

## HOME BREW PIN DIODE QRO QSK SYSTEM

Don Huff, W6JL, 11 March 2016

The CW Q-signal "QSK" means, "I can hear you between my signals". As such, it requires that the receiver be available for reception whenever the key is not down. This means being able to hear the other station at full sensitivity between code elements (dits and dahs). There is thus no perceptible "delay" with true QSK operation, per definition. Without QSK (especially if the transmitter is locked on), the operator is transmitting blind, and has no knowledge if the other station is copying or not. There could be QRM, QRN, and QSB and the receiving operator has no way to interrupt the sending station who keeps on blindly sending to no one. This results in a big waste of time, and much frustration. So unnecessary! At the very least, a foot switch should be provided for T/R, if not even a VOX-operated T/R relay is available. I consider listening whenever the key is not down to be an essential part of good operating practice on CW.

My operational concept for QSK for my 600W rig is to keep the receiver running at all times. The QSK operation is transparent to the operator. My own signal is present in the receiver at the same level and frequency as the other station I am talking with. This also greatly facilitates knowing that you are exactly on the other station's frequency (over  $\frac{3}{4}$  of all the stations that answer my CQ's are not on my frequency, and they are all using modern transceivers, so there is a general problem somewhere with operators' knowledge of how to use them). In my rig there is no receiver muting, or inserted side-tone, as is done with commercially-made ham transceivers. There is thus no evident transition when the rig is switching back and forth from R to T and vice versa. If you want transparent QSK like this then you have to build it; you cannot buy it. Here is how I built mine.

The Problem: We want to transmit when the key is DOWN, immediately. We want to receive when the key is UP, immediately.

The Solution: Leave the antenna connected physically to the transmitter and receiver at all times. No physical switches or relays are thrown to go from transmit to receive, or vice versa. How do we accomplish this? By providing, immediately upon key-down, a very large attenuation, or insertion loss, from the antenna (which is connected to the transmitter output of course), back into the receiver input. This way the receiver's input circuits are never overloaded by my own 600W transmitted signal. I've found that this is not as difficult to accomplish as it may seem at first glance.

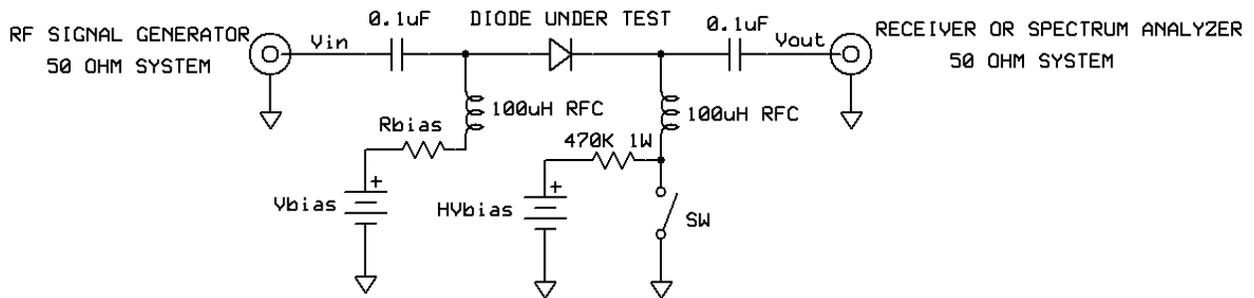
### **PIN Diodes**

The key to achieving the above is to use a very interesting characteristic of certain diodes: they switch very slowly, so slowly that at RF frequencies, they never turn "off" during several RF cycles, even with DC forward bias currents much smaller than the peak reverse RF current flowing through them. Such diodes have doping in their silicon junction that consists of a layer or stack of a P type, followed by a central I (for Intrinsic) type, and then an N type silicon. This P-I-N structure is called a PIN diode for short. (Are PIN diodes new? Nope, they've been around since the 1960's or earlier). In such a diode, the recombination time (the time for holes and electrons to recombine in the intrinsic (I), center region, is quite long, several microseconds or more. This makes the diode unusable as, say, a rectifier, at RF frequencies. But they work very well as rectifiers at low frequencies, such as 60 Hz line frequencies. Recall that with a normal diode, you must supply sufficient forward bias to be greater than any reverse signal current. With a PIN diode, it is only necessary to forward bias it at a DC current level much smaller than any peak (in either direction) current at RF frequencies (1 MHz or higher), that may be passing through it. Even though the peak reverse RF current may greatly exceed the forward DC bias current, the diode is so slow that it cannot react to this large "turn off" current, and remains in the "ON" condition throughout the RF current cycle. This is just what we want for an RF switch: easy to turn "ON" with a relatively small DC forward bias current, and easily

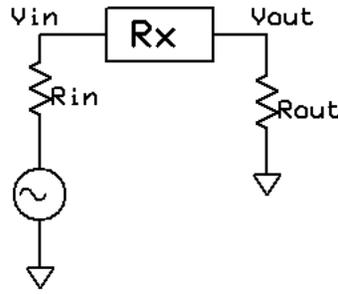
turned "OFF" by a large reverse DC voltage potential across the diode.

When the diode is "ON", it appears as a very low resistance. When the diode is "OFF", it appears as a small capacitance, a few pF or less. With these diodes as a key element, it is a relatively simple matter to configure them to work as a Single Pole Double Throw (SPDT) RF switch, as used in my QSK setup. The "arm" of the SPDT switch is connected to the antenna, and the two "contacts" are connected to the transmitter and receiver. I use a separate transmitter and receiver instead of transceiver operation but the same principles apply to a transceiver QSK switch.

Let's look at the above description of a PIN diode's operation in a schematic form and see how it is utilized. Below is a schematic of a test jig I built to test PIN diodes for forward and reverse characteristics at a given frequency. This is followed by a derivation of  $X_c$  and hence  $C_x$ , in terms of an insertion loss measurement.  $R_x$  and  $C_x$  in the equations below are the equivalent resistance and capacitance, respectively, of the PIN diode under test. (Note: The analysis uses high-school level mathematics, plus Ohm's Law and electric circuit knowledge that is tested on the FCC General Class and higher exams, so *should* be familiar to any HF-licensed amateur).



TEST JIG FOR MEASURING PIN DIODE INSERTION LOSS



Above, test jig's simple equivalent circuit.  $R_x$  is the equivalent series impedance of the PIN diode under test. In a 50-ohm system,  $R_{in} = R_{out} = 50$  ohms

$$Insertion Loss_{dB} = 10\text{Log}\left(\frac{\text{Power out}}{\text{Power available}}\right)$$

$$= 10\text{Log}\left(\frac{\frac{V_{out}^2}{R_{out}}}{\frac{V_{in}^2}{R_{in}}}\right) = 10\text{Log}\left(\frac{V_{out}}{V_{in}}\right)^2 = 20\text{Log}\left(\frac{V_{out}}{V_{in}}\right)$$

Using Ohm's Law, and since this is a simple voltage divider,

$$\frac{V_{out}}{V_{in}} = \frac{R_{out}}{R_{out} + R_x} = \frac{50}{50 + R_x} \quad \text{And, from preceding above,}$$

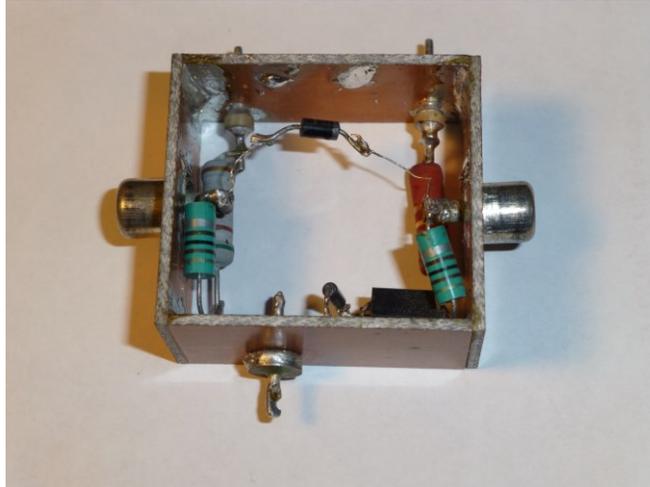
$$Insertion loss_{dB} = IL_{dB} = 20\text{Log}\left(\frac{V_{out}}{V_{in}}\right) = 20\text{Log}\left(\frac{50}{(50 + R_x)}\right)$$

To solve for  $R_x$ , we take the antilog of both sides of the equation:

$$10^{-\left(\frac{IL}{20}\right)} = \frac{50}{(50+R_x)}, \quad \text{rearranging, } 50\left(10^{-\frac{IL}{20}}\right) + R_x\left(10^{-\frac{IL}{20}}\right) = 50$$

$$\text{Solving for } R_x: \quad R_x = \frac{50\left(1 - 10^{-\frac{IL}{20}}\right)}{10^{-\frac{IL}{20}}}$$

If  $R_x$  is a capacitance with reactance  $X_c$ , then  $R_x = |X_c|$ , and  $C_x = \frac{1}{2\pi f X_c}$



Above: my complex, elegantly-built, PIN diode test jig

The diode is turned ON by a forward bias voltage as shown in the test jig schematic. The DC forward bias current is set by the bias supply (subtracting the forward bias voltage on the diode) and the series biasing resistor values, according to Ohm's Law. The diode is turned OFF by a relatively high reverse bias voltage, also shown. (As long as the reverse bias voltage is larger than the forward bias power supply voltage, the diode is OFF; it is not necessary to remove the forward bias voltage. It is the cathode-to anode voltage seen by the diode itself that matters). The magnitude of the reverse bias voltage required depends on the maximum power going into the ANT port when the transmitter is keyed. For example, for 550 watts into a 50 ohm antenna load, the RMS voltage is found from  $Power = \frac{(Volts\ rms)^2}{Rload}$ , and, rearranging,  $(Volts\ rms)^2 = (Power) \times (Rload)$  from which  $Volts\ rms = \sqrt{Power \times Rload}$

