $8.6\,mA - 5.8\,mA = 2.8\,mA$ is available to charge C7. As V9 rises, more current will be available. The slowest V9 can move is

$$I = C\frac{dV}{dt} \qquad (36)$$

$$2.8\,mA = 100\mu f\frac{dV_9}{dt}$$

$$\frac{dV_9}{dt} = \frac{2.8\,mA}{100\mu f}$$

So

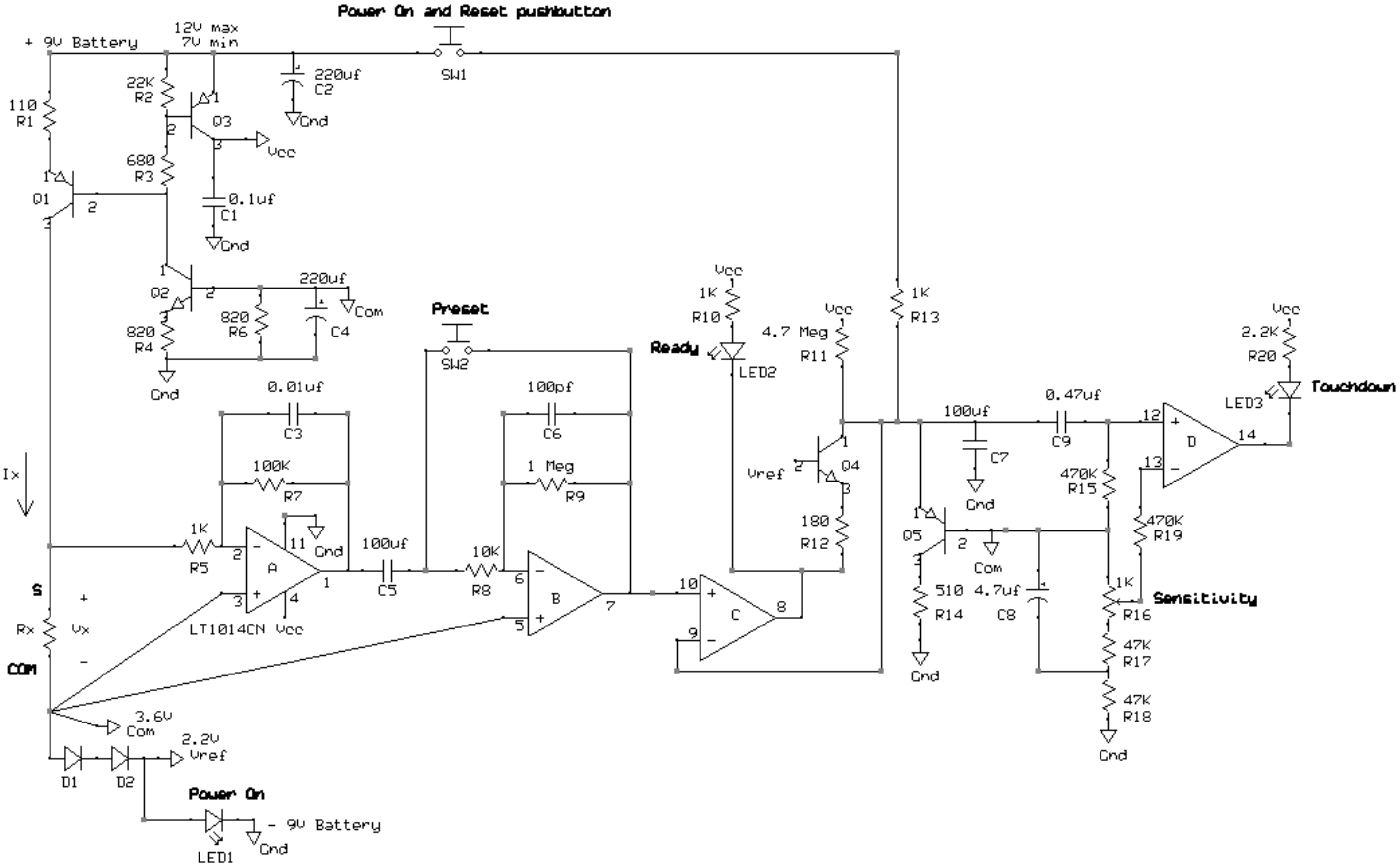
$$\frac{dV_9}{dt} = 28mv/ms$$

If I assume this slowest rate exists for the full range of V9 which is from 0.65V above COM to 1.3V below COM, it will still take less than $\frac{1.3V+0.65V}{28V} = 80\,ms$ to move V9 into position. This is occurring while the button is being pushed. The user will barely notice this delay.

Therefore, when the Power On and Reset button is released, the circuit will be following V7. If V7 is stable, the circuit will be ready to detect touchdown.

If V7 is not stable, further time will be needed to establish the new negative peak. The more noise mixed in with $V_{x,}$ the more time it will take to stabilize.

The Power On and Off Sequence

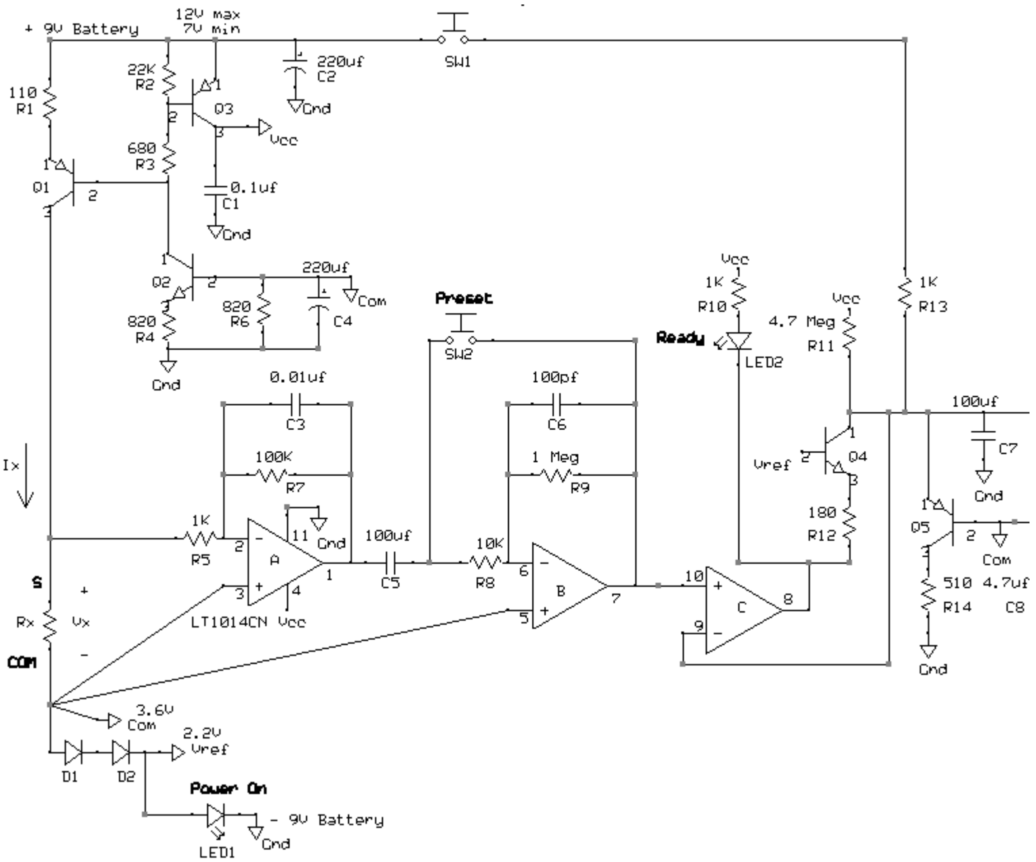


Back on page 31 I discussed most of the power control circuit.

To review, if I_x is flowing out of Q1, it flows mostly through our diode string formed by D1, D2, and LED1 and through R6. That causes a voltage at the COM node with respect to ground. This voltage is applied to the base of Q2 which causes a collector current to flow. This collector current flows through R3, generates a voltage, which is applied to R1 via Q1. The current through R1 is essentially I_x.

If I_x is not flowing, there is no voltage drop across our diode string, Q2 is off, and no voltage can be developed across R3 or R1.

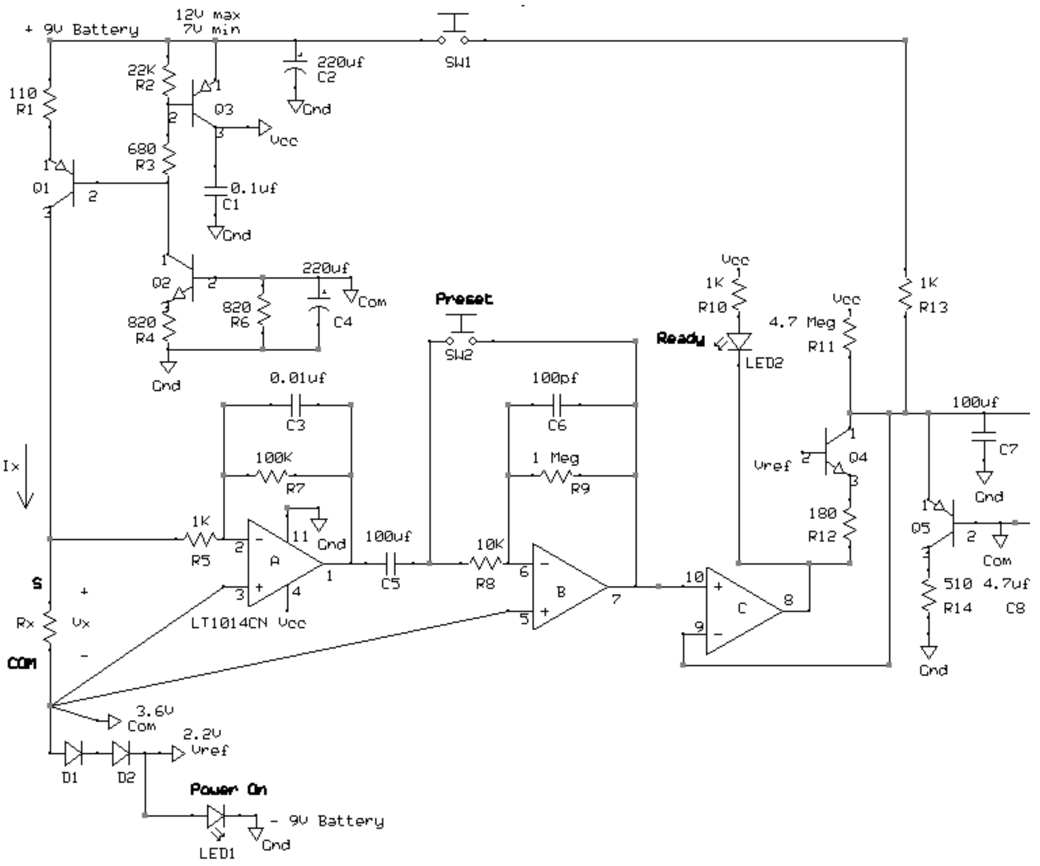
So we understand our ON state and our OFF state. How do we move between them?



The easy transition is from ON to OFF. By disconnecting one of the test probes, I_x loses its path. This stops the current into our diode string and bleed resistor R6. C4 will hold up this voltage for short disruptions in I_x but then the COM node drops to ground. That stops the current in Q2 so the voltage across R3 and R1 go to zero.

Moving from OFF to ON is not straight forward. If V_{com} is greater than about 0.65V, Q2 will start to conduct. This will cause I_x to become non-zero and that, in turn will cause V_{com} to rise a bit more. This positive feedback will continue until the diode string is stable. I just need to start the cycle and stand back.

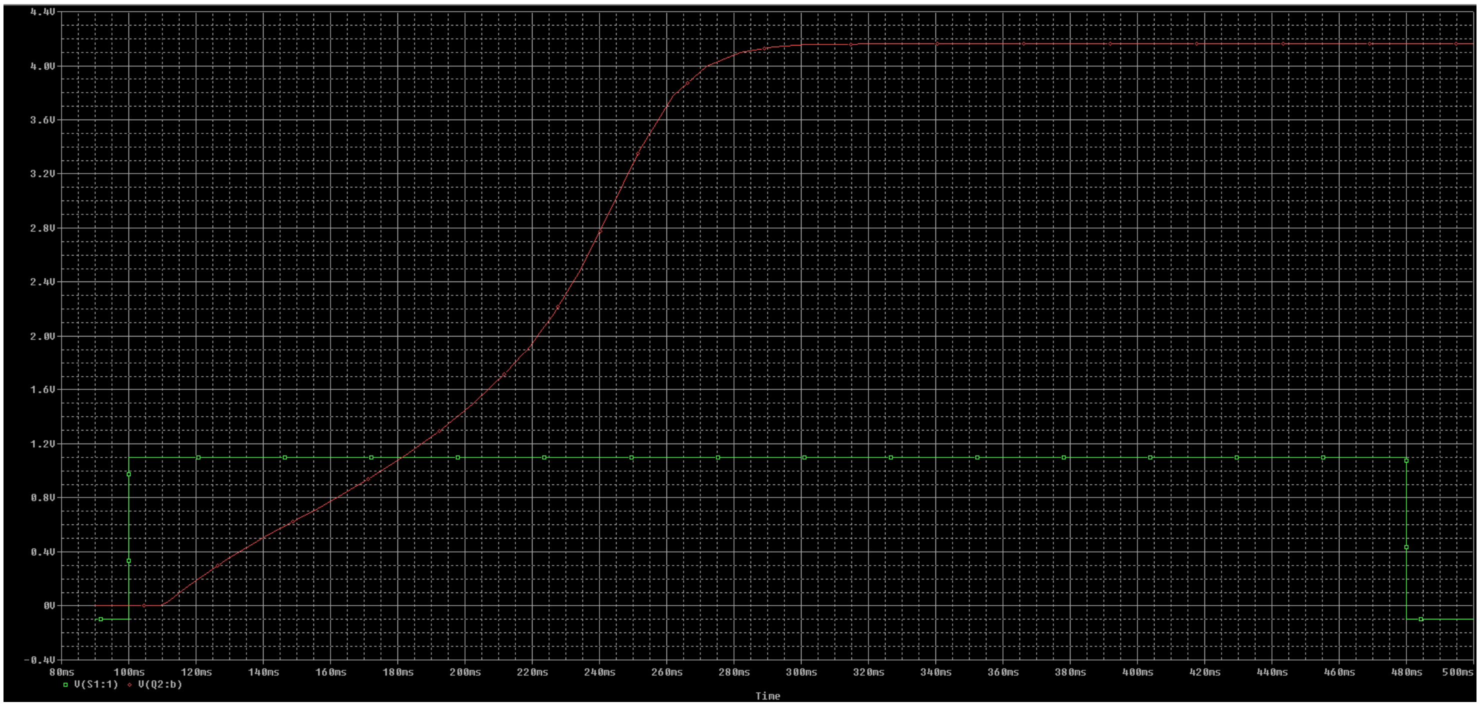
So how do I pull up on the COM node without adding another pushbutton or a second set of contacts on an existing pushbutton? This is where R14 comes in.



