Making a Heathkit Mohawk RX-1 Useable

By Thomas Bonomo, K6AD

The first ham radio station I saw as a kid was the Heathkit Mohawk RX-1 receiver with the matching Apache transmitter. Those big green high-tech boxes sure made an impression! So, when I spotted a good cosmetic-condition Mohawk to complement my Marauder transmitter, I jumped at the opportunity.

The Mohawk, which was introduced as a kit in 1958, was Heath's first ham band-only receiver, and was the first Heathkit to sport the now-famous "Heathkit green" colors. This receiver offers a number of interesting features. Heath separated the usual single knob RF gain control found on most receivers into two controls, an RF Gain and an IF Gain control. The ability to control the gain of the front-end stage separately makes it possible to reduce cross modulation from extremely strong signals, while maintaining good receiver sensitivity. The Mohawk also had a selectivity switch control allowing bandwidths of 5, 3, 2, 1 and .5 KHz.

I devoted many days to the restoration of my Mohawk. The results, however, disappointed me, so I took to the forums. The views of other owners ran along these lines: "What a junky receiver . . . .." "Terrible AGC . . ." "Worst audio of any receiver I own." "Product detector really distorts." "Drifts too much to keep up with."

The crummy performance of this receiver is probably the reason so few of these receivers have survived despite the great number sold. But with a few "reversible" modifications, it has the potential to be a wonderful receiver. The purpose of this article is to outline those modifications necessary to correct the most glaring deficiencies.

Even if you don't own a Mohawk, the theory presented in this article can be applied to other vintage receivers. The modifications are compartmentalized, so you can evaluate the results as you proceed. Threshold Automatic Volume Control (AVC) is addressed at the end of this article (rather than being grouped with the other AVC modifications), simply due to the larger nature of the topic. Just a note regarding the terms Automatic Gain Control (AGC) and Automatic Volume Control (AVC). The AVC circuitry in the Mohawk is intended to operate as AGC typically does. Since the Heathkit literature uses the term AVC, AVC will be used in this article in lieu of AGC.

Modifications philosophy

Everyone has their own philosophy when it comes to vintage radio modifications. With classic gear, I prefer to keep the original design as intact as possible by implementing only easily reversible "no holes" modifications. This means restricting mods to engineering "refinements" rather than wholesale change: no tube lineup changes, no additional tubes requiring chassis punching, no IF strip chips from National Semiconductor, no mechanical filters from Collins, etc.

I do, however, believe in using a handful of solid state components, when necessary, on the bottom side of the chassis, where they can be hidden from view. Adding some solid state circuitry was unavoidable to achieve the "high priority" performance and usability enhancements I desired, but my goal was to end up with an improved Mohawk, not a Collins 75A-4 inside a Heathkit case.

Design problems

These were the performance issues that annoyed me the most about the Mohawk:

* *Lousy AGC system*. It was too easy to get blasted listening to this receiver.
* *IF GAIN control lacks the benefits of a threshold-type system*. The IF GAIN control cannot be used to effectively reduce background noise and static crashes without also reducing the audio level of desired signals.
* *Severe SSB distortion*. SSB signals were almost always severely distorted unless the IF and RF gain controls were turned way down. Even then, intermodulation distortion was irritating.
* *Severe AM distortion*. Strong AM signals were distorted, which became worse when the ANL was turned on.
* *Useless ANL*. Excessive audio distortion rendered the ANL unusable.
* *Inaccurate S meter*. SSB signal readings were totally inaccurate when compared to a calibrated CW signal.
* *Poor audio fidelity*. The audio had too much bass and sounded muffled. Summertime static crashes on 75 meters sounded more like a train rumbling through my shack than atmospheric noise.
* *Horrendous VFO drift*. On 75 m this radio would drift nearly 3.5 KHz in the first hour in the cabinet (2 KHz when open on the bench)! The drift was much worse on the higher bands.

AVC system problems

The Mohawk's AVC and audio distortion problems are interrelated, and both stem from the fact that the receiver's high gain is not properly controlled by the AVC system. The distortion forces you to turn down the front panel's IF and RF gain controls to clean up the audio. While this helps, it renders an already weak AVC system useless. As you tune across the band or listen to nets you'll frequently get blasted by strong signals. So the Mohawk, as designed, offers you clean audio with no AVC, or lots of distortion with very weak AVC. Turning the AVC switch on and off demonstrates how weak the AVC system is — only very strong signals show any AVC reduction.

If you examine the Mohawk's AVC circuit in Figure 1, you'll notice a number of design problems:

1. Heath used "delayed" AVC, wherein the incoming signal must reach a predetermined level before the AVC detector tube produces AVC output. The purpose of delayed AVC is to allow for a better signal to noise ratio by permitting maximum gain on weak signals. Heathkit set the AVC signal threshold very high (look at the +10 V bias on the cathode), so only very strong signals generate AVC control voltage.
2. Heath only applies AVC to the front-end RF section and the first IF amplifier. There is no AVC applied to either of the second conversion stages (50 KHz IF amplifiers).
3. The Mohawk's AVC time constant is very fast — 10 ms — with no difference between rise and fall times (i.e. no fast attack, slow decay characteristic). The short time constant allows background noise crashes to pop up in between words and syllables.

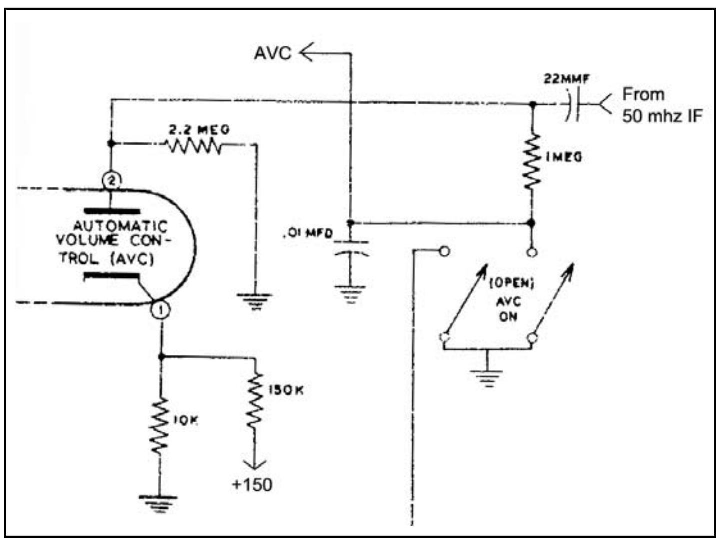

Fig.1. Mohawk AVC detector.

One wonders why Heath designed the AVC system this way. I could find no good receiver of similar vintage in which the AVC system was designed in this manner. What *were* they thinking? However you look at it, the Mohawk's AVC system is just a bizarre piece of engineering.

Fast attack, slow decay circuit modification

The modifications shown in Figure 2 provide a fast attack, slow decay AVC characteristic and increase the time constant from 10 ms to about 3 seconds. D1 rapidly charges C2 to produce a fast attack, while the decay time is controlled by the time constant produced by C2 and R1+R2+R3. The time constant resistors serve double duty as a voltage divider for the S meter amp (discussed later in this article). Changing the value of the charging capacitor C1 from 22pf to 100pf reduces its reactive impedance from 140K ohms down to about 32K ohms (using the formula for capacitive reactance: $X_C=1/2\pi fC$ @ f=50KHz), thereby reducing the time necessary to charge C2.

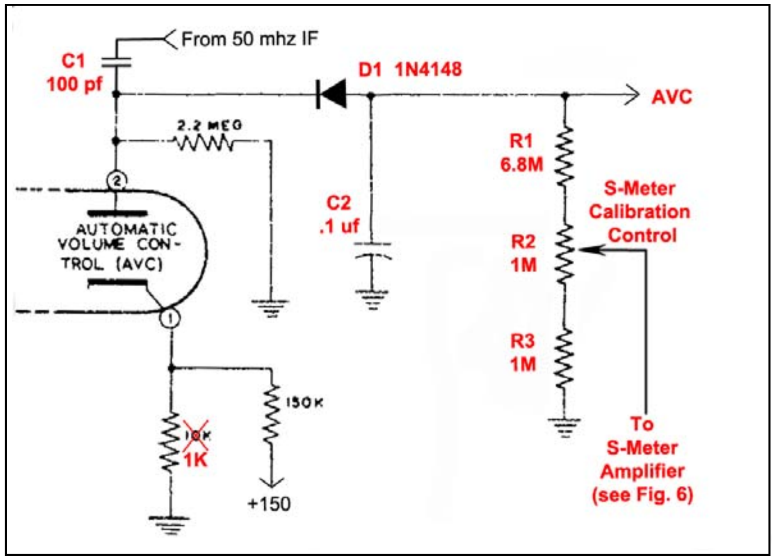

Fig. 2. Simple fast attack AVC circuit. Note added or changed components are labeled in red.

