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Re-greening the Heathkit IG-18: IG-18 #1
Revisiting and improving a favorite project from many years ago

The **Heathkit IG-18 Sine-Square Audio Generator** (and the IG-5218, SG-5218 and other Heath/Zentih/Schlumberger variants) is a classic piece of test equipment that gave many audiophiles and techies their first experience of a decent quality, wide-frequency-range, low-distortion sine-wave oscillator at an affordable price. I bought and built one when Moses was still alive, then watched through the years as people worked various changes and some improvements on the basic design.

The IG-18's strengths:
1) A simple and elegant design with great economy of parts and generally good-for-the-time performance, designed at the mid-morning of Solid-State.

2) The Bridged-T frequency selector/filter circuit has four decade ranges and uses only five tight-tolerance capacitors to do the work of the 8 that would be needed in a Wien Bridge type of filter, or the 12 needed for a Twin-T type. This design combines economy (tight-tolerance capacitors are expensive) with decent selectivity. The filter only needs 5 caps because the two caps for each range have values that are in a 10:1 ratio. This means that the bridging cap for one range becomes the pillar cap for the next, etc.

Because the caps in the filter have a 10:1 ratio, the notch depth (which becomes a peak in the amplifier response, because the notch is in the negative feedback loop) is about 15-16dB, depending on the exact ratio, or a factor of about 6X (for exactly matched resistors and an exact 10:1 ratio of capacitors, the notch attenuation = 2/(2+ratio) = 0.1666... = -15.6dB, and the peak is then 15.6dB. This formula works for any ratio of caps as long as the two resistors have equal values). This means that the positive feedback loop needs to set the amplifier gain to approximately the same 15.6dB amount in order for oscillation to occur.

3) The use of an incandescent lamp for AGC (automatic gain control), as pioneered by Bill Hewlett, gives really amazing performance compared to other costlier and more complicated schemes -- one simple part replaces many other parts and works about as well. The lamp has a positive temperature coefficient -- as more current flows through it, it gets hotter, and its filament resistance increases. In the IG-18, the lamp is the feedback leg of the positive feedback loop, with the ground leg being a 750 ohm potentiometer that coarsely adjusts the amount of feedback; the lamp's resistance variation with output level supplies the fine control. This maintains the operating point of the system at a good place for both stability and (relatively) low THD.

4) The two discrete frequency selector switches, combined with a ganged-potentiometer vernier, provided frequency selectivity that other commercial gear of the time didn't -- most of them were capacitance-tuned, like the tube-type HP 200CD, with a single dial; they could be finely tuned but generally not quite as finely as the Heath design, where the vernier control gives a total span of 1%.

The IG-18's weaknesses:
As chronicled by Reg Williamson, writing in the Audio Amateur magazine in 1971 (the PDFs of the article are here -- please read his article about the mods), the original IG-18 had many problems. Reg was the first I'm aware of to "green" the IG-18, and his comments and mods led Heath to make some design changes after the article appeared, and to make more changes in the SG-5218. The weaknesses Reg

identified were/are (some were fixed by Heath):
1) No "zero" or off position of the vernier control, leading to frequency errors of the output compared to the dial markings (on the old originals, this can be fixed by disassembling the vernier pot and knifing, filing, or dremeling a break in each of the pot's resistance elements near the ccw end, so that the wiper loses contact with the element, making an open circuit). Heath fixed this promptly.

2) Several incorrect part values, including particularly one resistor and one capacitor that contributed to poor distortion performance. These were not fixed for a long while.

3) Insufficient power supply filter capacitance.

4) Early units had a poor grounding scheme for the square-wve output which led to distortion and noise problems in the sine-wave output from the square-wave circuitry.

5) No buffering of the meter circuit from the main output, also leading to rectifier spikes and higher distortion. Heath never addressed this issue.

6) Generally insufficient open-loop amplifier gain, leading to poorer THD.

Reg Williamson's greening (reportedly) fixed most of these problems, and he improved the amplifier circuit by making a Darlington pair for the voltage amplifier/output driver stage Q3, which (supposedly) significantly improved the overall performance by increasing the open-loop gain.

The plan:
I'm going to take an IG-18 and revisit the greening a bit at a time, and then do a bit more, to see how far I can go without completely redesigning the unit (I'm also going to completely rebuild another IG-18 using Bob Cordell's outstanding state-variable oscillator design from the '80s, for which I've done a PC board layout -- I'll build a page for IG-18 #2 when it's finished. The Cordell design is capable of THD levels below 0.0005% (5ppm) at 1kHz. Many have done other designs using some of the IG-18's parts; Morrey's rebuild, detailed in the Audio Amateur, was very extensive but there were reportedly some problems with it as well.

Why do this?
Simple pleasure. Experimenting is fun, and making things better is satisfying. I'm also interested in the results that Reg got -- I think they may be a little too good, given the lamp AGC system and the relatively low-Q filter.

The measuring gear used to evaluate performance and modifications:
-- An HP 334A THD Analyzer, with a distortion + noise measurement floor of about 0.01% and an ability to measure distortion and noise products beyond 5 MHz. This is a very wide-band instrument that does not have any low-pass noise filtering built-in; it does have a 400Hz high-pass filter which is effective for reducing or eliminating hum products.

-- A Core 2 Duo 2.67GHz PC running the ARTA Spectrum Analysis software that has an awesome set of features and controls. Input is through the on-board Intel HD Audio chipset, which in my experience has a THD-only floor of around 0.0018% (18 parts per million) or better, and a maximum sample rate of at least 96kHz and maybe even 192kHz, and 24 bits of resolution. The ARTA software can measure the true distortion only, by computing the RMS sum of the multiples of the input signal, and/or can measure the

RMS THD+noise -- very, very nice. This setup will give very good evaluations throughout the audio range, and a bit beyond, with 5+ times the distortion resolution of the HP 334A.
-- A nice Tektronix 7603/7A26/7A22/7B53 100+MHz analog oscilloscope.
-- An HP 3458A 8-1/2-digit DMM. They don't get any better than this.

Test levels:
Output levels will be at a full 10VRMS -- if the unit is stable at all frequencies -- into a 10k ohm load, with attenuator and fine amplitude both full cw (or nearly so) to get the 10V level. Past experience for me has been that with its low impedance output stage, this unit does not care very much about a 600 ohm load -- the output level with the attenuator set to full 10V output drops a couple of volts due to the 150 ohm internal resistor in series with the output attenuator, but otherwise, nothing else changes.

THD levels will also be checked at 1VRMS by setting the attenuator to 3V (+10dBV) and adjusting the fine control for 1V output, also with a 10k ohm load -- this tends to show up hum and noise problems. The bias control will be set for symmetrical clipping of the output waveform or for best stability or for lowest THD, with a best-case balance among the competing alternative settings.

The feedback control will be set for the desired output level with lowest distortion while also maintaining stability at every output frequency from 10Hz to 111kHz. I'm not that interested in frequencies below 10Hz, but if it's stable there, it most likely will be stable at 1Hz too. Then I'll try a feedback setting that gives 5VRMS output to gauge the effect of lower voltage and current on the THD.

The steps:
1) If not already done, fix the grounding for the outputs and level controls, and
  a) Add Reg's emitter-follower meter buffer (albeit with a small change).
  b) Fix the vernier pot if it needs it (it didn't -- this is a later model than I thought).
  c) Depending on spikes present in the THD+noise signal, this may be the time to add a switch to disable the square-wave generator section. It's probably best to just separate the power line for the Schmitt trigger from the sine generator and put a switch in that. One could just switch off the input signal, but that would leave one of the Schmitt trigger transistors in full conduction -- a waste of power.

2) If not done by Heath, and it is needed, increase power supply filtering, then
  a) stiffen regulation by adding a Darlington transistor to the regulator pass transistor.
Possible mods here would be to either
  b) change the value of the zener diode and make the second transistor a voltage gain stage, rather than just a Beta increaser; or
  c) use an LM317. We'll see.

3) If not already done,
  a) change the value of R3 to 56k, add a 500uF or larger cap around R11, and increase C5 to 10 uF (Reg's value) or 100uF (Heath's value in the SG-5218).
  b) Then maybe Beta-match Q1 and Q2, substituting as needed.
  c) Change Q4 and Q5 to higher power devices.

4) Change Q3 to an MPSA64 Darlington transistor (the alternative is to do what Reg did, and splice in a 2N5087).

5) Add a 100 ohm pot in series with D1 and D2 to adjust the bias of the output stage and see if there is any improvement in THD due to lessening of crossover spikes in the output stage. This may require higher-power devices for Q4 and Q5.

6) a) Replace R4 with an N-channel junction FET connected as a constant current source (with a 5k ohm pot from source to ground) to adjust the current for the differential pair of input transistors, or
b) try using FETs for Q1 and Q2.

7) Do the same as 6) for R12, the 10k ohm load for the voltage amp stage.

First, the original IG-18:
Starting point -- the original design as built by Heath, in a unit recently bought on eBay. Serial Number looks like 09308 but it's hard to read.. Like my first IG-18, this is one of the old "brown" ones with brown and tan knobs, panel, and covers. The black metal end caps and handles that hold the covers on have been replaced with black-painted pieces of wood. The wiring is nicely done and looks a bit more factory than a typical kit, but there's plenty of rosin residue, so it was a kit.

All of the sine oscillator parts are original as per Reg's example -- R1 is 4.7k; C1 is 0.68uF and is a Tantalum epoxy-bead unit. The lamp has the markings SYL 90V -- it appears to be a 90V @ 30mA bulb; a currently available equivalent is the type 90MB with miniature bayonet mount. This is the first IG-18 I've seen or heard of that has the lamp type actually on the lamp. Heath hasn't ever, to my knowldege, disclosed this information.

The Schmitt trigger has a 2.7k resistor added to the bottom of the circuit board paralleling R26, 1.6k, the collector load for Q8, the square-wave output transistor. In addition, there's a 5pF disc cap collector to base on Q8, and a transistor connected as a diode around C8, the 250uF coupling cap feeding signal to the base of Q8. The square wave just doesn't look that great to me, with lots of edge issues and a best risetime of 20nsec. I'm going to undo those mods to go back to square 1.

THD, 1kHz, 10V = 0.12%, independent of load. Nearly all 2nd Harmonic. Contrary to my past experience with two other units, there's no sign of hum or EMI problems. THD, 1kHz, 1V = 0.1%. There is some spiking/notching in the distortion waveform, with some odd order products showing up. No hum at the lower level either. The stability is marginal below 1kHz, but OK above. The feedback control setting is very dicey. All this is most likely caused by the **way** too-small 0.68uF decoupling cap for the feedback.

Step 1) I think the grounding scheme used by Heath is exactly backwards. For minimum hum and noise, the square-wave attenuator/output ground wire needs to be connected where the center-tap of the power transformer connects, and the sine-wave attenuator and output ground wire needs to be connected to where the square-wave attenuator's wire was connected -- on the ground trace at point L.

With the grounds fixed, the notches in distortion are <u>not</u> from the square-wave generator. I dremelled the supply traces to the Schmitt trigger and ran a power lead over to the sine generator section, disabling the square-wave generator. THD @ 1V unchanged.
a) Added the meter buffer and spiking disappeared completely, THD @ 10V = 0.1% -- not much improvement. Will have to add the meter damping capacitor at some later point, and possibly Reg's meter compensation network.
b) Vernier pot fine, with a true zero position.
c) A switch for the square-wave generator will come later, if ever.

Step 2) Checked power supply ripple with stock 300uF/60V caps; the output voltage is 39.8V, a bit low -- there is no ripple in the output from the pass transistor, but there is 40mV p-p of output signal. coming back from the sine-wave oscillator.

a) Adding a Darlington transistor made no improvement in supply cleanliness. The fix might be a real feedback-controlled regulator, but not now, if ever.

Step 3) a) Changed R3 to 56k, the value Heath used in the SG-5218 upgrade to the IG-18 (Reg used 47k); I'm not honestly sure what it is this base-to-base resistor does other than damp the Q of the filter somewhat -- it is essentially connected in parallel with the Bridged-T filter plus the feedbck resistors. I increased C5 to 100uF, added a 470uF cap around R11. LF stability improved. THD+noise, 1kHz, 1V = 0.016% with HP 334A; I got the same readings from ARTA Spectrum Analyzer, so the HP is reliable at this level, not that I doubted it. Good LF operation to below 10Hz, then signal just disappears at 3Hz and below; HF operation not yet checked.

Incorporated **Step 4** in this step and changed Q3 to Darlington MPS-A64. This worked OK at low frequencies, but there was no stable operating point above 10kHz and the output waveform was very distorted at higher frequencies. I put the Heath PNP back in, and still could not get a stable operating point above 70kHz. Removed the 470uF cap bypassing R11, and obtained full bandwidth with good stability. THD+noise, 100kHz, 1V = 0.12%. Lead dress of the wires to and from the frequency selector and range switches could be causing excessive stray capacitance problems. The MPS-A64 is only rated for 30V and it's possible it was working on the ragged edge. I will try a 2N5087 like Reg did.

Actually tried an MPS6729, which I have a lot of, and used a 100k resistor from its emitter to the + supply, where Reg used 220k. Works well. Tried the 470uF bypass on R11 again -- had to remove it -- with it, distortion was 1% at 100kHz. Removing it gave better stability and dropped the distortion 10-fold. This result is a bit counter-intuitive, given the lower open-loop gain without it; but with no cap, there is local AC emitter current feedback due to R11, the 150 ohm emitter resistor. Perhaps the 2N5087 is the best choice. I have a couple. ...

The 5087 was much worse than the MPS6729 as a Darlington device. I changed R3 from 56k to 100k with an immediate improvement in 1kHz THD and stability. I dug around and found an MPS8599 and replaced the Heath Q3 with it -- no Darlington. This works. It does not benefit from a Darlington second device. It's THD, 100kHz, 1V = 0.03%, the best reading at 100kHz I've gotten so far, and the 1kHz, 1V THD = 0.010%, with reasonable settling time and good stability. Unfortunately, the output level for this performance is 5VRMS max -- any higher and the distortion climbs rapidly. At 5V out, the settling time at 1kHz is long and the lamp is very sensitive to vibration -- other frequencies, higher and lower, settle faster.

Step 5) Replaced Q4 & Q5 with higher power devices, ECG 373 and 374 -- standing no-signal current is 13mA, sufficient to completely prevent crossover distortion, so no bias pot is needed. These devices are overkill to a significant degree. I'm going back to lower power, higher Beta transistors.

Step 6) b) Tried matched JFETs as the diff amp devices. ... The result was OK, but not as good as with the Heath 2N3416s; more noise, more distortion -- just not enough gain. The 56k R4 is a pretty long tail for the diff pair, but maybe it's time to try the JFET current source. ... Found a broken switch wafer on the 1's decade switch, which is why the lowest settings didn't work. Fixed that, which required a major disassembly. Ran the 10V and 1V THD+noise tests again at 10V out: 100kHz, 10V = 0.13%; 100kHz, 1V = 0.11%. 1kHz, 10V = 0.021%; 1kHz, 1V = 0.022%. All a bit better than expected at a basic 10V output from the oscillator.

Have decided to put a 47V zener in the power supply -- everything in the system should be OK with a roughly 45-46V supply output. This may give the voltage amp and the output stage some breathing room. Will at the same time use the original devices for Q4 (2N3416) & Q5 (2N2306) to let Q3 have a bit more dynamic range. These transistors are rated at 500mA and 625mW, so there's plenty of margin for them in this circuit. Not sure why Reg felt he had to use larger devices. Supply voltage is 45.3V. Everything is doing just fine.
-- THD+noise, 100kHz, 10V = 0.067%; 1kHz, 10V = 0.02%.
This level of performance is perhaps 2-5X better than Heath's original -- I have no idea how it compares to the SG-5218.
After tweaking the diff. pair bias control for lowest distortion at 100kHz, the THD at 100kHz = 0.058%; 1kHz = 0.025%.
Readjusted feedback for 5VRMS max. output:
-- THD+noise, 100kHz, 5V = 0.034%; 1kHz, 5V = 0.011%. Half the output, half the distortion, hmmm...
As before, settling time is fairly long at 1kHz, about 3-5 seconds.

See update note on grounding, above... I reconnected the square-wave generator, and made a ground run change -- I ran the attenuator/output ground back to the point where the power supply center tap connects to the oscillator PC board -- the square-wave generator has a negligible effect on sine distortion, especially with the square-wave output controls turned down.

The spectrum analyzer software shows that 60Hz is down about 76dB in the output, and multiples are nearly 100dB down, so hum just isn't much of a problem for higher frequency operation. **UPDATE --** I checked the output by setting the range switch to X100, the 1's switch to 0 and the 10's switch to 10, which raises the impedance of the frequency selector by 10X. The 60Hz compnent was much worse, and now I remember the issues I had with my first IG-18. This problem is EMI, not poor supply filtering. The cure in the other unit in which I put the Cordell oscillator circuit was to use a nice Avel-Lindberg toroid power transformer.

Perhaps more could be done with lamp choice -- maybe a lower resistance lamp, such as an 1819 or 1829 in conjunction with a resistor, which would be a less mechanically fluttery solution, leading to better stability and perhaps lower THD. Or perhaps a higher voltage lamp like the SC3 or 120PS (wire leads) or 120MB (miniature bayonet base), or a 4W 120V night-light bulb might work better. I think the thermal mass of the lamp needs to be higher than that of the 90V lamp.

Note that this unit is quite sensitive to the capacitor ratios in the range switch -- a slight change of ratio in range switching easily leads to stability problems and higher than expected level shifts. This is also true for the two resistor values in each pair -- they really need to be matched to better than 1%, but this is less critical than an accurate capacitor value ratio. This is because the resistance ratio changes the frequency slightly, but the capacitor ratio changes the overall gain level and feedback margin. The caps need to be matched to better than 1%. My biggest cap, the 5uF, was 5.05uF, so I used that "505" value for all the others except the smallest, where wiring C comes into play.

Access to an impedance bridge would be helpful here -- the second IG-18 I got just doesn't have a happy place on all the ranges, with big output level shifts and squegging on one range or another, depending on the feedback and bias control settings. Turns out two of the caps are pretty far off.

I think I'm done with this one for now. Good experience, and fun too. I will proceed with putting the Cordell oscillator in the second unit along with a nice, quiet power transformer (see IG-18 #2 on this site). In addition to trying the constant-current source replacements for the resistors I mentioned, another possible mod for the IG-18 is to do what HP did in the 239A oscillator and the 339A test set, both of which use a Bridged-T design -- use 8 capacitors and set the caps in a 100:1 ratio. This gives a notch/peak of around 34dB, meaning that the positive loop needs a "gain" of 34dB too. This significantly enhances selectivity, which should help lower distortion and improve stability -- but the open-loop gain of the amp may not be high enough to allow this with low distortion. Then you're looking at using an op-amp for the oscillator and needing a buffer amp to isolate the oscillator from external conditions, and now you're in a whole different arena, very far from easily improving this old warhorse...

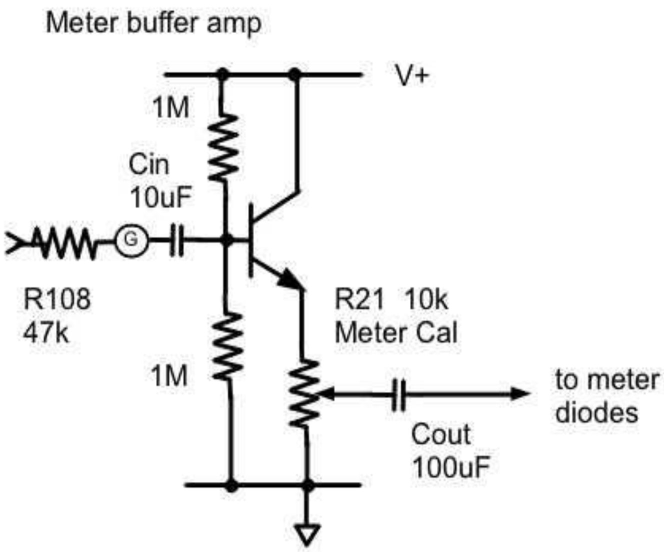
Summary:
I don't know how Reg Williamson got stable operation at high frequencies using a 2N5087 as a Darlington with Q3, and bypassing R11. Bypassing R11 does improve 1kHz results but significantly increases distortion at 100kHz, and really negatively affects HF stability. Adding the 2N5087 Darlington just made a world of hurt at 10kHz and above. **UPDATE --** a post on diyAudio forums from Bill T. indicated that this mod really didn't work and Reg retracted the idea of using it.