Plugging in 550W and 50 ohm load yields

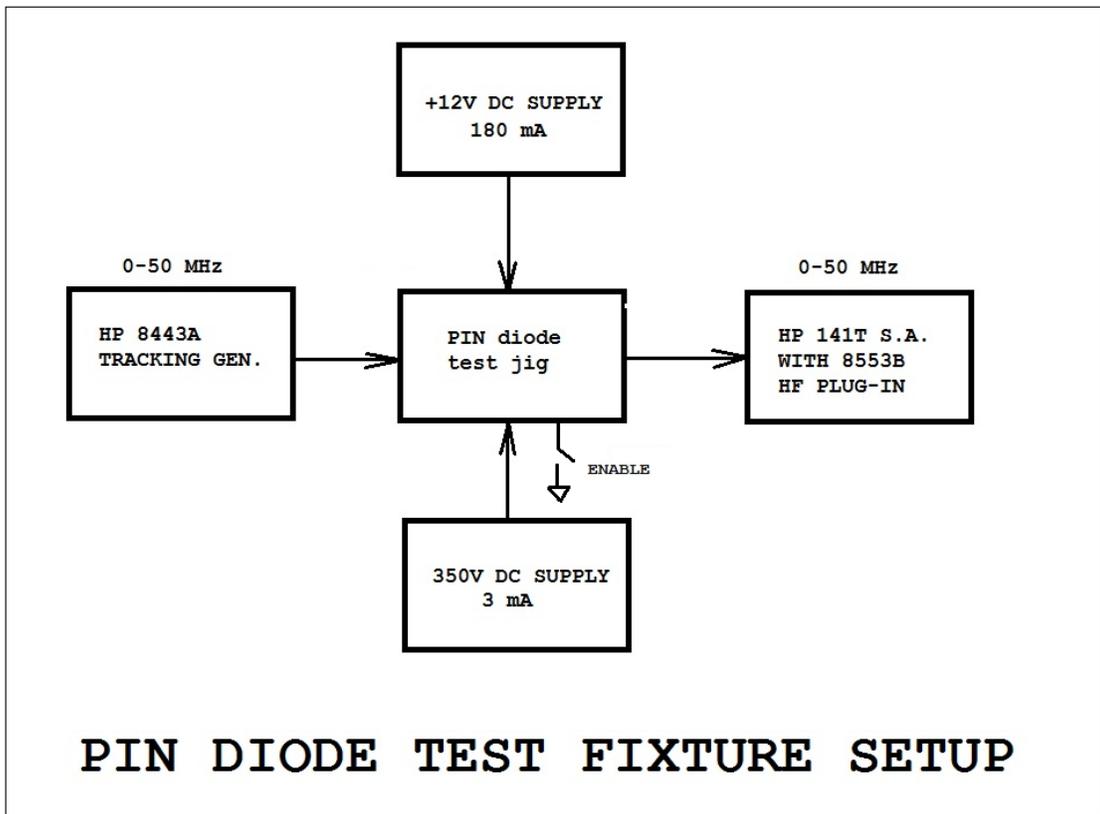
*Volts rms = 166V rms, which is 234V peak.*

In my case in my PIN switch I use the 350VDC screen voltage already available in the amplifier, for my PIN diode blocking bias.

In the OFF condition, the diode acts as a small series capacitor, and the diode's attenuation in a 50 ohm system at any convenient frequency can be easily measured with simple equipment (an RF source applied at the left side, and a 50 ohm

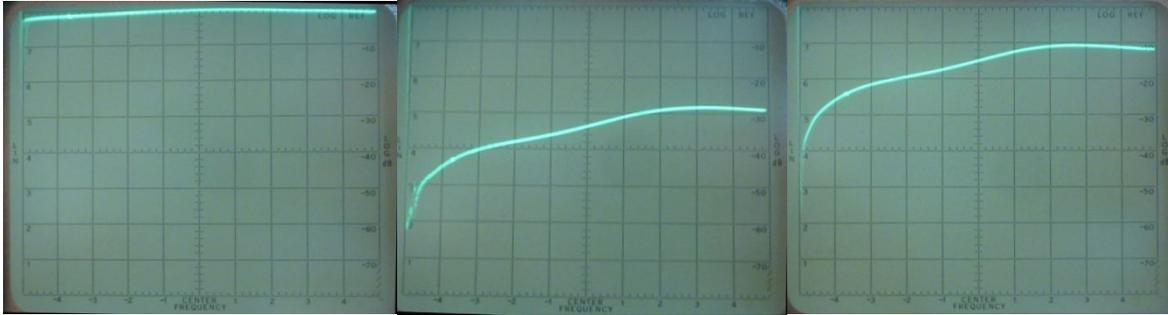
load with RF voltmeter (or receiver or spectrum analyzer) at the right side). The test jig is also useful for measuring diode temperature rise with high power RF from the transmitter driving the circuit, using a 50 ohm high power dummy load. This circuit is also easily simulated in any circuit simulation software program, such as the free LTSpice, [http://www.linear.com/designtools/software/?gclid=CLKJ2fPRnbYCFc5\\_QgodGGoA3g#LTSpice](http://www.linear.com/designtools/software/?gclid=CLKJ2fPRnbYCFc5_QgodGGoA3g#LTSpice) using Spice models (also freely available online) for the diodes under consideration. This makes it easy to compare the simulated and actual results.

Let's look at a typical measurement result using the test jig. In this case, let's measure a 1N4007 1KV 1A rectifier diode. I will use my 1960's vintage HP141T spectrum analyzer, now 50 years old and still a very useful and cost-effective measurement tool. I could just as well have made the measurements with a homebrew RF voltmeter of any kind, or an oscilloscope. The test setup is shown below. The HP141T has a companion tracking generator, the HP8443A. Its output tracks the instantaneous sweep frequency of the 141T making it easy to make measurements of most any RF two-port (input and output) device, in dB.



In the above diagram, the ENABLE switch effectively shorts the HV DC to ground at the cathode of the diode, via an RF choke, as shown in the test fixture schematic previously. (The 350V is current-limited by the 470K resistor, so excessive current does not flow). This is the condition for the diode being in the ON state. When the enable switch is opened, the forward DC bias current is removed, since the cathode of the 1N4007 diode rises to +350V, overwhelming the +12V forward bias on its anode. The diode is now back-biased with  $350V - 12V = 338V$  DC. The swept response in either case is easily observed on the screen of the spectrum analyzer. Below are screen shots of three different test conditions: Forward bias (diode is ON), reverse bias (diode is OFF), and with zero bias (diode is OFF, but there is no reverse bias applied, and its capacitance will be somewhat higher than with reverse bias applied).

The results are shown in the CRT images below (10dB per vertical division, and 5 MHz per horizontal division). The top graticule line represents a loss of 0dB through the diode, and each line below is 10 dB of additional loss. The traces indicate a sloping response vs frequency in the OFF condition, as we would expect for a small capacitance equivalent for the diode under test. For the ON condition the diode is a virtual short circuit for RF and so very little attenuation is shown (it is actually less than 0.1 dB. This can be verified in several ways, the easiest being to simply measure the output voltage vs input voltage using a sensitive RF voltmeter or oscilloscope. The spectrum analyzer, being a logarithmic device, lacks sufficient resolution to measure fractional dB accurately). For the ON condition, the trace is along the top reticule line, indicating 0 dB. For the OFF condition, the trace shows about 28 dB of attenuation above 30 MHz, increasing to 40 dB at 10 MHz, to over 60 dB at 1 MHz.



PIN diode ON                      PIN diode OFF                      PIN diode w/zero bias  
 No attenuation                      30 to 60+ dB atten.                      10 to 30+ dB atten.

With zero bias, the attenuation varies from 10 dB at high frequencies, to about 30 dB of attenuation at the bottom of the HF spectrum. So it can be seen that reverse bias increases the OFF attenuation of a single diode by 30 dB. What is the equivalent capacitance of the OFF diode at, say 25 MHz? The attenuation at 25 MHz is about 34 dB as shown on the center spectrum analyzer trace above.