To produce a 3 second AVC decay, I chose R1 = 6.8M, R3 = 1M, and R2 = 1M (R2 is the S-meter sensitivity pot discussed later in this article). If you desire a different response time, change the value of R1 and R3 rather than changing C2 (higher values of C2 increase the width of AVC delay spikes by increasing the time needed to charge C2; lower values lead to AVC overshoot problems, caused when the spikes themselves overcharge C2).

Other AVC rise time improvements

If you test different values of capacitance for C2, it will became clear that AVC voltage rise time is critical. Slow rise times will increase the audible "pop" heard on the leading edge of a strong signal, which is particularly irritating on SSB. To digress for a moment, the effect is similar to transient intermodulation distortion created in an audio amplifier which uses large amounts of negative feedback. As a signal passes through the amplifier, the internal stages of the amplifier operate at full gain until the signal reaches the output where it can then be fed back to reduce the input signal. Until the feedback takes effect, internal stages can reach saturation. When the feedback finally takes effect, the output is reduced. The short interval where no feedback is present produces output "spikes" (which is one reason why audiophiles don't like solid state amplifiers which use lots of negative feedback). The time needed for a signal to reach a specified output level is known as the amplifier's "slew rate" and is usually measured in micro seconds. Low slew rates produce wider output spikes when lots of negative feedback is used.

AVC circuits exhibit a similar effect because receiver stages operate at full gain until the AVC voltage (negative feedback) cuts back the gain. The effect of AVC "kick-in delay" is audible pops heard on SSB signals. The longer the duration of AVC kick-in delay, the worse the pop. At the extreme, the pop becomes a "thump" lasting a verbal syllable or more in duration.

To break down the problem even further, there are two components which contribute to the delay in generating AVC control voltage. The first component is the time delay from the input of the receiver to the AVC pickoff point. The second is the AVC rise time component: how fast the AVC circuit can generate AVC control voltage from the time it first appears at the pickoff point. Ideally, we'd like a receiver with both a minimum delay and an AVC circuit with a very fast rise time. In a good SSB receiver, spikes caused by delayed AVC action are short enough in duration that they are not noticeable (ideally less than 0.2 ms).

The first component of AVC delay in the Mohawk — from the antenna input to the AVC pickoff point at the output of the 50KHz IF transformer — was measured at 0.7 ms. This is more than most receivers, probably because Heath used LC circuits rather than crystal or mechanical filters for bandpass selectivity. There is nothing that can reasonably be done about this component of the AVC delay.

The second component of the problem in the Mohawk — AVC rise time delay — was greatly aggravated when Heath designed the AVC circuit with a super slow rise time (1M X .01µf = 10 ms) and by RC decoupling the AVC voltage to the grids of the RF amp and first IF amp with time constants that are way too large (100K X .01µf = 1 ms). The fast attack circuit configuration addresses most of this problem, but additional improvement can be realized by reducing the decoupling time constant on the grid of the first IF amp, C3, from .01 to .001, and on the RF amp by removing C4 (see Figure 3). C4 is located inside the front-end section on the top side of the chassis. It isn't easy to get to, but the easiest way is to remove the small side panel and clip the lead of C4 which goes to ground on the RF amp (the other end of C4 is very hard to get to without tearing everything apart). If you feel ambitious, remove the RF front-end section from the chassis and replace C4 with a .001 instead of just removing it. I decided it wasn't worth the extra work and instead found an easily accessible place to add C5 inside the RF front-end section. C5 ensures that any RF noise picked up by the long AVC line is not coupled into the RF amplifier. Measurements and listening tests confirmed no difference with C4 removed, but adding C5 is cheap insurance.

As with most AVC systems that utilize high impedances, you'll want to make sure the RF amp and IF tubes don't have any grid emission (easily checked on most good tube testers), because this will shift the AVC line positive and ruin an otherwise good AVC response.

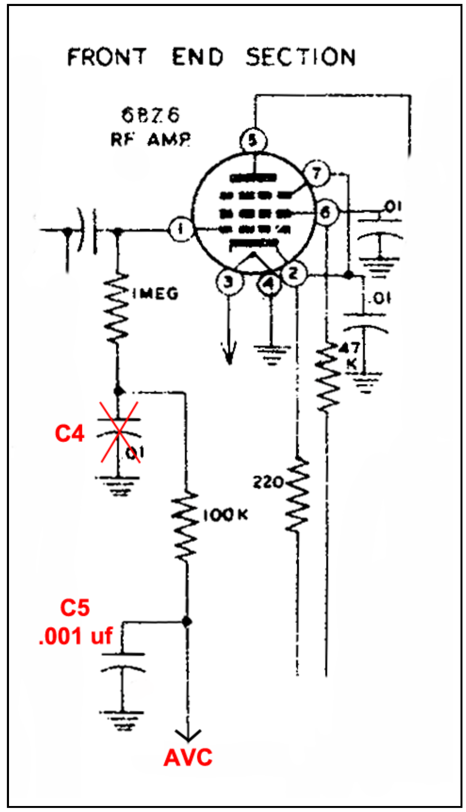

Fig. 3. RF front end

Adding AVC to both 50KHz IF amps

Heath omitted adding AVC control to both of the 50KHz IF amplifiers. Extending the AVC line to the 50 KHz IF amplifier grids dramatically improves AVC action. A small terminal strip can be added near each IF tube using the screw which mounts the IF tube socket to the chassis. On each IF amp, remove the 1M grid resistor from ground and connect it to the AVC line.

The addition of AVC control voltage to all IF amplifiers in the Mohawk appears to be in conformance with accepted design practice at the time. It is hard to understand why Heath omitted AVC from these stages. Certainly the number of additional parts required is minimal, and the improvement in performance is dramatic. Maybe they never listened to many *real* SSB signals on this receiver when it was designed back in 1958.

Adding an AVC clipper and amplifier

There is a tradeoff between obtaining minimum AVC delay (which minimizes the audio spikes caused by delayed AVC action) and clean, undistorted audio. The problem is that you want to use a high value of C1 in order to charge C2 rapidly and therefore minimize width of the audio spikes, but higher values of C2 load the IF transformer while the AVC diode is conducting, which adds distortion that is particularly noticeable on AM. The suggested value of 100pf represents a compromise that produces AVC delay-induced audio spike widths of about 3 ms with reasonable audio. 51pf produces spikes widths of about 5 ms with better audio, while 270pf produces spike widths of 2ms but the audio is much more distorted (especially noticeable on AM). Fortunately, there is a solution to this dilemma.

Adding an AVC amp between the last IF transformer and the AVC circuit will reward you with minimum achievable AVC delay-induced audio spikes and excellent audio. The problem is that the output impedance of the IF transformer is not sufficiently low to drive the AVC circuit, which by its nature needs a low impedance to charge C2 quickly. The amplifier isolates the last IF transformer so that it is not loaded down by the AVC detector while it is charging C2. You could do the same thing using a tube, but it is much easier to implement a reversible mod using an FET, where it can be easily hidden underneath the chassis.