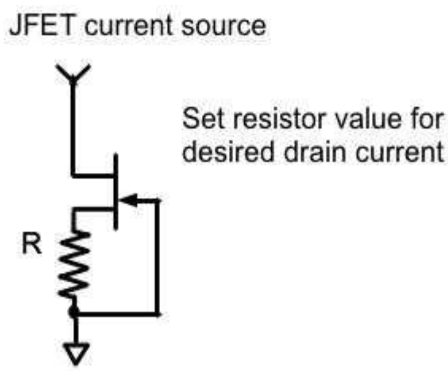
As to Reg's 1kHz distortion of 0.003%, maybe Reg misplaced a decimal point? I really can't explain his results -- 0.03% is altogether reasonable, but 0.003% isn't, based on my experience.

If I come back to this unit, I'll try the JFET current sources in place of R4 and R12.

So, the important things:
1) Change R3 to at least 56k -- 100k worked better for me.
2) Change C5 to 100uF/35V.
3) Buffer the meter circuit with an emitter follower! This is so important for low distortion and good high-frequency performance that it can't be overstated. I cut a trace and used the meter cal pot as the emitter resistor, and added a 100uF cap across the cut trace between the emitter and the meter's diode rectifiers. This saves a resistor if you're willing to select the base bias resistor.
4) Check the range caps and pad as needed to get the output frequency to be in exact 10X multiples as you switch the range switch, without changing the frequency switch settings. The smaller, HF caps will be slightly smaller than the others due to wiring capacitance. I used a 470pF cap paralleled with an 8-50pF variable for the X1k range.
5) For lowest THD, set the max. output to 5VRMS with the feedback control. At lower settings of the attenuator and with a 600 ohm load, the maximum output will be 2.5VRMS.
6) Try a current source for R12. Best overall performance will likely be with a current between 2 and 5mA.

UPDATE -- 10-2-2010 I did try the JFET current sources. As I thought, replacing R4 with the FET made no difference in THD at all -- the 56k resistor is sufficiently large to get the job done. Replacing R12 with the FET, however, made a difference. I used a 2SK246, a 50V FET, and optimum performance overall came at a constant current through Q3 (and the output bias diodes) of approximately 5mA, produced by a 120 ohm resistor from source to ground, with the gate grounded and the drain connected to the base of Q5. The bias setting affects THD as well. There is a balance between best 1kHz and best 100kHz THD. Now, with 10VRMS out, 1kHz = 0.016%, 100kHz = 0.068%. Readjusting feedback for 5V output, and resetting the bias control yields 1kHz = 0.014%, 100kHz = 0.028%. Most likely, a different current level through the FET would give best 1kHz THD at the lower output level. Clearly, the current source helps most at high frequencies.



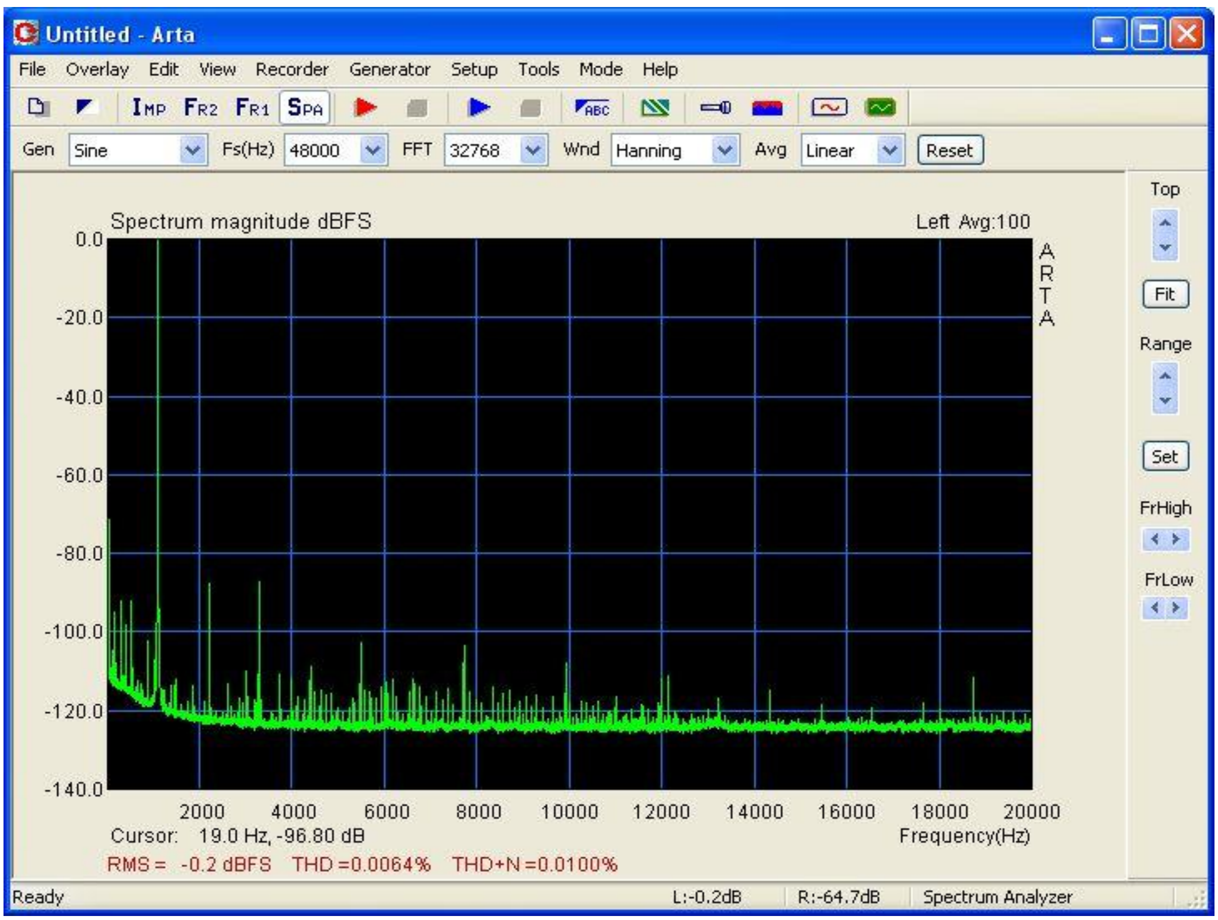


UPDATE -- 10-9-2010
Somehow, in fooling around, I killed the constant-current FET used in place of R12. While I was measuring another FET, which I set a 3.5mA, I decided to replace the diff amp pair with Beta-matched PN2484s as well. This got me to thinking about the amplifier in general -- how best to improve the linearity of the amp without making huge changes, like adding a current-mirror pair to the collectors of Q1 and Q2?

Reg tried increasing the open loop gain by bypassing R11. Through the years, working in low-noise audio gear, it has been my practice to get as much gain as possible in the input stage. I decided to try that here as well. I'm not much good with Spice -- I tried modelling this amp in LTSpice, and I can't figure out how to get a THD measurment with it, or even to show a sine wave on the output. So I plowed ahead with the usual tools -- soldering iron and miscellaneous parts.

My reasonining is that trying to get more gain out of Q3 will inevitably lead to higher distortion, not lower, because of non-linearity. I decided to increase the emitter resistor, R1, from 150 ohms to 1k ohm, which would add a lot of current feedback/emitter degeneration, and also raise the input Z of Q3. Then I increased Q2's collector load from 10k to 22k to get a bit more gain from the diff pair. And, thinking it couldn't hurt, I increased the power supply filters, C1 and C2, to 2200uF each.

The results were actually very good, better than I expected -- 10V, 1kHz THD = 0.016% and 10V, 100kHz THD = 0.056% (HP 334A). And with feedback set for 5V output, 1.1kHz THD = 0.0064% (ARTA). Here's the spectrum at 5V output (I raised the frequency so the spectrum lines wouldn't lay on top of the grid lines):



These results raised the question of whether further improvement could be gained from raising the capacitor ratio, similar to the HP 239A & 339A. Using 1uF and 22nF would make a ratio of 45 (recall that in the 239A and 339Aoscillators, HP uses a ratio of 100) and put the output near 1kHz, so I used those values on the X10 range. The amplifier had enough gain and the feedback control had the range to get stable operation, but the THD was slightly worse than with the stock Heath 10:1 ratio.

Oscillator parts changed/used: Q1 and 2: PN2484
Q3: MPS8599
D7: 47V 1W zener diode
C1 and 2: 2200uF 50V
C5: 100uF 35V
R2: 22k 1/4W
R3: 100k 1/4W
R11: 1k 1/4W
R12: replaced by N-channel JFET and 220 ohm 1/4W
Make sure that the tuning caps have values accurate to better than 1%. Measure all five, find the one with the highest value, then pad the others with small parallel caps to get the same digits.

Meter buffer parts added:
Qm: 2N3904 NPN
Cin: 10uF 25V
Cout: 100uF 25V
Cmeter: 100uF 25V
Rb1 and Rb2: 1M 1/4W
R108: 47k 1/4W

Adjustments Setting the bias and feedback controls worked best for me this way: Set the oscialltor to 100kHz. Fiddle with the bias and feedback pots until you get a stable 100kHz output of 10V with the attenuator and output level controls full up. Once 100kHz is good, then check at 1kHz. If you have a distortion analyzer, you can get a small improvement by slightly adjusting the bias control.

For lowest distortion with good stability, set the feedback control for slightly more than 5V output, then check the 100kHz output for stability and tweak the bias control if needed. You might be able to further improve THD by setting the output for 3V, but stability and settling time may be problematic.

The new combination of parts values has led to better overall stability and settling time is now acceptably short at all frequencies at 5V output.

Given the relatively low sensitivity of the lamp feedback, I think this performance is better than good. So three transistors and a large handful of parts has got the IG-18 working very well indeed. Not state-of-the-art, but real good.

12-9-2010 update
I hadn't turned this oscillator on since I finished the upgrades. I found a couple of things -- it didn't start up at 1kHz, and its 100kHz output was erratic. Moving the range switch fixed the startup -- as soon as I moved the knob, the oscillator started. Then I had to adjust the bias control to get a good 100kHz signal. The start-up problem persists over all settings of the various controls -- I have to move the range switch to get the oscillator to start up. Once going, it's fine. I can live with that, although I'd like to know the cause -- usually the transient when power is applied will start the oscillator. It may be that I've filtered the power supply too well and the start-up transient has too low a rate of change.

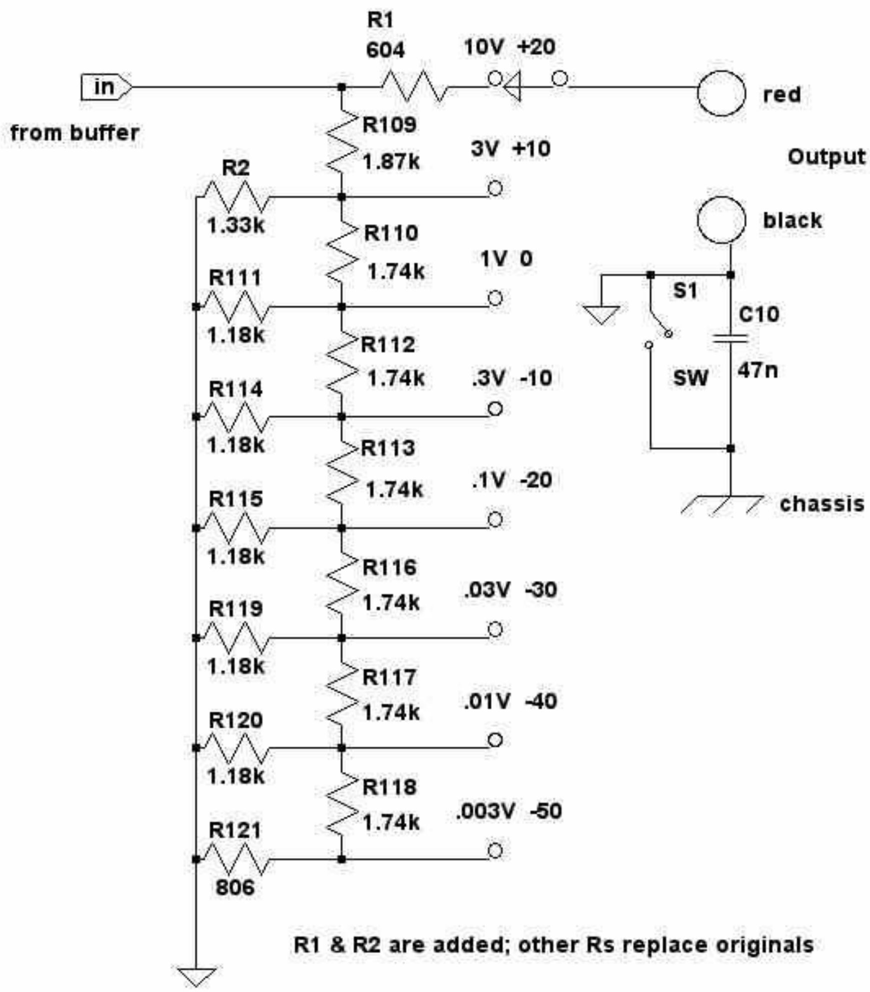
The other day I received an HP 8903E THD analyzer. This nice old unit has fully automatic operation, with auto level control and auto tuning, and a resolution of around 0.0015% 0.0009% at 1kHz, all of which is very convenient when you're fiddling with the feedback and bias controls. With it, I discovered that the setting of the feedback control didn't have an enormous impact on THD, but did, naturally, have a huge impact on output -- with feedback set for a maximum output ranging from 5V to 10V, feedback went from a low of about 0.008% at 5V to a high of about 0.012% at 10V -- so call the 1kHz THD 0.01% at max ouput.

What I did find is that the THD is very sensitive to the settings of both the 10dB step level control and of the variable output vernier control -- seems clear that a buffer amp on the output would be helpful. I also found that 60Hz hum in the output is a real problem at the lowest settings of the tens frequency switch, where the bridge impedance is highest. No surprises about the hum, but it is a bit discouraging if you want a good quiet oscillator. Fixing that will mean either replacing the power transformer or -- maybe better -- moving it out of the oscillator to an enclosure a few feet away.

5-20-2011 update
Recently I recieved an email from Richard Andresen, the author and publisher of the fine circuit simulation program 5Spice. Richard was curious about some modifications he had in mind, so we began exploring a few further mods. The idea was to get more open loop gain without sacrificing stability, with the end goal of lower distortion, especially at higher frequencies. We decided that Richard would work on a recently acquired IG-18, check for large problems, and if all seemed well, I would duplicate the mods as a check on variability, and then I would use my high-resolution distortion measuring gear to make a final evaluation of performance.

This work is now in progress. I hope to have more to say soon.

5-23-2011 update
BTW, here's an update to ther attenuator switch that uses the 1% resistor values from the HP 239A -- a very good fit for the IG-18 in any incarnation, and using the internal 600 ohm load switch to switch the ground between a floating ground coupled to the chassis with the Heath 47nF cap and a hard chassis ground is very useful. Note that the load on the output of the oscillator is a minimum of 1200 ohms if the 10V output is terminated in 600 ohms -- this load is OK for the IG-18 although it does increase THD somewhat. But into a high-Z load, no problem -- and the output Z is 600 ohms at any setting of the attenuator switch.


R1
604
10V +20
in
from buffer
red
R109
1.87k
3V +10
Output
R2
1.33k
black
R110
1.74k
1V 0
R111
1.18k
S1
C10
R112
1.74k
SW
47n
.3V -10
R114
1.18k
R113
1.74k
.1V -20
R115
1.18k
chassis
R116
1.74k
.03V -30
R119
1.18k
R117
1.74k
.01V -40
R120
1.18k
R118
1.74k
.003V -50
R121
806
R1 & R2 are added; other Rs replace originals

From: http://www.moorepage.net/IG-18-2.html

Rewritten 6-29-2012

Re-greening the Heathkit IG-18: IG-18 #2

Bob Cordell's State-Variable Oscillator, the BIG-18

With IG-18 #1, I explored the relatively easy mods to an IG-18 (or SG-5218, or IG-5218, and probably other Heath-Zenith-Schlumberger models which are identical or similar, like the IG-1272) to lower noise and THD without spending too much time or money. It ended up pretty well, with 1kHz THD+noise at around 0.014% - 0.02% depending on feedback level. That's pretty good -- much better than the original and 10dB to 20dB better in the audio range than the HP 204B and 204C units I once had.

But I bought a second old IG-18 in order to redo a mod that I first built some 20 years ago -- to replace the Bridged-T sine-wave oscillator with Bob Cordell's State-Variable oscillator design. Bob's design is built around the old but terrific NE5534 opamp. This type of oscillator has been used by Tektronix (notably in the TM500 Series SG-505, also with NE5534s), and I suspect by Krohn-Hite in the 4400A (Nope, it's an LM318; I'd change this IC, although the LM318 is a very good amp), and other commercial designs, such as the oscillator in the HP 8903A and B (also using 5534s).

Many will ask, "Why re-build an old product so extensively?" The answer for me is that the old product has a great chassis, very good switches and mechanical gear, a decent power transformer, has a meter for the output level, and is generally well-shielded. These days, those are the parts that either are expensive, hard to get, completely unavailable, or, in many cases, just plain unattractive -- if you can find them. So for me, the iG-18 is a very good candidate for a new life of greatly enhanced performance.

Bob's design, published in Audio magazine in the mid-'80s, has well stood the test of time and he has a great article in a PDF on his website about building a complete THD Analyzer, of which this oscillator is part. Around 1990, I built Bob's design as two boxes -- analyzer and oscillator -- and the pair had, for their time, fabulous performance, better than almost anything commercially available, at least for the analyzer section. The pair I built measured 1kHz THD at a little under 0.0005% -- but I couldn't confirm that result with any other gear I had access to. Was the result real? The short answer turns out to be "Yes." Bob's article thoroughly discusses the merits of the SV design and I won't repeat any of that here. If part IDs are needed for clarity, I will use Bob's number system. See his THD analyzer article for the schematic and reference numbers.

I first started this rebuild before I had the capability to do high-resolution distortion analysis. Even when I began, well over three years ago, I only had an HP 334 THD analyzer, and with a resolution limit of about 0.01%, it just didn't have the resolving power to tell me anything about the actual performance of this design. Then I bought an HP 339A and subsequently an 8903E -- and I found that those instruments, good as they are, also could not reveal the true performance of this design. They both resolve to about 0.001% at 1kHz -- not anywhere near good enough, as it turns out.

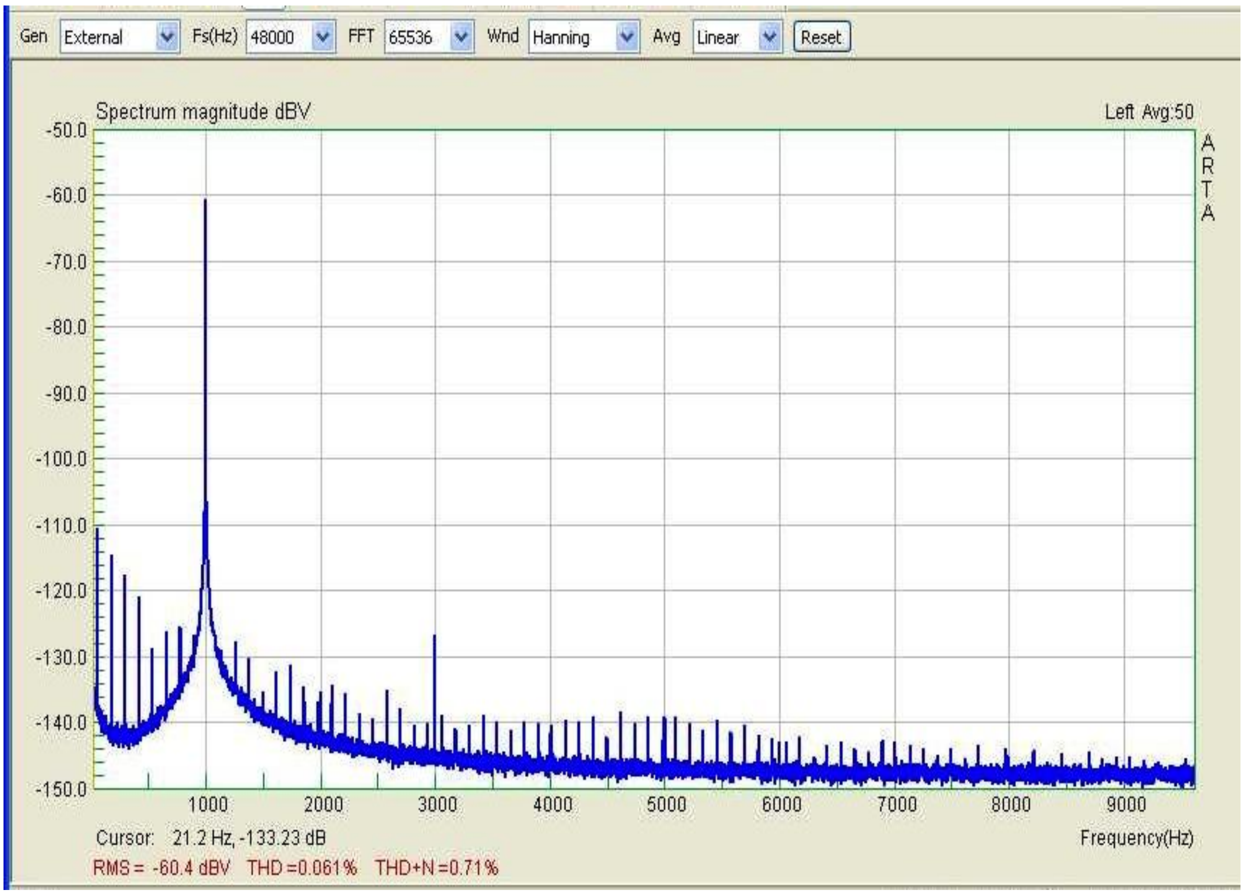
So, what can you expect from this BIG-18 oscillator?