When V_{com} is zero and the Reset pushbutton, SW1, is pressed, Q5 clamps V9 at around 0.65V and takes most of the current flowing in R13. The current tries to flow out Q5's collector but R14 causes V_{c5} to rise and saturate Q5. So most of the current flows out the base. This base current dumps into the COM node and pulls it up. In a very short time V_{com} is above 0.65V. This causes Q2 to start up. As current flows in the collector of Q2, I_x comes to life. I_x dumps into the COM node and causes it to rise. The spiral of rising V_{com} and rising I_x continues until V_{com} is stable.

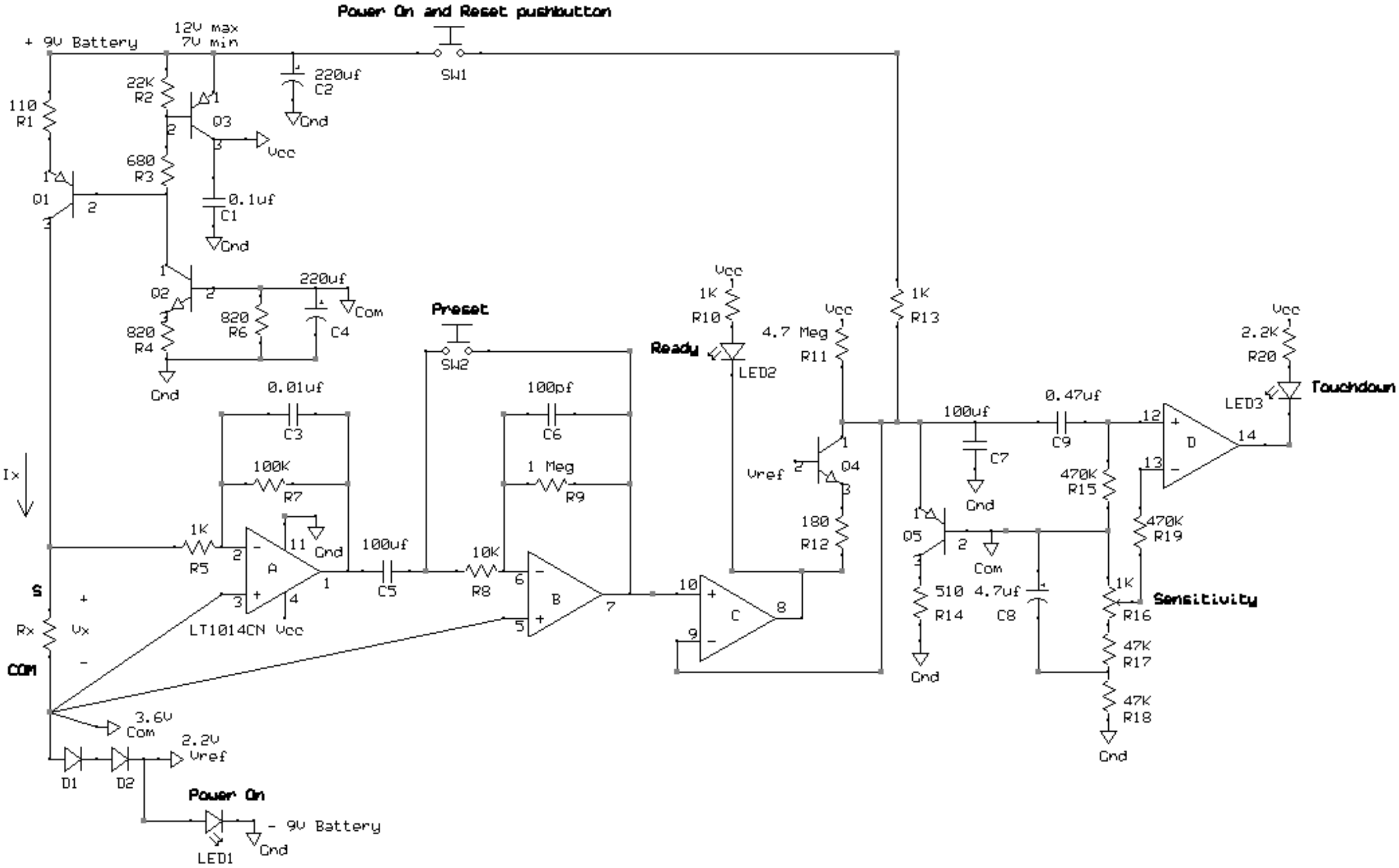
So in order to power up the circuit, I just needed to add R14 and rename the Reset pushbutton "Power On and Reset".

The detailed analysis of this power up sequence takes many pages and has many sub-state. I didn't think it was worth the time explaining it. However, it was fun to do and does involve a negative resistor and an exponential that heads towards infinity rather than an asymptote.



Shown above, in green, is the Power On and Reset button being pushed. The red trace is V_{com} rising from zero to its final value. The event takes less than 300 ms and when done, the Power On LED turns on to confirm success.

Time Constants

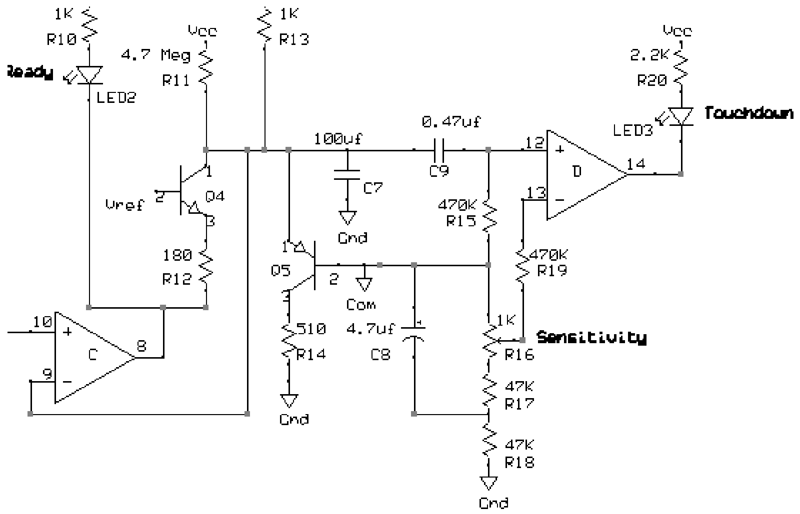


A detailed look at the time response of this circuit will be done with a simulator. But to design the circuit, it is essential to reason it out.

When touchdown occurs, V_x will instantly drop a tiny amount. This will cause an instant drop in the current through R5. That drop in current flows through C3 and R7 to drive V1 to a smaller value. The time constant is $R_7 \times C3 = 100K \times 0.01\,\mu f = 1\,ms$.

V1 feeds into the second gain stage. It first sees C5 with its time constant of $C5 \times R8 = 100\mu f \times 10K = 1\,second$. So it will be invisible to the transition. The next time constant involves C6 and R9. It yields $1\,Meg \times 100\,pf = 0.1\,ms$. Being ten times faster than the first gain stage, I can ignore it.

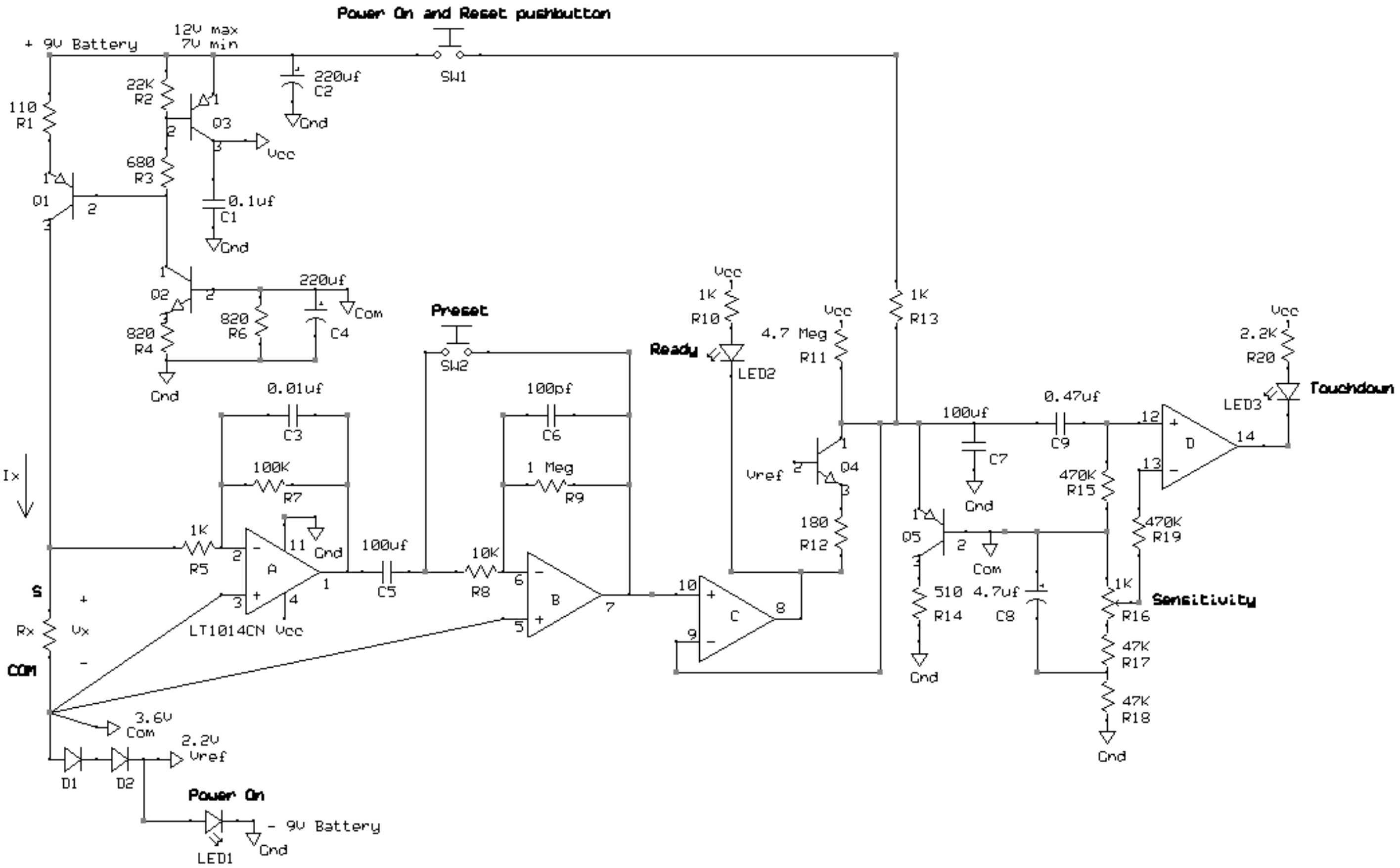
C7 really has no time constant since it moves as a function of V7 when V7 falls.



The last time constant involves C9 and R15. It is $0.47\mu f \times 470K = 220$ ms. Five of these time constants equals 1.1 seconds.

When C7 is yanked up during Reset, V12 rises too. Assume V9 is held at 0.65V above COM for more than 1.1 seconds. Then C9 will charge up to 0.65V. This will put the left end of C9 at 0.65V and the right end at zero with respect to COM.

Then when Q4 kicks in and pulls V9 down, V12 will fall and turn on the Touchdown LED. After V9 is stable, it will take as much as 1.1 seconds for V12 to return to zero and turn off the LED.



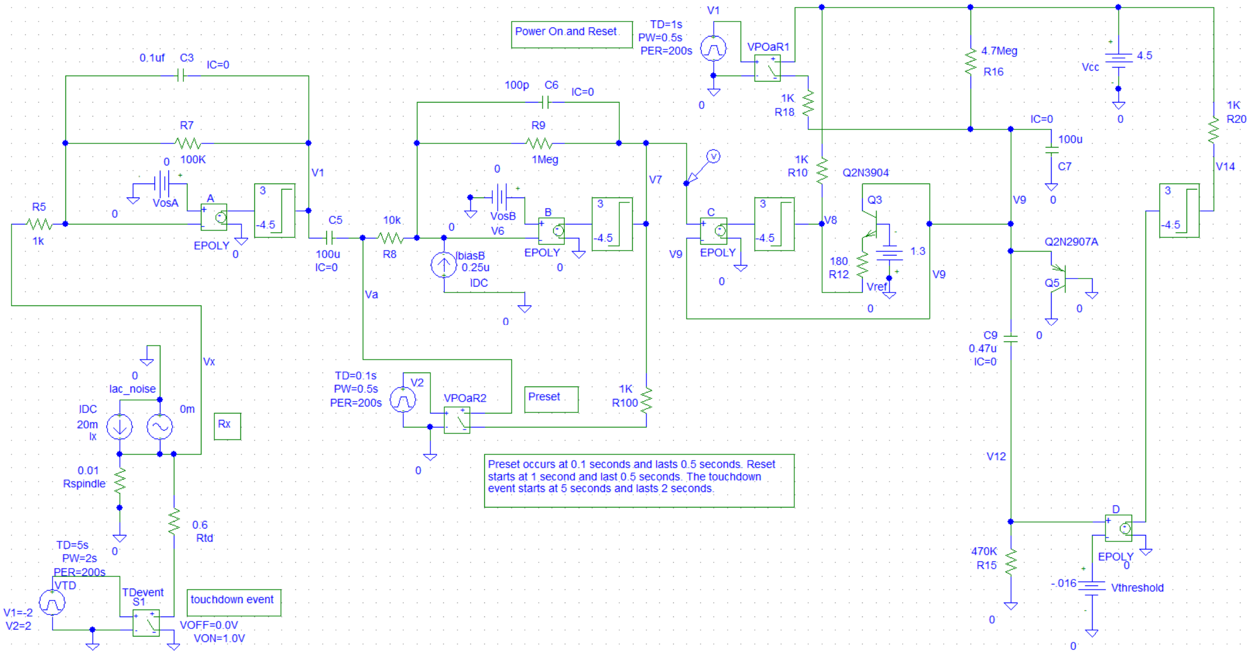
When a touchdown event occurs, V1 will drop exponentially with a time constant of around 1 ms due to the first gain stage. This transition will sail through the second gain stage because C5 is so slow and C6 is so fast. The transition will similarly pass through C9 because it is so slow. V12 will initially drop by the same amount as seen on V7. After the initial transition, V9 will very slowly rise with a time constant of $R11 \times C7 = 4.7Meg \times 100\,\mu f = 470\,seconds$. C9 with R15 will see this slow rise as not moving and V12 will rise back to zero with its time constant of 220 ms.