An alternative approach, would be to buffer the output of the AVC diode with a source follower to charge C2. However, the Mohawk lacks a negative supply and rectifying the filament voltage, even with a doubler, does not supply enough

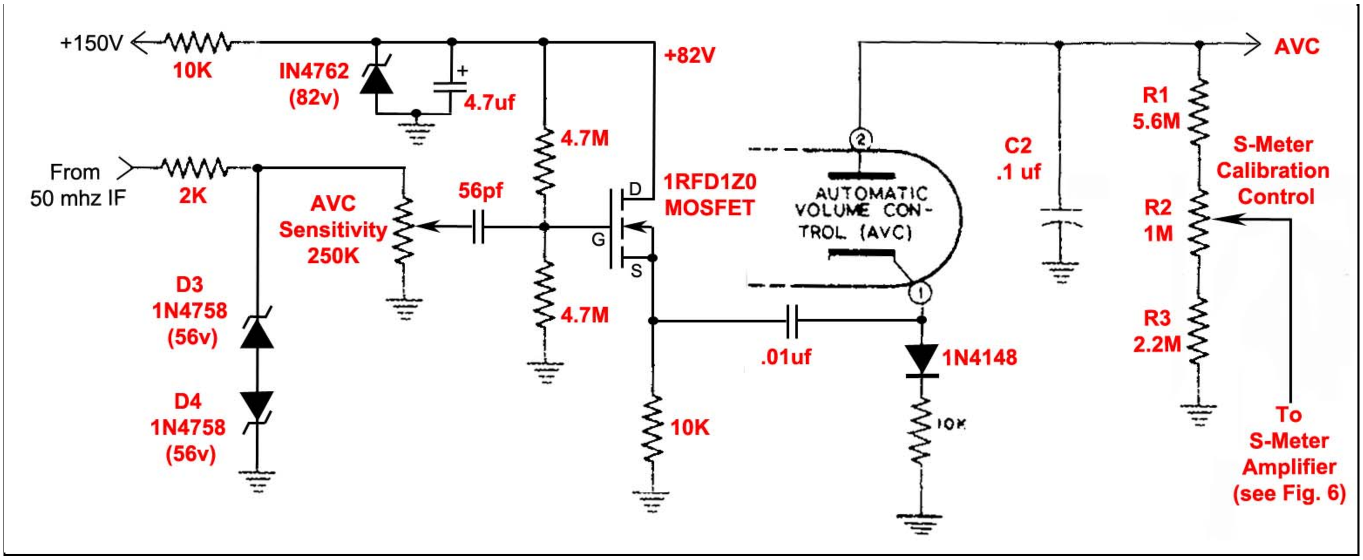

Fig. 4. Fast attack AVC circuit with limiter and amplifier.

voltage to do the job since strong signals generate an AVC voltage of over -28 volts. With circuit drops and headroom, you'd need at least a -35 VDC supply.

Figure 4 shows the new AVC amplifier. It is designed as a common drain, source follower and has a very high input impedance and a low output impedance — perfect for driving the AVC diode. This circuit configuration is largely independent of device parameters so you can use nearly any N channel JFET or any N-channel MOSFET (depletion or enhancement mode) that you have on hand, provided it has a minimum breakdown voltage (BV_{DSS}) of at least 100V. I used an International Rectifier IRFD1Z0 N-channel MOSFET simply because it was the only one I had on hand that met the required breakdown voltage requirement.

D3 and D4 limit the large AVC voltage spikes (which can exceed ± 150 volts!) at the input to the FET amplifier to ± 56 volts, preventing its destruction. The placement of the limiting diodes also limit the spike voltages seen by the AM and product detectors, further reducing AVC-induced audio pops by nearly 60%. As before, if you want to change the AVC time constant, change R1 and R3 rather than C2.

The AVC Sensitivity potentiometer should be adjusted so that strong signals do not reach the limiting point created by D3 and D4. This can be done by listening to a very strong signal (+40 db/S9) on 75 meters (with the IF and RF gain controls fully on) and adjusting the sensitivity just to the point that no audio distortion is apparent. Better yet, if you have the test equipment, inject a very strong signal at the antenna (-27 dBm) and look at the output of the 50KHz IF amplifier transformer. The limiting action at ± 56 volts is easy to see with an oscilloscope. My Mohawk produced an AVC voltage of -26 volts at this calibration point. When you are operating, you will need to back off the RF gain control to avoid distortion for signals stronger than the calibration point you choose. By calibrating the AVC system in this manner, AVC spikes will also be limited to the amplitude of the calibrating signal.

We've attacked the AVC-induced audio spike problem on two fronts: we've made the widths as narrow as possible, and we've limited their amplitude. With the MOSFET amplifier in place, the spikes caused by AVC delay are reduced to about 1 ms duration without any loading of the IF transformer, so the audio will be much improved. This isn't too bad considering that receiver delay alone contributes 0.7 ms of the spike width. Some pops are audible on strong VOX signals, but the overall improvement of the Mohawk's AVC system is dramatic.

S-meter improvements

If you calibrate the S-meter to a CW or AM signal, the readings on SSB will be quite stingy. The problem is that the S-meter is configured to measure the output of the diode detector, and the meter movement is not able to keep up with the momentum of the SSB voice peaks. Since the modified Mohawk (with an AVC amp) already has everything needed to meter AVC voltage directly, you avoid the need to build a peak-hold amp. Instead of feeding the S-meter amp with the output of the diode detector, it is fed from the voltage divider formed by the time constant resistors used in the AVC circuit in Figure 2 or 4. Now SSB voice peaks and a CW calibration signal give the same meter readings.

When calibrating the meter, it makes a difference which band you select for calibration, because the Mohawk has much more gain on the lower bands. I chose to calibrate mine on 20 meters, which means that the meter will be stingy on the higher bands, while the RF/IF gain controls needs to be backed off on 40, 80, and 160 meters for the same reading. Unfortunately, the meter doesn't respond as calibrated on its face, regardless of the circuit used. An S unit on this meter is roughly 4 dB, and depends on signal strength, front panel settings and the band selected. Because the meter is "backward" reading (off=full scale) I failed to come up with a *simple* way to make it more logarithmic. Using S7 (-79 dBm at the antenna input) as the calibration point on 20 meters produced the most satisfactory result on my unit. But give signal reports with a grain of salt.

Product detector improvement

Heathkit's product detector was another poor piece of work — equally useful as a distortion generator as much as a product detector. Distortion was improved by changes to the AVC system, but it was still unacceptably high. With an antenna input signal of -65dbm and the IF and RF gain controls fully on, the output of the product detector was clearly being driven into non-linearity as can be seen in the oscilloscope waveforms in Figure 5. The input signal to the product detector can be reduced by changing the grid resistor on pin 7 of the 6CS6 product detector from 470K to 22K (see Figure 6). Just this one simple change resulted in much cleaner SSB audio, but to my ears it still had high amounts of intermodulation distortion.

Heath tried to do everything in one tube which acts as both BFO oscillator and product detector. There are two problems with this design, both contributing to lousy audio. First, oscillator "pulling" in the tube results in very distorted low frequencies, giving it a rough sound. If you examine the output of the product detector on an oscilloscope, you will see that as the audio goes down in frequency, it changes from a sine wave to a series of spikes. The oscillator is literally trying to "sync up" to the frequency of the incoming IF signal, which begins to distort the waveform at about 600Hz. Second, the BFO waveform at the output of the product detector was a sawtooth rather than a sine wave, which resulted in tons of intermodulation distortion. This gets on your nerves quickly —it sounds like a bad speaker or something resonating and buzzing in the audio.

After much experimentation, I can say I'm not a fan of self-excited product detectors that operate at low IF frequencies like the Mohawk's 50KHz second IF. The best the existing design could deliver was a triangular waveform, which was an improvement over the sawtooth wave, but it still led to the generation of distortion products, so I redesigned the thing.