Here are two plots of its performance, using OPA1641 for amplifiers IC1 and IC2, LT1468 for IC3 and IC4, LME49710 for IC5 and IC7, NE5534 for IC8, and an OPA134 for IC6.

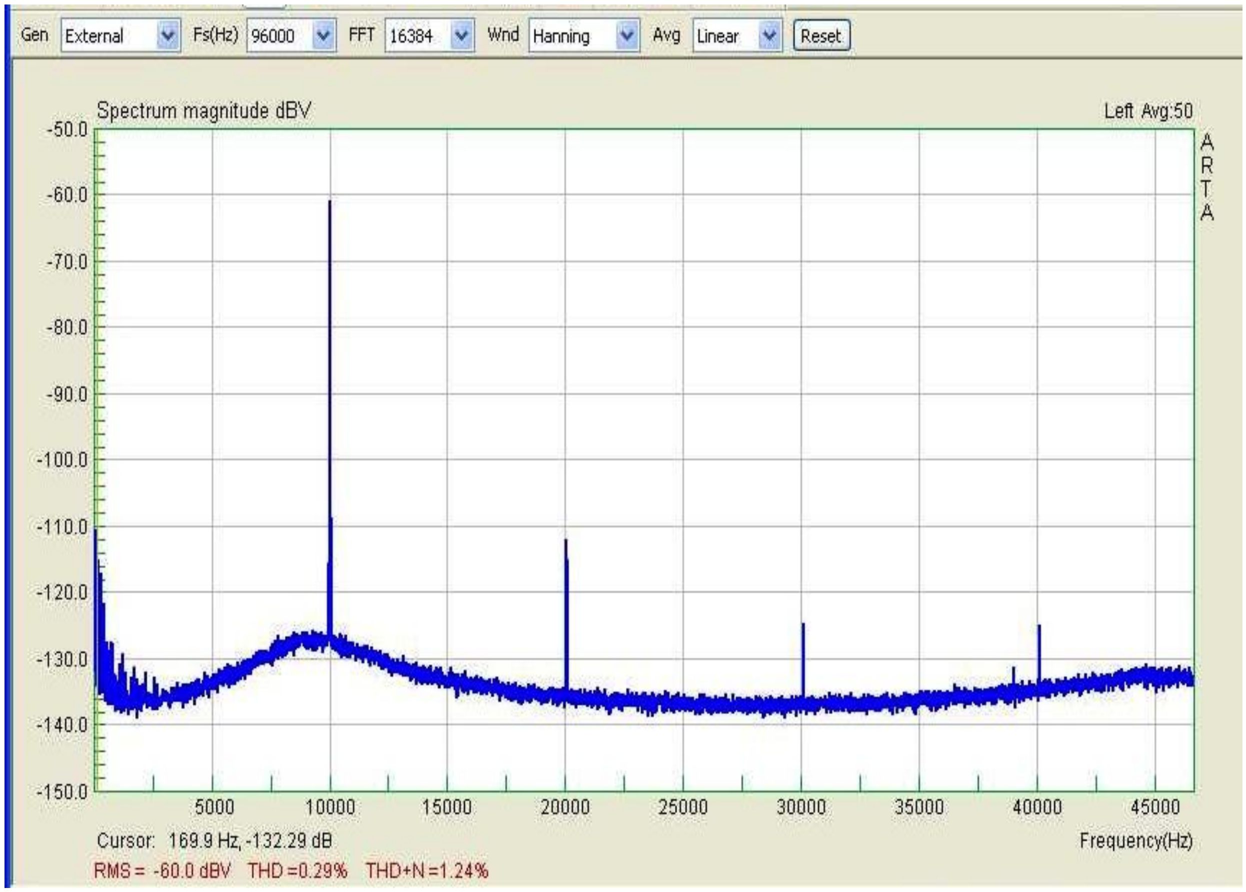
The first plot is the spectrum of its output of 10VRMS (!!) at 1kHz into a medium impedance load. The BIG-18's 600 ohm output drives a 30k ohm level pot, followed by an Active Twin-T notch filter to suppress the fundamental while passing harmonics unchanged. This is followed by an E-MU 0204 USB sound module, which then supplies data to a PC and the [ARTA Audio Analysis](#) software's spectrum analyzer function.

In the first spectrum, the 2nd H. is pretty low among the line-frequency harmonics at around -135dBV (there's a smaller power-line harmonic just to the left of the 2kHz spike), while the 3rd H. is about -126dBV; the calculated THD is 0.000061% -- that's 0.61 parts per million...

The second plot has easily-seen harmonics every 10kHz, with a few spurious noise spikes. The calculated THD is 0.00029%. The rise of the curve above 30kHz is due to the ADC's Sigma-Delta sampling system.



Spectrum of 1kHz output. Note the powerline-frequency related products at 120Hz intervals.

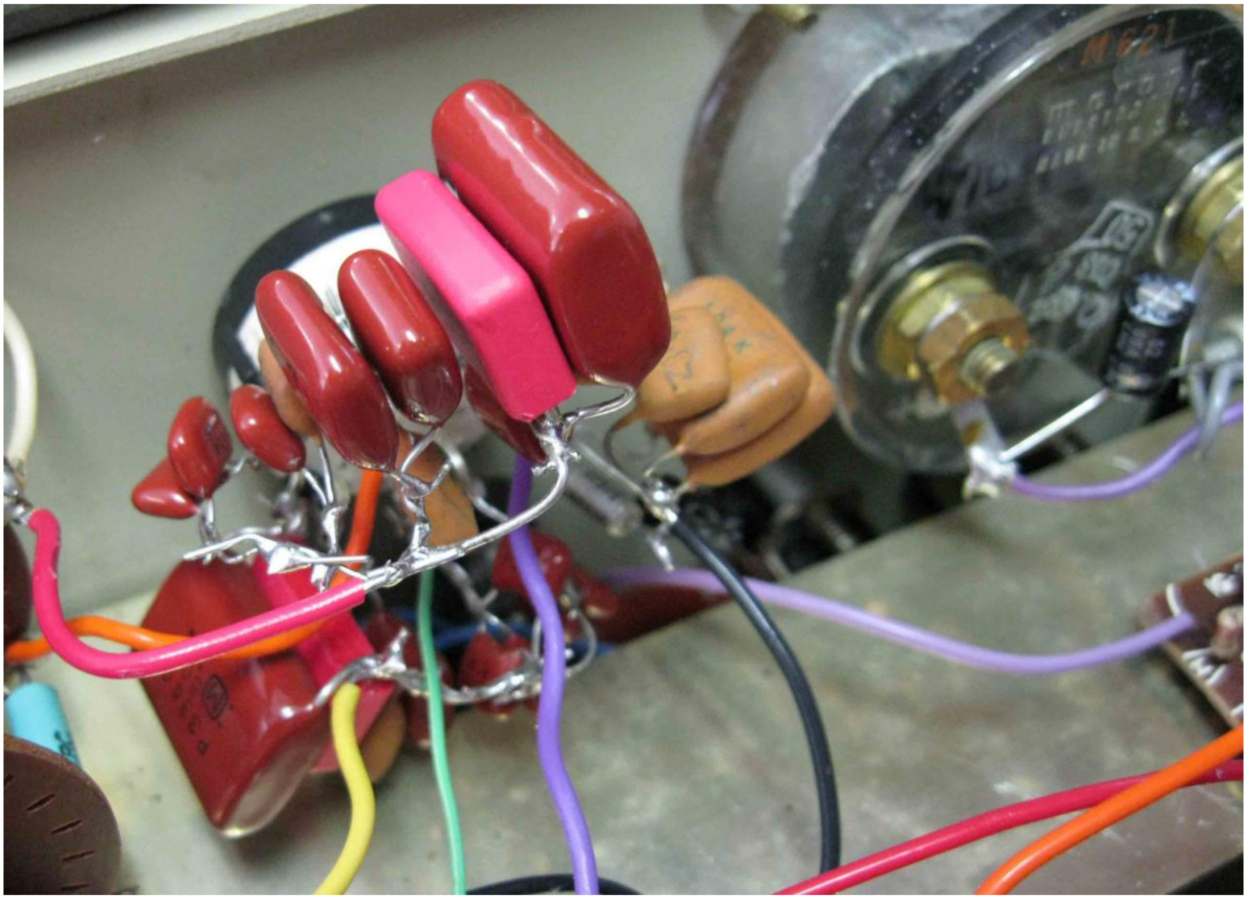


Spectrum of 10kHz output. The sampling frequency of 96kHz limits the spectrum to about 47kHz, but the higher-order harmonics don't amount to much -- the ones you can see are the prominent ones.

Caveats --
In the measurement area of parts-per-million, almost anything can change everything, so the results above are not guaranteed. For example, I found that using some ceramic caps in the tuning cap section of another oscillator design resulted in a huge increase in distortion, so parts materials matter, as do the actual parameter values of the various parts. I used mostly 5% carbon-film resistors on the oscillator board -- now I would use all 1% metal-film resistors. I also used a mix of polypropylene and mylar film caps for the tuning caps, and now I would use all polypropylene (or polystyrene if I could get the values needed).

Reconfiguring the IG-18
When I first put this oscillator into an IG-18, I used the existing Heath frequency selector switches and resistors -- I recommend doing that. I only needed to rewire the Bridged-T resistor switches to eliminate the junction points of the two resistor sections of each frequency step. That's because the state-variable design needs the two resistors (and also the two capacitors) of the two integrator sections to be independent, having an R and a C as part of each integrator stage.

In addition, I replaced the range switch because the switch needs (at least) 3 poles and the Heath Range switch only provides 2 poles. A 3-pole, 4-position switch is required for the 8 caps of the two filter sections, and for the 4 time-constant adjusting caps needed for the AGC feedback loop. Plus, if you use it, a fourth pole is needed for Bob's frequency trim pots, R1 thru R4 -- I didn't use them since I used selected parts for tight tuning accuracy. 3-pole, 4-position switches are readily available at low cost at Mouser and on eBay, but 4-pole, 4-position switches will cost more. The cheap switch I used has a plastic shaft, which was ideal for filing a flat onto for the Heath knob to fit over.



The new and very messy range switch. Note the 100uF damping cap on the meter.

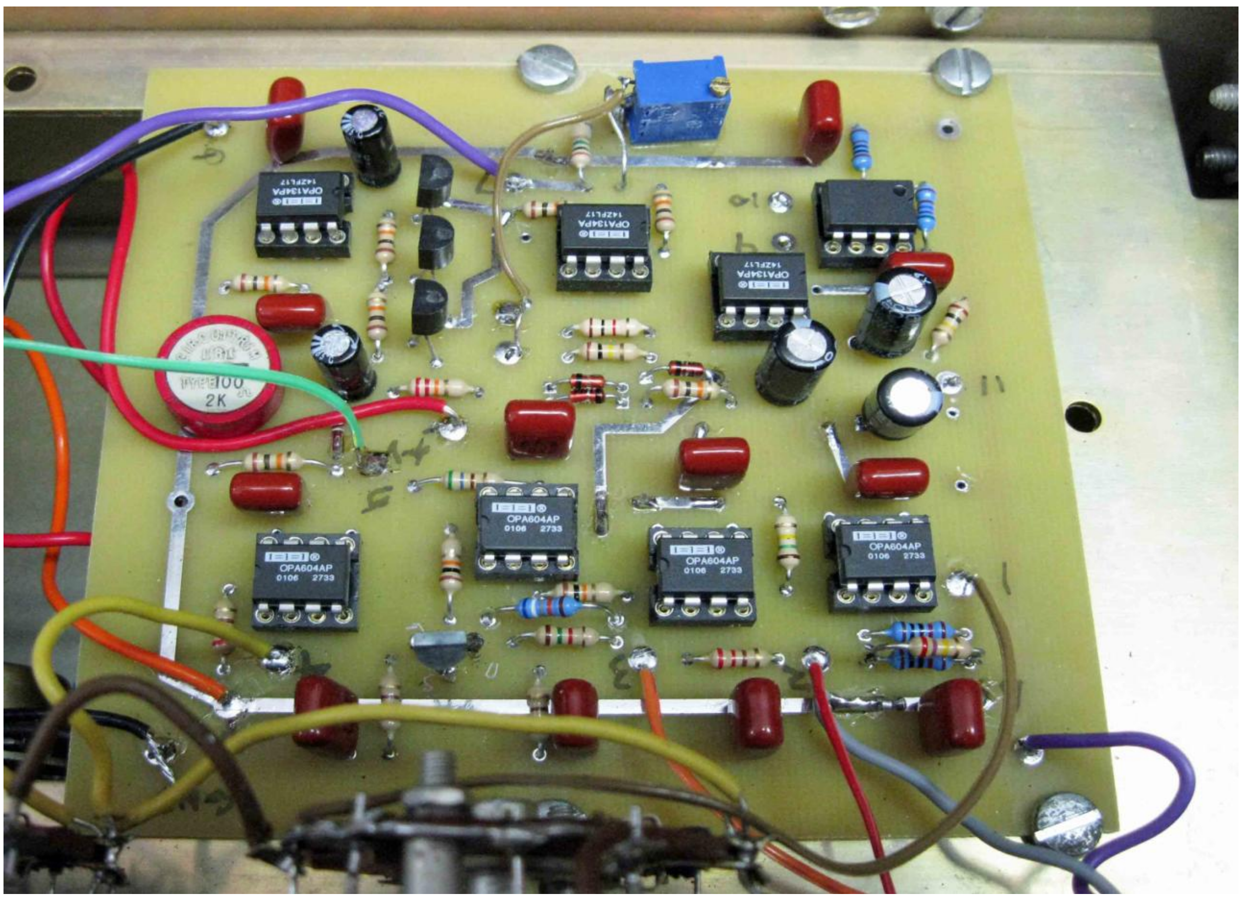
The IG-18 power transformer has two primary windings so that it can be used with 220V-240V power. This means that you can put the primary windings in series, as used for the 220V connections, and connect the 120V lines to transformer pins 1 and 4, and strap pins 2 and 3. This simple tactic cuts the secondary voltages in half, and using a full-wave bridge rectifier lets you have +/- 24VDC supplies -- ideal to supply the LM317 and LM337 regulators that supply the plus and minus 17VDC that I used in order to get the full 10VRMS output.

However, I thought -- wrongly -- that the Heath transformer was too noisy with lots of EMI hum, so I bought an [Avel-Lindberg](#) Y236004 15VA toroid with 18VAC secondaries. While not needed, and an extra cost, the toroid does offer very low radiated fields.

Regarding hum and noise, be sure to remove the neon bulb pilot light (just pull it out of the red plastic housing) and also remove it's associated wiring carrying 120VAC. Old neon bulbs make great sources of oscillation and noise in nearby circuits. If you want or need a pilot lamp, use a red LED and limiting resistor run off of one of the regulated supplies, and stuff it into the original red housing.

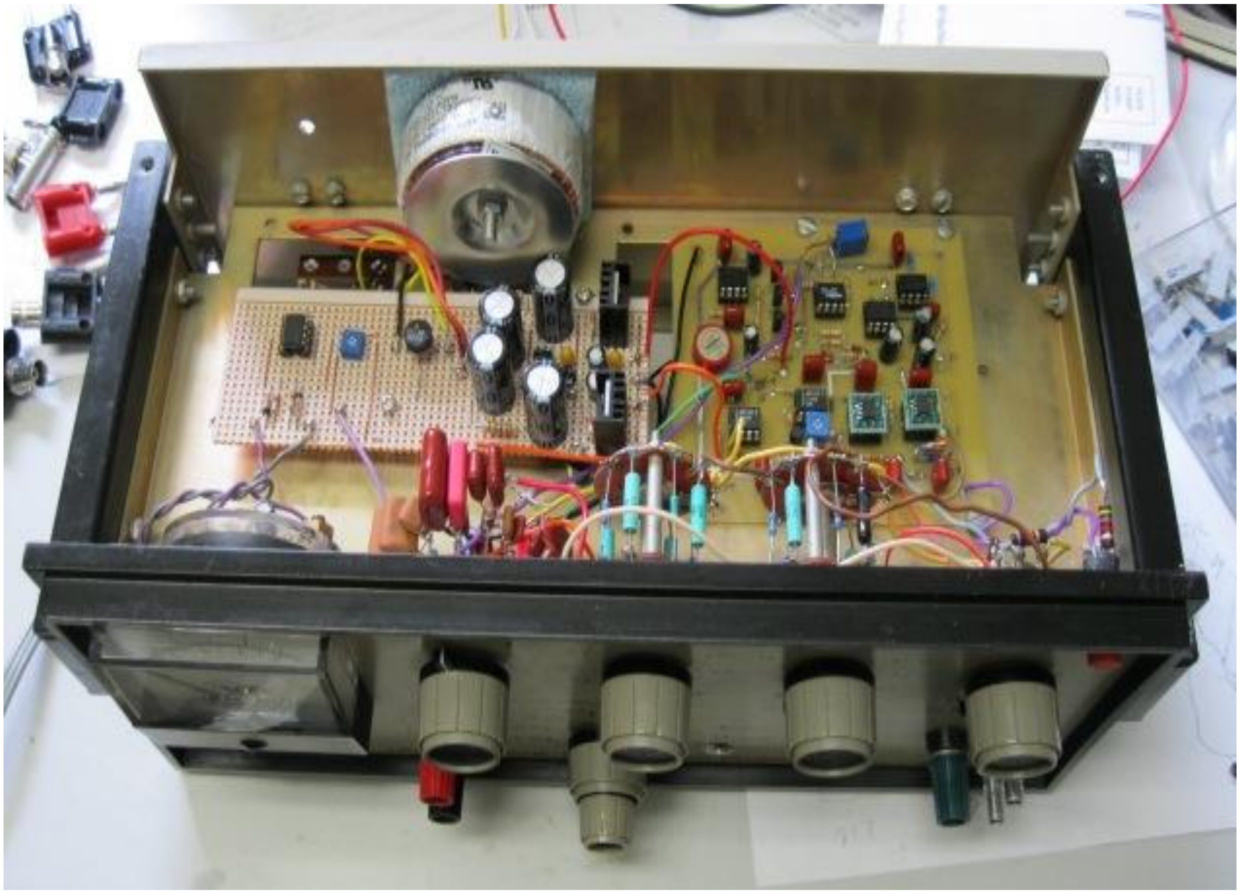
I drew up a two-layer PC board layout using [ExpressPCB's free software](). It is easy to use compared to other packages I looked at, but it does have limitations. I couldn't figure out how to leave lots of copper on the board -- I couldn't find a "flood" tool that fills areas to be left unetched even though I'm sure the tools are there to do that. The other issue is that the files it creates are not industry-standard, like Gerber files, and so you're stuck using ExpressPCB to make the boards from the layout you create. I have the ExpressPCB board layout in their format -- if you would like it, email me and I'll email it back

I wanted low cost, so I had four prototype boards made (no silkscreen for parts location) -- one for the IG-18, and three for online friends. ExpressPCB delivered the four boards in 6 days -- pretty great. Given the size of the boards and the layout, they ended up costing $27 each -- an order for four was only slightly more expensive than the minimum order of two.