Note that I chose the time constant for C9 and R15 to be small so that I would not have to build a reset circuit for it. The trade off is that the Touchdown LED does not stay on for a "long" time. I do believe it is long enough to easily see. How long it stays on depends on the magnitude of the drop at V_x. As long as V12 is less than V13, the LED stays on.

The bottom line is that I have chosen time constants so each stage stays out of the way of the rest. That sure simplifies hand analysis.

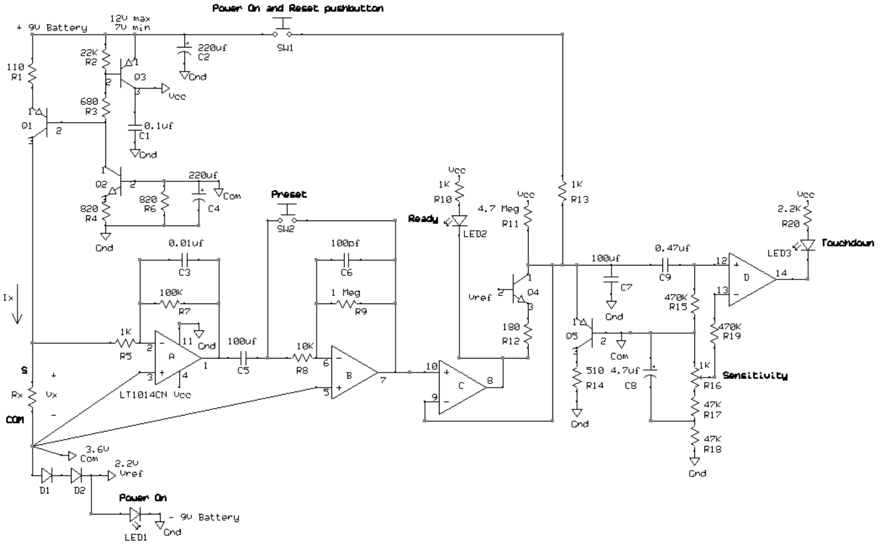
Simulating the Circuit

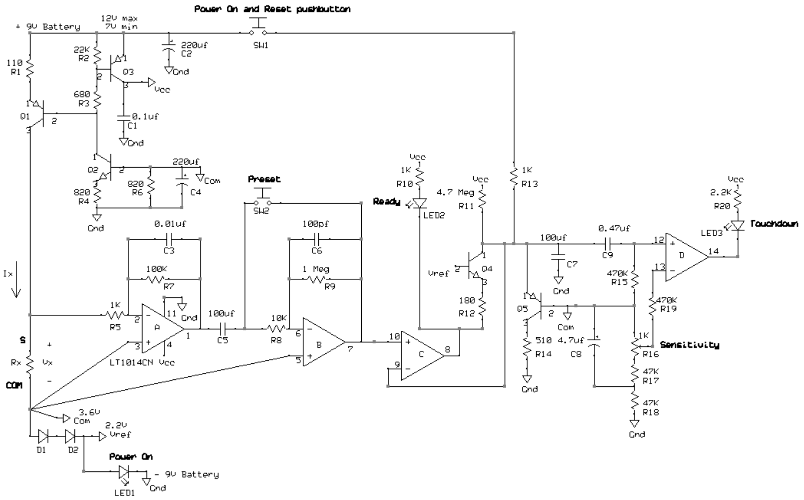
I found it helpful to use a simulator, PSpice, to see what should be going on. It lets me control every aspect of the circuit. Rather than use built in op amps, I formed them from voltage controlled voltage sources to give me gain, Limit blocks to give me my saturation voltages, DC supplies to give me input offset voltage, and DC current sources to give me input bias. In the model shown here, I removed the input offset voltage and input bias current sources that do not effect circuit operation. R11 has not been updated for this section.



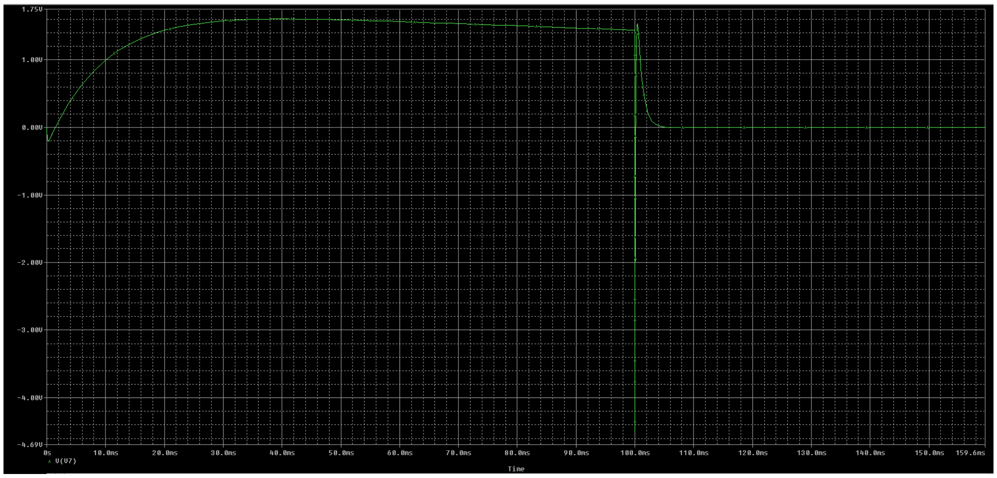
Preset occurs at 0.1 seconds and lasts 0.5 seconds. Reset starts at 1 second and last 0.5 seconds. The touchdown event starts at 5 seconds and lasts 2 seconds.

I did not simulate the difference between the COM node and ground. This study was done on the bench with the real circuit.

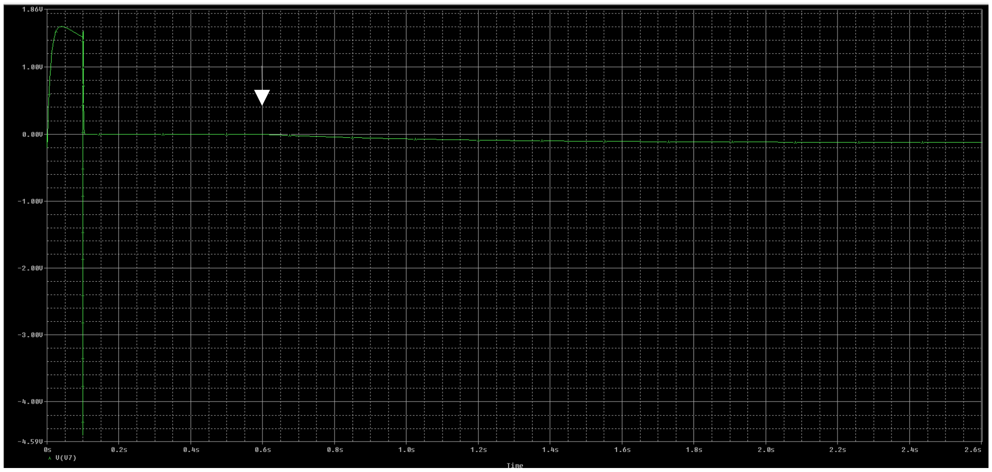




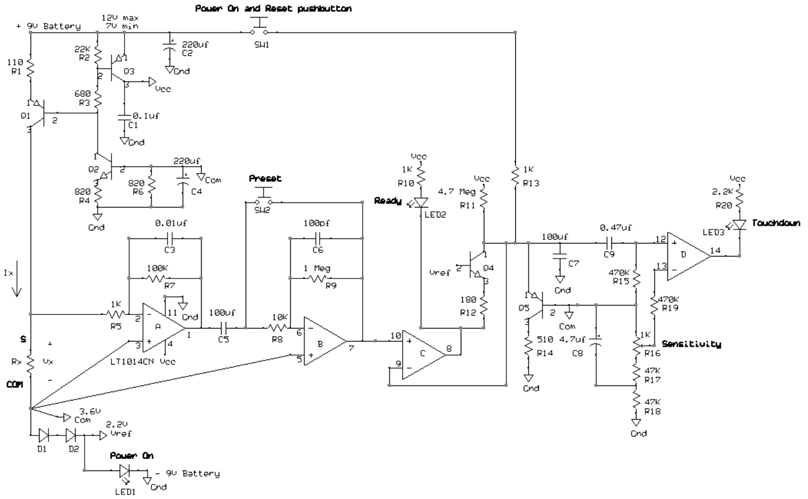
This is a trace of V7 during Preset. Power is applied at time equal 0 and the Preset button is pushed at 100 ms. It is released at 500 ms.



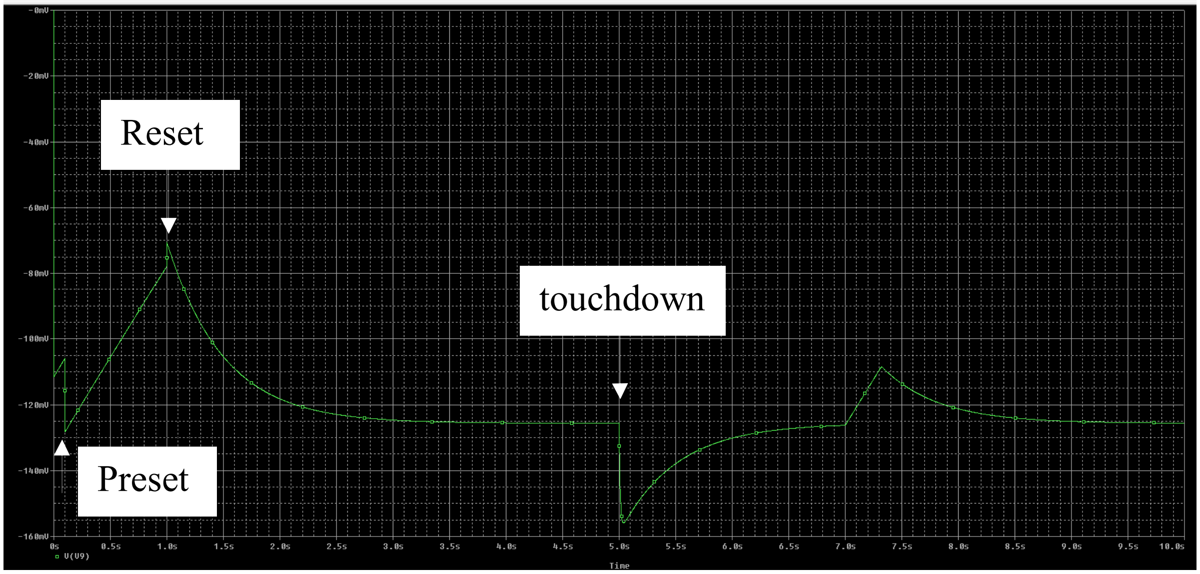
You can see V7 quickly rise up to almost 1.75V and sit there while C5 starts to charge. Then at 100 ms the Preset starts and V7 is moved to near 0 in less than 10 ms.



This is more of the same trace. At 600 ms the Preset button is released and V7 falls to its final value with a time constant of 1 second. By 5 time constants it is at about -120 mV. This voltage is due to the input bias current of the second gain stage.



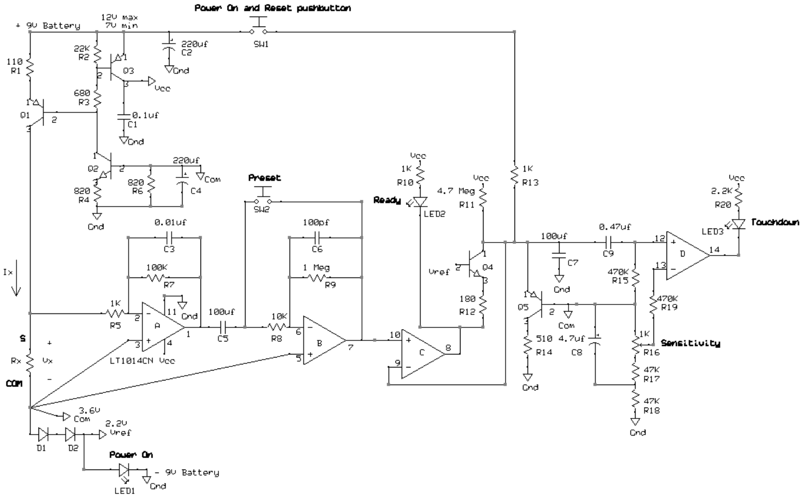
This trace shows V9. Recall that Preset, which charges C5, started at 100 ms. The button was released at 600 ms although V7 was close to its final value by 105 ms. You can see V9 drop at about 100 ms in response to the drop in V7. V9 then starts a slow rise until 1.0 seconds when the Reset button is pushed. This abruptly pulls V9 up. It does not go all the way to 0.65V because Q4 is pulling it down. V9 follows the fall in V7 until the touchdown event at 5.0 seconds. It lasts until 7.0 seconds.