The new product detector and BFO are shown in Figure 7, which still uses the stock 6CS6. The addition of a MOSFET corrects both problems with Heath's design by separating the BFO and product detector functions. In addition to

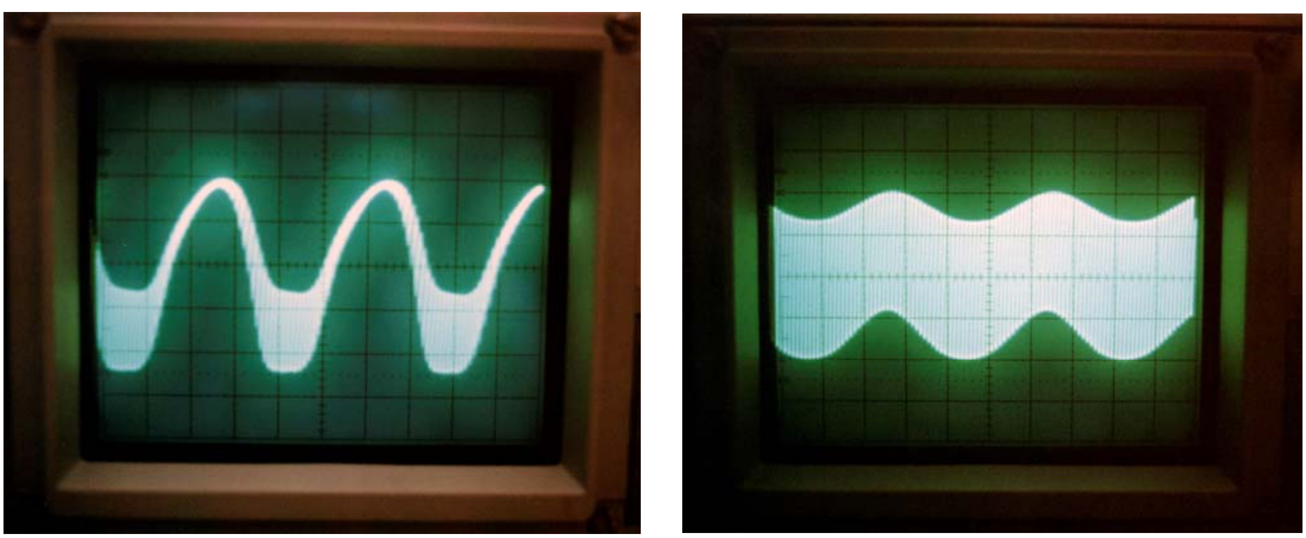

Fig. 5. Product detector output. Reduced input greatly improves linearity. Left trace shows non-linear output; right trace shows improved linearity.

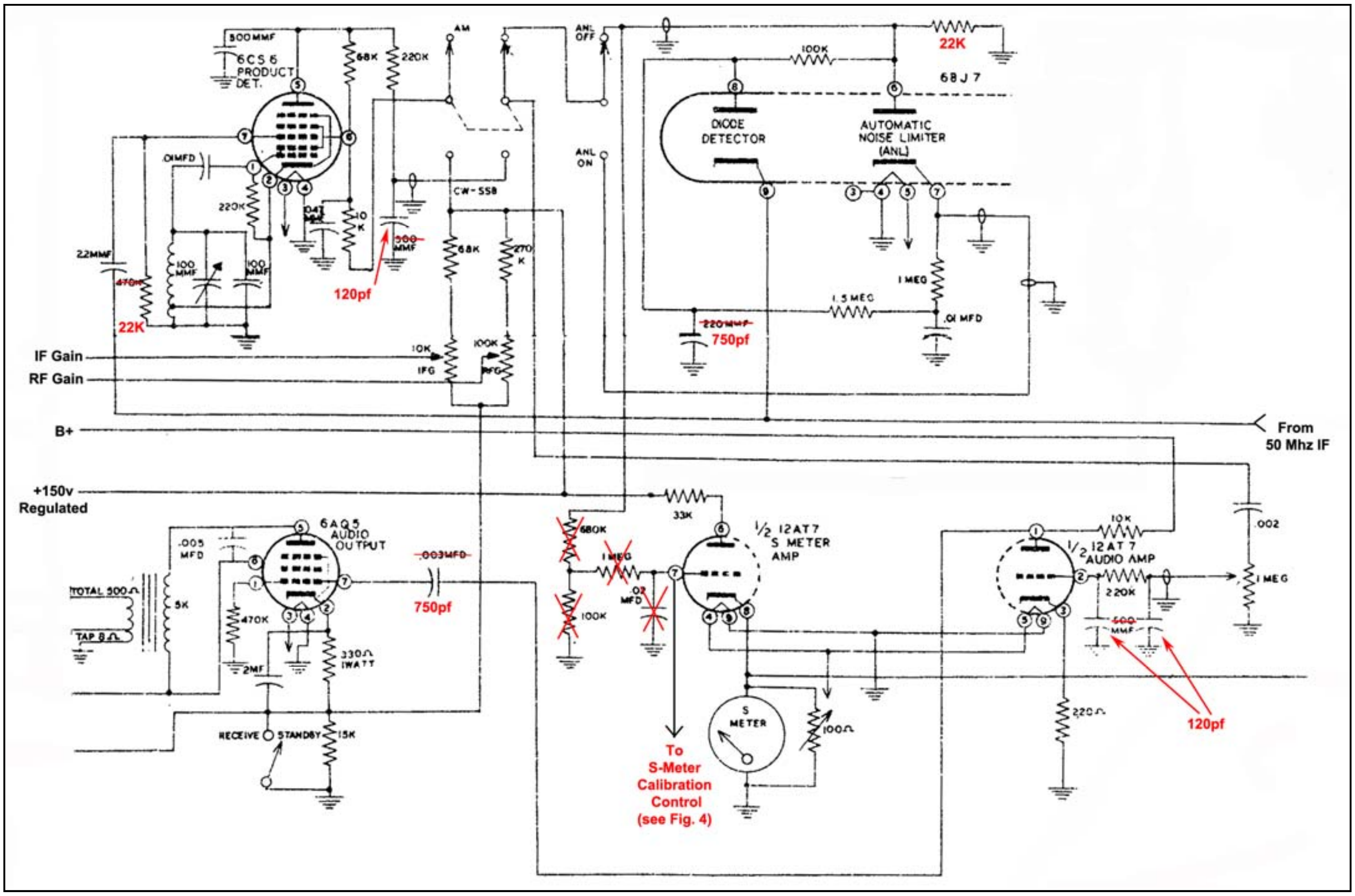

Fig. 6. Product detector, S-meter amplifier, ANL and audio amplifiers.

reducing oscillator pulling, you will notice that the 50KHz output at the plate of the product detector is a nice clean sine wave rather than a sawtooth, vastly reducing intermodulation distortion (a sawtooth wave contains the fundamental and lots of odd harmonics which each mix with the incoming IF, creating unwanted distortion products).

For convenience, I used the same MOSFET type as for the AVC amp, an International Rectifier IRFD1Z0 n-channel MOSFET. You can use nearly any N-channel type you have on hand, provided the breakdown voltage is at least 100V. The 62K biasing resistor may need to be selected depending on the particular MOSFET that you use. Values which are too high will lead to clipping of the bottom portion of the cycle (as seen at the MOSFET's source) while values which are too low will result in oscillator starting problems. I built the entire thing on one 7 lug terminal strip mounted next to the product detector on the bottom side of the chassis (there is plenty of room).

Notice that the input and BFO signal grids have been switched from Heathkit's design. These signals are injected into the product detector tube on different grids to take advantage of the tube's differing cutoff characteristics for each grid to produce a more linear output. The result is a product detector that produces *great* audio.

AM and Automatic Noise Lock problems

There is noticeable distortion of AM signals which becomes completely unacceptable when the Automatic Noise Limiter (ANL) is turned on. This circuit could never have worked as designed. Examination of the AM detector and ANL circuits in Figure 6 reveals a rather standard, but simple design which only works in the AM position (it has no effect in the SSB/CW position). Observation of various waveforms revealed that the ANL circuit was the culprit which was causing AM distortion, even when it was turned off!