Using the previously derived equations, we can now calculate equivalent series Rx (and, since the OFF diode is effectively a capacitor with capacitive reactance  $X_c$ ) as follows:

$$R_x = |X_c| = \frac{50(1 - 10^{\frac{-IL}{20}})}{10^{\frac{-IL}{20}}} = \frac{50(1 - 0.02)}{0.02} = 2450 \text{ ohms}$$

The equivalent Cx is then,

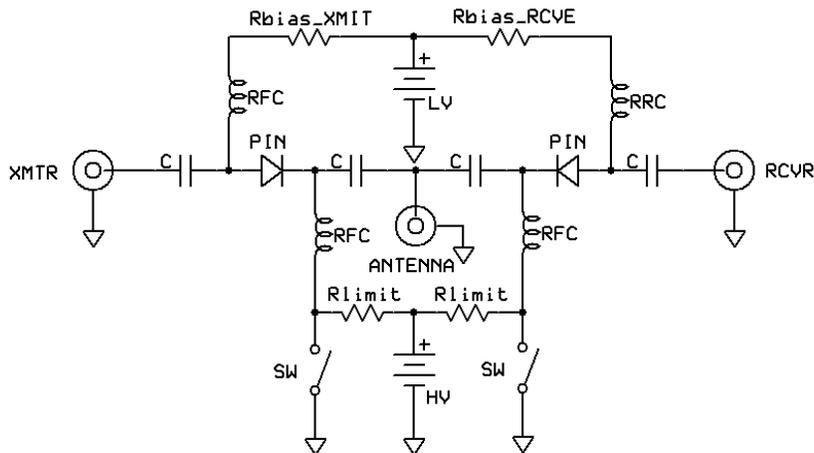
$$C_x = \frac{1}{2\pi f X_c} = \frac{1}{2\pi(25\text{MHz})2450\text{ohms}} = 2.6 \text{ pF}$$

So the 1N4007 appears as a 2.6 pF capacitor at 25 MHz when in the OFF condition with 350V of back-bias applied. This is a fairly large capacitance for a PIN diode, but, hey, we can only get so much for \$0.10, and this is one reason why the 1N4007 is not advertised and marketed as a PIN diode :o). (We will see further that this capacitance is not a significant handicap, since we will cascade series and shunt diodes to achieve very high attenuation in the OFF condition for the receive path).

It turns out that one of the ways of achieving high voltage breakdown in silicon *rectifier* diodes is to use a relatively thick I region between the N and P type silicon diode ends. Depending on the exact diode chosen, it can sustain several hundred watts of RF power passing through it for reasonably short periods of time. All this for less than perhaps \$0.20. Sweet! This is all we need for a QSK RF switch, at up to several hundred watts or more of RF power. (Note: Small DO-41 packaged diodes like the 1N4007 would never be used in a commercial QRO QSK design; the power rating is too marginal for just any user. But for a single operator/homebrewer, who understands their power and thermal time-constant limitations, they serve very well, and have for me for thousands of hours of operating, at 550W, thank you. Again, we are talking about conscious abuse of parts, thought out beforehand. And if one ever does fail, the 10-cent replacement cost is bearable too. This is one of those design freedoms that homebrewers have but designers of equipment for sale do not). We will see later on that using a 1N5408 3A diode enables high power with a single diode in the transmit path.

All that is left now is to configure a circuit that allows using these diodes to obtain as little forward loss, and as large a reverse loss, as required. For this we can use relatively high DC forward bias current, and high voltage reverse bias.

Below is the basic circuit of a SPDT PIN diode switch.

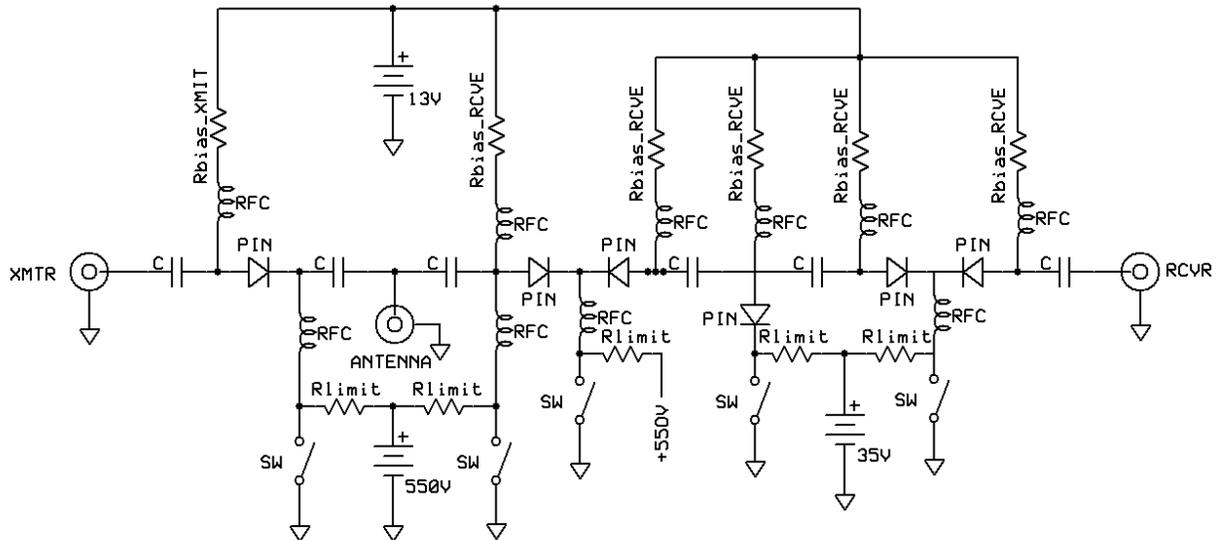


**SPDT PIN DIODE SWITCH**

In this circuit, all series capacitors have low reactance and all shunt RFCs have high reactance at the RF operating frequencies. The circuit works by forward biasing the left-hand PIN diode switch for transmit (left hand Switch is closed), which supplies forward bias current to the left hand (XMIT) PIN diode, turning it ON. Thus, the ANTENNA is now connected to the XMTR. At the same time, we open the right-hand switch, which enables HV reverse-bias to be applied to the right hand PIN diode. (The high voltage is greater than the peak RF voltage applied to the ANTENNA from the XMTR, thus assuring that the right hand PIN diode remains OFF at all times during XMIT). For receive (key up) the reverse occurs: The left hand switch is opened, which back-biases the left-hand PIN. The right hand switch is closed, forward biasing the right-hand PIN and enabling the ANTENNA-to-RCVR path with very low attenuation. The capacitors are there to isolate the DC levels of the diodes from the XMTR and RCVR ports. The RFC's are there to prevent RF voltages from having a low impedance path to ground through any of the biasing circuits.

For a really good PIN diode switch, we want a large OFF attenuation, and a small ON attenuation, in each path. (We especially want a large OFF attenuation in the RCVE path when the key is DOWN, to protect the receiver, at the very least, which is sensitive to nanovolt signals!). To achieve large OFF attenuation we can use multiple diodes, in series and in shunt instead of just the single right hand diode shown, so that the combination results in the creation of a large attenuation due to the low series capacitance, and low series resistance, of OFF and ON PIN diodes, respectively. In other words we can form a ladder attenuator using several diodes to achieve high insertion loss from ANTENNA to RCVR. We must also isolate the sections of the attenuator by suitable shielding so that leakage paths around the diodes do not degrade the attenuation from the diodes themselves. Below is the elemental schematic of a SPDT arrangement that I have found capable of achieving, by measurement, over 130 dB of receiver isolation at the receiver port, while withstanding 600W of RF power at the amplifier/antenna port. I simulated several arrangements of series/shunt PIN diodes using LTSpice and found to my surprise that it was apparently fairly easy to achieve very high

isolation in the receive path, in excess of 160dB, at least in the pure world of simulation where physical constraints (such as enclosure leakage, etc) are not included in the results. I found that, theoretically at least, it only requires a series-shunt-series arrangement of low capacitance PIN diodes, as shown below. All series capacitors are low reactance at HF, ie, 0.1uF.



### SPDT HIGH-PERFORMANCE PIN DIODE SWITCH

So now we have our desired circuit arrangement. But how do we switch things around in a real system? We need to control the turn-on and turn-off of each of the diodes at the right time, in the right sequence. For example, when in RCVE (near zero attenuation from ANT to RCVR), the series diodes are ON, and the shunt diode is OFF, in the RCVE path. When the key goes DOWN, we want to immediately turn the receive path series diodes OFF, and the shunt diode ON. How do we do this? We use transistors (actually, mosfets) in place of the switches shown, to switch the forward bias current and reverse bias voltages on and off at the proper times. I have found it useful to base all of the switching on the state of the key line itself. That way, the key controls everything. When the key is UP, the voltage on the key line is approximately 13V in my transmitter. When the key is DOWN, it grounds the key line, so the key line voltage is

essentially zero. We can use this to control all the switching in the complete T/R switch.

Later on in this narrative is a complete working schematic of my PIN T/R switch, showing these transistors and all other components. Although it may look complex at first, it is still basically the same concept discussed above. Referring to this schematic now, let's walk through this circuit and see how it works. Print this schematic out so you can refer to it while reading the following:

Starting at the bottom left of the schematic first, the KEY LINE drives the gate of mosfet Q5 via an adjustable timing circuit consisting of D5, R10, and C10. This provides instantaneous shut-off of Q5 when the key goes DOWN, but a delayed drive to Q5 when the key later goes UP. D5 is forward-biased when the key goes DOWN, discharging C10 rapidly through the grounded key line, and cutting off Q5 immediately. When the key later goes UP, diode D5 is back-biased, and capacitor C10 charges much more slowly through R10, delaying the turn-on of Q5. This delay on key UP delays the switching of the entire T/R switch from Transmit back to Receive, to allow the RF envelope from the transmitter to reach zero before enabling the receive path. When the key goes DOWN, immediately the T signal at the drain of Q5 goes high. Thus, everywhere in the circuit where there is a T signal connected, that signal goes high immediately upon key DOWN. Looking at Q12, we see that its gate is driven high by T, turning it on immediately (within microseconds) of key DOWN, and thereby turning D2 ON. The XMIT path is now enabled.