Reset

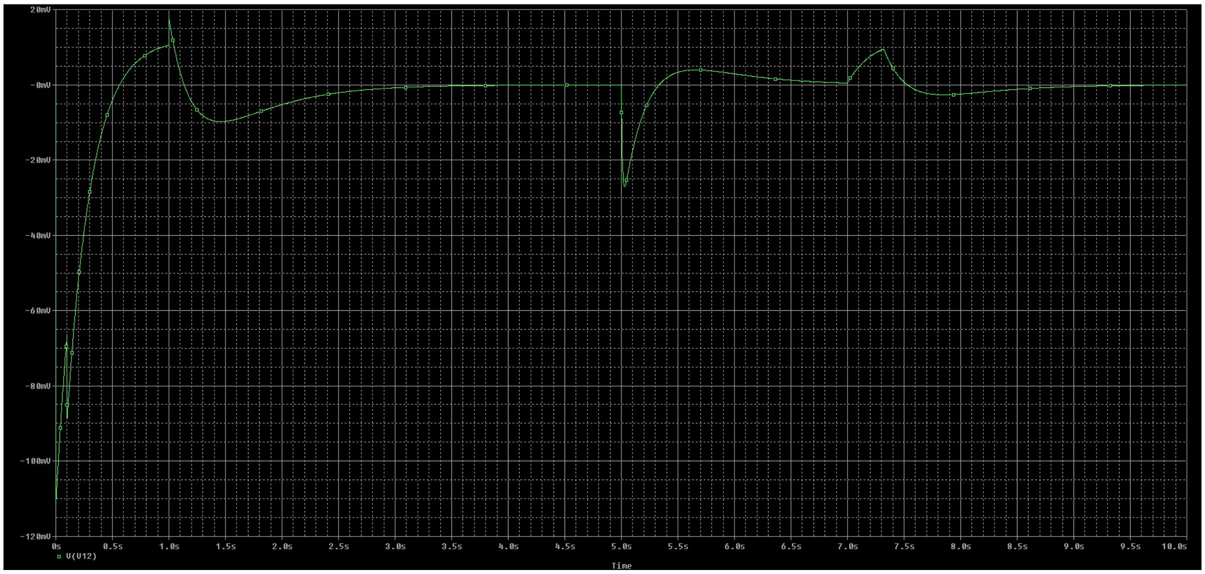
touchdown

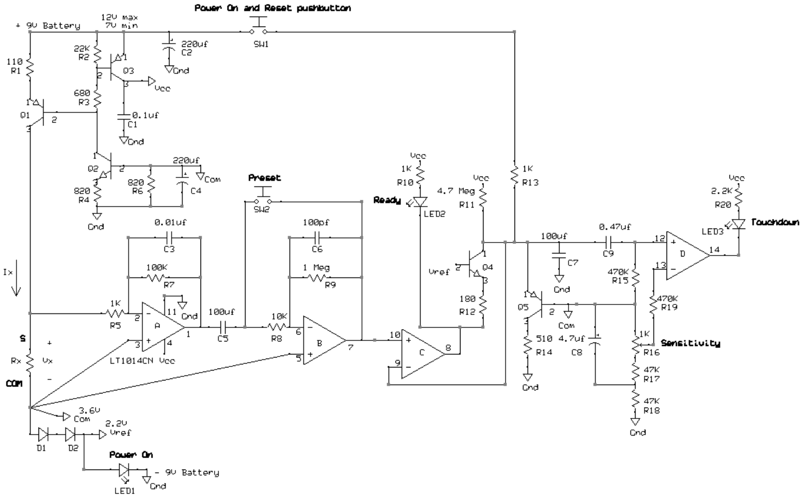
Preset

V9 drops by about 30 mV which is consistent with the discussion on page 13 related to minimum spindle resistance and maximum touchdown resistance. When the touchdown event is over, V9 kicks up a little as C5 adjusts and then settles back to its original value within a few seconds.



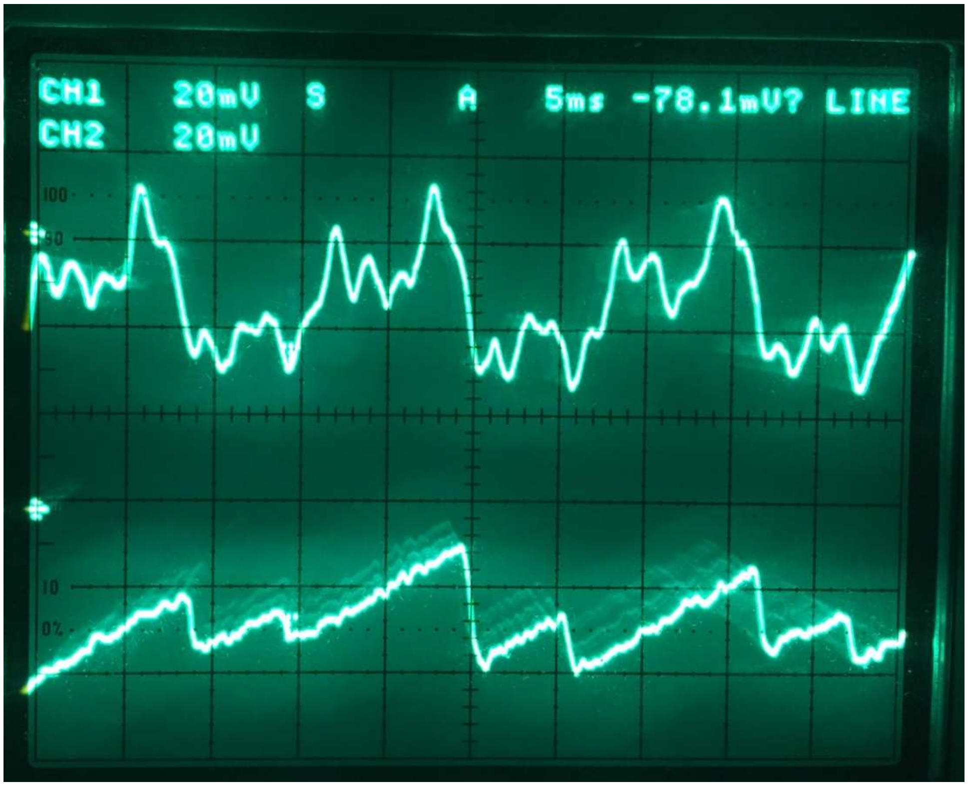
Now we are looking at V12. The Preset of C5 at around 0.1 seconds caused V9 to drop so you see this drop on V12. Then we have the Reset at 1.0 seconds which causes a rise in V9. V9 has settled out by 3.5 seconds. Since the time constant of C9 and R15 is much smaller than C7 and any of its time constants, V12 causes little added delay before becoming stable. It is ready for the touchdown event at 5.0 seconds. You can see that the negative spike is almost 30 mV. We did blunt the spike slightly by using the smaller C9/R15 time constant.





Here is the final trace, V14. When V14 is low, the Touchdown LED is on. We can see that the Touchdown LED initially is on when Preset is taking place. The next time it goes on is after Touchdown. The Touchdown LED is on for only 100 ms but this is plenty of time to see it.

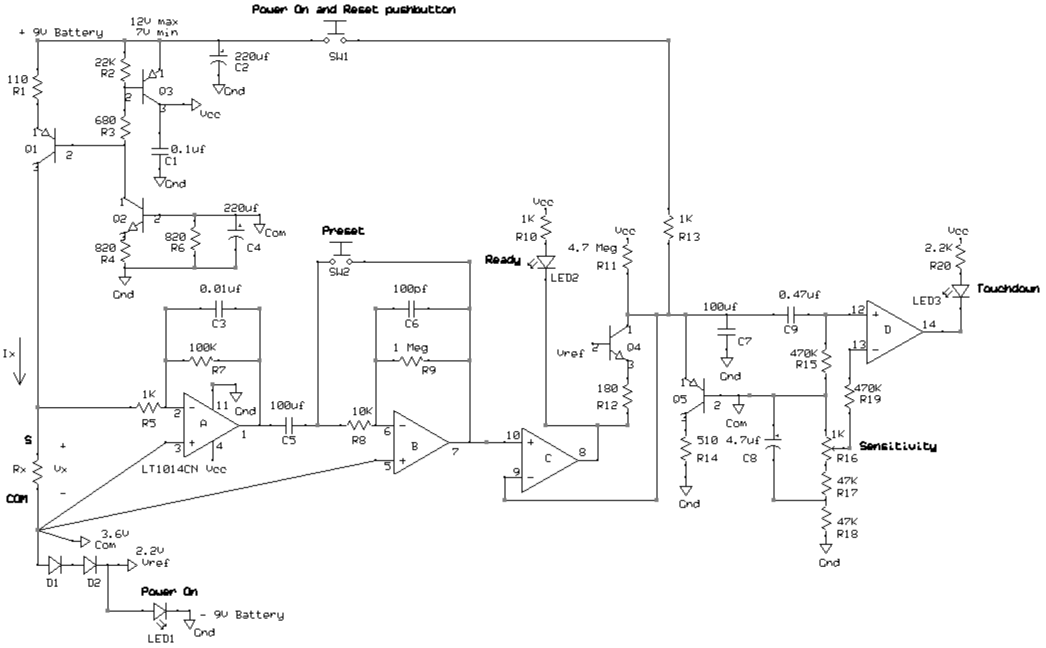
Scope Pictures

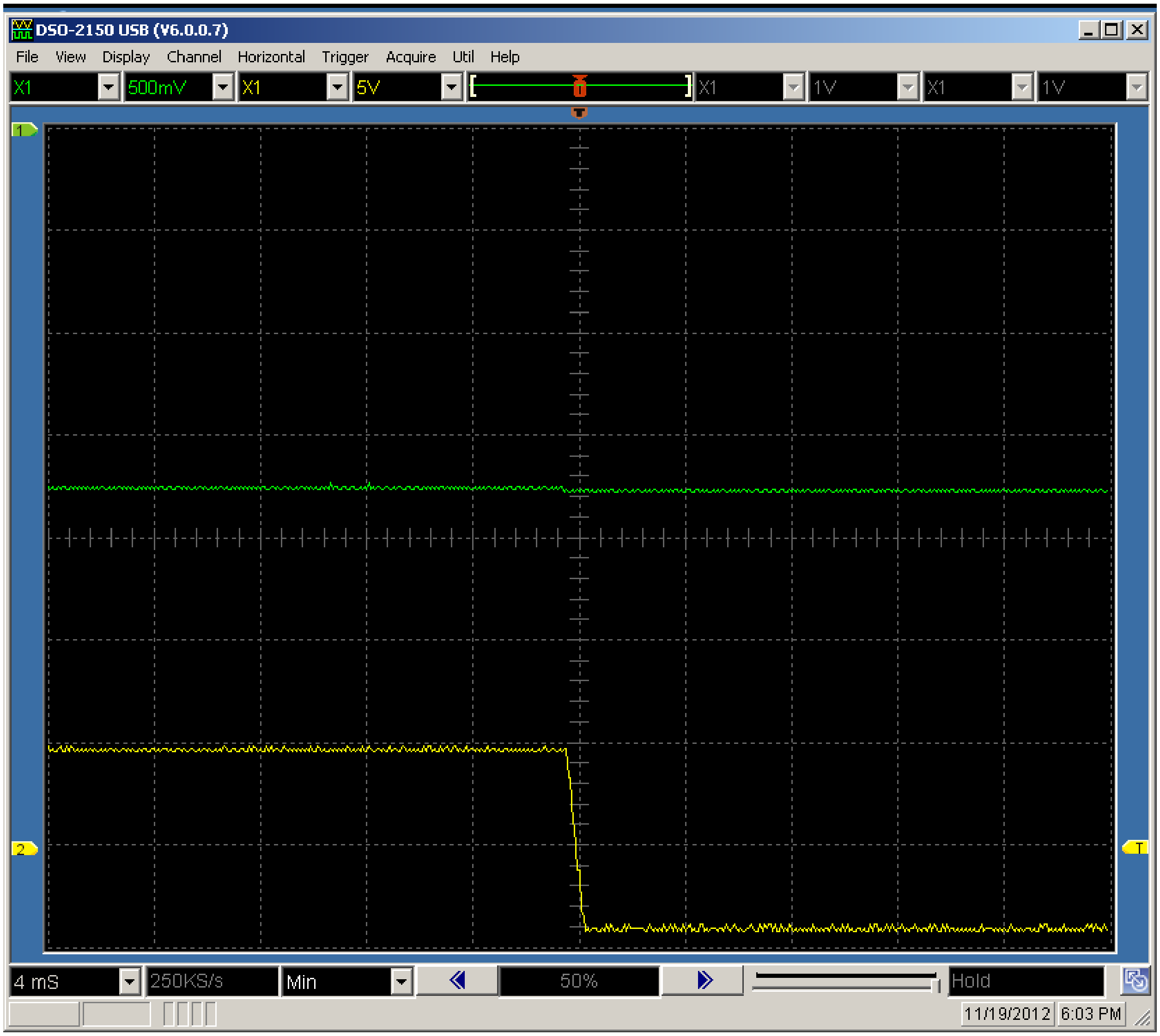


Test conditions:
R_x = 0.01 ohms
No induced AC noise.

The top trace is V_7 at 20 mV/division. The bottom trace is V_9 at 20 mV/division. Time base is 5 ms/division. The "+" on the left side indicates zero volts for each trace.

You might be able to see how V_9 is following the negative peak of V_7. When the two traces share the same zero volt line, it is more obvious.

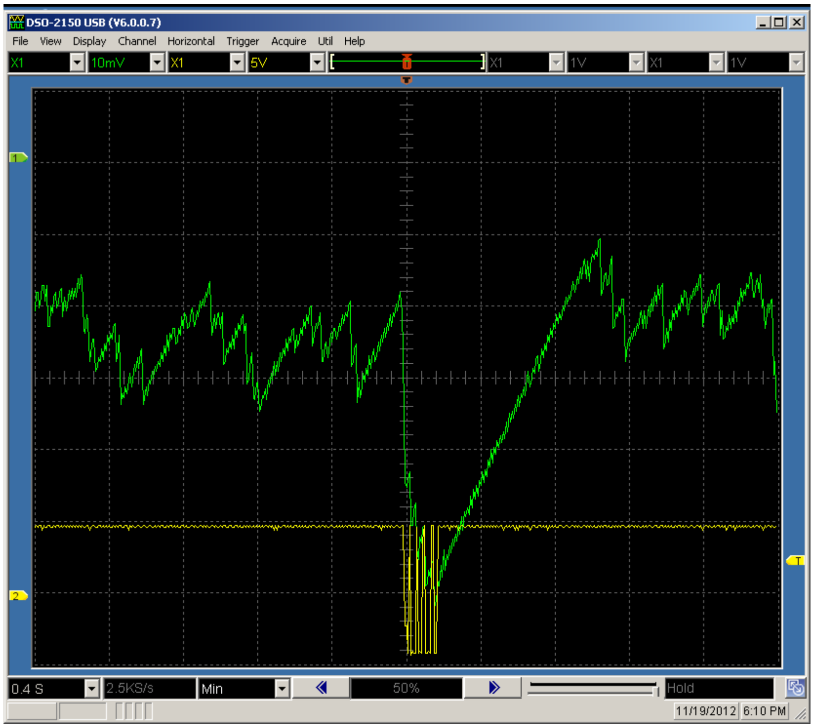




Test conditions:
R_x = 0.01 ohms $R_{touchdown}$ = 0.6 ohms
Induced AC is above the maximum of 1.4V: 1.7V peak at V_7

The green trace is V_7 at 500 mV/division. The yellow trace is V_{14} at 5V/division. Time base is 4 ms/division.