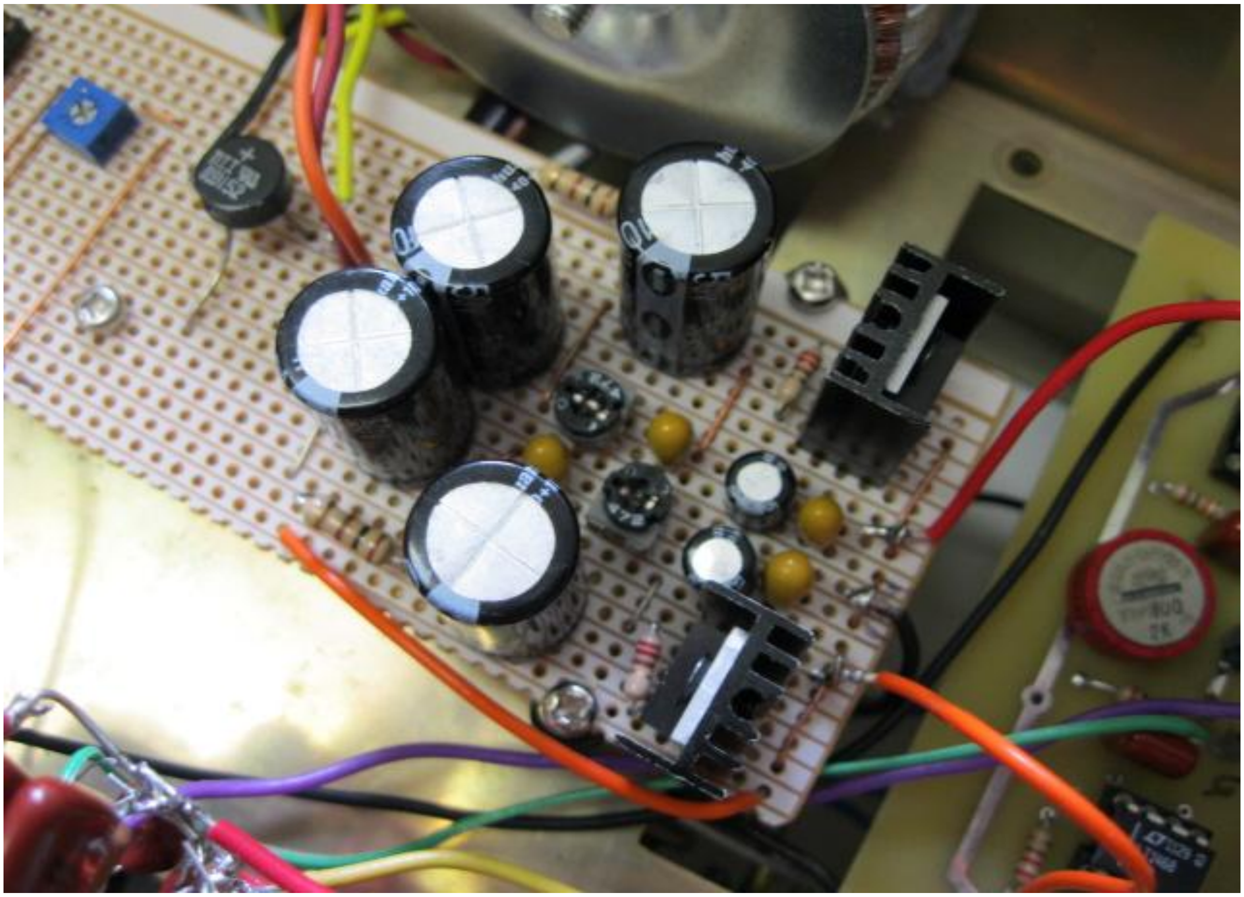


The oscillator board -- this older photo shows the board populated with different opamps than I now recommend using. The blue 10k pot at the top replaces Bob's 6.8k R19, and adjusts the reference DC level for the AGC loop, which then sets the output level of the SV oscillator. This makes a handy level trim.

Putting the parts in was easy and the board tested OK by using a small three-output supply for power; I kludged up the two Rs and two Cs for a single-frequency test. All of the Heath boards came out, as well as the switches, in order to make reconfiguring easier. It really isn't any fun removing parts that were installed just the way that Heath recommended, with leads tightly twisted together on switch contacts, so rewiring the switches was tedious.



Top view of the oscillator, showing the osc. board, right, and the power supply & meter board, left. The toroid power transformer is on the back panel (see the bottom of the page for an important change in location). If you use the Heath power transformer, there's still room for a power supply/meter amp board in the chassis. Note the two OPA1641 on the adapters at bottom-right of the board.



Close up of the power supply, with the DC voltage adjust pots between the caps.

I built the power supply on strip (Vero) board for the rectifier bridge, filters, regulators, and the meter amp, and mounted it over the holes where the Heath transformer and power supply had been. I put the Avel toroid on the rear panel. Lastly, I mounted the oscillator board where the Heath oscillator board had been.

AGC issues

Cordell's design uses a 2N4091 as the JFET for the AGC loop's voltage-variable resistor. These are virtually unobtainable, but PN4091s, and PN4092s are still at Digi-Key and other distributors. The FET requirement is simple -- an Rds(on) of 50 ohms or less, which is pretty low for small-signal FETs, many of which have values in the hundreds of ohms. I'm using a PN4092 for the JFET -- it seems to put the gate voltage in a better range than the PN4091, which was quite close to pinch off for me.

THD Analysis

In order to tell how well the oscillator is working and in order to make adjustments for lowest distortion, you need a very good THD analyzer or spectrum analyzer. As mentioned, I bought an HP 339A and it is good, but not good enough -- it has a measurement floor set by the level of 2nd H. distortion in the notch filter, and that limits its performance to about 0.001% at 1kHz. This same limit and cause exists for the otherwise excellent HP 8903 series. I'm guessing that it is common-mode failure in the notch filter summing amps, but I'm not completely sure.

I initially used the Intel "High Definition Audio" chipset in my PC, which has 24-bit converters with a maximum sample rate of 96kHz. With that, I used the ARTA Spectrum Analysis software. It isn't clear to me exactly what the limits of performance are for this hardware & software combination, but it looks to be around 0.0005-0.001% for THD only, with THD+noise dependent on a host of external and internal factors. That's pretty good -- except for limited bandwidth, it's comparable to many commercial analyzers. For comparison, my old, now sold-off set of the Cordell oscillator and analyzer had a 1kHz combined floor of under 0.0005%, depending on low-pass filtering and wind direction -- now, remind me again, why did I sell those?

But, as mentioned above, I now use an E-MU 0204 USB sound module for input to the PC and ARTA. By itself, the 1kHz residual of the E-MU is in the 0.0002-0.0005% area, which is excellent. Used with the Active Twin-T notch filter, the results are spectacularly better, as shown the the two spectra at the top of the page. And it offers a maximum sample rate of 192kHz, allowing measurements out to about 90kHz -- good enough for 10 or 20kHz depending on the strength of the higher harmonics.

First run

Oscillation did not occur -- well, it did occur on a couple of ranges, but erratically and with lots of distortion. Checking wiring, I found that I had mis-wired the range switch all the way around by one position. Fixed that and everything was good. I started out using the op-amps Bob specified -- NE5534s for oscillator and output, and LM318s for AGC gain. Performance was good but not spectacular -- somewhere in the muddy area of "Is the oscillator good?" and "Just how well are the analyzers actually working?" I substituted OPA134s for the NE5534s, and the 1kHz THD measured by the PC & ARTA went down by almost two. that was good news, and since then, the news has only gotten better and better.

Checking THD at the 100 setting of the 10s decade (using the OPA134 and 604 amps) resulted in higher THD than when using the 10 setting and the next higher range. This meant that the integrators were not so happy with the 1k ohm resistance that the Heath design has on the 100 setting. I decided to rewire the resistors in the range switches, doubling the values -- Bob's design has a minimum R of about 3k ohms, with frequency steps on ISO 1/3-octave frequencies. That was a lot of work, and I would not now change the frequency resistors to larger values -- modern opamps seem to be perfectly OK with the 1k resistors at the 100 frequency setting. But I don't intend to un-rewire the switches! If you keep the existing Heath resistors, the range caps need to be "16" values, from 1.6nF to 1.6uF -- see the next section.

Rs and Cs

The Heath resistors are switched combinations of 10k, 5k, 3.33k, and 2.5k for the 0 - 100 switch, for a span of 10k to 1k in equal frequency steps, and 100k, 50k, 33.3k, and 25k for the 0 to 10 switch, for a span of 100k to 10k. The vernier is a dual 1M pot with each section shunted with 82k. Shifting Rs up by two meant I would need some 20k and 6.66k, and 200k and 66.6k resistors, and reuse the 10k and 5k, and 100k and 50k resistors. 20k and 200k are standard 1% values. 6.65k and 66.5k are also. That's close enough -- low by 0.15%, if exact -- and, if necessary, they could be padded.

Since I had previously sorted and assembled the caps to better than 1% using parallel combos and a 3-1/2-digit C meter, the caps were going to be a chore, and I wasn't sure I had enough of various values. The oscillation frequency is given by f = 1/(2*Pi*R*C). For the Heath resistors, the C values are 1.6uF, 160nF, 16nF, and 1.6nF, with two of each needed.

I had ordered "1.5" series 2.5% caps from Mouser, as well as "1" series values for padding. Changing the resistors up by two meant I could use half of the Heath 1% resistors, and add four other 1% values to each switch. But the caps would need to be "8" series -- not so great. But "3.3" and "4.7" could be paralleled, and given 10% tolerances, I could get pretty close. It turned out that "8.08" was the magic number, given the caps I had on hand, and I got the necessary set of 8 caps through a lot of fooling around and sorting.

Op-amps... oops
After making all the changes, I now discovered that the OPA134s did not like this set of Rs and Cs above about 60kHz. I'm not sure why. The 5534s were just fine, but I didn't want to lose the lowered THDs from the OPA134s. I had a number of OPA604s, which have better than twice the gain-bandwidth-product of the 134s and are unity-gain stable, and they were as happy in-circuit as the 5534s, but with essentially the same THD as the 134s. My mental assessment is unproductive of why the 134s were not happy with 808pF of C and the higher-frequency smaller Rs -- I'm guessing too much edge-of-band phase shift plus maybe some problem with stray capacitance, but I really don't know.

By the way, as noted earlier, Bob's design uses an extra pole on the range switch to provide connecting an on-board 6.8k resistor, R5, to a switched pot for each range, in order to trim the frequency for range-to-range consistency. Because I had pretty accurate cap values, I just used a 10k resistor in place of all that, then padded it with 270k in parallel to get the output frequency near to "1" given the slightly large Cs I used -- the close matching of the Cs made it good enough. If you want real frequency precision, then use a 4-pole range switch and Bob's four-pot trim system.

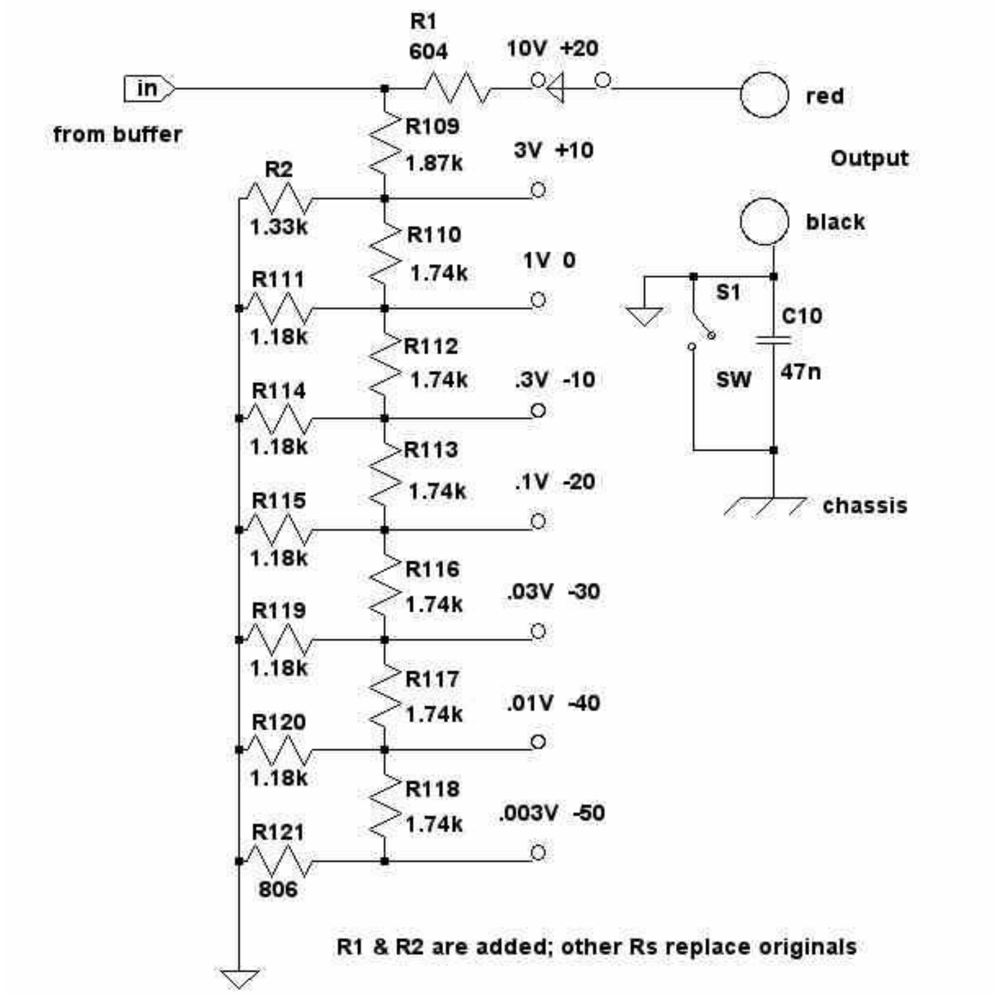
Since then, I've bought a variety of good opamps that offer particular strengths for the various circuits. See the next **Opamps** section below.

Output attenuator
Heath's design for the output attenuator provides a reasonably high resistance load to the oscillator on the 10V/+20dBV and 3V/+10dBV ranges and prevents the front panel switchable 600 ohm load from being used on those ranges, again to avoid output loading. I think that was because they didn't use a buffer amp between the oscillator and the load -- at the time, a good buffer was hard to find.

The Cordell design does have a good buffer amp. I rewired the attenuator to put its vernier pot between the oscillator and the buffer, per Bob's design, then connected the output of the buffer directly to the switched attenuator. I realized that the 5534 buffer didn't care very much about loading, especially at lower signal levels from the vernier pot, so I removed the 150 ohm, 2.4k and 390 ohm resistors of the first two positions. I initially wired the output directly to the first position and made an 1155 ohm shunt resistor (30k//1.2k) and two 1708 ohm (39k//1.8k) series resistors, giving the attenuator better accuracy both unloaded and loaded by 600 ohms. The result is that if there is no external load, the oscillator sees a roughly 2.4k load from the attenuator. The resistor combos above for 1155 and 1708 ohms could be used for the rest of the attenuator, too, but at the time I didn't want to work that hard.

Attenuator upgrade
Then I decided I wanted a **good** attenuator. The following schematic shows the attenuator 1% resistor values as used in the HP 239A and 339A. These values are a very good fit for any IG-18 incarnation -- especially this one -- and provide much better accuracy than the Heath design does, both terminated and unterminated. The load on the output buffer amp is a minimum of 800 ohms if there is a 600 ohm load terminating the output of the 10V range, and a load of almost 2.5k if the output is not terminated.

Using the 600 ohm load switch for a floating or hard ground is very useful. I made the mod and highly recommend it.



Further work

For now, there is no square-wave capability. That is fairly easily fixed by using a CMOS hex inverter and a few resistors, but I'm going to skip having a square-wave output in this version -- the boards are pretty full now and I have other sources for square waves.

More seriously, the output is slow to start up at very low frequencies. I used a larger time-constant cap than Bob specified for the X1 range, which may be part of the problem. I'm using 5.5uF (a paralleled combo) and the settling time is quite long, with the output doing a lot of flopping around before getting steady, but with 1uF, it just didn't wan't to settle at all. Using the PN4092 for the JFET (where I initially had a J105) seems to assure start up -- it takes about 8-10 seconds to settle, but it does settle, even at 1Hz. 5.5uF may simply be too large.

I was traveling for a couple of days and I had the thought of trying to see if having the output amp circuit's currents run through the board ground trace as I had it laid out was causing THD or stability issues. While I was at it, I also decided to put in a 2M pot to replace the 750k R13 signal feedback resistor that goes from the oscillator to the gate of the AGC FET and see what the best setting is. I had asked Bob about the value, and he couldn't remember how he arrived at it. Switching the grounds had no impact on THD as far as the PC and ARTA could show, so I'm using the board as laid out.

The 2M pot around the FET, however, made a difference, Using the ARTA spectrum analyzer, I found the best setting at 1kHz was very near 700k ohms, with a very significant and very sharp drop in 2nd harmonic level, and less large drops in the other even harmonics, but no change in the odd harmonics. Changing the pot's setting by a few kohms made a 6dB difference in on-screen 2nd harmonic level. Clearly the FET's channel resistance modulation consists primarily of even-order harmonics.

Opamps
The opamps used in the state-variable section, U1 thru U4, are very important to overall performance, as is the output amp U8. I tried various combinations of OPA134, OPA604, NE5534, LM318, LME49710, LT1468, and OPA1641 in the U1 - U4 positions. The best combination turned out to be OPA1641 for U1 and U2, and LT1468 for U3 and U4. And counterintuitively, the NE5534 turns out to have the lowest distortion as the gain-of-6.6 output amp, U8. I thought that the OPA1641 or LT1468 would be better. The LME49710 was as good as the 5534, but had a very-low-frequency pulsing, and since the 5534 was as good, at least at 1kHz, I just used it. 100kHz may be another matter -- high-resolution distortion measurements at 100kHz are still beyond my reach. BTW, the Active Twin-T notch filter works at 100kHz, so with that and a good hardware spectrum analyzer, such measurements are practical -- but I don't have a good hardware SA, and they are still spendy.

I found that with the extra resolution of the Twin-T notch filter and the spectrum analyzer, I could see that the AGC loop's rectifier balance, adjusted by the 2k pot I used (1k R24 in Bob's circuit), actually does affect the level of the 2nd harmonic distortion a moderate amount at 1kHz and below, so I set it for lowest 2nd H. level. The main output from the output buffer amp, which has a gain of roughly 6.6, is a little noisier than the output at connection E5, the output of the state-variable section, but does not have significantly more THD, and that's very nice. The other adjustment is the JFET 2nd H. cancellation, and I adjusted this for minimum at 10kHz, then went back and made sure the rectifier balance pot was still right for 1kHz.

I settled on using LME49710s for U5 and U7 and an OPA134 for U6, although these are not crucial to low-frequency distortion performance -- again, 100kHz may be another matter. The high input Z of the OPA134 helps avoid AGC input filter level sag so that filtering is efficient.

After finishing IG-18 #3, now called IG-339A, I had some PN4091 and PN4092 JFETs. Bob Cordell used the 2N4091 for the AGC FET in the original. The IG-339A works really well, so I decided to see if this version could be improved, too. I had been using the Fairchild J105 for the JFET. The PN4091 worked better, but the gate voltage was very close to pinch-off. I subbed in a PN4092 and it worked the best of all for quick settling time and overall stability.

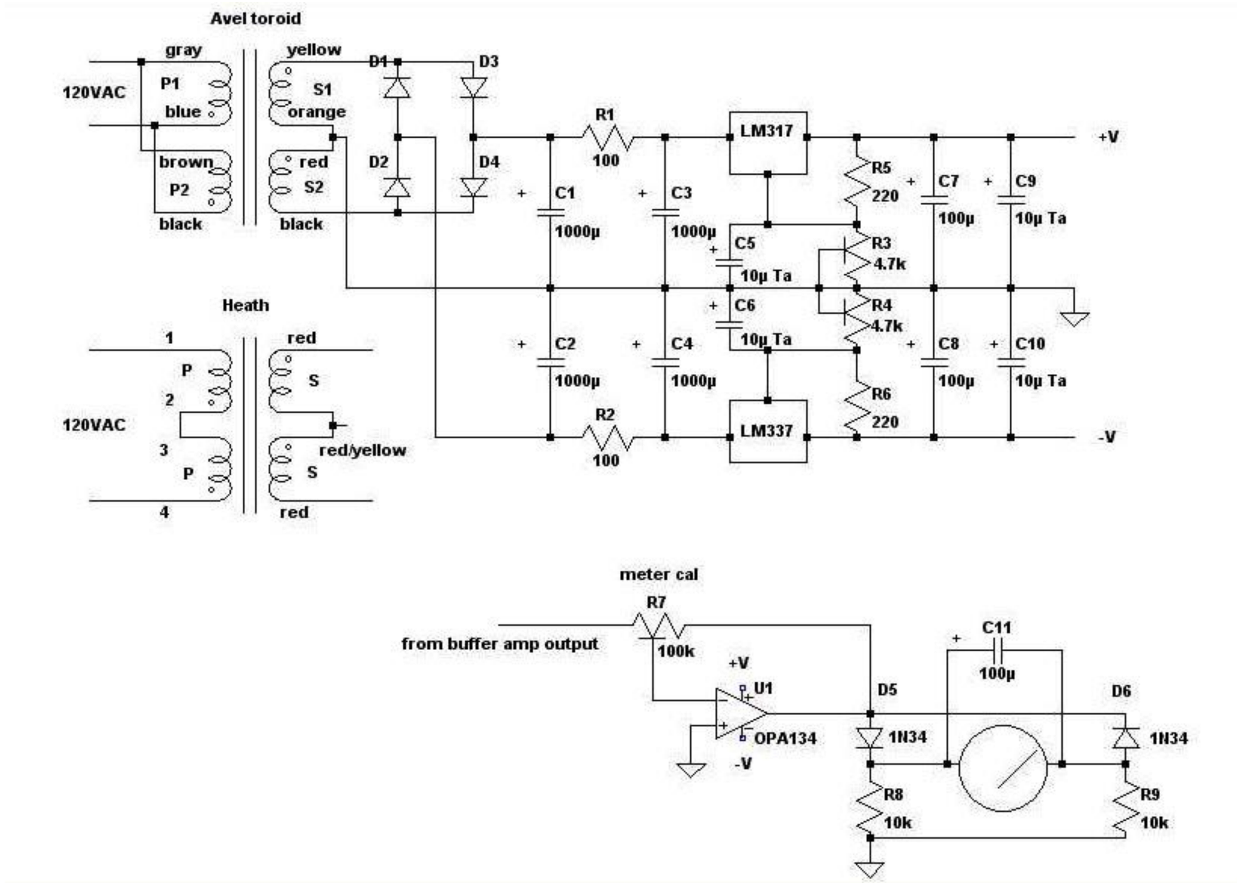
I modded the Heath front-panel 600-ohm load switch to be the floating/chassis ground switch as in the HP 239A attenuator diagram. This change let me hard ground or float ground through 47nF as needed. You can see how low the power line noise is in the 1kHz spectrum, with peaks under -110dBV.