At the same time, T also drives, through isolation resistor R30 and feed-through capacitor C41, the TX-2 signal, which drives Q1, turning on the shunt PIN diode D6, thereby grounding the receive path at that point. At the same time that T goes high, R is driven low via Q11, which is turned ON by Q5 turning OFF. The R signal drives RX\_1 and RX\_3 low also. RX\_1 turns Q7 OFF, enabling high voltage blocking bias for PIN diodes D7, D3 and D8. D7 and D3 are in series to provide up to 2KV of combined breakdown voltage to allow for even an open circuit to exist at the antenna, with the amplifier attempting to put full power into that open circuit. Finally, dual PIN diode D9 is turned OFF by RX\_3 going low and mosfet Q17 turning OFF. PIN diodes

D8, D6, and D9 see no high voltage from the XMTR at any time, so do not require high voltage back-biasing. I use +34 VDC, obtained by a voltage divider from the +550V supply.

All of the foregoing, upon key DOWN, is completed in less than about 50 microseconds (uS). The transmitter does not start ramping up the RF envelope until over 100 uS has passed since key DOWN, so there is no hot switching of RF at the start of a code element.

Dual PIN diode D9 (63 cents each, in unit quantities!), has only 0.34 pF of OFF capacitance for each diode, and a 100V breakdown rating; such a deal. It makes an ideal last-section series diode PIN attenuator, but it is in a very tiny SMT package.

When the key later goes UP, the above process is exactly the opposite, except that the key UP level change is delayed to Q5 by the aforementioned variable timing circuit. Thus, the switching of the diodes from XMTR back to RCVE is delayed about 9 milliseconds, to allow the RF envelope of the transmitter to fall to zero, with a bit of extra time margin provided. (The transmitter envelope actually falls to zero in a linear ramp, requiring about 5 milliseconds). While the key is UP, the forward bias on the diodes in the receive path is set by resistors R11, R12, R18, R19, and R22, to just a few mA for each diode. This is all that is required to assure they are in a low-resistance state. The receive-path attenuation is well under 1 dB.

Note that there is no possibility of there ever being a simultaneous T and R signal in the T/R switch. This is essential, because if that happened, full transmitter power could be applied directly to the receiver path circuits, with undesirable results. So far, in several years of using homebrew PIN diode switches, that has never happened in actual operation on the air. But it has happened once, while testing and probing the circuit, because I slipped with the probe. The result was destroyed PIN diodes D6 and D9, but, surprisingly, no damage to the receiver input circuits whatsoever. I would not count on this result every time, however :o).

So much for the receive path (not so hard to understand, so far, eh? You need to know basically Ohm's Law, and how common devices work). Next let's look at the XMIT path, when the key is DOWN.

The transmit path via D2 requires very low loss in XMIT, but only moderate loss in RCVE (30 dB or so is adequate). This is because in receive, we only need to isolate the output of the transmitter just enough to prevent it from interacting with the receive signal path from ANT to RCVR. This is easily accomplished with only a moderate attenuation of say 20-30 dB in the XMIT path, when the key is UP. This is achieved by using a single robust diode in the XMIT path as shown. In this case a 1N5408 3 amp silicon rectifier diode is used. It too has a PIN structure. It has a larger OFF capacitance than the 1N4007, but also has greater power handling capability (larger junction area, and heavier leads to conduct heat out of the diode junction). (Actually, I have used parallel 1N4007's in the transmit path with complete success for years, at least at the 600W level).



By now the arrangement for switching D2 ON and OFF should be familiar, as it is similar to the switching of the series diodes in the RCVE path. However we must deal with much larger voltages and currents in the XMIT path. Consequently, there is a larger value and physical construction of the RF chokes to isolate the RF path from the switching circuits, and also larger wattage resistors in the DC biasing circuit because the forward bias current is much larger in the XMIT path diode D2. How large? By using 550W from my transmitter in the diode test jig previously described, I measured the temperature rise of the 1N5408 diode vs forward DC bias current, while driving a 50 ohm dummy load. It turns out that about 150mA or above results in only small RF heating of the diode. Below that and there is a fairly rapid rise in temperature as bias current is reduced. So I have chosen to use about 175mA of DC bias current for the 1N5408. Things like this require only the simplest of test equipment (my temperature-sensing index finger, in this case).

Let's now look at how the transmit path forward bias circuits work in more detail. In the XMIT path the larger bias current requires consideration of power dissipation in the bias circuit components. When the diode is ON (key is DOWN), as previously discussed, switch Q12 is ON, and +13V is applied to the path R5 L1 D1 L2 and Q12. Since we want 175mA to flow in this path, the total resistance must be approximately 13V/175mA or about 74 ohms. The RF chokes I am using are small pi-wound units with about 10 ohms of resistance each. The mosfet Q12 has an ON resistance of about 10 ohms (see its data sheet). This adds up to 30 ohms. The remaining 44 ohms is supplied by the 42 ohm bias resistor R5. The current through the bias string is 175mA, so the dissipation in each resistance is  $(I^2R)$  or (175mA) squared, times the resistance. For R5 this works out to 1.3 watts. A 3W metal oxide resistor is chosen so as to have a safe operating margin, although a 2W resistor is adequate. The dissipation in the RFC's and in Q12 is similarly calculated to be .31 watts each. This is low enough to be dissipated with the components in free air, especially since these dissipations only exist while the key is DOWN. In RCVE (key is UP), the dissipations in the transmit path forward bias circuit are zero, of course.

Let's now look at the reverse bias circuit. This is similar to that of the RCVE path. We actually do not need to have high voltage for back biasing D2, since the XMTR is OFF when this path is OFF, but since the high voltage is already available, we can make things simple by leaving the +13V forward bias string in place, and applying the much larger HV bias to D2 cathode as shown, switched to ground in XMIT by Q12. The current limiting resistor R24 is chosen to limit the HV current flowing in XMIT to a conveniently low value; in this case 470K ohms is chosen, limiting the current to about 1.2 mA (550V/470K). The dissipation, key DOWN, in R24 is then  $1.2\text{mA}^2$  times 470K ohms, or 0.64 watts. A 470K 2W carbon resistor does the job nicely. (We want to use a physically large resistor anyway, when we are subjecting a single resistor to several hundred volts, so that we have a long voltage breakdown path length along the resistor body).

The transistors that switch the 550V bias high voltage (Q12 and Q7) need to withstand this voltage of course. I am using 600V devices (FQN1N60C mosfets), which I had on hand. They are a bit marginal voltage-wise, but their *typical* breakdowns are higher than 600V, and, again, we are building just one of these T/R switches, to be used only by the designer/builder. So, hey, we can cut any corners we feel comfortable with.

We always need to ascertain the voltages and currents applied to components in order to choose the proper ones. Not just any capacitor, resistor, or inductor will do, especially when considerable RF power and high voltage is involved. It is instructive to consider the RF voltage present on the XMIT path components when the key is DOWN. If we have say 600W of transmitter power driving a 50 ohm antenna load, then the RMS voltage is easily calculated from  $P=E^2/R$  by rearranging the terms, finding that  $E=\text{Square root of } (P \times R)$ . Substituting real values, we get  $E=173\text{V RMS}$ . Now this voltage is applied across the RFC's and bias resistor in the XMIT path. With the values chosen (240uH), and at the lowest expected frequency of use being 1.8MHz, the reactance is  $2 \times \pi \times L \times F$ , or 2714 ohms of inductive reactance. This results in an RF current flow of  $173/2714$  or 64mA RMS. This is a substantial current, which also flows through the bias resistor and any other resistors before reaching the first bypass capacitor to ground, which in this

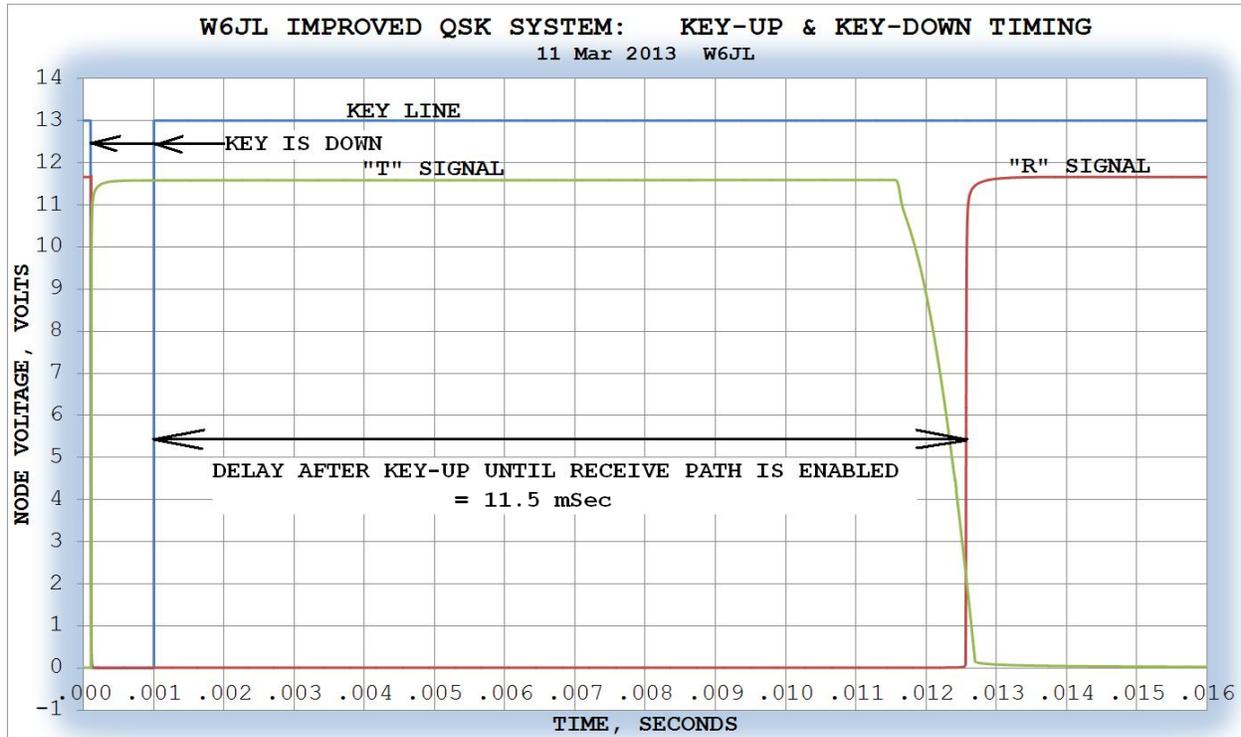
case is C17. This capacitor must safely handle this RF current, through it to ground. Fortunately even a small 0805 surface mount capacitor can do this easily. The dissipation in R5 due to RF current alone is about  $(.064)^2$  times 42 ohms, or 0.17W, which is small compared to dissipation due to the DC bias current of 175mA, discussed above.