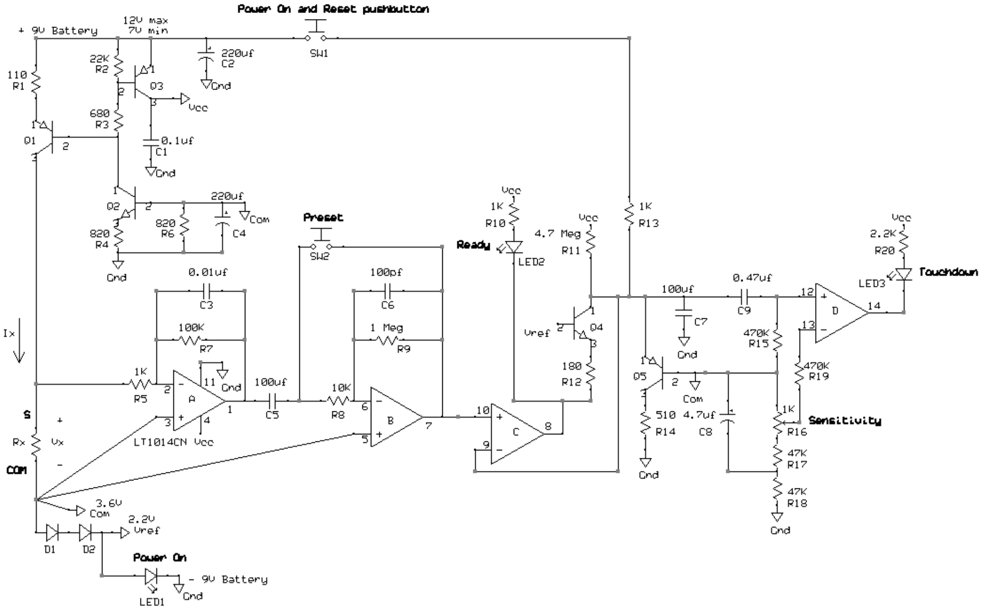
This is a touchdown event at above the maximum induced AC. You can see a small drop in V_7 just before V_{14} goes low. When V_{14} is low, the Touchdown LED lights.

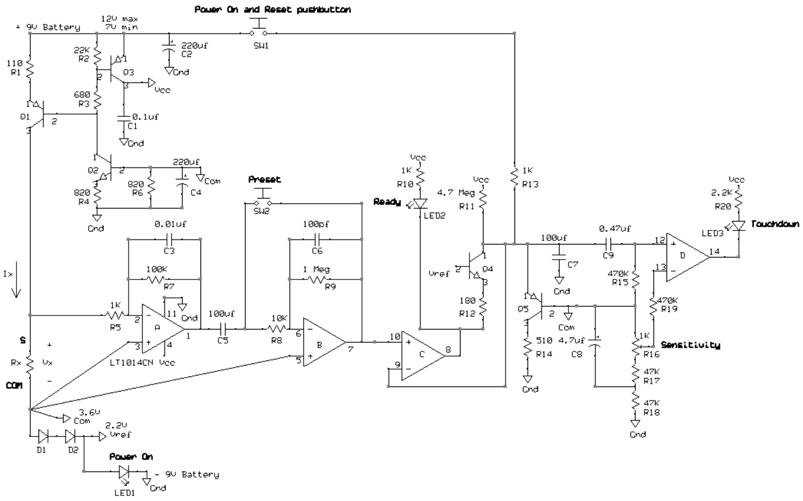


Test conditions (same as last picture except no AC noise):
$R_x = 0.01$ ohms

The green trace is V_9 at 500 mV/division. The yellow trace is V_{14} at 5V/division. The time base shows 0.4 sec/division but there is a bug in the scope's software and this is really 2 seconds/division.

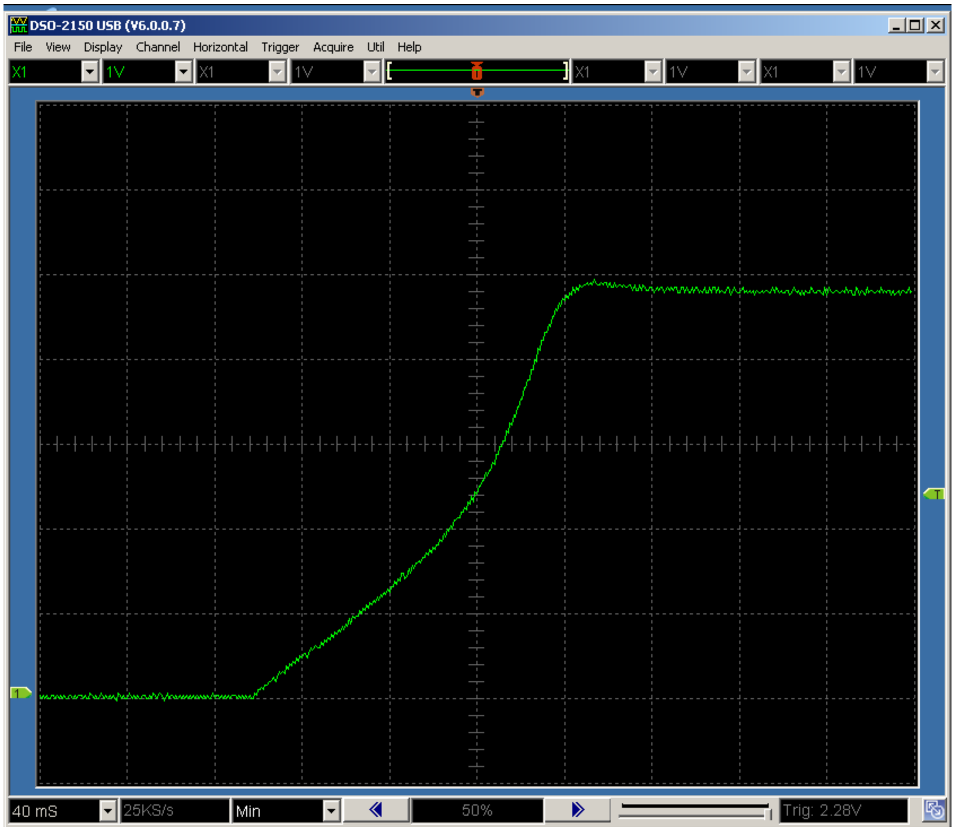
You are looking at a touchdown event with around 15 mV of random noise induced into V_x. The drop in V_9 is around 32 mV as predicted on page 13. V_9 drops low in response to a negative going noise spike and then drifts positive after the spike goes away. The rate of rise is around 2.8 $divisions \times \frac{10\ mV}{division}$ = 28 mV in 1.3 divisions $\times \frac{2\text{ seconds}}{\text{division}}$ = 2.6 seconds which comes out to $\frac{28\ mV}{2600\ ms} = 11\ mV/second$. Back on page 53 we calculated a rise of 11.5 uV/ms or 11.5 mV/second. The touchdown LED flashes for about 1 second which is consistent with what I see on the bench.



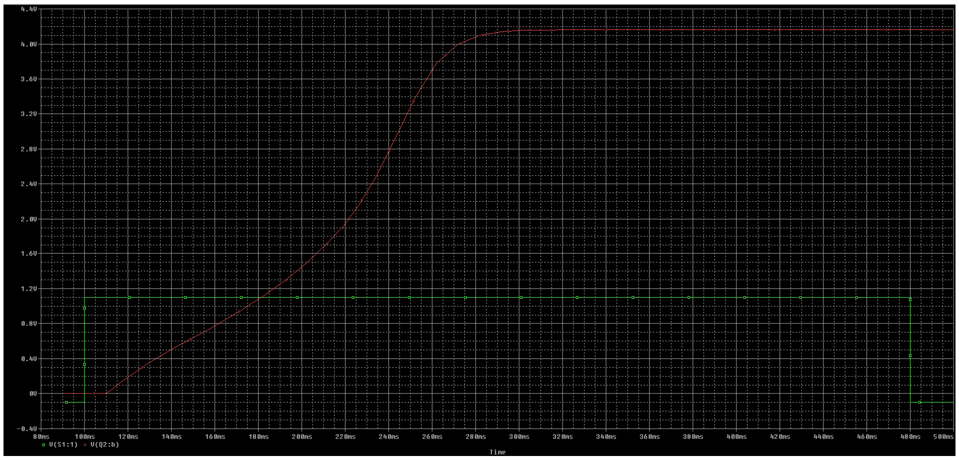


The rest of the scope pictures were taken on the prototype.

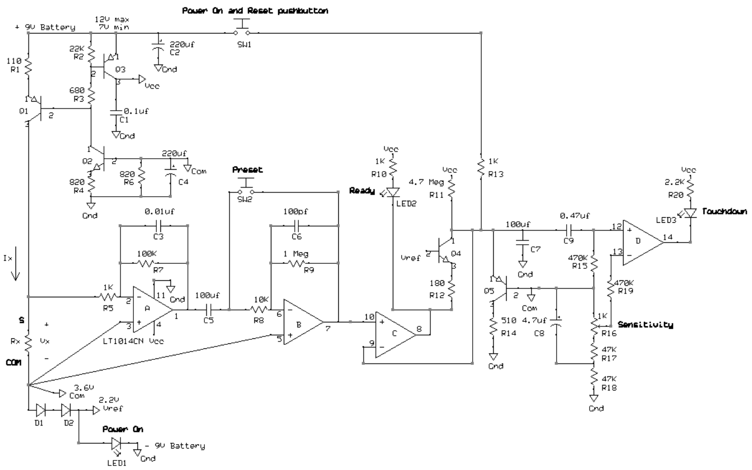
This scope picture is of V_{com} with respect to ground. Below it is the simulation. The final value of V_{com} is different due to the LED I am using in my prototype but the wave shapes are similar.



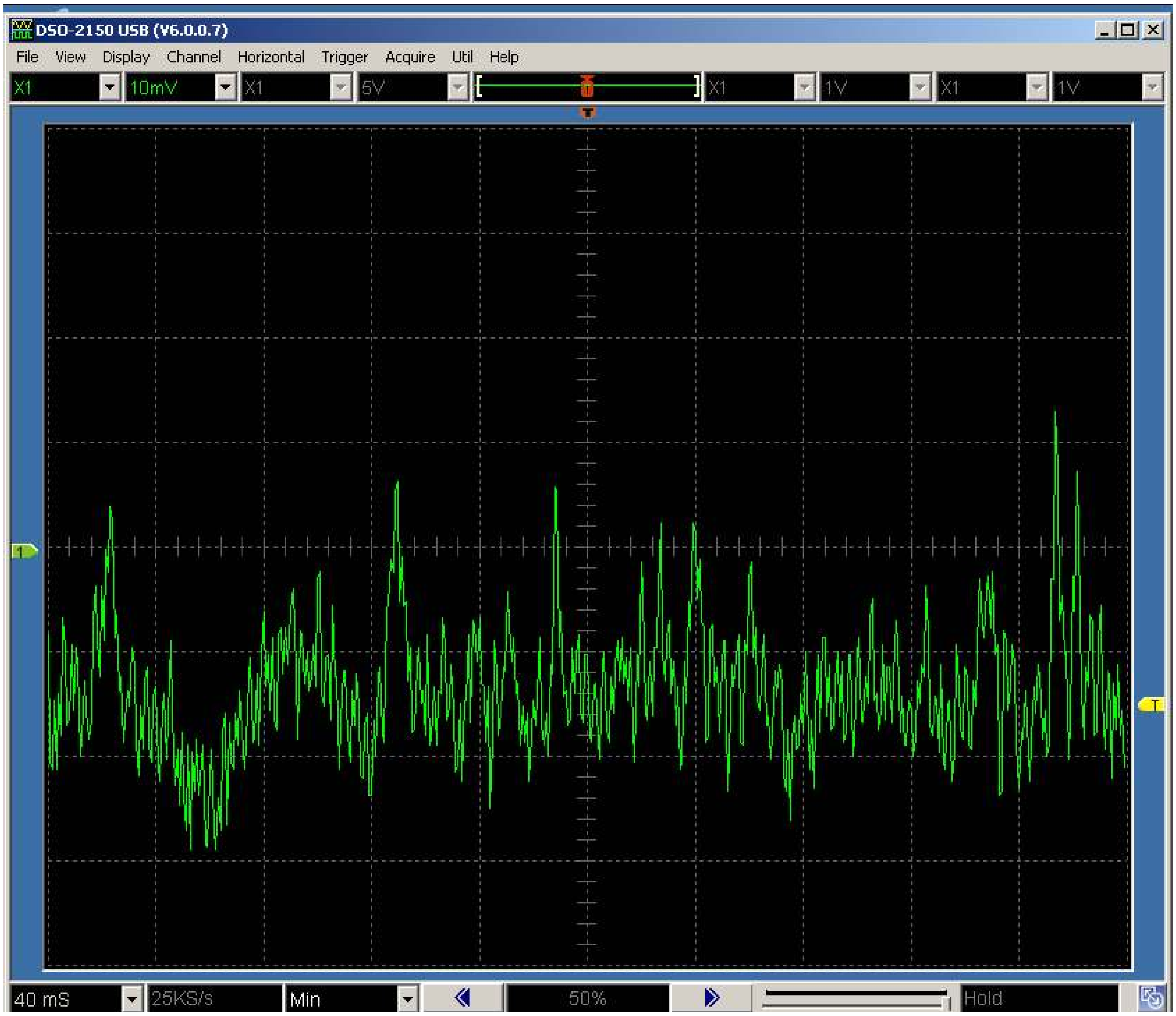
V_{com} comes up in about 160 ms in this scope picture.