Lowest distortion?
Achieving the full 10VRMS output of the original IG-18 required higher supply voltages than usual -- 17V versus 15V -- and higher gain from the buffer amp, so some distortion performance is sacrificed compared to Bob's design. If you can live with around 1.5VRMS output, then using the usual 15V

supplies and a unity gain output buffer amp will give the best results, with an up to 6dB improvement at 1kHz. And you'll get lower distortion with 15V supplies and a gain-of-4 buffer amp that gives you 6.3VRMS open circuit output or +10dBV when driving 600 ohm loads.

Circuit details

The oscillator circuitry is mostly as it is in Bob's article, with the exception of the few changes that I've made. The power supply and meter amp circuits are shown below:



The inverting configuration of the meter amp means that the amp can have gain or loss as needed for proper adjustment, and the large value of the pot (at reasonable settings) does not load the output of the buffer amp. Noise of this amp is not a factor, so the large 100k pot works fine. The input leg of the pot connects to the main output point of the buffer amp -- I found it convenient to connect to the junction of the 604 ohm and 1.87k resistors, the input point of the output attenuator.

I deleted Bob's switches S1 J & H, and made R5 10k, then padded that to get the output frequency set to 1kHz-- a 20k pot could be used for R5 to give any needed frequency trim. If you don't use Bob's frequency trim pots, it's good to check for the frequency on all four ranges and split the largest spread in two, for the smallest overall error.

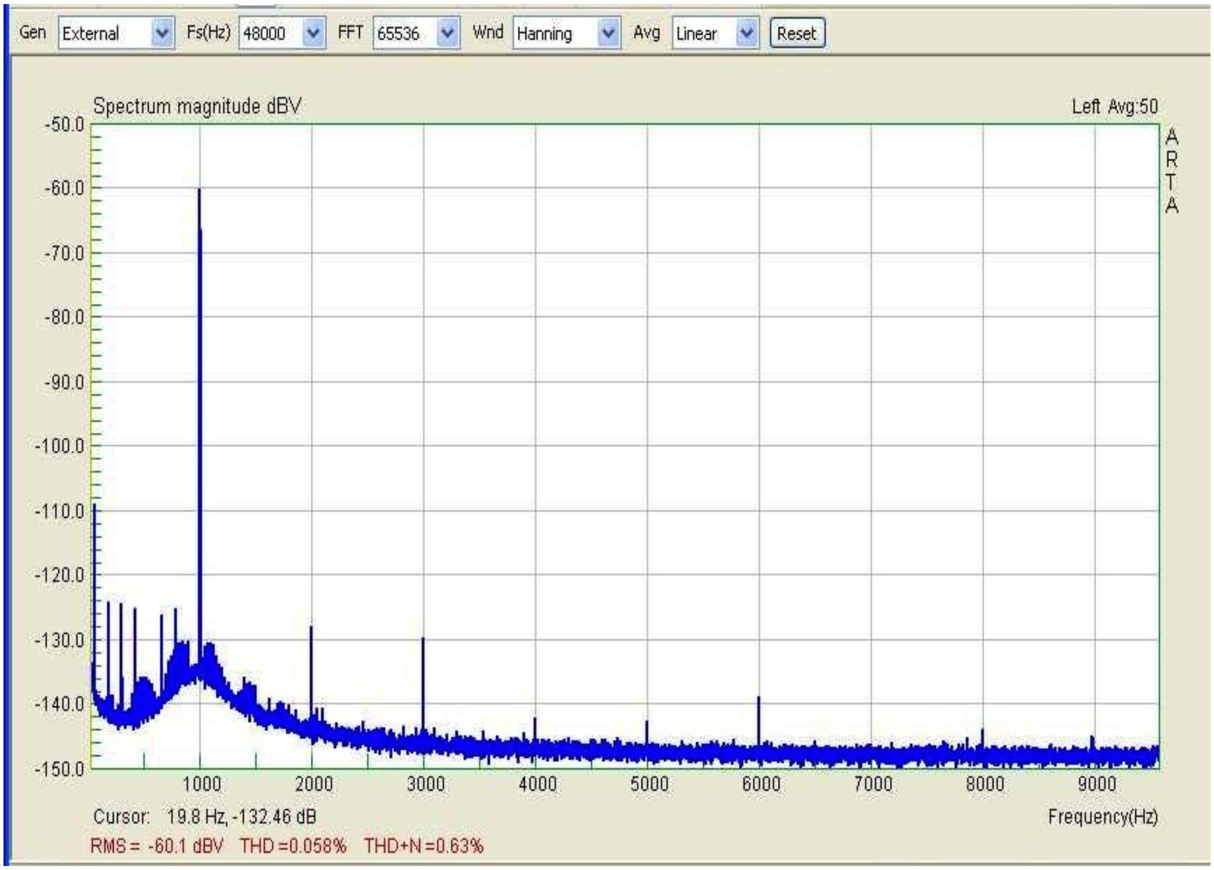
If you use the opamps I did, then you won't need the compensation caps C1 thru C3 -- but I did need a 2pF cap around R5 for stability at 110kHz.

Putting a scope on the junction of the emitters of Q3 & Q4 and adjusting R24 for equal size peaks of the two phases of the rectifier output, as Bob suggests, is probably the best way to get the best performance on all ranges.

I set the output buffer amp up to have a gain of about 6.6, which turns the roughly 1.6VRMS output of the oscillator into a full 10VRMS at the output. I used 5.76k for R32 to increase the amp's gain. I deleted switch S2A and R29, and used the Heath level pot in place of R30. And, of course, I did not use Bob's output attenuator, in favor of the HP design.

One last thing...
I've been fiddling with audio for almost 60 years, and in all that time, there's never been a minute that I didn't hate hum. All through this process, I've been dealing with power-line EMI issues -- you can see the low, but obtrusive and obvious line-related noise spikes that extend throughout the spectra at the top of the page. Yes, I'm using a toroid power transformer that has low radiated fields, but low doesn't mean none. That line noise has been bothering me for a long time now. A couple of years ago, I bought some mild steel electrical junction boxes in various sizes -- these provide OK magnetic shielding on the cheap. One of them was perfect to hold the Avel toroid, so I finally dug it out, bored some holes in the back of the BIG-18, mounted the toroid inside the junction box with the bolt through the back, and mounted that assembly to the rear of the oscillator. Here's what it looks like:

Here's what the spectrum looks like for a 1kHz signal at 10VRMS:



Compared to the 1kHz plot at the top of the page, most of the line noise artifacts are reduced by over 10dB or more, and are limited to low-orders. I like this improvement a lot. And the spectrum at the top of the page isn't as bad as it got, because I made those plots with the top cover of the oscillator removed -- putting that back on concentrated the magnetic fields and raised the noise level by more than 6dB -- so the improvement is even greater when the oscillator is in ready-to-use condition.

UPDATE -- I got around to adding a junction box cover plate between the box and the BIG-18's chassis. This reduced the power-line noise even further. While this simple magnetic shielding is far from perfect -- there are gaps and holes -- now all of the line-related noise spikes are below 1ppm, and most are well below that. But be aware that everything in the measurement chain affects or responds to any relative position, source, and strength of noise. EMI is pervasive, as is RFI, and a lot of fooling around is needed to get reliable and repeatable measurements. I'd like to be able to take the computer out of the chain, for example, but it is so convenient and helpful that for me, that is a non-starter.

Even though I didn't do it (so far, anyway), I like Bob Cordell's on-line suggestion of using a "wall-wart" transformer or small switching supply with pre-regulation, located at the wall outlet, to deliver positive and negative filtered and regulated power to the set of regulators inside the oscillator chassis. This separation of EMI noise sources would be very helpful.
###

From: http://www.moorepage.net/IG-18-3.html

Re-written 5-21-2012, updated 6-11-2012
Re-greening the Heathkit IG-18: IG-18 #3
The IG-339A oscillator, an HP 339A oscillator in new old clothing

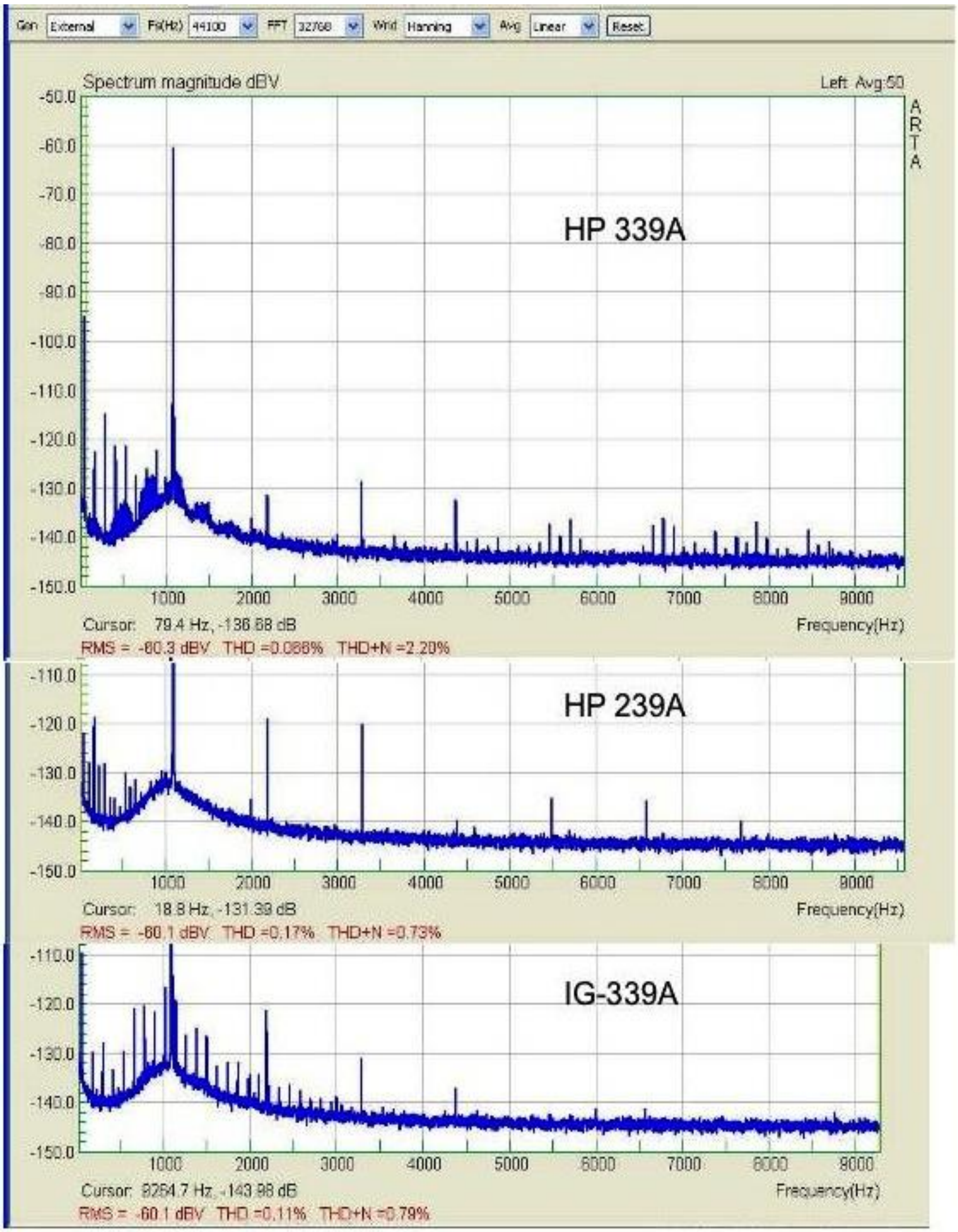
If you've visited this page before, you know I intended to describe the modification of a Heath IG-18/SG-5218 to significantly improve distortion performance. That project is now well underway completed. This page is almost as much about measurement and analysis as it is about modding an old Heathkit to work better.

UPDATE 6-11-2012
Where is the IG-339B mentioned on this page, below?
I've created a separate page for the B version of this oscillator. The IG-339B uses the 10:1 capacitor ratio of the Heath IG-18's range switch, which somewhat simplifies the construction of the oscillator. Click the link above to the B version page.

Simple magnetic shielding for the power transformer
Please note the end section of the B version page that deals with magnetic chielding of the power transformer. That shielding will also help this A version build a LOT.

Fixing a previously unnoticed and peculiar noise problem
I discovered a signal contamination problem by accident, and it was serious -- my PC somehow interferes with the Active Twin-T notch filter and the E-MU 0204 USB sound module by generating noise spikes starting at exactly 1kHz and proceeding up in frequency at 1kHz intervals. These noise spikes were hiding the true levels of 1kHz test signals and especially the harmonics at 2kHz, 3kHz, etc. The solution was to physically separate and re-orient the various boxes -- this makes my measurements much less convenient, but solves the noise spike issues.

What can you expect from this build?
I've edited some spectrum plots for easier viewing to compare the performance of the HP 339A, HP 239A, and the current IG-339A. The IG-339A's performance sits comfortably between the 339 and the 239 at 1100Hz. I moved the test frequency up slightly from 1kHz in order to be sure that I didn't have the signal and noise contamination discussed above.



Several things to note here -- the line frequency harmonics in the 339A and the IG-339A are spaced at 120Hz, while those of the 239A are spaced at 60Hz. I don't know exactly why; the 239 is the quietest of the three at these frequencies. All three spectra were taken under the same conditions of output level

(6.6VRMS), sampling frequency, and FFT sample size. I used the Active Twin-T notch filter to lower the fundamental level to -60dBV so that the calculated THDs can be compared. A little of the 1kHz-spaced products can still be seen in all three plots -- greatly reduced, but not eliminated.

I had modded the HP 339 a while ago with a pot to control the cancellation of the 2nd H. in the AGC JFET, the same as has been done to the IG-339A. I have not made this mod to the HP 239 yet. The 2nd H. cancellation is not as good in the IG-339A as in the HP 339. The two 339 variants have much lower 3rd H. than the 239 -- this may be due to the small design differences between the AGC systems of the 239 and the 339.
###

Introduction
Quite a while ago, my online friend Larry Burk sent me a link to Steve Lafferty's IG-18SL mod project page detailing his mods to use the circuitry of the oscillator section of the HP 339A Distortion Measurement Set in the IG-18. I'm going to call my unit the IG-339, with two versions: IG-339A and IG-339B.

I had previously written in 2011 that I was planning on modding an IG-18 to the circuitry of the HP 239A -- the 239A and the HP 339A oscillators differ only in some details of the automatic gain control (AGC) loop. But they are essentially identical and also measure nearly identically. So when I looked at Steve's great article, I was very interested in how far we had been going down the same road, including re-wiring the power transformer primary to get the right secondary supply voltages.

Then Larry told me he was etching and drilling a few circuit boards for Steve's project and offered to send me one, That's when I decided to use the 339 circuit instead of the 239. I really **hate** making circuit boards. Many thanks to both Steve and Larry for their good work. Please read Steve's pages carefully, since they have very good and clear descriptions of the changes to the IG-18.

On my other IG-18 webpages, IG-18 #1 and IG-18 #2, I modified Heath IG-18s bought cheaply on eBay. I made a few relatively simple mods for lower distortion in IG-18 #1, and made much more extensive mods for very low distortion in IG-18 #2.

Both projects ultimately were successful in achieving their goals. But evaluating those results requires some very accurate analysis equipment. The HP 339A, for example, cannot resolve the distortion of its own oscillator -- it has a measurement floor of about 0.001% which is the result of a relatively high level of 2nd Harmonic distortion in the notch filter circuitry. So how can distortion be determined?

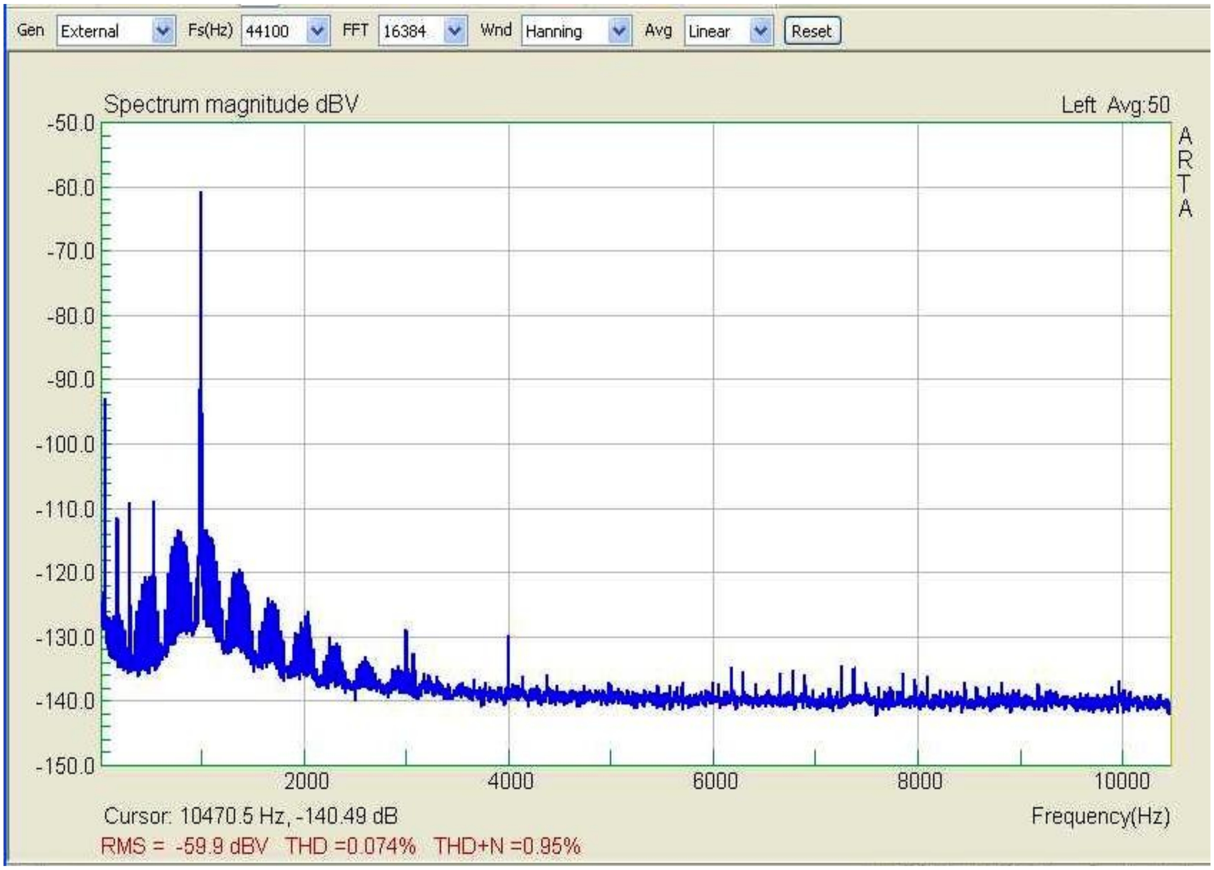
Test equipment for this project
A high resolution spectrum analyzer is essential, whether a software analyzer or a hardware analyzer. I use the spectrum analyzer function in the ARTA Audio Analysis software, which runs under Windows on a PC. Input to the PC and ARTA now is from an E-MU 0204 USB sound module. This unit has 24-bit ADC and DAC converters and a maximum sample rate of 192kHz, so it serves for my testing at up to 15kHz fundamentals with analysis of harmonic distortion products up to the fifth. It also has extremely low self-distortion, potentially under 0.0002%. The EMU is fed from a level control pot driving an Active Twin-T notch filter, which removes the fundamental input signal without significantly altering the harmonics in that signal. This improves the dynamic range of the distortion measurement to more than 130dB for audio band signals.