Before turning to the solution, it is helpful to first understand the operation of the ANL circuit. As seen in Figure 6, the cathode of the ANL diode remains at an average voltage equal to the output of the diode detector. In order for the ANL

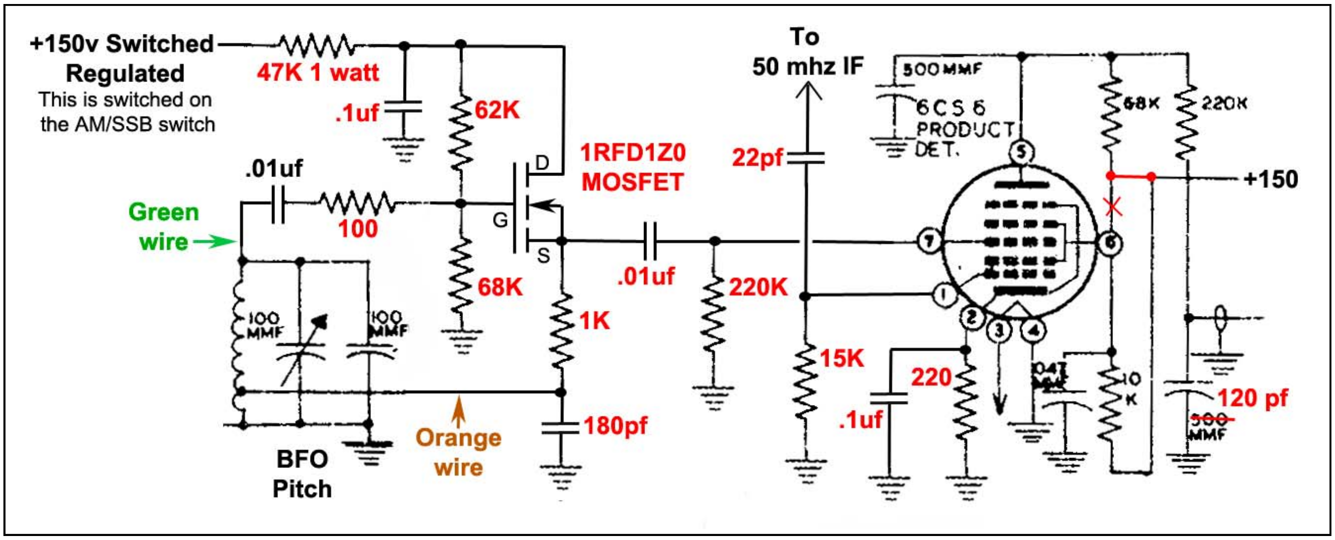

Fig. 7. Modified BFA and Product Detector.

to work properly, the peak-to-peak input signal on the anode of the ANL diode must remain less than the average potential seen by the cathode. The result is that the ANL diode is always on and the cathode voltage follows the anode. When properly designed, if the peak amplitude of the envelope of the received signal exceeds twice the carrier amplitude (100% modulation), as it would in the case of a noise spike, then the potential at the anode will become less than the bias on the cathode, the ANL diode will become non-conducting, and the audio frequency output will fail to reproduce the excess negative voltage.

Take a look at the input voltage divider on the plate of the 6BJ7 automatic noise limiter in Figure 6. The diode detector feeds the ANL plate through a 100K resistor which forms the top half of the voltage divider. The maximum theoretical value for the bottom half of the divider would need to be 100K in order for the plate to see a maximum signal of twice the average carrier amplitude appearing on the cathode (ignoring relatively small source impedances). Note that the bottom half of the divider is formed by the resistors feeding the S-meter amp. Since you know how the circuit is supposed to operate, you can readily see that Heath completely miscalculated the resistor divider on the ANL anode — Heath used 780K for the bottom half of this divider! As designed, the anode gets way to much signal, resulting in the diode turning off during a substantial part of each audio cycle — thus the severe distortion when the ANL is turned on.

Now if you'll remember, there was still a fair amount of AM distortion even when the ANL was turned off. As it turns out, it was the differential loading on the AM detector by the ANL diode as it turned on and off during each audio cycle (which it shouldn't have been doing) which was causing the AM distortion. Make the ANL work properly, and both forms of distortion will disappear. These are easily fixed by making these changes: remove the 680K and 100K series resistors from the plate of the ANL diode which formerly served as the voltage divider for the S-meter amp. Add a 22K from the anode of the ANL diode (pin 6) to ground (much less than the theoretically correct value of 100K due to circuit losses, source impedances, etc.). Change the 220pf on pin 8 of the diode detector to 750 pf. With these changes in place, the noise limiter works surprisingly well considering its simplicity. You'll be real happy with the AM audio too, considering that the Mohawk's configuration really only allows one sideband for AM reception.

How could Heath have miscalculated a bias point so critical to this circuit's proper operation? It is clear that it could never have worked as shown on the schematic, and it would be difficult to accept that Heath's engineers simply couldn't hear this much distortion or fail to notice that the ANL just didn't work. Perhaps the ANL anode divider resistor accidentally got changed when they decided it should also feed the S-meter divider. Or perhaps Heath knew about this mistake and released a service bulletin I've just never discovered.

Fidelity improvements

Now that your Mohawk is producing undistorted AM and SSB signals, let's address the receiver's poor audio fidelity. Take a look at the input filters on the 12AT7 audio amp and at the output of the product detector in Figure 6. Using the formula $f=1/(2\pi RC)$ to calculate the -6db point for audio rolloff reveals that Heath began tapering the high frequency response at 677 Hz! *No wonder* it sounds so muffled. Maybe Heath thought rolling off all the highs made this receiver sound quieter (which it did, but the loss of fidelity). Changing the two filter capacitors on the input of the 12AT7 audio amp from 500 pf to 120 pf and one of the filter capacitors on the output network of the product detector from 500 pf to 120 pf (see Figure 6) produces much crisper audio with a high frequency rolloff beginning at 2.82KHz.

The audio will still have too much bass, which can be corrected by reducing the value of the coupling capacitor on pin 7 of the 6AQ5 audio output stage from .003 mfd to 750 pf, beginning a 6 dB/octave low frequency rolloff at 450 Hz. Now the fidelity is decent.

VFO Drift

The VFO drift in the Mohawk is atrocious: on 75m, nearly 3.5 KHz in the first hour in the cabinet and over 2 KHz when open on the bench! The news doesn't get better on the high bands, where drift is far worse. I'm amazed there is still chrome left on the main tuning knob, since I needed to give it a tweak every few minutes. Leave the room for an hour, and the Mohawk could drift up to a completely different QSO! Heath just didn't seem to pay to any attention to temperature compensating the VFO. Perhaps Heath's engineers only listened to AM signals where drift is not as annoying as with CW or SSB.

As shown in Figure 8, the heart of the Mohawk VFO is a 12AT7 twin triode which combines the functions of a Hartley oscillator and cathode follower. Oscillator "pulling" is eliminated by using half of the tube as a cathode follower to couple the oscillator to the mixer stage. Unlike most receivers, in which the VFO generates the same range of frequencies for each band, the Mohawk VFO generates *a completely new range of frequencies for each band*! With its 1.682 MHz 1st conversion stage, the VFO always operates 1.682 MHz above the desired frequency. This means the VFO must cover a range from 3.4 to well over 30 MHz.

This is why the drift gets worse as you go up to higher bands. If you think of the oscillator drifting a fixed percent with temperature over time, it is apparent that the number of cycles it drifts is directly proportional to the frequency generated. Since SSB reception is sensitive to drift in terms of the absolute number of cycles, not percent, the drift problem in this receiver just gets magnified as you move up to higher bands.

Another problem with Heath's design approach is that the short-term stability of the oscillator becomes worse on the higher bands. On 160 m, short-term drift is barely noticeable. But on 10 meters, within just a couple of minutes you will see short-term swings of nearly 150 Hz. (which in terms of % isn't too bad for a variable oscillator at this high frequency). This made taking measurements on the high bands a real pain.

The hunt for drifting components

I tried heating and cooling various individual components, but everything associated with this VFO seemed to cause it to drift. Since I wasn't making good progress with the VFO mounted in the radio, I disembowel the entire oscillator section from the radio, where it would be easier to isolate and test components.