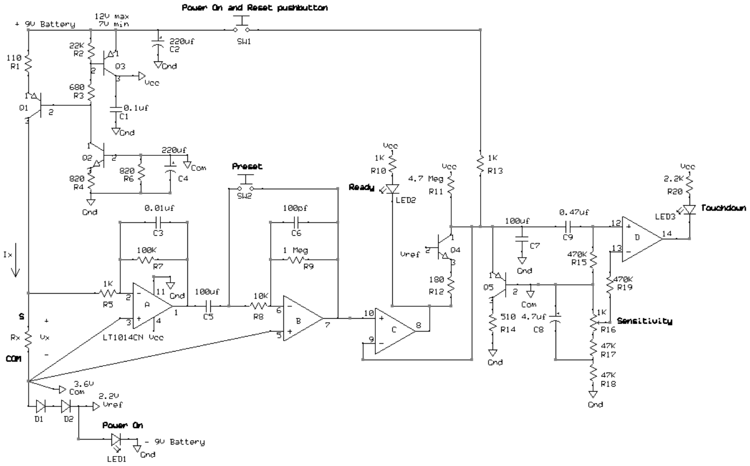


V_{com} comes up in about 170 ms in the simulation.



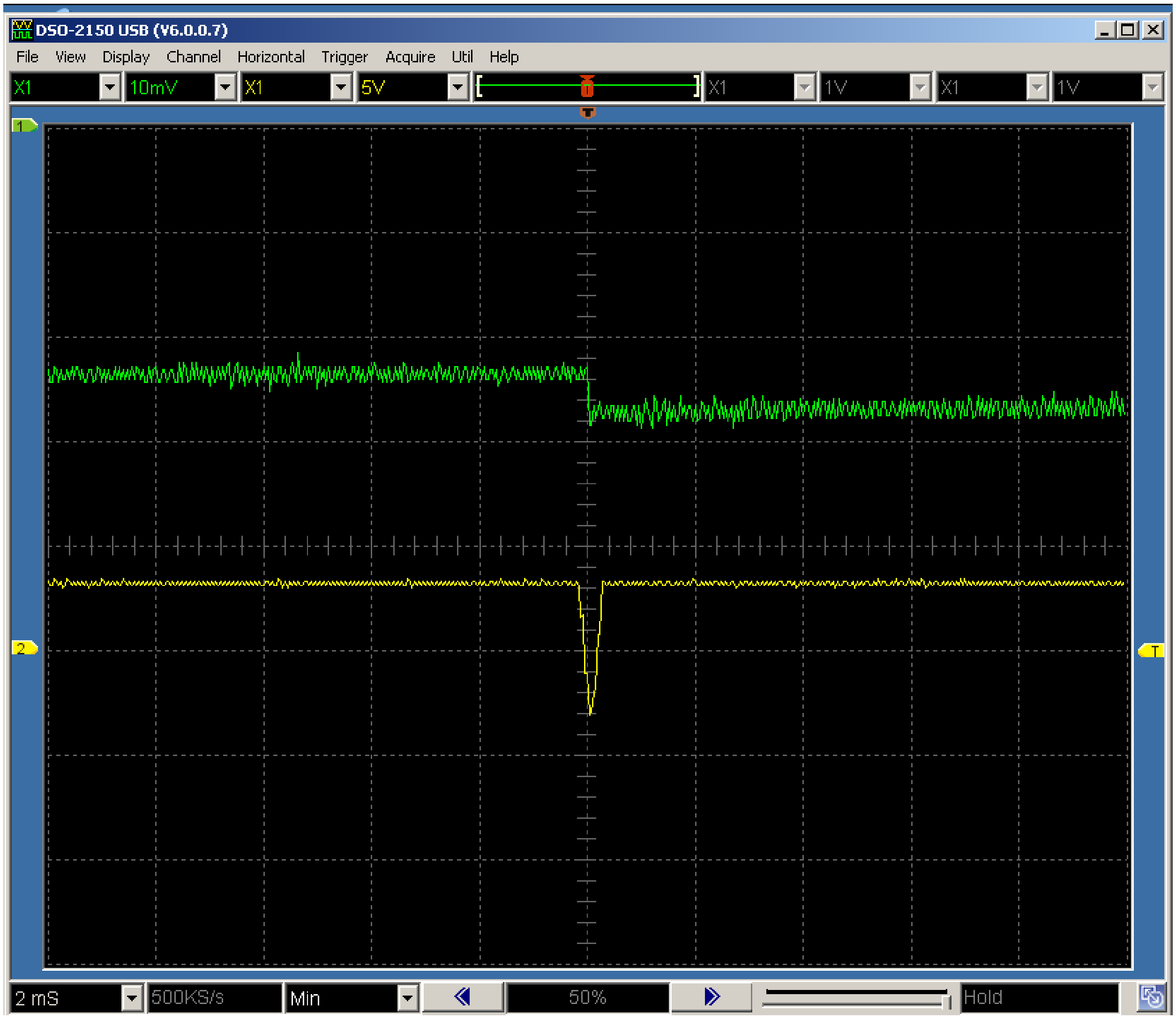
Here we have V7 with respect to COM. The vertical is 10 mV per division. The horizontal is 40 ms per division. The noise you see is mostly picked up from V_x and not internal to the circuit. There are spikes in this signal that are too fast to be seen with a time base of 40 ms per division.

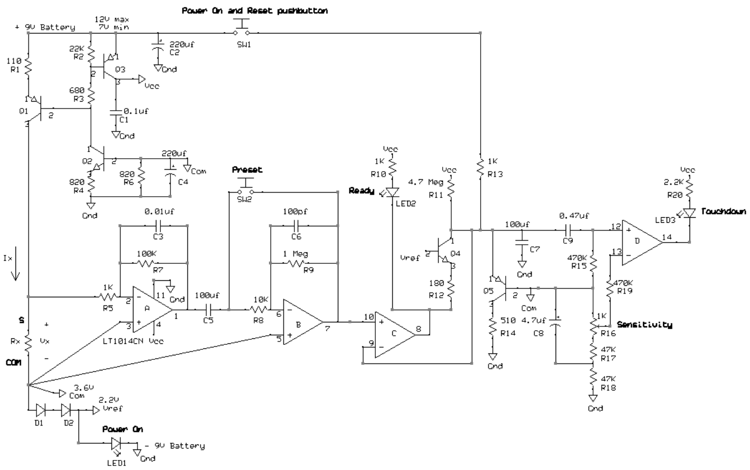




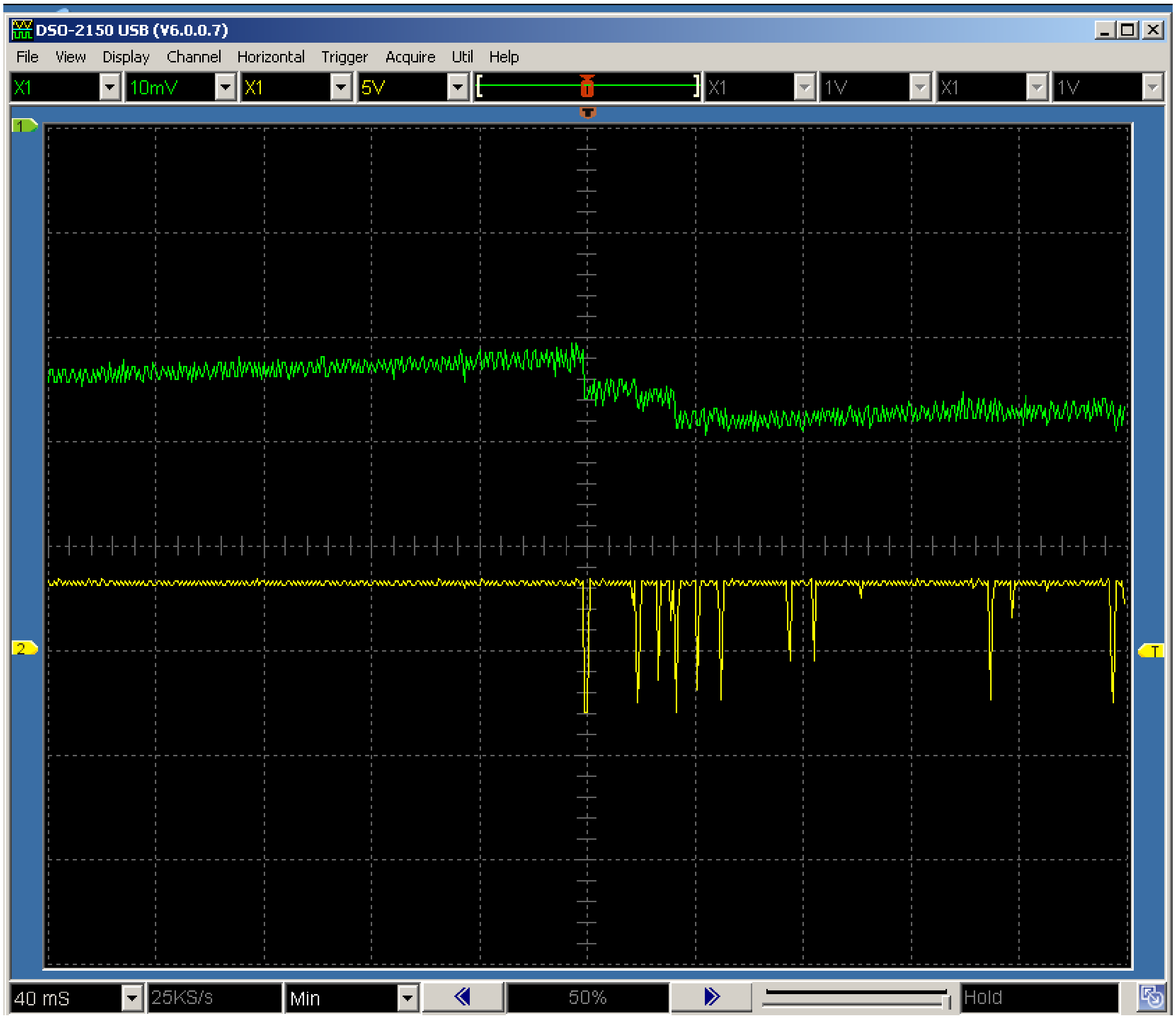
This picture shows V9, in green, and V8 in yellow. Both are with respect to COM. When V8 dips low, you see a sudden drop in V9. The rest of the time V9 is barely rising. The noise you see on V9 is most likely from the scope's power supply. It is about 1 mV peak to peak. This channel is set to 10 mV per division. The time base is 2 ms per division.

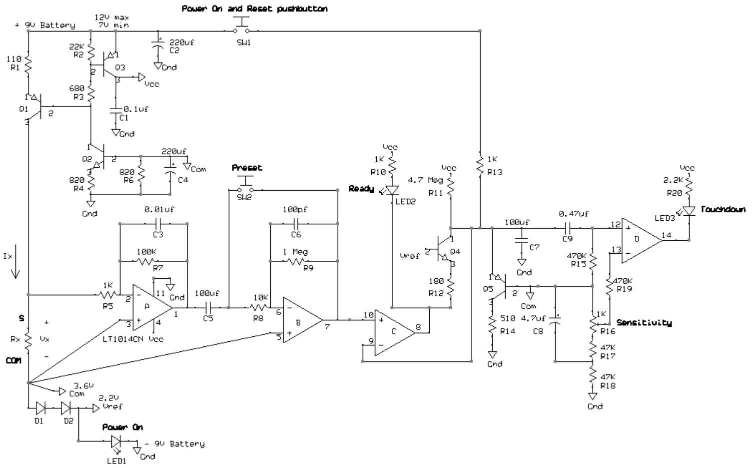
The yellow trace is a lot quieter because the vertical is 5V per division.