Any measurement errors?
Nothing important, once I cleared up the mystery about spurious signals and noise issues connected to the relative placement of the various parts and pieces of the test equipment and the computer. Please see the linearity information on the [Active Twin-T notch filter]() page.

As I've noted on other pages here on the site, higher harmonics than the 3rd are generally very small in very low distortion oscillators, so the roughly 90kHz upper frequency limit (approx. 1/2 the sampling frequency) of the EMU is not a concern. However, noise is, and the noise spectrum of the EMU begins to rise significantly above 40kHz, so resolving higher harmonics above that can be problematic. Finally, I use the Active Twin-T notch filter for highest resolution by notching out the fundamental while leaving the harmonics essentially untouched. Additionally, the extra 20dB of gain available in the filter helps to raise the distortion products above the noise. Given the very low self-distortion of the EMU (after you carefully optimize the input level to take full advantage of the resolution) you may wonder why you need the Active Twin-T. You'll see shortly.

So, why IG-18 #3? If we can get close to HP's level of performance with a modded IG-18, well, that's just great news. And what level of performance would that be? Here's a spectrogram of the output of the HP 339A oscillator at 1kHz at full output of 6.7VRMS (mine runs a little high -- cal should be 6.32VRMS for the essentially unloaded output), into the medium input Z of a level pot and the Active Twin-T notch filter. The 0dBV reference level is 1VRMS, and the input signal has been set to 1VRMS with the level pot used with the Twin-T filter.

I tuned the notch filter for 60dB of attenuation of the fundamental, so that the spectrum analyzer's calculated values for THD would have a meaningful reference level. To get the actual percentage levels, divide the displayed percentage by 60dB = 1,000. This means that the actual RMS THD level is 0.074%/1000 = 0.000074%. That's why you need the Active Twin-T. BTW, I prefer a linear frequency scale for distortion plots because it really helps to find products in a lot of noise if you know where they should be.



Note that the 2nd H. is buried in modulation artifacts and noise, and that the 3rd and 4th are around -130dB. The THD+noise level is higher because of the presence of line-frequency products and because of the actual shape and level of the noise curve at low frequencies. I generally disregard the THD+noise figures because of the spuriae that don't have anything to do with actual non-linearity.

I don't care what you compare it to, this is just an excellent oscillator. And the HP 239A offers essentially identical results. It's a shame that the analyzer section of the 339 doesn't even come close to the performance of the oscillator. Ideas, anyone?

IG-18 on steroids

Oddly enough, you can think of the oscillators in the 239 and 339 as more modern versions of the IG-18. All three use a Bridged-T bandpass filter for frequency selectivity. All use two banks of switched resistor decades and a dual potentiometer vernier for frequency control. All have four ranges and all span a total frequency range from <10Hz to >100kHz. All use relatively simple amplification schemes. But then the differences begin to appear.

The range capacitors in the IG-18's Bridged-T frequency selector/filter circuit have a 10:1 ratio between the bridge and pillar positions -- please see the IG-18 #1 page for full details.

But in the 239/339, the capacitors have a quite large ratio of 100:1, which means using 6 capacitors in the 339 to do that work of 8 (as described in IG-18 #1). And in the 239, there are actually 8 caps, which makes the push-button range switching used in that unit's design simpler. The 100:1 capacitor ratio means a filter notch and resulting amplifier peak of 34dB, and that means much more gain and bandwidth are needed from the amplifier, especially at 110kHz. But the benefit is higher filter selectivity, or Q, which means higher frequency selectivity, resulting in higher rejection of distortion products, which in turn means less distortion.

The 100:1 ratio is about as high as I would want to go, given the resulting very wide spread of capacitor values -- and to also yield reasonable values that work with the existing Heath bridge resistors. In the 239/339, the caps range from 5.6uF down to roughly 56pF (roughly, because of stray C effects). HP chose to use larger resistor values that require smaller cap values in order to minimize the loading on the HA2625 amplifier by the bridge network. With newer opamps like the LT1468, OPA1641, and LME49710, this wouldn't seem to be a problem, so I'm keeping the range resistors as-is in the IG-339A, and using larger caps.

The HP 339 AGC system uses a half-wave peak-detector acting as a sample-and-hold, an amplifier, an integrator, and an N-channel JFET voltage-controlled-resistor (Teledyne VCR2N) as a variable resistor in the AGC feedback loop for level control. The higher gain of this AGC system yields much tighter amplitude control and better operating point control, and therefore lower distortion than the lamp system of the IG-18 can achieve. An old friend used to call this kind of improvement "hanging a barn door with electronics." Gotta love the lamp, until you want ultimate performance.

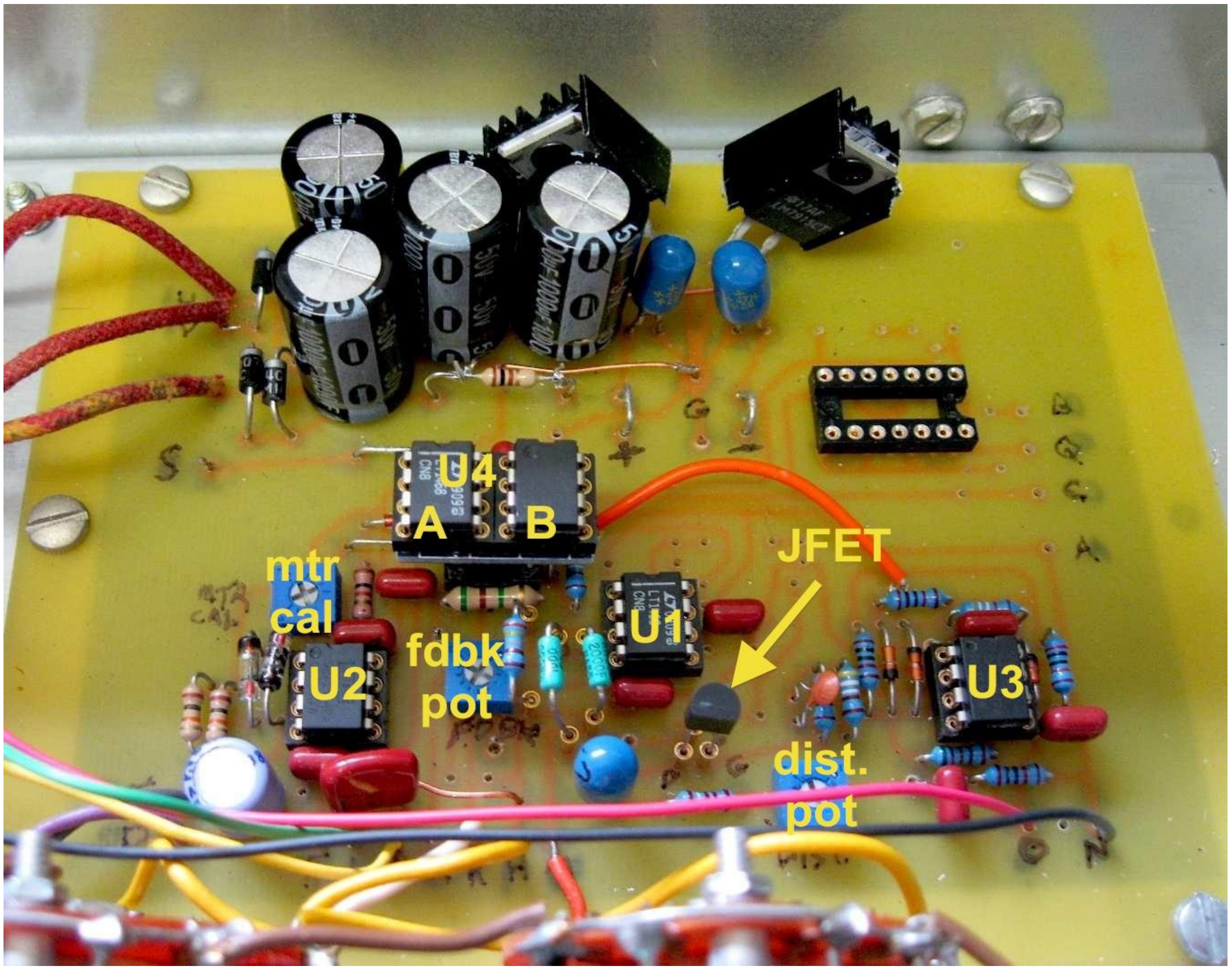
Some general construction notes

Given the IG-18 chassis, power transformer, and various switches and controls, and a proven working PC board, either version is worth building. These days, it's the mechanical stuff that's hard to find.

In my other web pages about the IG-18, I cast doubt on the Heath power transformer, accusing it of being a source of hum and noise. This is **not** the case -- it is very quiet, although line noise can be reduced in this circuit; but there is no need for a different power transformer unless you need to run off 240V line power.

I'll use Steve's part IDs just for the sake of simplicity. I chose to physically turn the power transformer around 180° so that it's secondary wires can reach the PC board without splicing in more wire. This may mean moving a bit more of the power cord into the chassis, depending on the length of the existing cord from the Heyco retainer to the other wiring.

Here's a pix of the PC Board with some parts called out. Note the three resistors in the middle, next to the feedback pot, mounted in socket pins -- these are the feedback leg and ground leg resistors associated with the JFET. The 1000uF power supply caps I used just barely fit. The blue "bulbs" are Tantalum caps. Note the dual-to-single 8-pin DIP socket for U4:



For the IG-339A, I added two decks to the range switch and used eight caps for the range switching, and built a separate board to hold the four larger pillar position range caps and one set of the AGC filter caps. See the detail pix below, in the Range switch section.

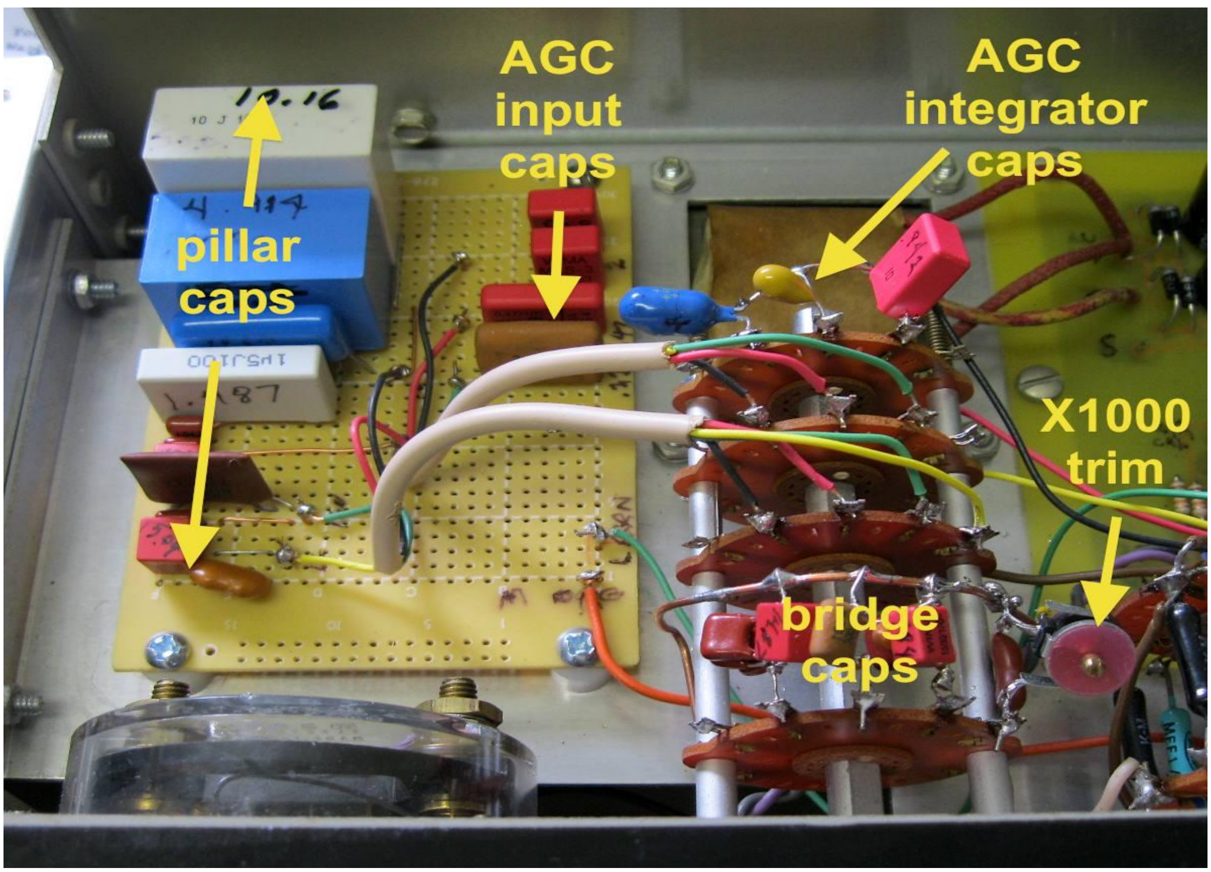
Power supply -- I increased the four power supply filter caps from 470uF to 1000uF, and changed R1 and R2 from 36 ohm 1W to 100 ohm 1/4W. I intend to use a separate +5V regulator for the square-wave section, which results in lower loading of the +15V supply. The higher value caps and resistors mean less ripple at the inputs to the +15 and -15 regulators, giving them less work to do in ripple rejection. Although I didn't do it yet, I recommend using LM317 and LM337 regulators for the +15 and -15V supplies -- especially for the negative supply. The 7915 negative regulator just does not perform anywhere near as well as the LM337 used with the appropriate resistors to set the output voltage. It's noisier and has worse ripple rejection, not to mention poorer regulation.

The rear panel of the IG-18 makes a very good place to mount the regulators, but be sure to insulate both/all of them from the panel to avoid grounding problems.

Parts HP didn't use -- Steve didn't indicate why he chose to isolate the drive to the AGC rectifier with a unity gain buffer -- the HP 339 and 239 don't use it and obviously don't seem to need it. But it doesn't hurt anything either, since this is not AC feedback but DC. But it is essential that the rectified signal not have a low-Z discharge path, so U4B really should be a FET-input opamp, so that the filter Cs 21-24 only discharge through the 5.1M R36. Steve makes this amp half of a dual and calls for an OPA2134 -- an excellent choice, but note that 1/2 of this amp is the output buffer, and an LT1469 may be better -- but there are issues -- I'll talk about this later.

Range switch -- The Heath range switch is a lot longer than it needs to be, which is good, because the easiest way to add the switching for the AGC filtering is to add two decks to the range switch. A handful of small spacers (brass, aluminum, or plastic tubing -- see your local hardware store) and a couple of decks scavenged from another old switch or two, or taken from new ones, will let you easily mount the needed decks. I had two complete Heath range switches, so I just took the decks from one and added them to the other. Most people won't have that option, but the switches that Heath used are a fairly common size with 30° rotation detents, which is the usual amount for many, many switches.

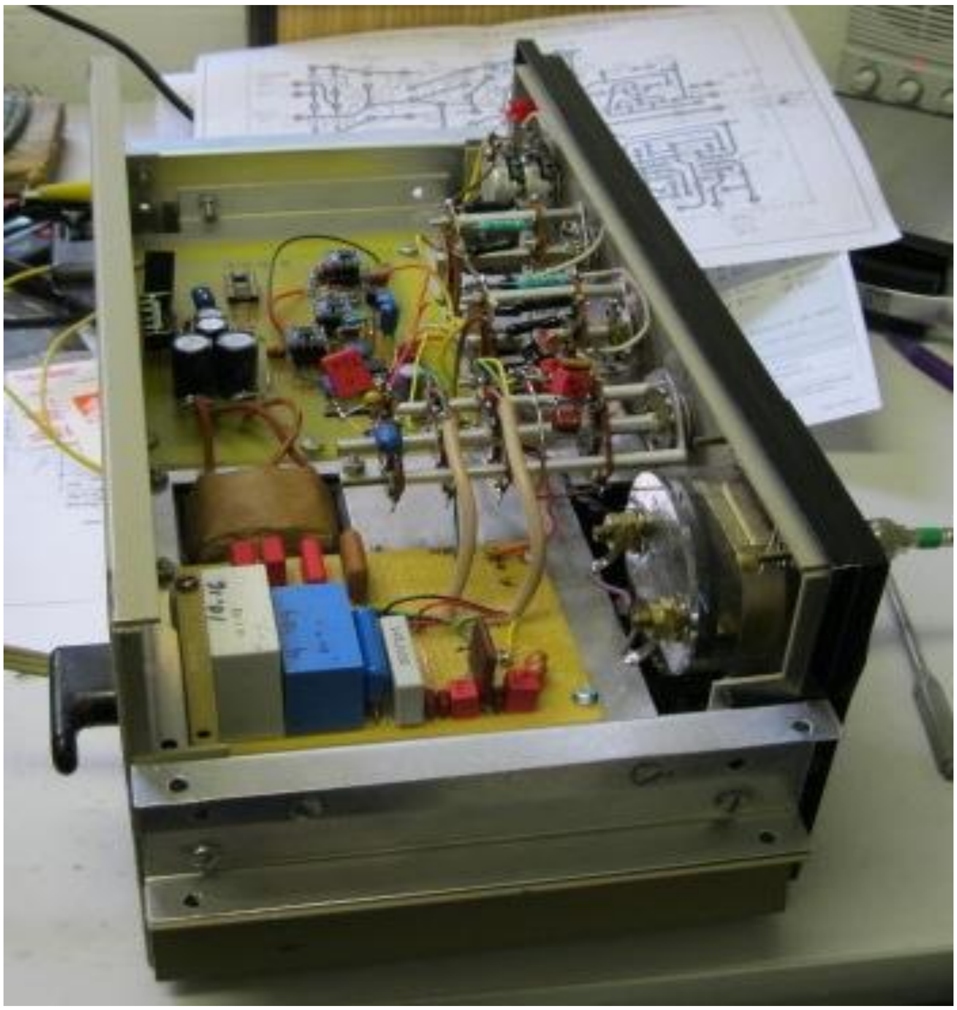
Here's a pix of the modded range switch as a guide. The four large pillar caps (3 caps for each one) really take up space. There are only 3 caps for each of the AGC filter sets, because one of the caps for each set is mounted on the PC board:



Cap values -- The biggest difficulty for most constructors will be selecting cap values. The needed caps are in the "16" range -- all are very close to 16 in value, with different decimal points. These need to be accurate to 1% or better. They are fairly easily made from parallel combos of 15 and 10 series caps. All of my pillar section range caps ended up being combos of three parts to get the accuracy and matching needed. A reasonably accurate cap meter or impedance bridge is essential -- the range caps must be selected to 1% or better for good results and stability, since any changes in the cap ratios mean changes in the circuit gain. An inexpensive 3-1/2-digit digital cap meter will work fine, since absolute accuracy is not so important, but ratio accuracy is. I use both a digital meter and an old but very accurate ESI 250DE RCL Bridge.

The IG-339A runs

I finished the main board and switches and made a board to hold the bigger film caps for the pillar portion of the bridged-T and for the film cap input filters for the AGC rectifier. The smaller bridge caps for the bridged-T are on the first deck of the 4-deck range switch and the AGC integrator caps in the U3A circuit are on the fourth deck. You can see how well the turned-around power transformer works out:



Then I powered it up and it oscillated at 1kHz. Of my two JFETs that work, a J105 and a 2SK152, the 2SK152 seems to be the best, and I was able to get decent operation at 100kHz, letting me adjust the trimmer cap I used in parallel with a 100pF silver mica cap to make the 160pF smallest tuning cap. But neither of these JFETs are the right parts for this unit. Larry Burk pointed me to Electronics Goldmine, who have some of the Teledyne VCR2N JFETs used by HP, which have a nominal resistance range of 20 ohms to 60 ohms -- should be ideal for this circuit, so I ordered a few. I also ordered some PN4091 and PN4092 JFETs from Mouser to see if one of them will work -- Steve Lafferty used the 4092. That's the good news.