Similarly we can find the XMIT path RF current into the ANT port from  $P=I^2$  times R. Rearranging terms,  $I = \text{square root of } (P/R)$ . This yields 3.46A RMS. Notice how nifty it is to be able to handle 3.46A RMS (which is 4.90A peak) of RF current by biasing the diode D2 to only 0.175A of forward bias! The reverse RF peak current is  $4.90/0.175$  or 28 times the DC bias current, yet the diode remains in the ON condition for RF, never having time enough to turn itself OFF during the half cycle period of the RF, which is less than a microsecond over the HF range. The recombination time of several microseconds in these diodes is more than long enough to guarantee the diode will never be turned off by the large RF reverse current flowing through the diode.

The 3.46A RMS current must safely be handled by any series capacitors in the circuit, these being C1, C6, C7 and C33. These were chosen on the basis of a precision temperature sensor, utilizing my favorite, the "finger" test. I felt how warm these capacitors got in the circuit by holding the key down for various times at 600W into a dummy load, then (AFTER releasing the key!), touching the capacitors with my well-calibrated finger thermometer, to see how hot they were. The Finger Test revealed that even small 1KV ceramic capacitors can apparently withstand this current safely, and so that is what I used, but I placed them in parallel just for extra margin (there were many of them in my junk box). The voltage rating was arrived mainly by making it much larger than the peak RF voltage applied to the antenna, which in this case of 600W is  $174*1.414 = 246V$  peak. I chose 1KV because of ready availability (they were in the junk box, right?), and some margin for mismatched loads.

LTSpice AC sweep simulations (if you haven't yet used LTSpice, I highly recommend you download this free software tool and play around with it!) already gave me the attenuation from each

section of the RCVE path. Not surprisingly, the greatest single attenuation occurs in the two series sections, especially the first one, since that section is shunted on its downstream side by the low ON resistance of D6 in shunt. This makes a very effective voltage divider if its input and output are well shielded from each other. Likewise the last series section needs to be suitably shielded, particularly across D7, the first series diode.



**Above: LTSpice transient simulation of keying of PIN T/R switch.**

This is a simulation run on the exact circuit contained in the previous complete schematic. The Key Line controls the "T" (transmit) and "R" (receive) control signals in the T/R switch. The "T" and "R" signals change immediately upon key-DOWN. Later, upon key-UP, the "T" and "R" signals remain in the Transmit state for a delay of about 11.5 mSec before disabling the Transmit and enabling the Receive path again.

So much for the circuit. We now have a PIN switch circuit that should work. Next we need to determine a PHYSICAL arrangement of parts that will compromise the performance of the switch as little as possible.

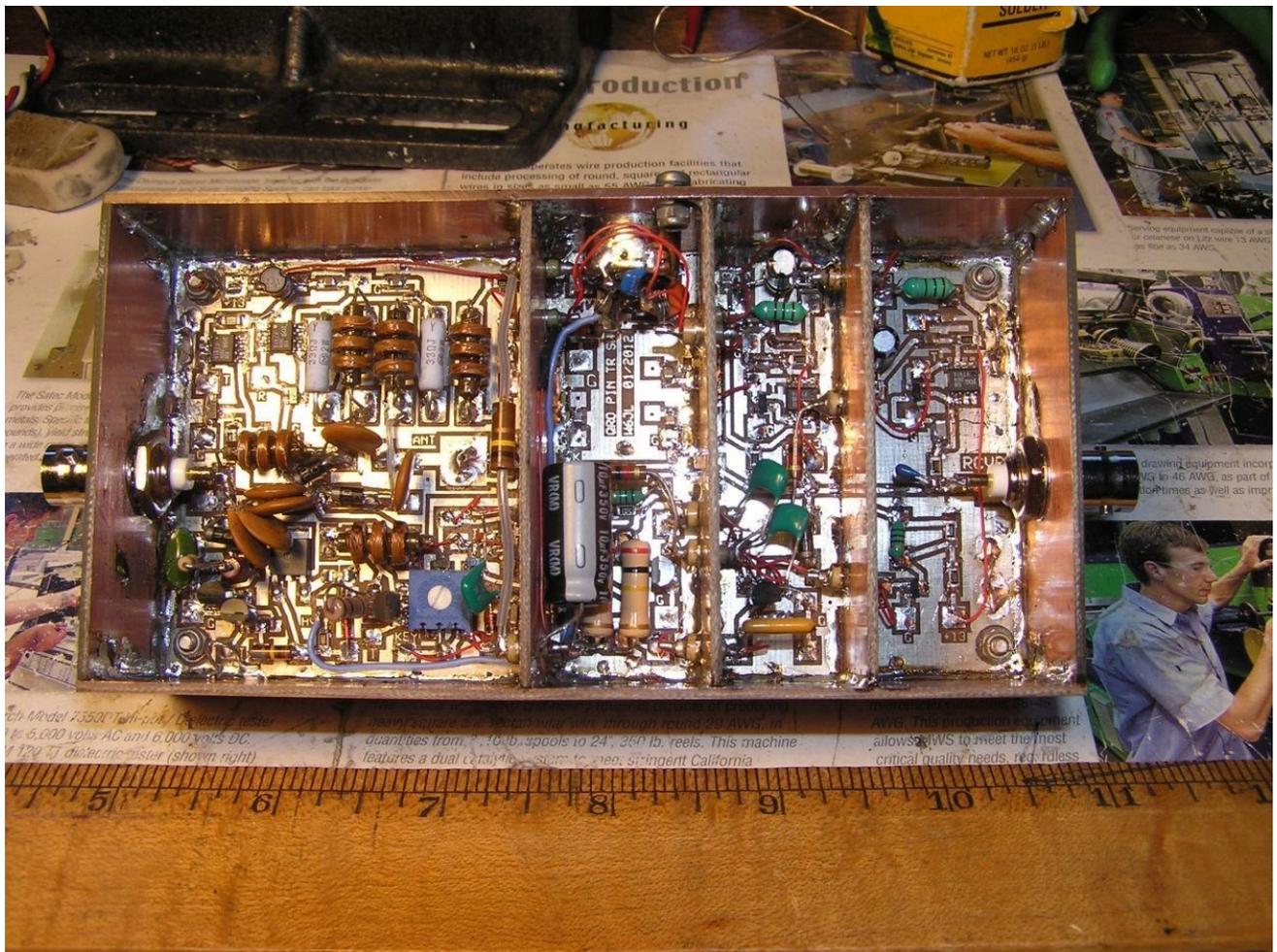
### Putting it together.

One of the beauties of homebrewing is we can decide to build something any way we want, out of several alternatives, with no real regard for it being practical (or even cost effective) in mass production. This is a great increase in the degree of design and construction freedom, compared to the designers and manufacturers of commercial equipment. This is one reason it is possible to obtain features and performance in homebrew rigs which cannot be found in commercial designs. Unlike in commercial ham equipment, whenever we homebrew, we are limited only by what is POSSIBLE. We need to capitalize on this flexibility advantage of homebrewing whenever we can. Thus we can use any physical arrangement that we feel we are capable of making. We are making only ONE, after all, optimized for ONE customer only: the builder/user.