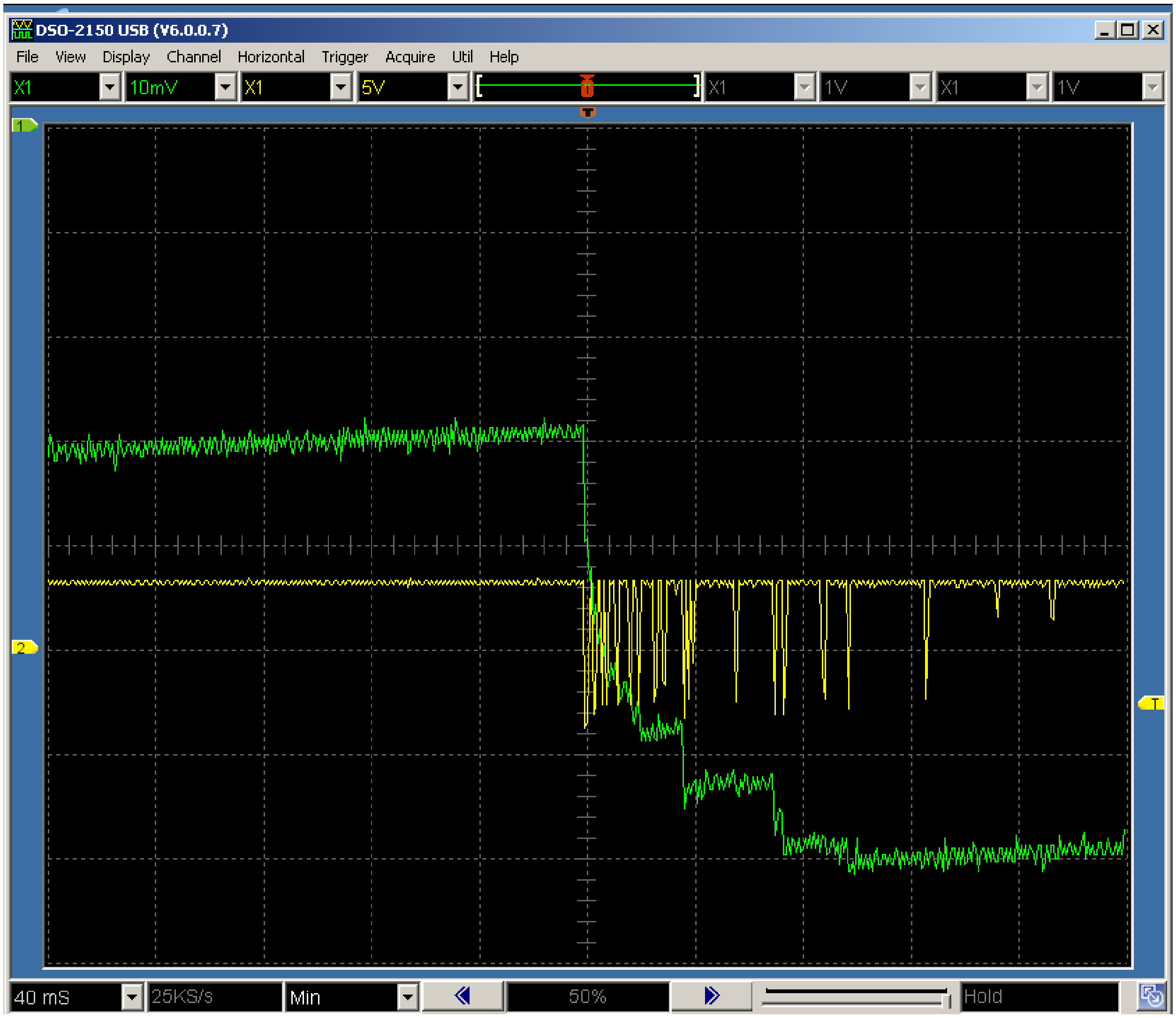


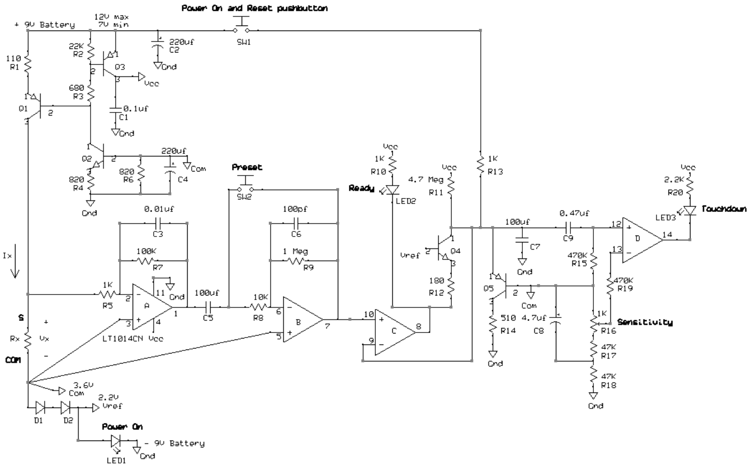
I am again looking at V9 and V8 but this time you can see multiple spike on V8 causing V9 to step down to lower values. The wider the spike, the more V9 drops.



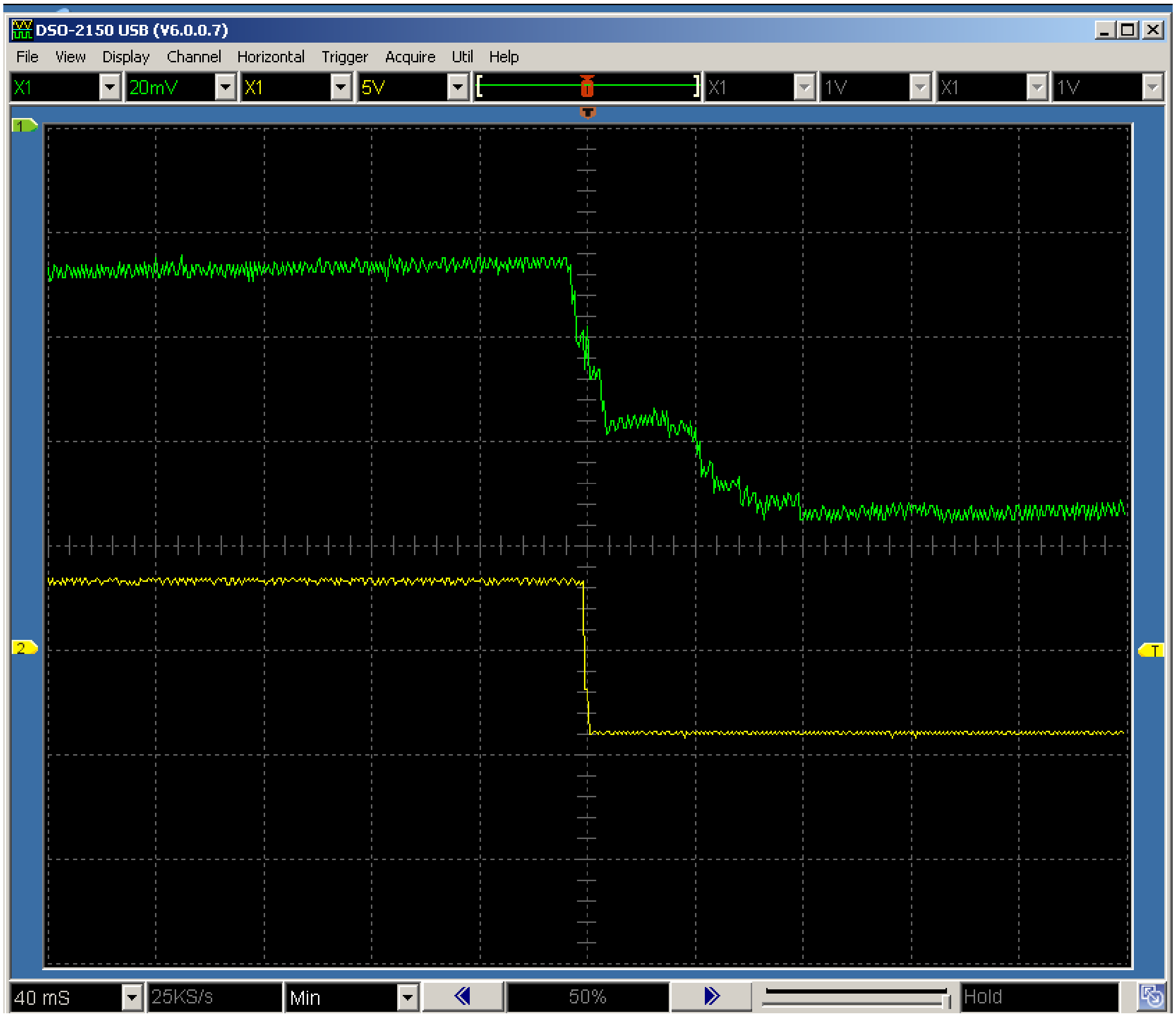


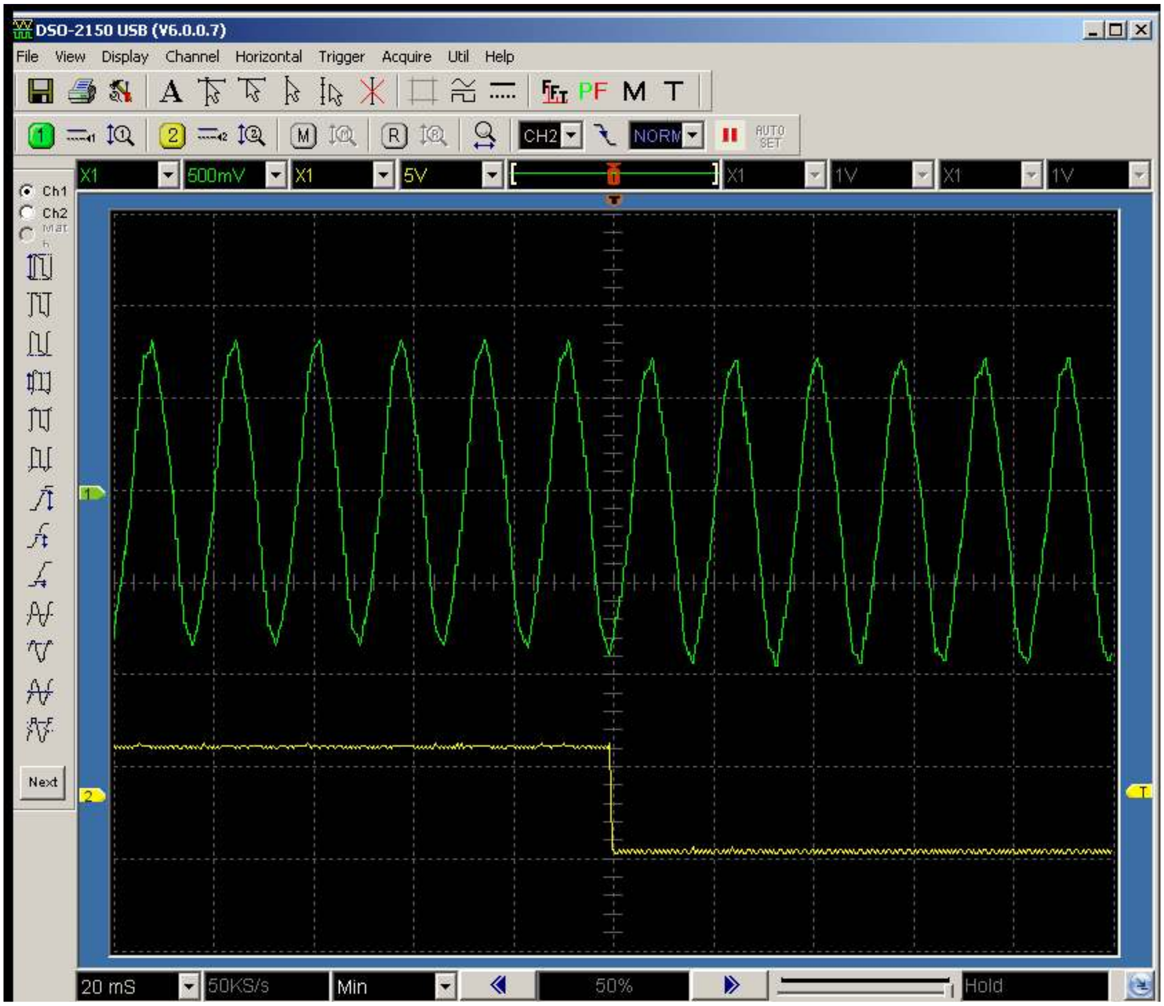
This time I am looking at V9 and V8 during a touchdown event. You can see V8 repeatedly pulling down on V9 in order to match the drop in V7. That first flurry of spikes is related to the actual touchdown event. The rest are probably reacting to noise that has gotten into the amplifier stages from V14 after touchdown.

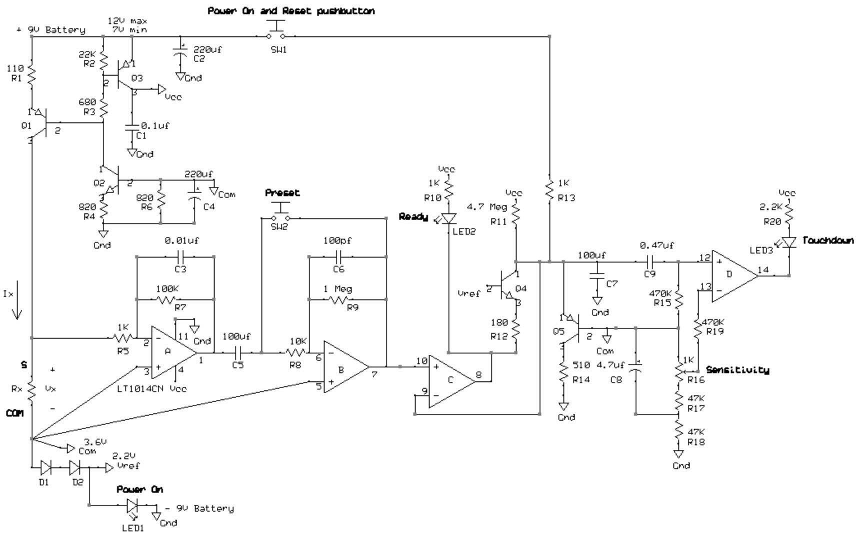




This scope picture is of V9 during a touchdown plus V14. When V14 goes low, the Touchdown LED goes on. V14 went low after V9 dropped about 30 mV. The subsequent drop is probably a shift in COM with respect to ground. It is harmless.

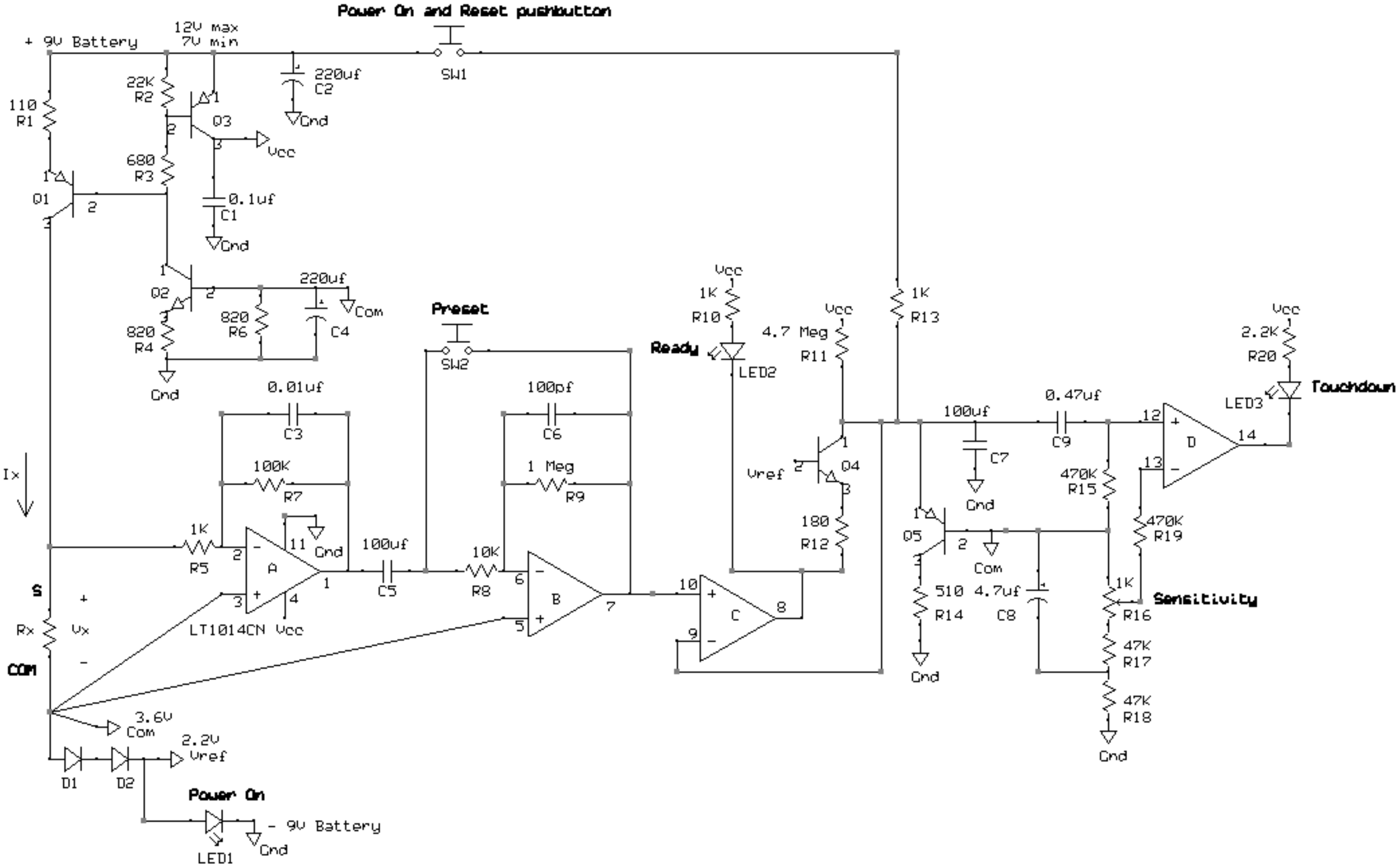






This scope picture shows V_7 in green and V_{14} in yellow. A 60 Hz noise current has been induced in the probes that generated this 0.9V peak signal. It translates to a 90 µV signal at V_x so is 30 times larger than the touchdown event.