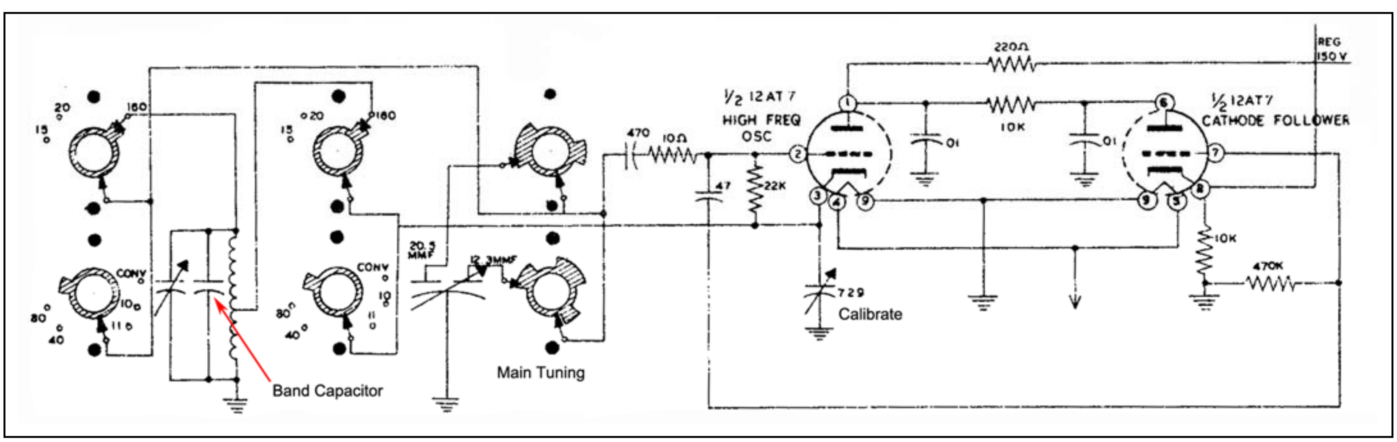

Fig. 8. HF Oscillator.

With the oscillator on the bench, I warmed it up for 2 hours. To ensure accuracy of the oscillator drift tests, the AC line voltage was regulated ± 1 volt. Heating and cooling individual components was now possible without affecting other parts. The following table, derived on the 75 meter band, shows that everything is drifting in the same wrong direction.

Main Tuning Variable	-1.610 KHz	
Calibrate Variable	- .100 KHz	
12AT7 VFO Osc.	- .650 KHz	
Band coil	-1.250 KHz	
Band capacitor	-1.820 KHz	
Total Drift	-5.42 KHz	(wow!)

None of the remaining HF oscillator components are included in the list, but I went thru the gruesome task. The good news is that none of the components in the "nearly-impossible-to-get-to, fully enclosed, soldered-in-place RF front-end" cause much drift.

The VFO tube contributed much less drift than one might expect. I measured the drift contributed by various cold 12AT7 in the warmed-up steady state oscillator. Surprisingly, not all the tubes even drifted in the same direction. While most tubes made the problem worse by drifting down in frequency as they warmed up, several drifted *up* in frequency, at least partially offsetting the drift caused by other components. Out of a batch of 6 tubes, the range was from -650 Hz. to +180 Hz. For drift, GE tubes seemed to be the worst, while Sylvania gave the best results.

Finding compensating capacitors

The remaining sections of this article require temperature compensating capacitors. These days, capacitors with a negative temperature coefficient are not as easy to find as they used to be. The type of capacitor to use is a Class 1 temperature compensating disc ceramic. The temperature coefficient is measured in capacity change in parts per million and carry EIA designations like N80, N330, N470, N750, N1500, etc. These drift in a negative direction with increasing temperature. NPO capacitors are temperature stable and so have no drift.

You may find it difficult to find the exact values with the exact temperature coefficient you need, but you can combine various values. A good source of capacitors (in years past, at least) has been Johnson Shop Products in Cupertino, CA. I'd suggest ordering numerous values so you can easily combine them to get the temperature coefficient you want.

For best results, the top and bottom sides of the chassis must be compensated separately, since drifting components are located in both places and heat reaches them at different points in time. Trying to do it all in one step results in the receiver first drifting one way, and later in the opposite direction. Doing the top and bottom of the chassis separately requires more work, but the results are worth the effort.

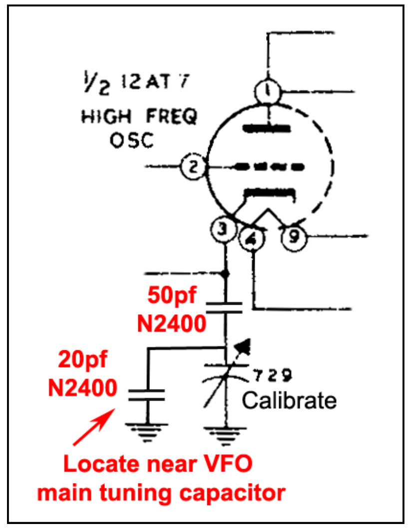

Fig. 9. VFO temperature compensation.

Chassis top side compensation

The problems created on the top side of the chassis were caused by the main tuning and calibrate air variable capacitors having a positive temperature coefficient. The circuit shown in Figure 9 adds negative compensation by adding two temperature compensating capacitors to the calibrate capacitor (at the cathode side of the oscillator). The calibrate and main tuning variables are physically close so this scheme compensates both of them in one stroke.

In addition to providing temperature compensation, this circuit also reduces the range of the calibrate capacitor, which I found excessive and hard to adjust. The original range of the trimmer was 7-29pf, while the range using the compensating circuit was reduced to 17-24pf. The reduced range makes it easier to set. This scheme isn't perfect, because Heath switches in two different sections of air variable, and they each contribute different amounts of drift. In addition, the drift will be a bit different depending on whether the air variable is fully open or fully closed, however, the bulk of the drift is still eliminated.

Experimentation revealed that the circuit of Figure 9 needed capacitors with a larger negative temperature coefficient to eliminate all the drift, but I couldn't find them. The N2400 types just don't quite do the job: the chassis top side drift component is still about -300Hz on 75m and about -1.5KC on 10m which is a big improvement.

Chassis bottom side compensation

The band coils and capacitors for each band are located on the bottom side of the chassis (only one band assembly is shown in Figure 8). Physically, the capacitor is located right on the coil for each band. Heath used precision mica capacitors for each of the bands, but with the wrong temperature coefficient. The band capacitors and coils each contribute to a drop in VFO frequency. Instead of using capacitors which offset the drift contributed by the band coils, the capacitors used by Heath had a positive temperature coefficient which contributed further to the problem. Replacing the band capacitor located on each coil assembly with one that has a small negative temperature coefficient can eliminate nearly all of the drift caused by this assembly. The values which gave the least drift in my unit are shown in the table below. Because this circuit is sensitive, your mileage will vary.

Band	C Required	Band Coil Capacitor (pf)	PPM Units (*C* X *TC*)
160	245pf	128-NPO + 110-N150 + 7-N750	21,750
80	120pf	102-NPO + 7-N750 + 11-N470	10,420
40	270pf	214-NPO + 56-N330	18,480
20	220pf	162-NPO + 47-N330 + 11-N470	20,680
15	245pf	186-NPO + 56-N330 + 3.3-N750	20,955
11	270pf	220-NPO + 47-N330 + 3.3-N750	17,985
10	165pf	80-NPO + 39-N220 + 47-N80	12,340
Conv.	75pf	36-NPO + 39-N220	8,580

On most band coils, I used a combination of three or more capacitors designed to provide the right capacitance at the right temperature coefficient. You will probably get reasonably good results using the values presented in this table, but if you find a particular band still drifts more than you want, you can have some fun fine tuning. It took about 5 hours to compensate each band through tedious trial and error (each test took about 1 hour, and there were usually 4 to 6 iterations for each band). I considered the result acceptable if the band coil assembly contributed less than -0/+300 Hz on the low bands and -0/+800 Hz on the high bands. A small positive drift is definitely better than a negative one as it helps offset the negative drift still remaining on the chassis top side.

Tips and techniques

When heating the band coil assemblies with a hair dryer or other heat source, provide only light heat and be very, very patient. Subtlety should be your guiding principle. This isn't an exercise using a heat gun and freeze spray. Take a systematic, standardized approach. Don't try providing heat by hand, because you'll get inconsistent results. I rested a hair dryer on a pile of books set to the low heat setting, 11" from the assembly. Provide low heat in this manner for at least 10 minutes. A rise of only 20° F will give you the answer you want. When you first begin heating the band coil assembly, the frequency will go up fast. Don't be alarmed — this is because the compensating capacitors heat up more quickly than the heavy ceramic band coil form.

To shorten the time required for the oscillator to re-stabilize when turning off the radio, which can take 20-30 minutes, I soldered the capacitors to the band coils "live". There aren't any high voltages in the band coil section of the radio. Just switch to a band you aren't working on when soldering the capacitor in place. Keep the temperature in the your room constant (you'd be surprised how a few degrees can throw off your measurements). Allow sufficient time for the band coil to completely cool after a heating cycle. Even using a small fan, these things take 25 minutes or more to re-stabilize (they behave as though they have memory).