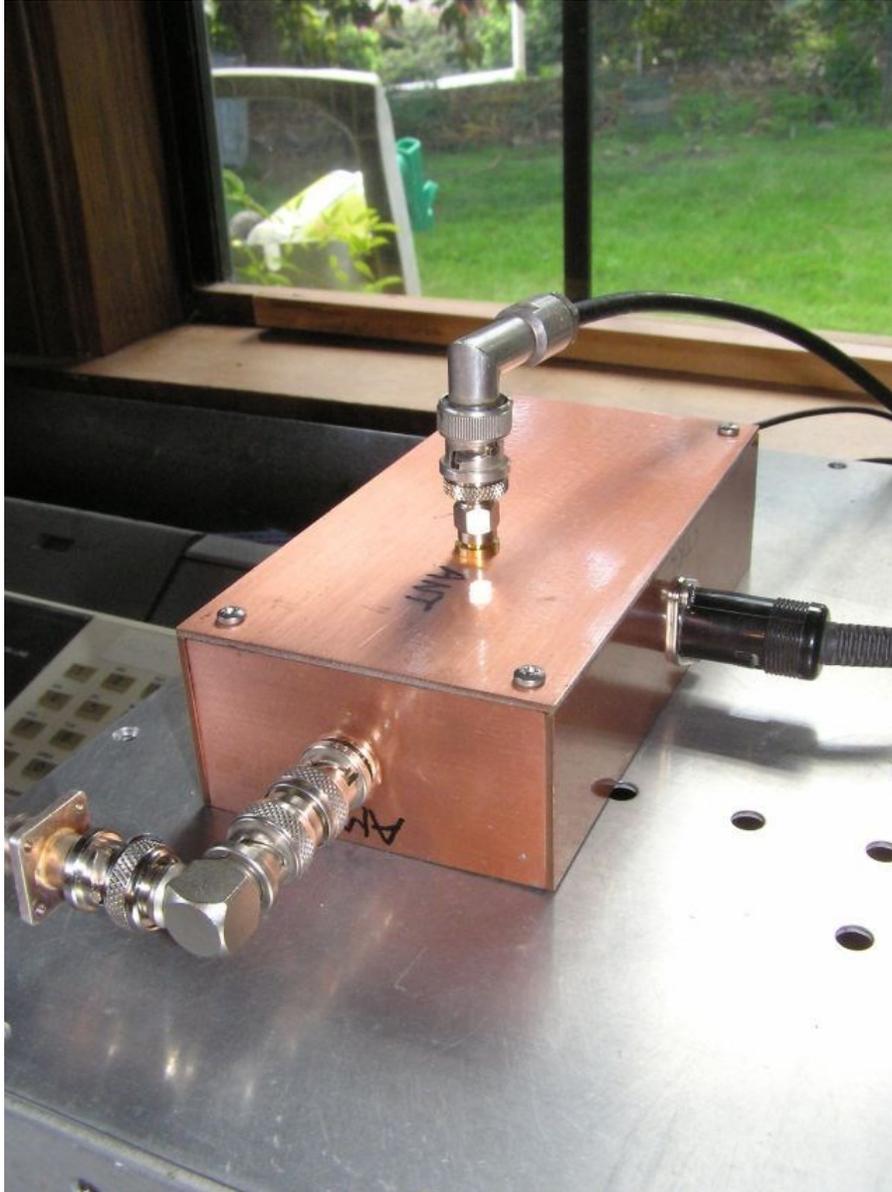
As an aside, it is a mistake to believe that all technically possible features and performance can be found in commercial ham equipment at any price; only what most buyers want to buy is included in such gear. If there is no *significant* market demand, it will not be included. This is as it must be for rig manufacturers. Most hams today have no real idea of what is technically *possible*, since they have little or no knowledge of electronics, and so most are "technically-uninformed" (that is to say, technically ignorant) buyers. The manufacturers carefully target their products to such buyers. This is one of the major drawbacks of being a so-called Appliance Operator; your choices are limited to only what is commercially available, with no knowledge of what that limitation may mean. Most hams, apparently, find themselves easily satisfied by what is available commercially. I am not one of them.

For my physical packaging, I decided to solder the entire PCB, containing the complete PIN T/R switch, into a copper-clad box, which itself contains solid soldered seams. (Once soldered, there is no going back!). Then, shield partitions, also made of copper-clad of course, would be soldered across appropriate areas of the box and the PCB to divide the box into RF-tight chambers. This was all decided BEFORE the layout of the (homebrewed, single-sided, solid copper on the back side) PC board, so that suitable clearance and solder/grounding traces

would be in place in the right areas to facilitate the soldering-in process. Also, a perimeter around the entire board was provided for soldering the board into the flush-fitting box, in the right position to allow all required external connections. Lastly, feed-through capacitors with series resistors and RF chokes were liberally used to help eliminate any RF leakage paths between the chambers along the various control and power supply lines. Finally a copper-clad lid with finger contacts was provided for the box. The photograph below illustrates better than words most of the above. This whole arrangement would be unacceptable for a production product, but completely do-able (and acceptable) for homebrew.



Above: PIN T/R switch, soldered in box, lid removed. Homebrew tin-plated PCB. XMTR port at left, ANT port on reverse side at center, RCVR port at right. The larger, pi-wound RF chokes are visible in the transmit path area in the left shielded chamber.

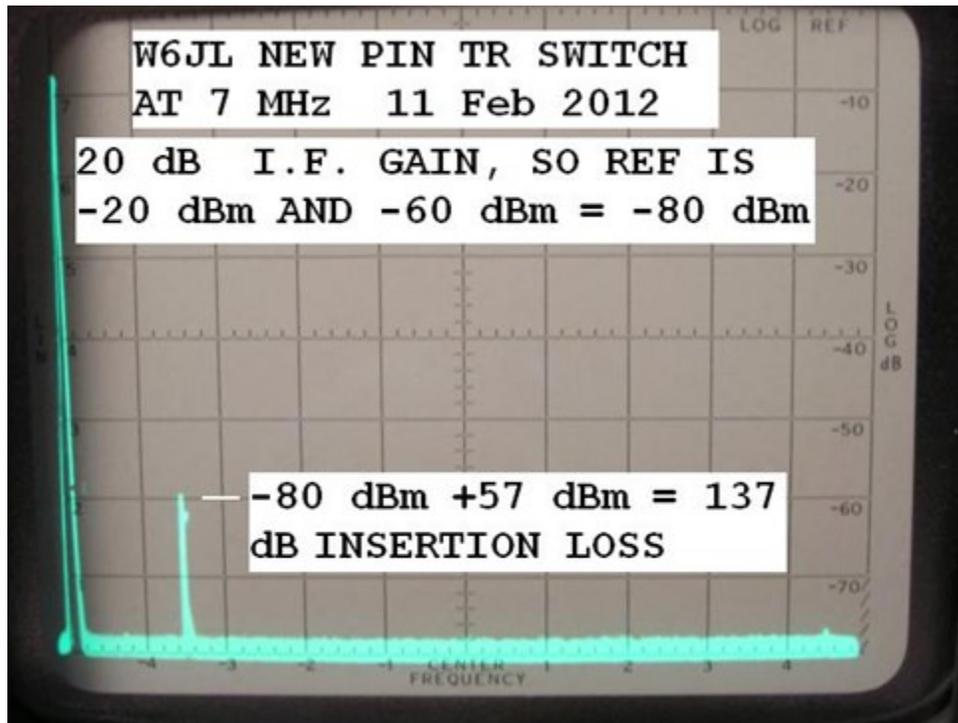


Completed QRO PIN T/R switch, connected and working.

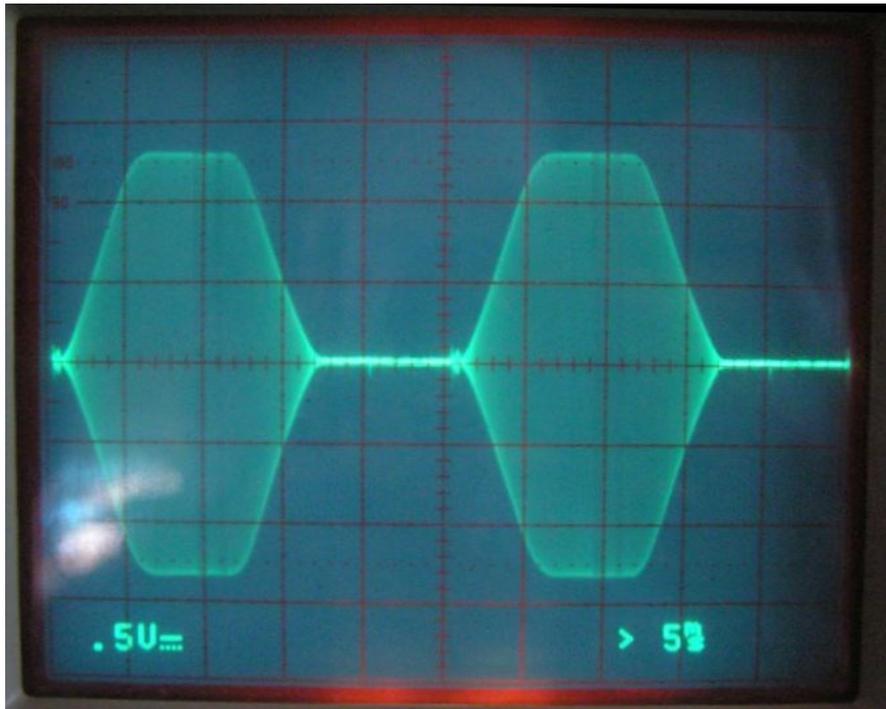
### **Measuring the receive-path insertion loss (key-DOWN)**

This turned out to be easier than I first imagined. I have a vintage 1970's HP 141T spectrum analyzer (they are available very inexpensively these days, but are still very useful pieces of beautifully designed test gear). A spectrum analyzer is nothing more than a calibrated wideband receiver (a receiver could actually be used, if well shielded). In order to obtain an easily measurable signal, I used my 500W transmitter as the

signal source, with a 50-ohm dummy load connected to the ANTENNA port. 550W is +57.4 dBm. So if the T/R switch had say 140 dB of insertion loss in the RCVE path, then the output signal at the RCVR port would be +57.4-140 or -83 dBm, easily measured by the 141T. I had to be VERY careful that when the transmitter was applying 500W that the T/R switch did not fail, because if it did, it would obliterate the input mixer of my spectrum analyzer! So I made these tests very briefly, just long enough to capture the response on the storage surface of the oscilloscope in the 141T. The result was an amazing -137 dB at 7 MHz, better than I had hoped for. At this level, we are getting into the realm of leakage from connectors, and even from the coaxial interconnecting cables themselves. In fact, these factors probably limit the attenuation in practice. It also means that with 550W applied to the antenna, my receiver will see only about 20 microvolts when the key is down. The input of the receiver can run at full gain always. Thus, I have achieved my goal of never overloading the receiver's front-end.