The bad news is that gain settings that give stable operation across all ranges also have distortion a little higher than desired -- around 0.0005% or higher, depending -- but not even in the same ballpark as the real 339. The bipolar opamps LME49710 and NE5534 do not do well for U1 in this circuit -- I suspect their low input Z loads the bridge, but I'm not sure that's the cause. Just for grins, I plugged in an HA-2625 for U1 (used in the HP 339 and by Steve) but without any of the compensation parts, and it was really bad -- it had OK level control and fairly stable operation across all ranges, but tons of distortion -- this is altogether strange -- it wasn't oscillating and clearly was in a linear part of its operating range. The FET-input OPA1641 and the bipolar-input LT1468 work the best, with the OPA134 close behind. I initially used an OPA1642 (dual 1641) for U4, the output buffer and the AGC difference amp. I really don't know now why the FET-input amps work all right, if not great, but the bipolar-input amps have very high distortion. Mystery.

This would be a good place to mention the possibility of using an OPA1641 or an LT1468 as an amp whose low-Z output can provide a virtual common ground for the positive and negative feedback legs of the bridge, in the way that Jim Williams used one in a Wien bridge oscillator in his Linear Technology App Note 43, Fig. 47 -- please note that Fig's 47 and 48 are swapped -- 47 is the right one. This amp potentially greatly reduces common-mode error, and may be a very good idea.

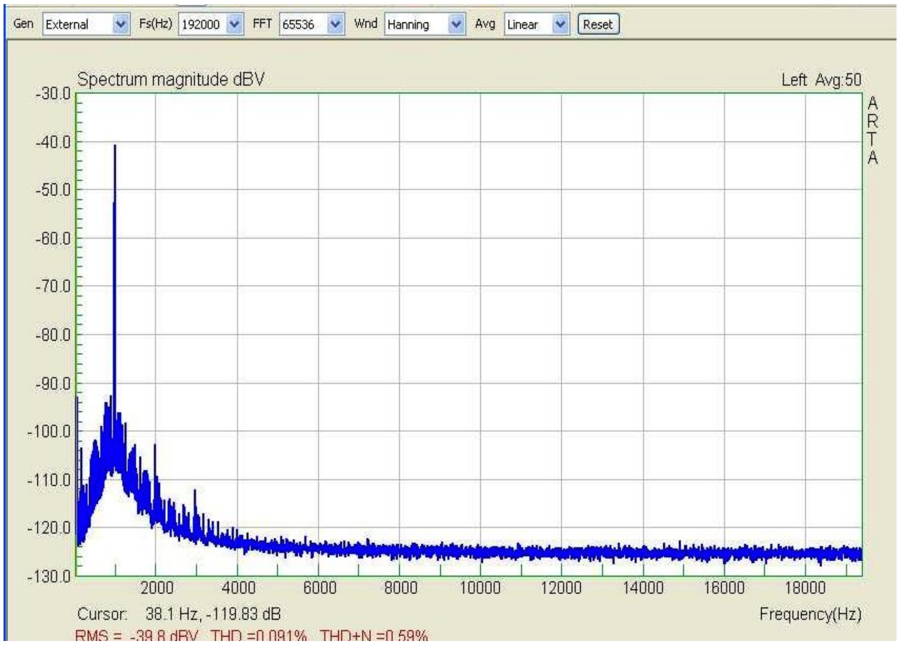
Parts arrive and results are good
Subbing in the VCR2N, PN4091 and PN 4092 revealed that the VCR2N and the PN4091 produce essentially identical results, and that the PN4092 works the best of the three. This is a little surprising but it's an OK result. Adjusting the feedback pot and the 2nd H. suppression pot (this pot is not in Steve's version -- it's been added by Larry and me) gave very good results at 1kHz. Then I added a 200pF cap around R50 in the AGC loop -- this resistor is 2k ohms in my unit, same as the HP value -- and that gave very good performance on all ranges.

Which opamps?
It was time to find out which amps would give the best overall results. All the bipolar amps except the LT1468 had already proven to be poor performers for some unknown reason, but I knew the FET-input amps all worked. Which amp works best? I swapped in various combinations of OPA134, OPA2134, OPA1641, OPA1642, and also the LT1468 -- I don't have any duals of this part (LT1469), but I do have a few dual-to-single converter sockets, so I could use two LT1468s for any spot where Steve used a dual.

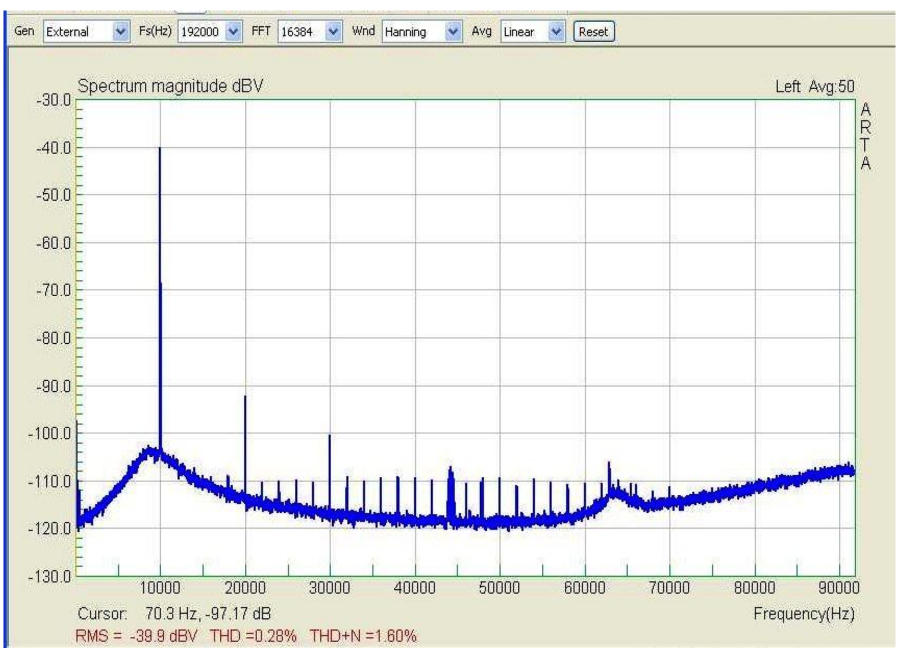
The results were clear -- the LT1468 is the best for U1 and for U4A, the output buffer amp. This was especially true at 100kHz, where my HP 339 measured the oscillator at 0.006% THD (-84.4dB) -- I don't have a hardware spectrum analyzer, so I couldn't make a high resolution measurement at 100kHz; but I have reason to think that the distortion is actually lower than that. My final opamp arrangement is:
U1 - LT1468
U2A & B - OPA2134
U3A & B - OPA2134
U4A - LT1468 or LME49710
U4B - OPA134

It probably would be most convenient to use a dual part for U4A & B, and the LT1469 may be fine there unless the lower input Z of this bipolar part causes problems at low frequencies due to bleeding the AGC input filter too quickly -- but it does work fine on the X10 range and above, and it seemed to work OK at 100Hz. The fallback is to use an OPA1642 or OPA2134.

Here's the spectrum at 1kHz, with 20dB of gain in the active Twin-T filter; this means that -100dBV is actually -120dBV. You can always ignore the cursor data in these plots -- I usually don't set it or use it, and it's just there. As you can see, the harmonic distortion products are all under -120dB, all under 1ppm:



And here are the results at 10kHz, all the same as above except for the sample rate, and including the extra 20dB gain. This means that the 2nd H. is at -111dBV, under 0.0003%, or 3ppm:



The 60Hz spike is at around -113dBV and the other line-related products are very small, which also reduces the modulation effects. I can't explain the 2kHz artifacts, all at the same level -- weird.

IG-339A wrap-up
The performance shown above results from a gate voltage on the PN4092 AGC JFET of about -0.7VDC at 1kHz, with an AC output of 6.75VRMS. This gate voltage will vary a lot with range switch settings, but should never be less than about -0.1VDC, and probably not more than -1.2VDC. The output amplitude variation is within +/- 0.05dB, but might vary a bit more, again depending on a variety of factors. I think it should always be within +/- 0.1dB -- but be sure your AC voltmeter is accurate at the frequencies being measured!

Because the output is metered, I didn't feel a need to accurately set the output level to 6.32VRMS (or anything else). If you want a different output level, you can vary either R34 or R35 to change the "reference" level that determines the output. Reducing the output level may result in lower distortion, particularly at 100kHz -- maybe.

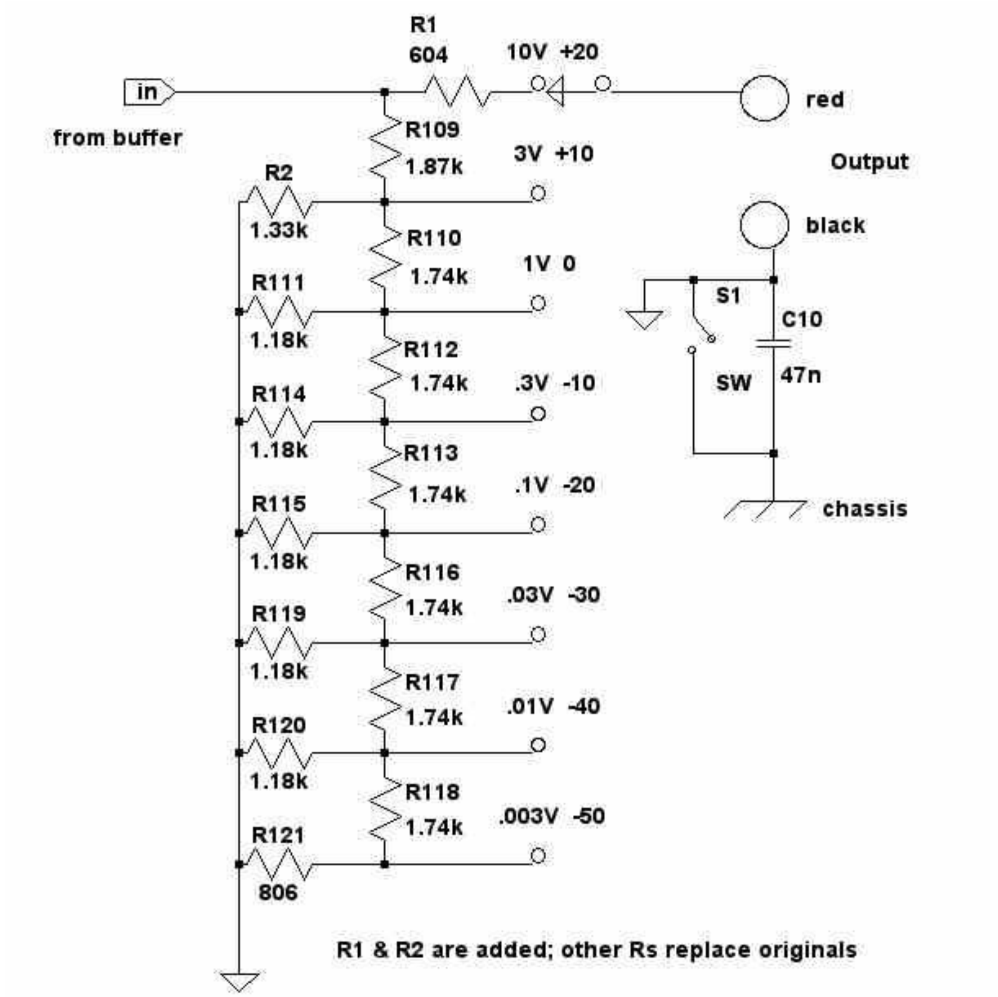
I briefly considered turning the output attenuator switch so that the max value is 3V (the 600 ohm terminated output level) and then setting the meter to full scale, and then making the lowest position an "off" position, but I decided that was overkill. But it would be easy to do. Alternately, you could make the unity-gain output buffer a X1.5 gain stage, raise the supply voltages to +/- 18V, and get a full 10VRMS output. This would be more attractive to me. All of the opamps used here can handle the higher supply voltages.

I used lowly 1N914s for the AGC network diodes and not the 1N4148 and Schottky diodes that HP and Steve used. Larry is sending me some Schottkys and I will try them when they arrive and see if they matter very much. I suspect that if they matter it will be at 100kHz, but since I don't have a hardware spectrum analyzer, it will devolve on others to figure out if they are really important.

I set out to try to answer the question of whether the performance of the HP 239 and 339 oscillators, which used the now-scarce HA-2625 opamps, could be matched using modern, readily-available, unity-gain stable opamps -- the answer is yes, and those amps are the LT1468 and the OPA1641. I've further established that the PN4092 works very well for the AGC JFET, and those are still available, too, although they are in limited supply. Once they are gone, some other form of AGC control or a totally different topology will be needed to equal or better their performance. I really wish I had a way to see high-resolution distortion performance out at the harmonics of 100kHz -- a lot could be cleared up.

Output attenuator mod
Here are the details on the output attenuator. This mod is not essential -- unless you want accurate attenuation and a known and accurate output impedance. All resistors are standard 1% values. This attenuator gives an output Z of 600 ohms at all settings, while presenting the unity-gain output buffer with an acceptable load even when the output is terminated in 600 ohms.

The triangle ground point and the top wire of the floating/chassis ground switch are connected to the ground wire on the attenuator switch that returns to the "clean" ground on the PC board at point M. The S1 switch is the re-purposed front-panel impedance selector switch for internal or external 600 ohm load. I used the Heath 47nF cap and left the chassis ground point of the cap where Heath put it, under the Z switch mounting screw, and soldered the bottom wire from the switch to the same solder lug the cap is soldered to. Seems to work fine, but I may run it's wire from the S point on the PC board. At the moment, line noise is low, well under -100dB.


R1
604
10V +20
in
red
from buffer
R109
1.87k
3V +10
Output
R2
1.33k
black
R110
1.74k
1V 0
R111
1.18k
S1
C10
R112
1.74k
.3V -10
SW
47n
R114
1.18k
R113
1.74k
.1V -20
chassis
R115
1.18k
R116
1.74k
.03V -30
R119
1.18k
R117
1.74k
.01V -40
R120
1.18k
R118
1.74k
.003V -50
R121
806
R1 & R2 are added; other Rs replace originals

From: http://www.moorepage.net/IG-18-4.html

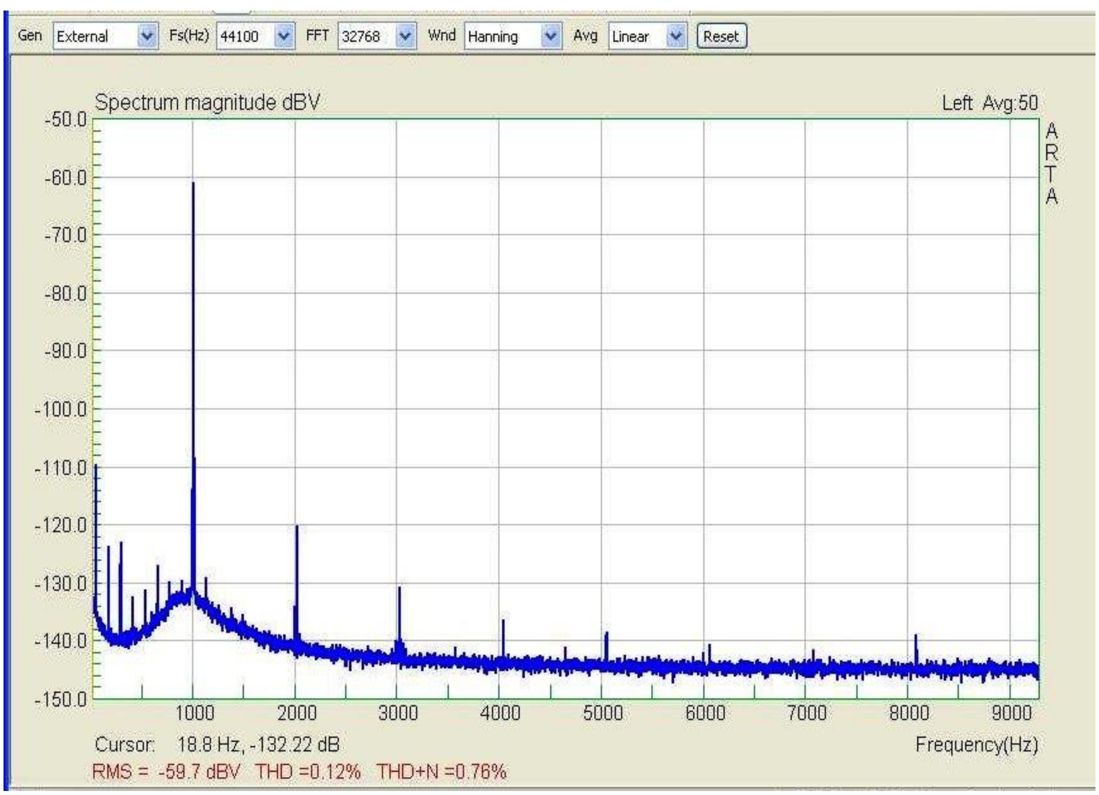
Created 6-14-2012
Updated 8-11-2012
Re-greening the Heathkit IG-18: IG-18 #4
The IG-339B -- an HP 339A oscillator using the IG-18's tuning network

What can you expect from this build? --
I have three spectra for you to view -- one at 1kHz fundamental, one at 10kHz, and one at 100Hz. The output signal from the IG-339B was 6.7VRMS for each frequency. This output was then adjusted to 1VRMS (0dBV) with an attenuator pot and fed into the analysis test system. The three plots were taken with the feedback and distortion cancelling pots unchanged from plot to plot, so I consider their settings optimized.

These spectra were plotted using my Active Twin-T notch filter to reduce the fundamental to -60dBV, while leaving the harmonics unchanged. The signal was then fed into an E-MU 0204 USB sound module connected to my PC. The ARTA Audio Analysis software then plotted the resulting spectra. The test system linearity is excellent over a 130dB dynamic range, and very good to -140dB. The sigma-delta analog-to-digital converter gives a rising noise slope above about 40kHz in the 10kHz spectrum, but the important harmonics show up well.

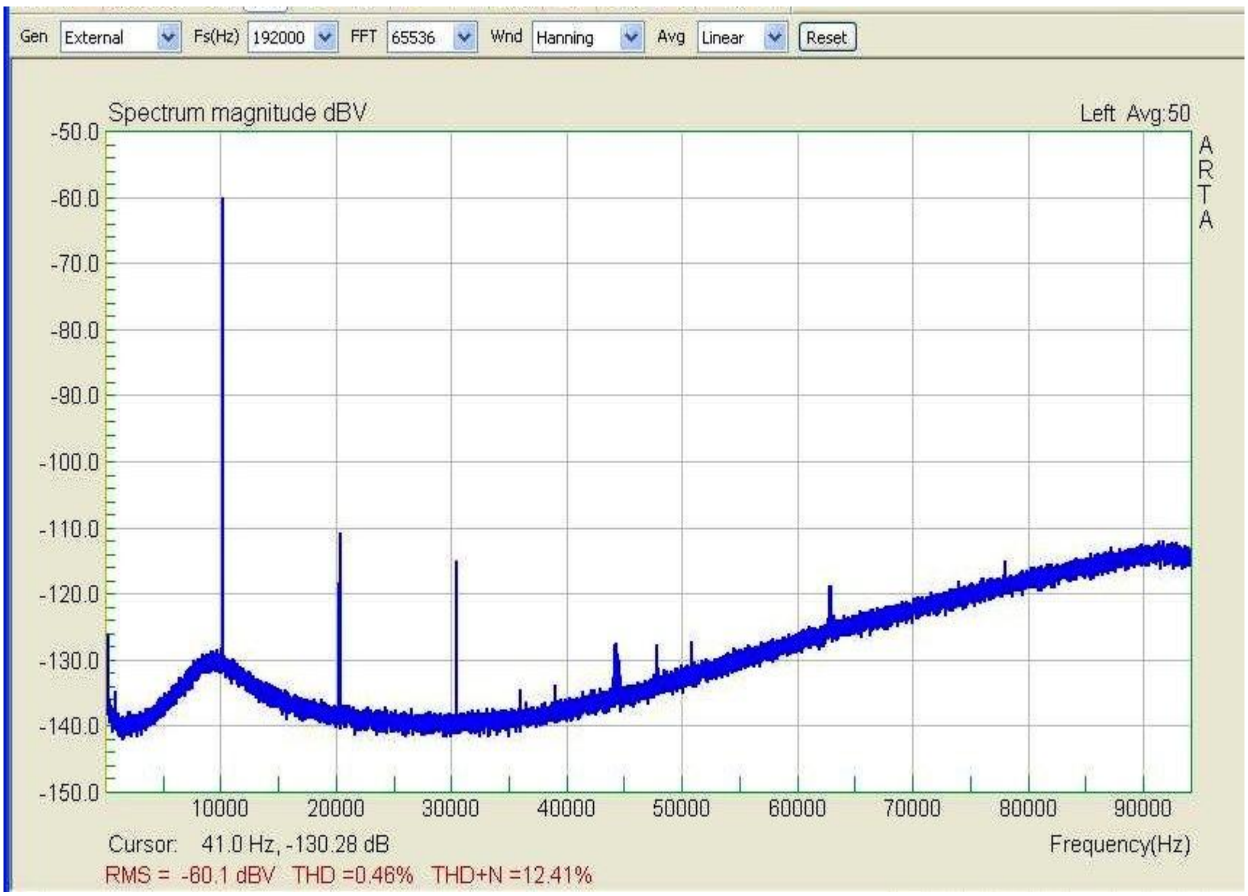
The system can capture harmonics up to around 88kHz, so it works fine for all important products in a 10kHz fundamental, and is OK at 20kHz, but not great, since the 5th H. is missing.