During bottom side heating tests, the top side of the chassis should be continuously cooled with a fan, so heat won't reach the air variable on the chassis. There is also a 1" X 2" square hole right under air variable which you should plug to keep warm air from reaching the air variable.

It is easiest to select temperature compensating capacitors by thinking in terms of “total part per million units”: PPM Units = (*temp. coefficient*) X (*capacitance*). When iterating during trial and error, it is easier to select one when thinking in terms of PPM Units. This eliminates much of the guesswork. You will find that a change of only 1,500 PPM Units will make a big difference. After selecting the compensating capacitor, use a combination of NPO values to make up the remainder of capacitance needed. It is helpful to have an LCR meter.

The acid test is trying your Mohawk in the cabinet and measuring the drift as before. You will probably find you need to make some adjustments(< 3,500 units) to get it just right (I took measurements every 5 minutes for 2 hours to be sure of the long-term trend). Remember: *drift is directly proportional to how much time and patience you have.*

The easiest way to measure the VFO frequency is to put the 12AT7 HF oscillator tube on a 9 pin extender. Hang your frequency counter on the cathode of the follower section, pin 8. The absolute frequency won’t be accurate because of the extender and probe capacitances, but the drift will be readily apparent. Use a 10X scope probe to avoid killing the oscillator.

Temperature compensating results

To reduce chassis heat, I replaced the 5V4 rectifier with a 5U4 solid state replacement and added a 110 ohm 5W resistor between the rectifier and first filter cap to bring the B+ back down to +225 volts. After making the recommended changes and selecting a 12AT7 for minimum drift, my Mohawk now drifts less than -325 Hz. on 75 meters from a cold start in the first two hours. This is an incredible improvement from the 3.5 KHz on the 75m band before modification. While drift on all of the low bands is now excellent, performance 15m and up isn’t as good — I measured about -1.2KC on 10m in the first two hours (but it was over 16KC before!!) Fortunately, almost all of this drift now occurs in the first 15 minutes. The short-term stability on 10 meters is still relatively poor, but design limitations prevent making reasonable improvements.

Receiver standby muting improvement

Receiver standby muting was another example of Heath not thinking this equipment through as a complete system. When the Mohawk is connected to an external transmitter like the Apache or Marauder, the receiver would come back “numb” after transmitting. If you look at the receiver muting line in the bottom corner of Figure 6 , the mute line is grounded in receive and floats to +44 volts in standby, cutting off the IF and RF tubes through the front panel gain pots. The mute relay contacts in the transmitter similarly ground this line in receive and let it float during transmit.

The problem is that Heath connected the transmitter’s mute relay contacts and the standby switch on the Mohawk in parallel when they should have been connected in series. The Mohawk’s standby switch grounds the line all the time in the receive position, allowing the receiver circuits to remain fully active while you are transmitting. Afterwards, the receiver will remain numb until the AVC voltage decays. The mute line from the transmitter is thus rendered useless.

You have two choices to fix this problem. The easy solution is to just leave the standby switch on the Mohawk in the “standby” position when using an external transmitter. The external relay contacts can then properly mute the receiver. The disadvantage is that the standby switch doesn’t work as labeled and you have to remember to keep it in “standby” which is non-intuitive.

The second option requires rewiring the standby switch. The benefit of rewiring is that the standby switch will work like it should have to begin with. Simply remove the ground from one side of the standby switch, and reconnect the wire from pin 6 of the accessory socket to the switch terminal from which the ground was removed (the switch is now in series with the external relay contacts via pin 6). When you disconnect the transmitter from the Mohawk’s accessory socket, you will need to plug in a jumper socket with pin 6 connected to pin 1 (ground). Now everything works as it should.

Threshold AVC theory and benefits

Threshold-type AVC systems can be applied to many boatanchor receivers with a noticeable improvement in performance. Manufacturers made the transition to the more modern type of AVC system at different times during the late 50s. The Collins 75A-1, and National HRO-50 and HRO-60 are examples of receivers that use the older linear-type

AVC system, and they benefit greatly from the simple modifications required to convert them.

A threshold-type AVC control system is employed in every modern ham receiver and is easily recognized by the S-meter reading that increases as the RF GAIN control is reduced. If adjusting the RF GAIN control on a receiver has no effect on the S-meter's reading (other than to reduce the signal strength of incoming signals), then that receiver likely employs a linear-type AVC control system.

If, on the other hand, adjusting the RF Gain control raises the S-meter a fixed amount as the RF gain control is reduced, then the receiver employs a threshold-type control system. This simple observation works because in most receivers the S-meter measures the AVC control voltage line (or its proxy) which feeds the r.f. and i.f. stage grids. The S-meter thus gives you a visual clue as to which type of system has been implemented. But the more important benefit of a threshold-type AVC system is that the RF GAIN control can be used to reduce background noise and static crashes without decreasing the volume level of desired signals.

The RF GAIN control on a receiver with a linear system will reduce everything — both signals and noise — in a linear fashion. The effect is not much different than using the volume control, and is usually only used to reduce strong-signal overload. It isn't effective at eliminating background noise because it tends to reduce the volume level of desired signals as well.

Receivers that employ a threshold-type RF GAIN control provide a threshold, or signal floor. Signals and background noise that fall below the manually set threshold are all reduced in direct proportion to the setting of the RF GAIN control. This can be observed by noting that signals that fall below the manually set threshold do not cause any movement of the S-meter, even though you can still hear them. Signals that are strong enough to exceed the manually set threshold, however, will generate sufficient AVC voltage to further reduce the receiver's gain by an amount that exceeds the threshold. These strong signals will cause the S-meter to kick above the manually set threshold. The threshold- type system is inherently nonlinear. Below the manually set threshold, there is effectively no receiver AVC, while above it there is.

Because of the threshold effect, the RF GAIN control can be used to reduce background noise and static crashes, while allowing strong signals to come through without noticeable reduction in volume. Receivers that use a linear-type AVC system cannot offer this advantage, because there is no threshold—all signals are reduced by the same amount.

Common circuit implementations

There were two different approaches engineers used to implement the RF GAIN control. The primary difference lies in how the AVC control voltage and the manual control voltage are fed to the r.f. and i.f. stages. When the RF GAIN and AVC control voltages are <u>each applied to a different tube control element</u>, it results in a "*linear-type*" control. Accepted design practice until the mid-50s was to apply the RF GAIN control voltage to the cathodes of the gain stages, while applying the AVC voltage to the grids of those same stages. The effect of feeding the AVC and RF GAIN control voltages to different tube elements is that they combine within the tube in a linear manner with respect to the output of the stage.

The other way of implementing gain control is to mix the dc control voltage from the RF GAIN control with the AVC

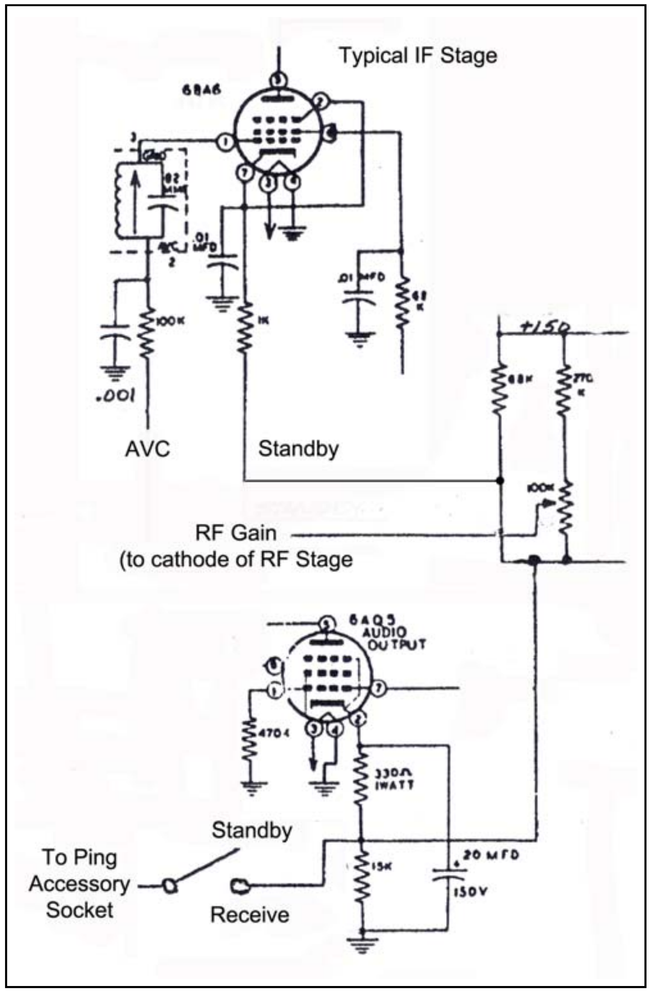

Fig. 10. Stock Mohawk uses different tube elements to control stage gain. The problem is that this results in a logical "AND" function, so there is not threshold effect.

control voltage from the AVC detector and then apply the mixed voltage to just one tube control element in each gain stage. This produces a "*threshold-type*" system. Accepted design practice beginning sometime in the mid-50s was to apply the mixed control/AVC voltage to the grids of the gain stages (no control voltage applied to the cathodes).