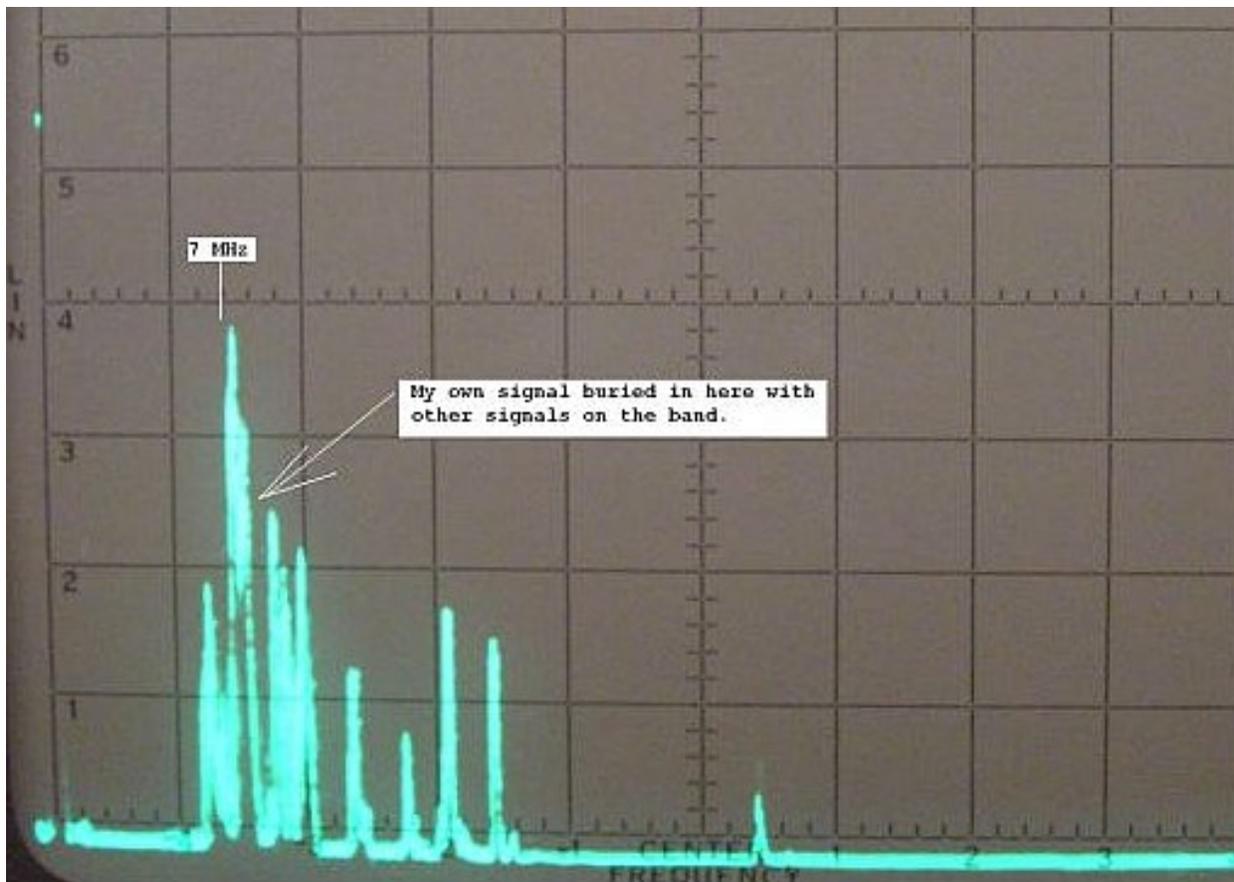
Receive-path insertion loss measurement, key-down. With 600W (+57 dBm) applied to the XMTR port, the signal at the RCVR port measures -80 dBm, as shown on this spectrum analyzer screen shot. The difference, 137 dB, is the receive-path attenuation.



Above, 550W amplifier output, RF keying envelope. As discussed earlier, the QSK switch has nothing to do with the shaping of the keying of the transmitter. That is done at the output of the DDS VFO/exciter. The amplifier, being linear, maintains that same shaping but at 550W output. Between these dits, the receiver is listening at full input sensitivity.

In actual use, my own signal in the receiver varies somewhat in strength as I rotate the yagi antenna, becoming somewhat weaker as the antenna passes through the front/side and front/back null areas. This indicates that at least some of my signal is leaking directly from the antenna, back into the receiver system, around the T/R switch.

An interesting facet of this system is that my own signal is of the same level as the stronger signals on a busy band. The following spectrum analyzer shot was made with the spectrum analyzer connected to the receiver port, while sending continuous dits from my own 550W transmitter, on 40 meters. Antenna is a 2 element yagi. My own 550W signal is lost amongst the others on the band:



### A common question

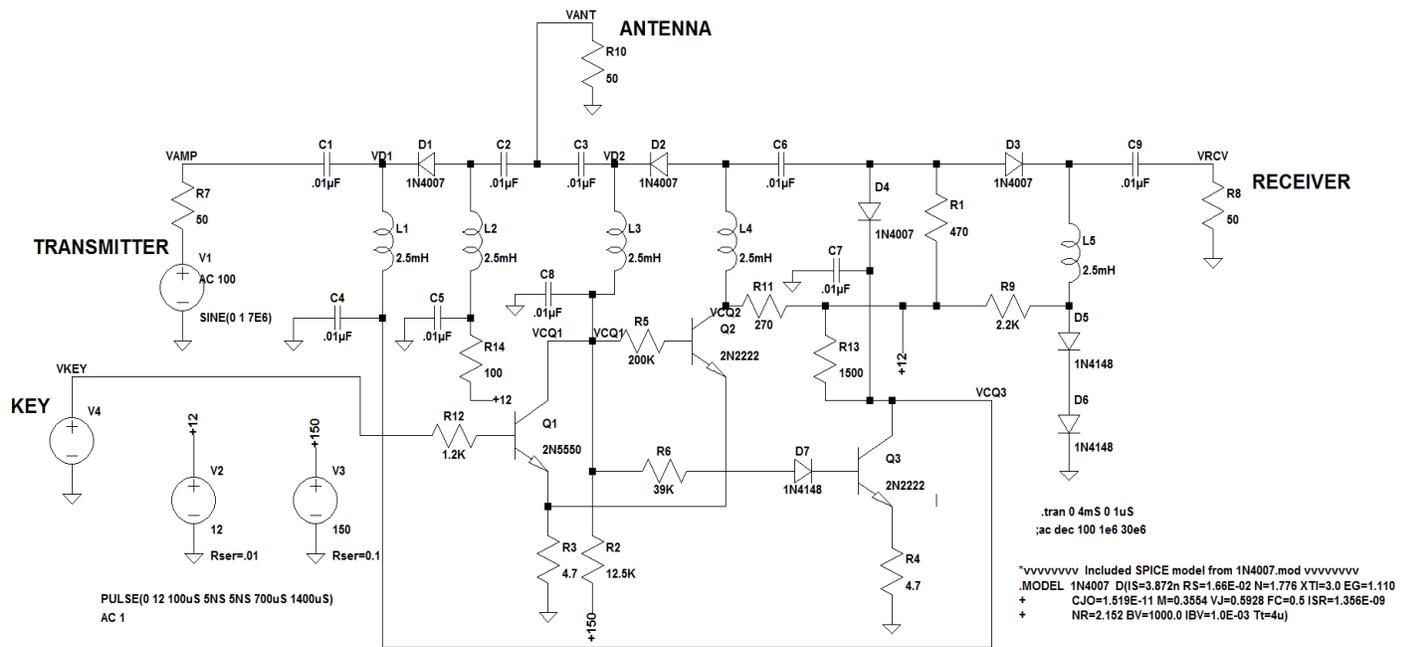
I have had hams on the air ask me that if QSK like this was so easy to do (so inexpensively!), and so useful, why don't commercial rig manufacturers provide it? This is a VGQ (Very Good Question). The answer is: they don't have to. No one asks for it, and the manufacturers have no loss in sales by not providing it. When you buy ham gear you get only what the marketing department of the manufacturer thinks that the majority of their buyers want. This, somehow, is never me. It is perhaps the primary reason I am a homebrew-only op, and I do not buy ham gear. It's actually a very easy decision to make.

Happy homebrewing, and 73,

Don, W6JL

## NOTHING NEW

The use of very inexpensive 1N4007 rectifier diodes as PIN diodes in homebrew QSK switches is certainly nothing new. The schematic below (which is a simulation schematic I have used for simulation of its operation using LTSpice) of a switch using these diodes, appeared in an article in the May, 1976 issue of "73" magazine. That was 37+ years ago. As you can see, the same basic principles used then are used today in my QSK switch. The principles and availability of PIN diodes goes back to the 1950's in the early years of semiconductors. So, there is nothing new about using inexpensive high voltage rectifier diodes as PIN diodes in this way, even down to the exact device, the venerable 1N4007, which is still in production.