Possible Future Improvements



Mark Cason has laid out a double sided and plans to sell both kits and finished product.

I have very limited experience with this circuit in working machine shops. It works fine in my shop and in two other shops.

The third shop is extremely noisy for two reasons. Plastic conduit was been used to carry 220V so grounding is not as quiet plus radiation from the wires is not enclosed as it would be with steel conduit. Secondly, a large CNC mill is present. The voltage spikes from the table drive motors is large enough to cause false touchdown indications.

One solution that does work is to raise the threshold voltage, V13, feeding the touchdown LED comparator (op amp D). This reduces sensitivity to the touchdown event and to the noise spike. In this case I am able to see the touchdown event and ignore most of the voltage spikes. More testing is needed.

Acknowledgements

I wish to thank Mark Cason for his hard work layout out the circuit ***many*** times. Just when I thought I was done, something else caused a modification to be made. His layout skills are truly amazing and it has been a pleasure working with him.

Thanks to Gregg C. for reviewing the circuit and suggesting key changes.

And a very special thanks to those people on various BBS that provided inspiration.

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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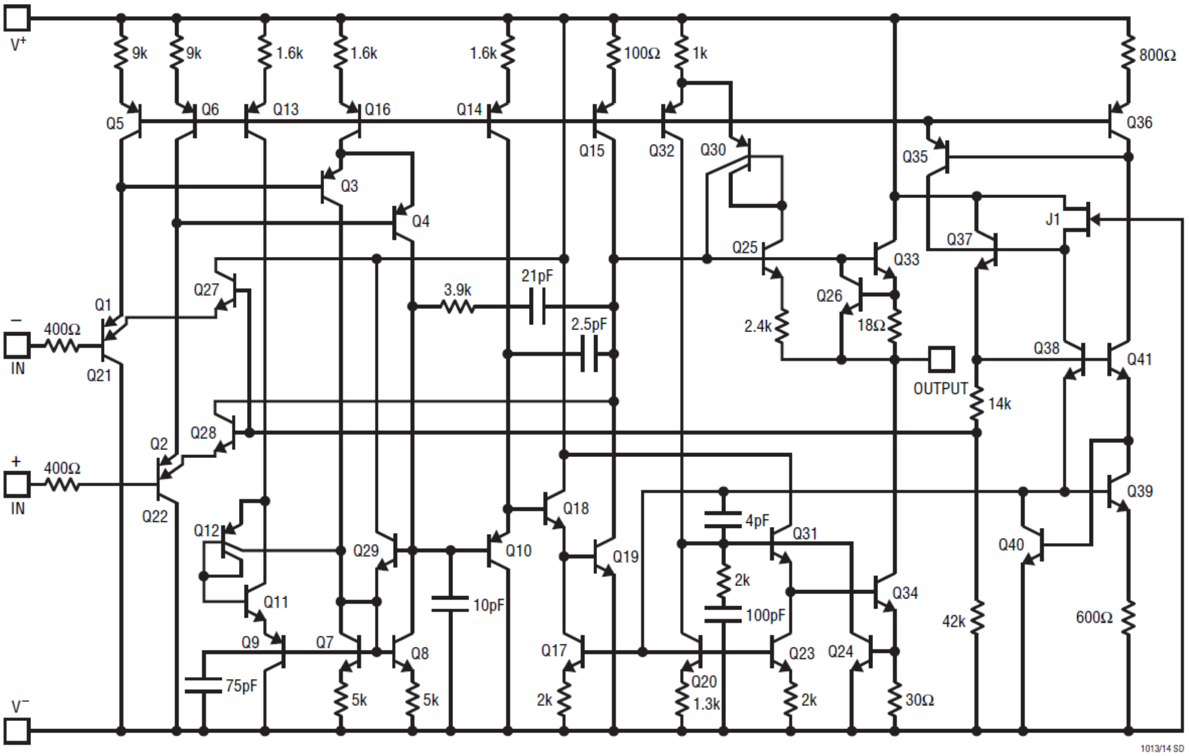
Appendix: The LT1014 Op Amp

SCHEMATIC DIAGRAM

1/2 LT1013, 1/4 LT1014



Here is a partial explanation of how the LT1014 op amp works.

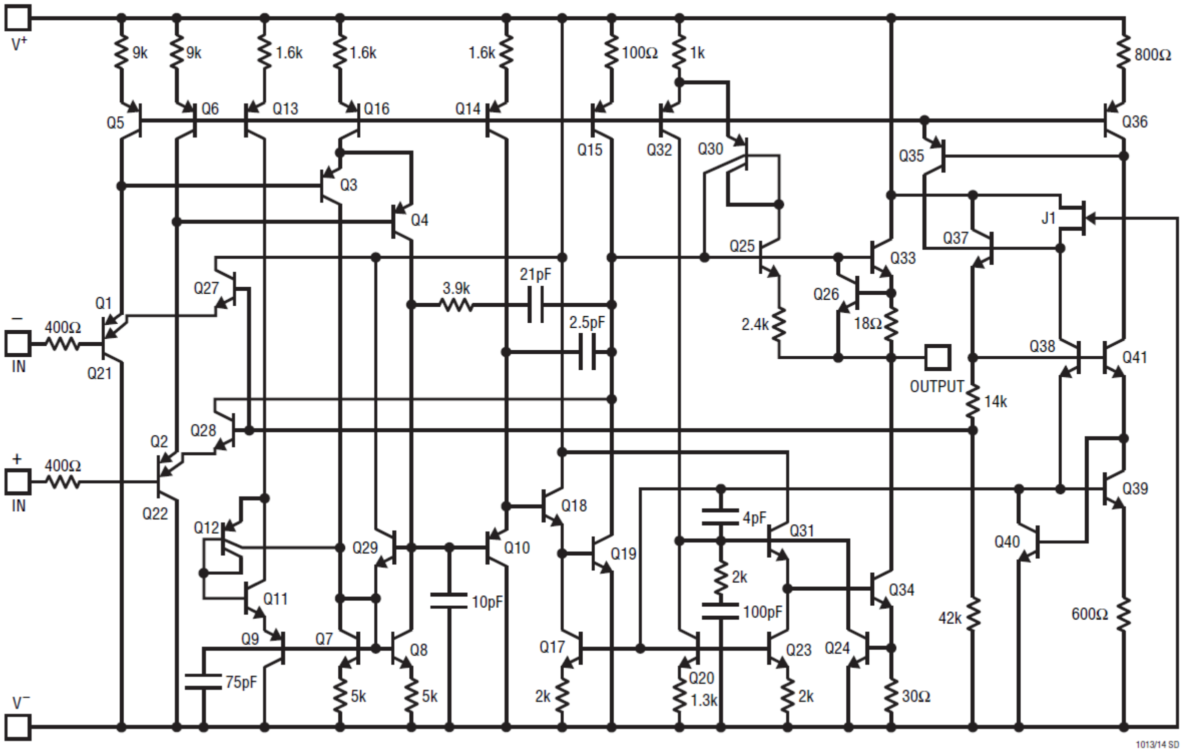
V$^+$ connects to Vcc which is around 9V. V$^-$ connects to ground because it is the most negative node in the circuit. Connected to the V$^-$ rail is Q24 which is on the lower right side of the schematic.

Q24 is normally off. It only turns on when the output tries to sink more than about 22 mA. Q34 pulls down on the output node when current must be sunk by the op amp. To do this, the voltage on the base of Q34 must rise. This is accomplished by causing the base of Q31 to rise.

Now let's start at the input and work our way to Q31.The input PNPs, Q1 with Q27 and Q2 with Q28, form a differential circuit. Notice that the base of Q27 is connected to the base of Q28. As the non-inverting input, IN$^+$, falls relative to the inverting input, IN$^-$, it pulls down on the emitter of Q28. This causes an increase in collector current flowing in Q28. This collector current is supplied by a current source involving Q15. As Q28 takes more current, there is less current available for Q19. The result is a rise in voltage at the collector of Q19. The base of Q33 is connected to this node. The rising of the base of Q33 causes the output node to rise. That can cause an increase in current flowing out of the OUTPUT node. If there is current flowing out of the OUTPUT node, then the voltage at the base of Q33 with respect to the OUTPUT node increase. This causes the current in Q25's collector to rise. More current in Q25's collector means more current flow in Q25's emitter which means less current flowing in the emitter of Q32. We therefore get less current flow in the collector of Q32. Less current flow in the collector of Q32 causes the base of Q31 to fall.

SCHEMATIC DIAGRAM

1/2 LT1013, 1/4 LT1014



As the base of Q31 falls, the collector current in Q34 falls. We therefore get an increase in Q33's emitter current while we get a decrease in collector current in Q34. Q33 and Q34 work in concert to increase the current flow out of the OUTPUT node. It can be shown that if the current is flowing into the OUTPUT node, the current through Q33 will fall and the current through Q34 will rise.

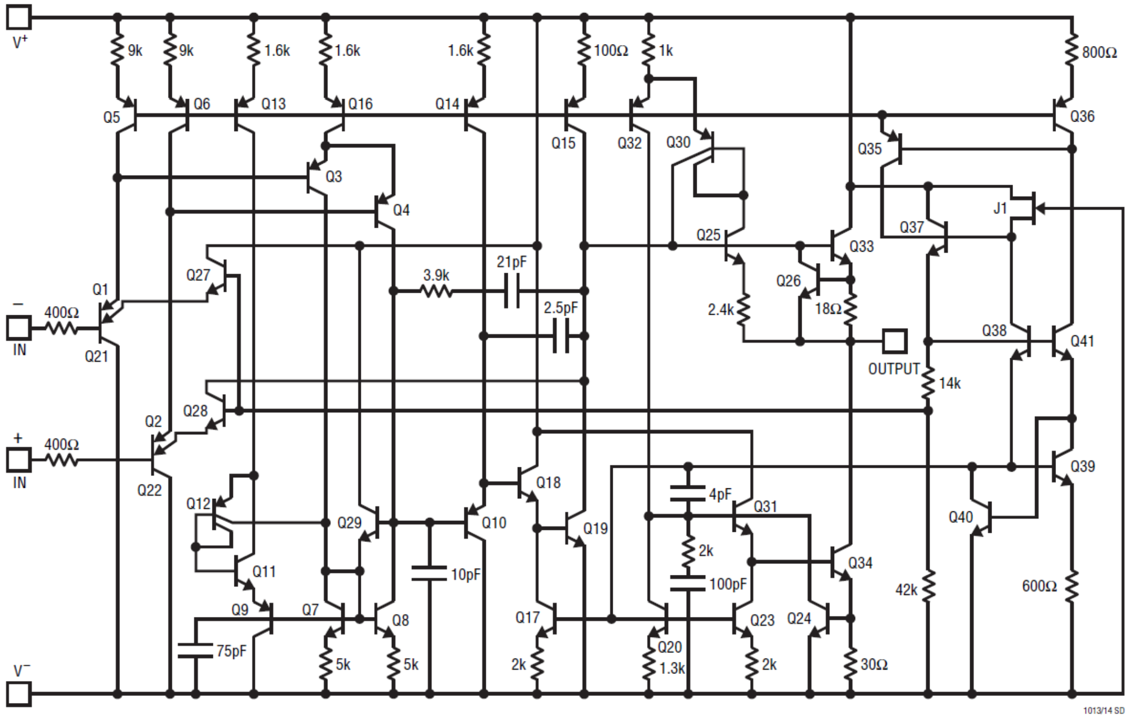
To sum up what we have found so far, as the IN^+ voltage rises with respect to IN^-, the OUTPUT node rises. Q33 works with Q34 to insure that the OUTPUT node rises for both the case of current flowing into or out of this node.

Consider what happens if IN^+ and IN^- rise the same amount. The base emitter junction of Q28 does not change so the collector current does not change. Therefore, the base of Q33 sees no change just as if IN^+ and IN^- didn't move at all. This is our Common Mode Rejection Ratio and is very large in this op amp.

Next we will look at power supply noise rejection. The top line is our V^+ node. The line parallel and just below it is our current source reference node. Q35 and associated circuits control the voltage between V^+ and this reference node. This control is not a function of V^+. The result is an array of current sources who's values are independent of V^+. The op amp therefore achieves a high power supply rejection ratio between V^+ and the output. In other words, noise on V^+ relative to ground is greatly attenuated before reaching the OUTPUT node.

SCHEMATIC DIAGRAM

1/2 LT1013, 1/4 LT1014



But what if the current is flowing into the OUTPUT node?

While IN$^+$ falls, assume IN$^-$ rises. This causes the base of Q1 to rise which cause the emitter of Q27 to rise. But the base of Q27 is tied to the base of Q28. We know that the base of Q28 is falling so with the emitter of Q27 rising, we get a reduction in the current flowing in Q27.