It is useful to think about the two different systems from the perspective of analog logic to understand why the systems sound so different. First, consider the linear-type AVC system. The linear-type circuit produces a logical 'AND' within the gain control tubes. In this case, the output is directly proportional to both the AVC voltage AND the manual RF GAIN control. Unless a signal is very weak, the receiver is always under the effect of the AVC voltage being generated.

Next, consider the threshold-type system. One of the most common approaches produces a logical 'OR' circuit. In this case, the control voltage will be either the AVC voltage OR the threshold provided by the RF GAIN control, whichever is greater. The AVC-generated control voltage has no effect below the manually set threshold.

Threshold AVC for the Mohawk

It is important to distinguish between receivers that have a single RF GAIN control, and those that have separate RF and IF GAIN controls. The gain of the r.f. and i.f. stages in most receivers are controlled from a single RF GAIN control to reduce cost. Heath's dual-control implementation did not capitalize on the advantages that separate RF GAIN and IF GAIN controls can offer (both systems were linear types).

Take a look at Figure 10 and you will see that the AVC control voltage was fed to the grids while the RF GAIN and IF GAIN control voltages were fed to the cathodes of their respective stages. The optimum implementation would have been to design the IF GAIN as a threshold-type control while leaving the RF GAIN as a linear-type control.

There are advantages of using a mixed configuration of control systems. The threshold-type system will allow the IF GAIN control to offer the advantage of being able to reduce static crashes without reducing the volume of stations which exceed the threshold. Retaining a linear-type control system for the RF GAIN control is handy for cutting down on front -end overload.

Figure 10 shows the RF GAIN and IF GAIN controls prior to modification, where the control voltages each feed a separate tube elements. Figure 11 shows a threshold-type circuit, which is composed of two distinct parts: a negative bias supply and a logical OR circuit. The negative bias supply is a voltage doubler consisting of R1, C1, C2, D1 and D2. Using the filament supply of 6.3 VAC, it will provide about -14 VDC. On one half of the AC cycle, C1 charges up through D1 to about -7.7 volts. During the next half of the AC cycle, C2 charges up through D2 to about -14 volts. The negative bias supply can be built on a small terminal strip.

The voltage doubler's -14 VDC is sufficient to substantially reduce the receiver's gain, but not enough not completely cut off the r.f. and i.f. stages. It would have been nice to have more negative bias, but the -14 dc supply allows the signal sensitivity threshold to reach +30db/S9, which is sufficient for most listening situations.

The actual threshold circuit is simplicity itself. The DC output of the IF GAIN control feeds diode D3, which is attached directly to the AVC line. As shown in Figure 12, it is diode D3, in conjunction with the 6BJ7 AVC detector diode, that forms the critical logical OR function. Whichever voltage is most negative, either from the IF GAIN control or the AVC

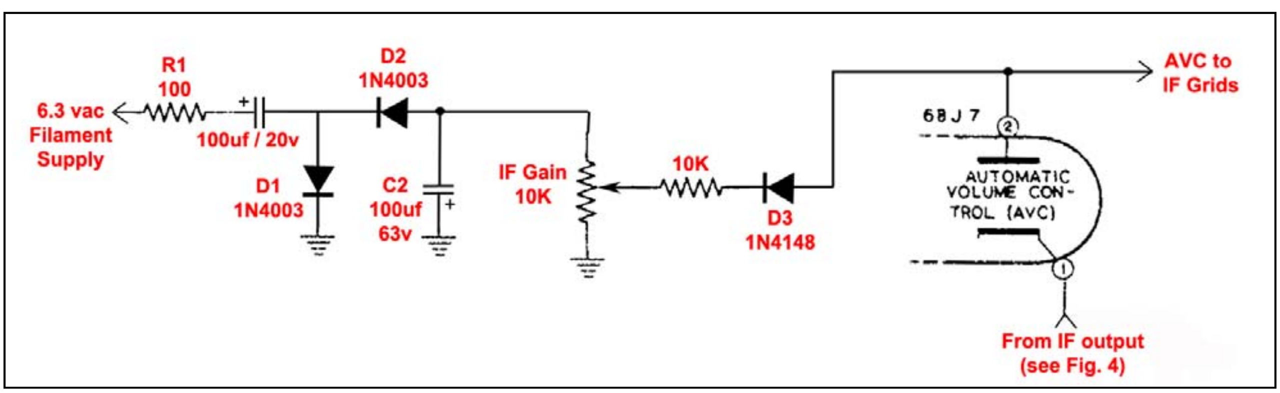

Fig. 11. Modified Mohawk combines AVC and IF Gain controls using a logical "OR" function to control only one tube element, thereby providing the desired threshold element.

detector, will be fed to the gain stages. Because the S-meter amp is attached to the mixed AVC line feeding the control grids, it will register the manual setting of the IF GAIN control.

The r.f. front-end stage shown in Figure 10 is a linear-type control system *only* with respect to the RF GAIN control. The other control voltage it receives is the mixed AVC/IF GAIN control voltage, rather than raw AVC, further enhancing the threshold effect.

Note that the line which fed the cathode of the 1682 KC i.f. amplifier (and a section of the selectivity switch) from the wiper of the IF GAIN control is now tied directly to the standby switch circuit. When the receiver is switched to standby, the cathodes of the 1682 KC i.f. amp and the audio output stage rise to about +40 volts, thereby cutting off these stages. Don't forget to trace the wiring from the IF GAIN control and the Standby/Receive switch to make sure they are wired correctly.

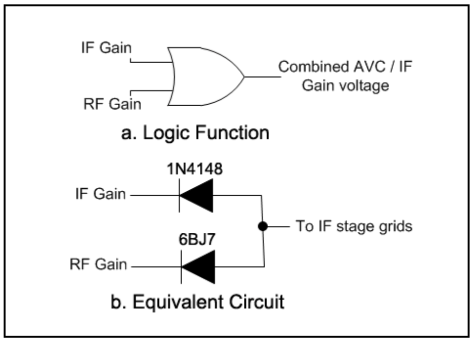

Fig. 12. "OR" implementation uses the original 6BJ7 detector diode and added solid state diode.

Schematic PDF with mods available

A modified Mohawk will offer surprising sensitivity, and the improved audio quality makes listening a joy. While it's selectivity won't match receivers with mechanical or crystal filters, the front panel's adjustable selectivity is convenient. The 5KC setting offers rich audio when band conditions permit. AM signals sound good, too, despite only one sideband being present.

I have scanned the modified Mohawk's schematic into a PDF. Email me if you'd like a copy. The schematic includes an audio feedback mod that contributes to an even smoother sounding receiver. This mod consists of a 1.5 M ohm resistor in series with a 0.47 uf capacitor from the 8 ohm tap of the audio output transformer to the grid (pin 2) of the 12AT7 audio amp. There is a modest reduction in audio gain, but there is more than enough in reserve to still blast yourself out of the shack.

Enjoy modifying your Mohawk, or applying these principles to other vintage receivers!

Tom is licensed as amateur extra K6AD, earned a B.S. in electrical from the University of Colorado, and was formerly an audio and video design engineer for Ampex Corporation

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