1N4007 PIN T/R SWITCH FROM MAY 1976 ISSUE OF "73" MAGAZINE

LTSpice simulation schematic

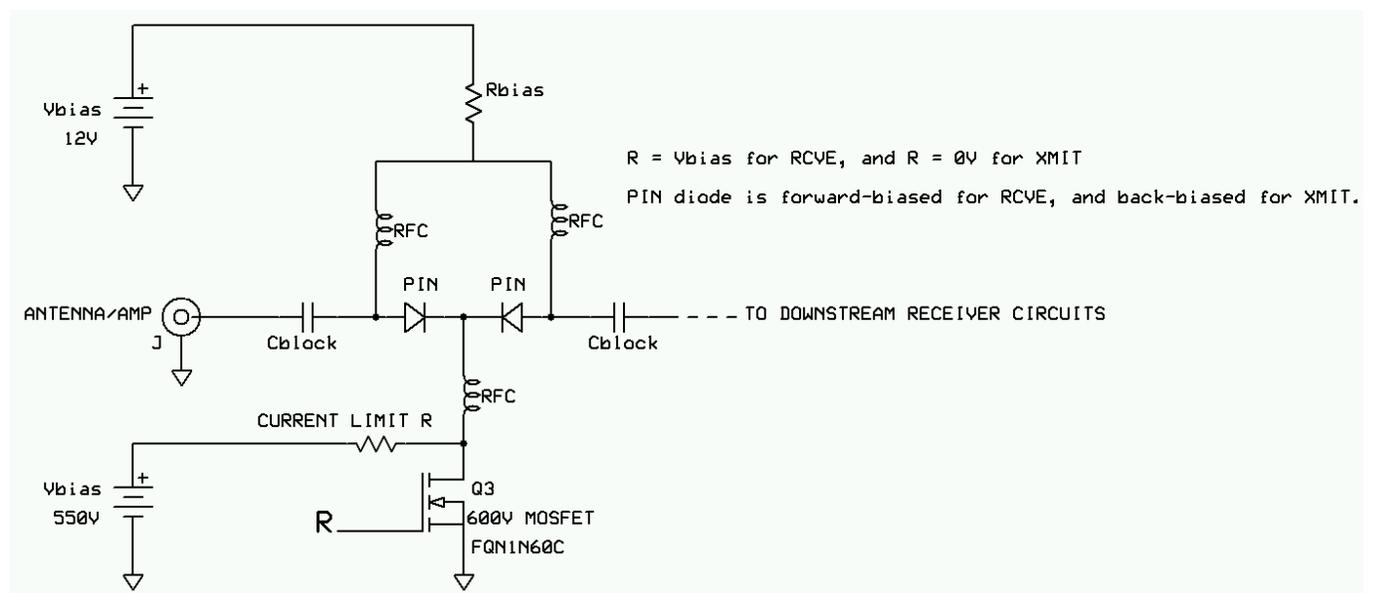
## WAYS OF SWITCHING BIAS FOR PIN DIODE SWITCHES

When we want to forward bias any diode, including a PIN diode, it only requires arranging the circuit so that the desired forward bias current flow from the DC supply through the diode.

See the schematic below. A resistor,  $R_{bias}$ , is placed in series to limit the current through the diodes according to Ohm's Law. The total current limiting resistance is the sum of the bias resistor(s), and the ohmic resistance of any RF chokes that are needed to isolate the RF path from the DC biasing path.

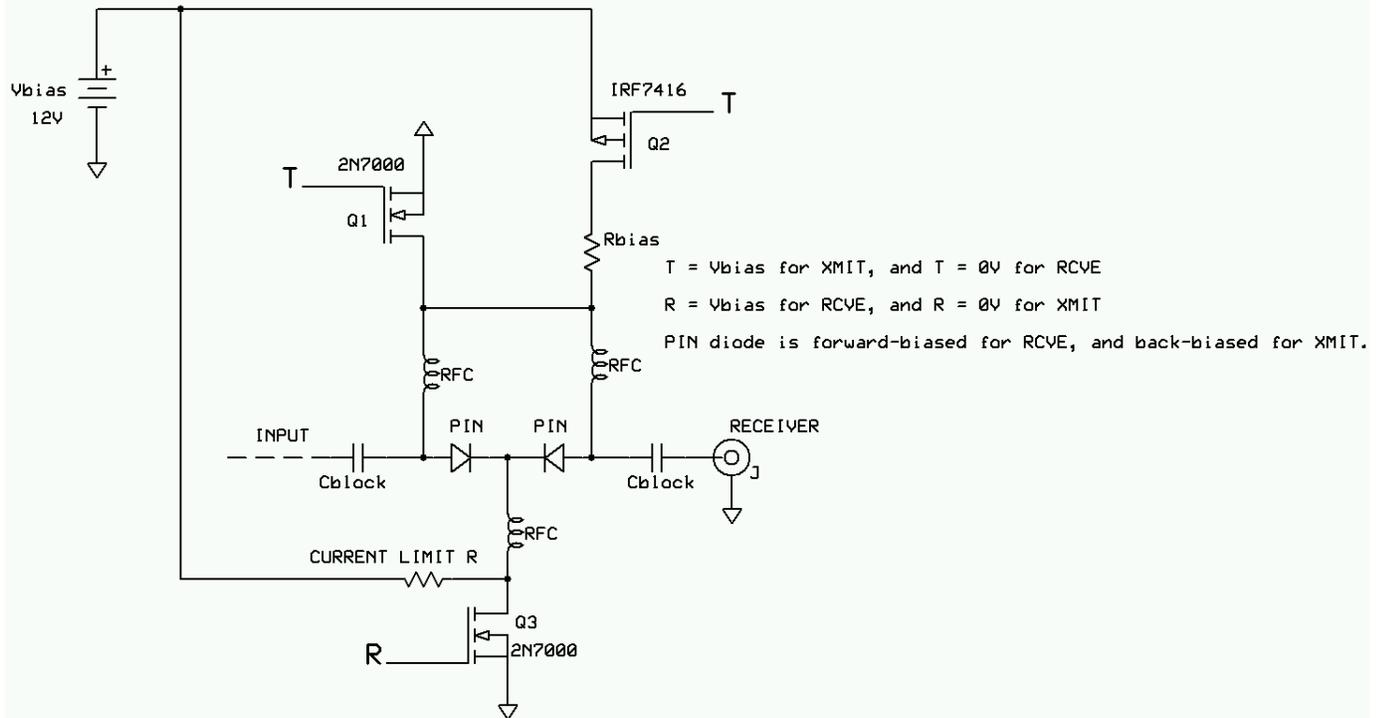
To reverse-bias the PIN diodes, the circuit is rearranged so that the same diodes will now see the desired reverse bias voltage across its junction. The rearrangement of the circuit can occur almost instantly by using the appropriate arrangement of transistors to accomplish the changeover. The choice of the exact arrangement needed depends on how high the reverse bias voltage must be for the particular T/R system.

The reverse bias voltage must be greater than the maximum peak RF voltage that the PIN switch will see when it is reverse biased (ie, OFF). For high power (QRO) applications, this may be several hundred volts. For reverse bias voltages much greater than the DC supply voltage used for the forward bias, it is not necessary to remove the forward bias from the diode. It is only necessary to supply sufficient DC reverse bias, taking into account the presence of the forward bias voltage. For example, if we need 538 VDC reverse bias and the DC supply is 12 VDC, then 550 VDC (the sum of the reverse bias and forward bias supply voltages) is required. See the following illustrative schematic:



For low-level signal paths, such as the signal path to the receiver, only a single low DC reverse bias voltage is needed (except for the first diode in a receive path, which sees the full RF peak voltage applied to the antenna when transmitting). The PIN diode forward DC supply must be turned off, and that side of the PIN diode connected to ground, in order for a reverse bias of voltage comparable to the DC supply voltage to effectively back-bias the diodes. Hence, the circuit requirements differ, depending on the available reverse-bias voltage. If the T/R switch is handling moderate power, such as 100 watts in the transmit path, then the reverse bias need only exceed 100V (the peak value of RF voltage across a 50 ohm load), plus some operating margin. This voltage is much greater than the usual DC supply for forward biasing, so the DC supply can be left in place, and the reverse bias easily increased to account for it.

For those cases where the required reverse bias is low (such as about equal to the DC forward bias supply), then it is necessary for the forward DC bias voltage to be removed (connected to ground) in order to have sufficient net reverse bias across the diode junction when in Transmit. The following circuit accomplishes this for low signal level receive paths, without requiring any voltage level greater than the bias supply voltage, by taking advantage of the complementary nature of N-channel and P-channel mosfet devices (there are no 'P-channel' equivalents for vacuum tubes!). I learned of this clever switching scheme from Jerry, K9JB, who uses it in his QSK system he has designed and built for his Collins S Line. His system provides him with the sweetest (and probably the ONLY) transparent QSK in existence for the S line.



The two N-channel mosfets (Q1 and Q3) are turned ON by a positive gate-to-source voltage equal to  $V_{bias}$ , but the P-channel mosfet (Q2) is turned ON by a negative gate-to-source voltage (ie, by pulling its gate to ground). Turning Q2 OFF is accomplished by raising its gate voltage to  $V_{bias}$ , thus making its gate-to-source voltage = 0. (Remember, all enhancement-mode mosfets require forward bias of gate-to-source in order to enable a low-resistance drain-source path for current to flow).

In the case of a complete QRO PIN T/R switch system, since the high DC bias voltage must be provided anyway, it is often simpler to obtain whatever voltage is needed for the low level receive path PIN diodes, by simply using a voltage divider to reduce the high voltage down to whatever voltage is required to back bias the low level path. For example, a lower voltage is easily obtained from a HV DC supply by using a high resistance voltage divider, say to 32 VDC. This then is adequate to provide 20V of back bias to any/all of the downstream PIN diodes in the receive path, without switching of the 12V forward biasing source. This allows use of inexpensive low-capacitance PIN diodes which have 100V or less reverse bias breakdown voltage ratings.