Here's the 1kHz plot:



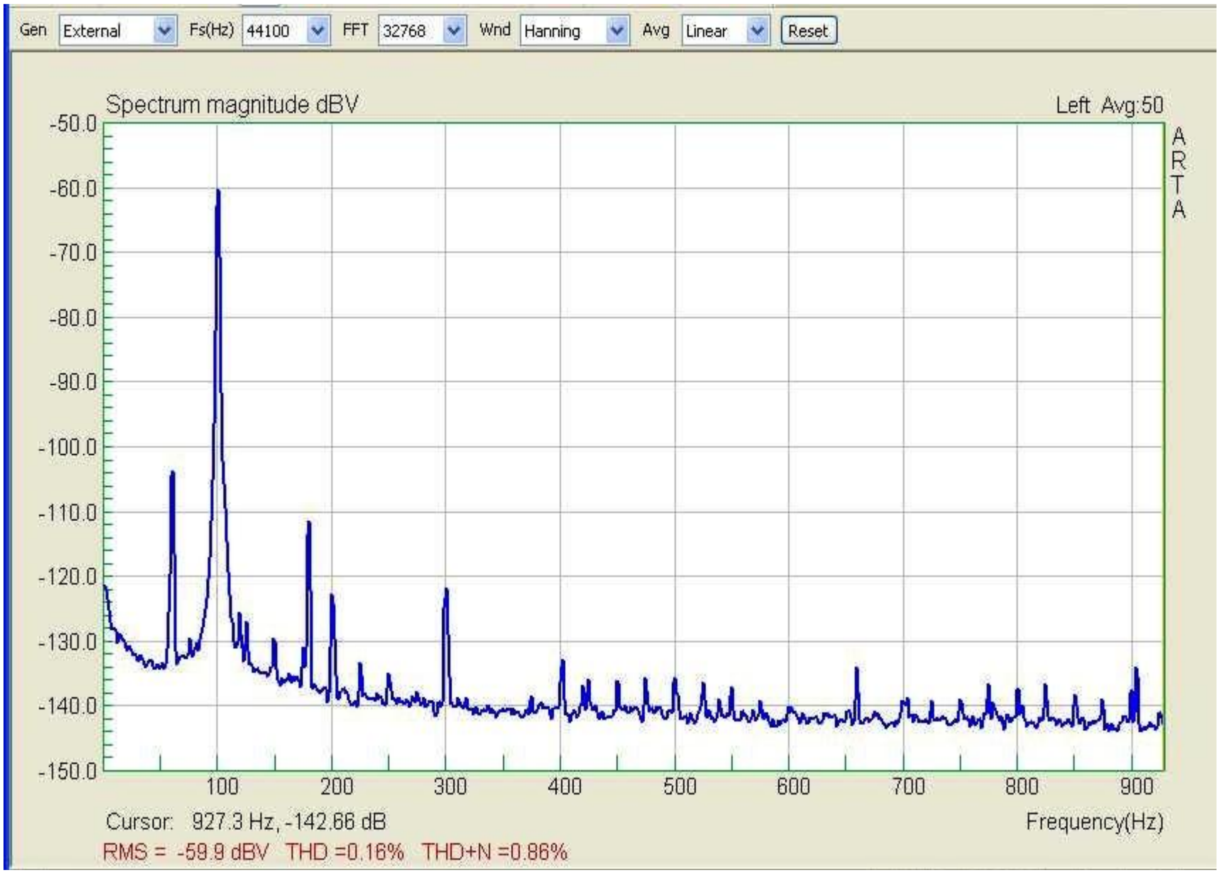
The calculated THD is 0.12%. Divided by 1000 to account for the -60dB fundamental, this means the distortion is 0.00012% referred to the 0dBV signal level, and that's really very good. The only significant harmonic is the 2nd -- the canceller adjustment has done its best, but it's not good enough -- if this peak could be reduced 6dB, then this version would be the full equal of the A version and the HP 339A.

Here's the plot for 10kHz:



The calculated THD is 0.46%, or 0.00046% referred to the 0dBV fundamental. Again, very good. And again, it's the 2nd H. that is the largest. The best canceller pot setting for 10kHz is also the best setting for lower frequencies.

And here's the plot for 100Hz, where line frequency products get in the way a bit:



The calculated THD is 0.16% or 0.00016% referred to the 0dBV fundamental. This number is inaccurate because the signal's 3rd H. at 300Hz falls under the line-frequency multiple of 300Hz. A slight retuning of the oscillator to say 110Hz would reveal that the distortion is about the same as at 1kHz or a bit lower even.

The goal --
This project was intended to explore the relation of the Bridged-T tuning capacitor ratio to overall performance. Much of the work on the IG-339A is germane to this "B" version of the HP 339A oscillator. Please see that page for discussion about the HP 339A oscillator and the build of an IG-18 that uses the circuit and tuning cap ratio of the HP 339A. Since the PC board I used for the two IG-339s is based on Steve Lafferty's IG-18SL mod project page, please read his excellent report on his mod of the IG-18 to the HP 339A circuit -- I use his part numbers, which are also mostly those used by HP, in discussing the variants.

Importantly, the HP design uses a very large 100:1 ratio of the tuning capacitors in it's Bridged-T frequency selective network to achieve a relatively high-Q peak in the oscillator's amplifier. This has some advantages in controlling the bandwidth of the oscillator to help reduce distortion. But it also requires much higher open-loop gain and unity-gain bandwidth from that amplifier, because the amp's gain needs to be around 34dB. This means that the feedback margin is lower at higher frequencies. With a maximum frequency of 110kHz, does this matter for this oscillator?

As noted on my other IG-18 mod pages, the Heath IG-18 uses a 10:1 tuning cap ratio for it's Bridged-T network, which only requires an amp gain of about 16dB, increasing the feedback margin. Using the

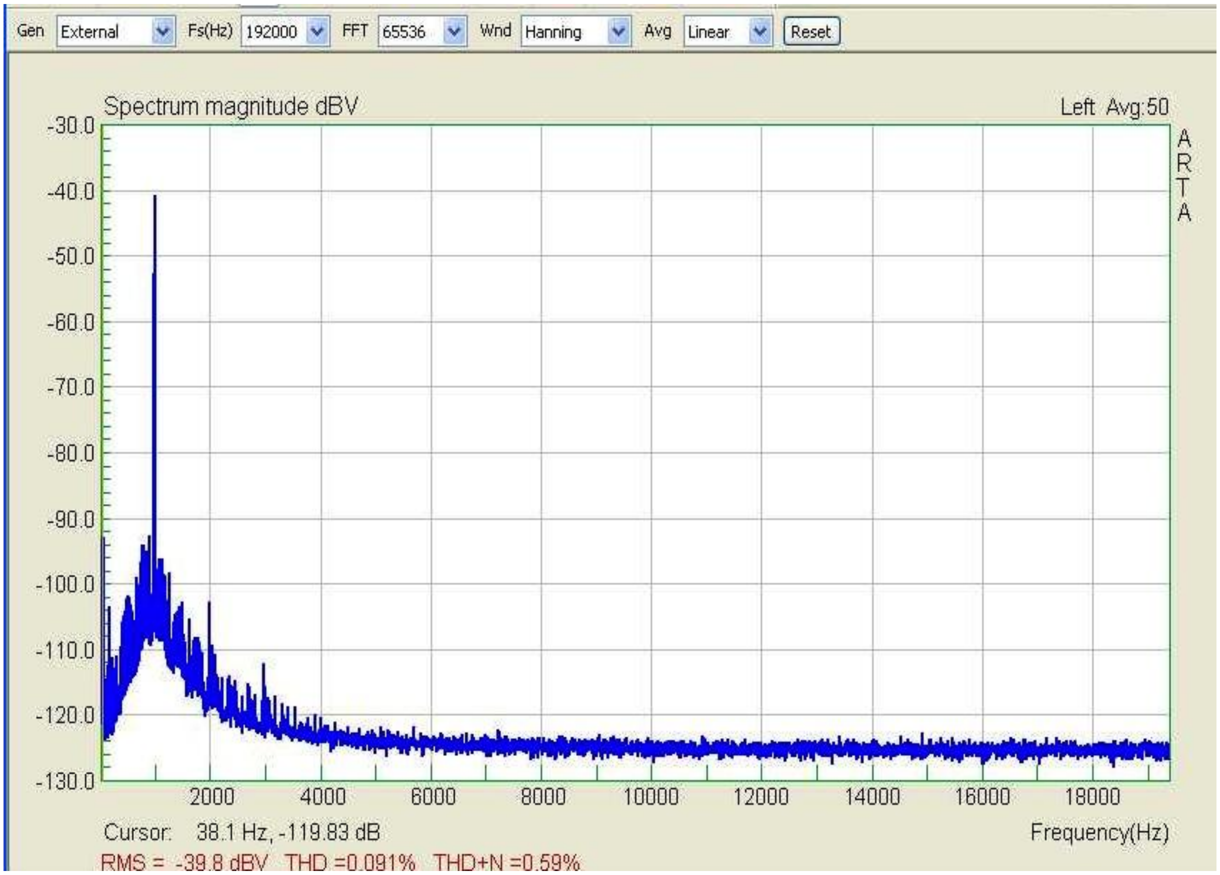
Heath 10:1 cap ratio provides some other benefits -- the Heath 2% tolerance caps can be used, as well as the Range switch with its clever switching that uses 5 caps to do the work of 8. However, the Range switch needs two more poles in order to do the AGC filter switching needed for the HP 339 circuit. But I will have suggestions to make about this presently.

A trial run --
The IG-339B 10:1 cap ratio system ran nicely from 300Hz to 100kHz during the build of IG-339A. Some resistor values on the PC board needed to be different from those in the 339 circuit used by Steve Lafferty in his IG-18SL mod because of the big change in gain. I added some socket-pin receptacles so that I could easily plug in different resistors and different JFETs, and all my opamps are socketed. I'll have more to say about the IG-339B parts values later.

Do the lower tuning cap ratio and the higher signal level on the AGC JFET really make a big difference? --
Having the 10:1 cap ratio implicitly means that the gain of the oscillator amp drops from 34dB to 16dB, from 51 to 6, which means a different, higher resistance range is needed for the JFET in the AGC system -- I've chosen to keep the feedback leg resistance pretty close to Steve's and HP's value of roughly 4k ohms in order to minimize loading on the oscillator amp and reduce distortion.

But this change in gain means that the ground leg resistance of the feedback network has to be much larger than in the IG-339A -- roughly 750-800 ohms instead of about 80 ohms, and it also means a higher resistance compliance range is needed for the JFET. It also means that a higher AC signal level will be on the drain of the JFET. It has been reported that lowest distortion results from the lowest possible AC signal level at the JFET's drain -- the effect of this is to minimize FET channel-resistance modulation. I can say that I find this *may* be true, but how *much* does it matter?

A partial answer --
Here's the IG-339A version at 1kHz, which closely matches the HP 339A. Note that the plot level has been boosted with 20dB of gain in the Active Twin-T, so the important number is the calculated THD, 0.000091% referred to the 0dBV fundamental:



The 2nd H. is about -123dBV, and the 3rd H. is about -133dBV. These products are each about 3dB lower in the IG-339A than they are in the IG-339B plot at the top of the page. That is not a huge difference in performance.

Given the results above, why did HP use such a large tuning cap ratio? --
An interesting question. Clearly HP engineers thought it mattered, but the excellent Krohn-Hite 4400 and 4402B use a Bridged-T circuit that has a 2:1 ratio of the bridge and pillar components, and their performance is said to be in the same area as the HP 239 and 339 -- I actually have a K-H 4500, but it's circuitry is somewhat different and it only gets into the 0.001% area at 1kHz. The answer to this question can only be guessed at. On paper it clearly seems to be a good idea, and if getting the last ounce of performance out of the oscillator is important then it seems to be the right thing, at least a low frequencies.

But the HP engineers didn't have the wonderful range of opamps available to us today -- the only amp that would work well was the incredible Harris HA-2625, and HP used it to great advantage. It may be that the HA-2625 works best with the large tuning cap ratio. This is a hypothesis that I won't be testing. Steve Lafferty used the HA-2625 in his mod, but that amp requires a boatload of compensation components for stability, and the great new amps like the LT1468 and OPA1641 do not require these elaborate networks.

Sorting out the final parts values --
Please see my IG-339A page and to Steve Lafferty's page for reference to the changes here. Steve used 5% resistors in his mod and changed some values quite a bit compared to those used by HP (the HP

339A circuit diagrams and manual can be downloaded as a big PDF file from the BAMA test equipment manual website; there's also a copy on the Agilent manuals website, but that copy is missing the oscillator section!).

I have used the HP parts values pretty much uniformly, unless I just didn't have the HP values; then I subbed in different values -- this only happened with AGC filter caps. Steve's mod uses several dual opamps, and that's fine, but I used a two-single amp to dual converter socket in testing to see which amps worked best in which locations. the opamps used in this version are the same as those for the A version, except I've ended up using an OPA1642 for U4. So this IG-339B actually is the IG-339A re-configured to use the 10:1 Heath tuning caps and different resistor values for R52 and R53 -- everything else is the same.

Keeping the existing feedback-leg components meant their center value resistance was 3.9k -- I use a 3.4k fixed resistor for R31 and a 1k pot for R30 -- which means for a gain of 6 (about 16dB), the ground leg resistance would need to total around 780 ohms. Right away, it was clear that because of the relatively high resistance compliance range needed for the JFET, the choice of JFET is important. I can happily report that the PN4092 works just fine.

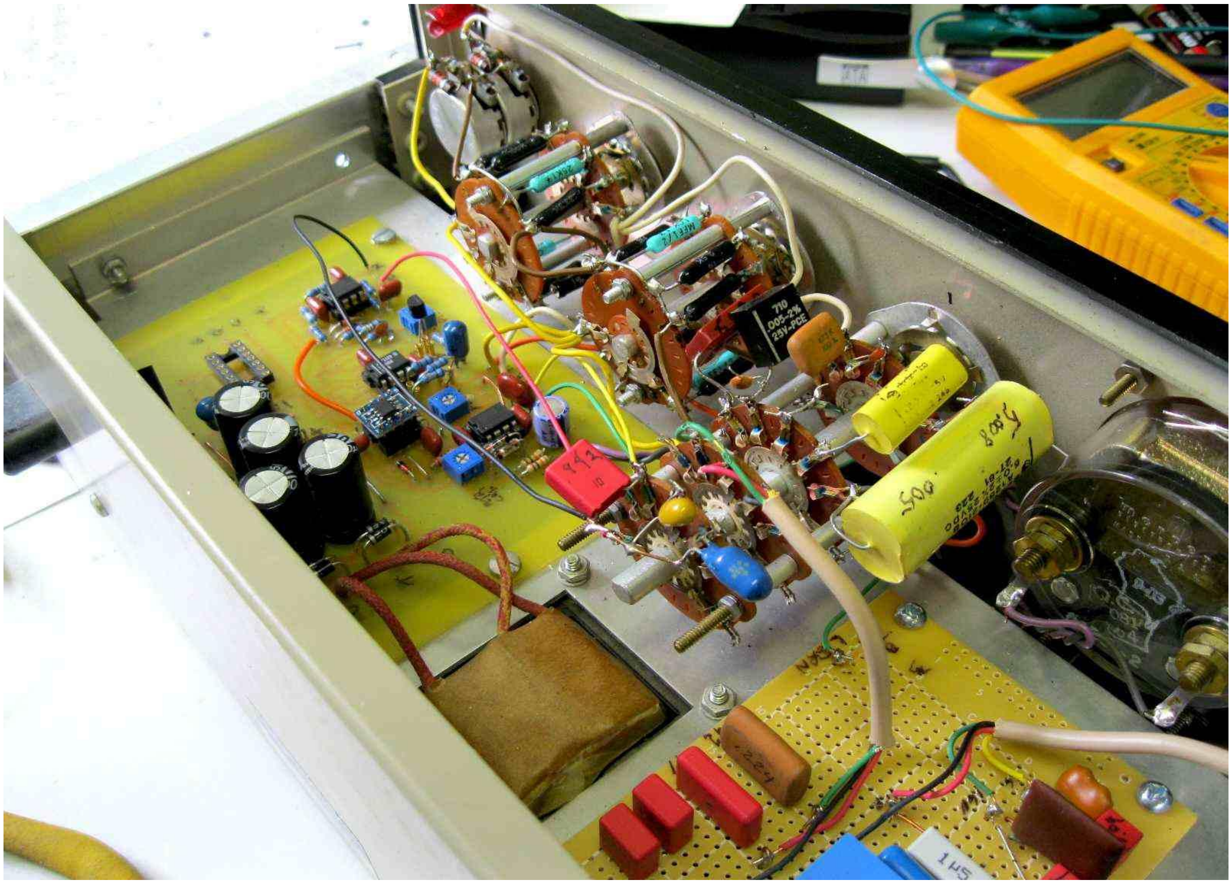
The resistors that limit the compliance range of the JFET, R52 and R53, obviously had to be larger. I initially thought that just raising them each by a factor of 10 would work, but after much fiddling, I settled on using 1.58k for each -- use 1.6k if you intend to use 5% resistors, but you may need one size bigger or smaller for either resistor depending on tolerances.

The distortion cancelling pot can reduce the level of 2nd H. distortion. It is placed in series with either R50 or R51, which are equal at 2.7k in Steve's build and 2.00k in HP's design, but one of them can be made smaller by 1/2 the pot's value. then the pot can find the best distortion cancellation of the JFETs channel resistance modulation, which reduces the 2nd H. distortion. You will need a spectrum analyzer to see this, but then you'll need the spectrum analyzer to actually measure the performance anyway -- whatever hardware THD analyzer you have probably won't be able to resolve it.

Building the Range switch --
As noted, the stock Heath range switch has two poles (well, not exactly -- it is a complex switch) and four positions, so the modded switch needs four poles -- two for switching the 5 tuning caps, as designed by Heath, and two for switching the two sets of AGC filter caps. I luckily had an extra Heath Range switch, so I just added its two decks to the existing IG-18 Range switch. But I had to reduce the spacing between decks. I had a bunch of aluminum spacers in my odds and ends, so I used what I had -- a trip to the local hardware store will likely yield up a great assortment of aluminum, brass, and plastic tubing and spacers that can be used.

The extra one or two decks can be scavenged from an old -- or new -- switch that has similar diameter decks with 30° rotation detents. A single deck with two poles and four positions will do nicely, IF the rotors align with the Heath frame so that the right contacts make and break, or you can use two decks, each with at least one pole and at least four positions, again making sure about contact alignment. Of course decks with more positions will make it easier to get an alignment that works with the Heath switch frame and original decks -- this is not a trivial issue, so two single pole, 12-position decks will give you the best shot at getting things to work. The alternative is to build a completely new Range switch, which is not as good since you will need two sets of four caps -- 8 in all -- instead of just 5 caps.

Here is a pix of the Range switch. See the chassis and PC board pix on the IG-18 #3 page for comparison. Note that the AGC filter caps are the same as in the IG-339A, and that the dual-DIP adaptor socket has been replaced by the OPA1642 opamp on a SOIC-to-DIP header:



Adjustments --
The adjustments are the setting of the feedback pot, the setting of the distortion cancelling pot, and setting the meter cal pot. The feedback pot should be adjusted at 1kHz for a DC voltage on the gate of the AGC JFET of about -0.6V. Then the range switch should be moved to each position and the gate voltage checked -- it should never be less than about -0.2VDC, but could be much higher, as much as -2V or more. The actual voltage depends highly on the precision of the cap tuning ratios -- the caps should be sorted to 1% value match or better.

For example, it is easiest to measure the Heath caps, find the biggest, then match the values of the others to 3 significant digits by adding smaller caps in parallel -- if your biggest is 508nF, for example, then pad the others up so their values are close to "508" too. With 508 as the value to be matched, for example, 1% matching will mean values from about 503 to 513.

I made the huge mistake in this unit of padding a couple of the Heath caps with some Chinese ceramic caps. The result was horrible 3rd H. distortion -- 20dB higher than now. I'm guessing they have a strong piezoelectric effect. So only use film caps (polypropylene is my fave, or mylar/polyester, or polystyrene) or silver-mica caps for padding the original Heath caps.

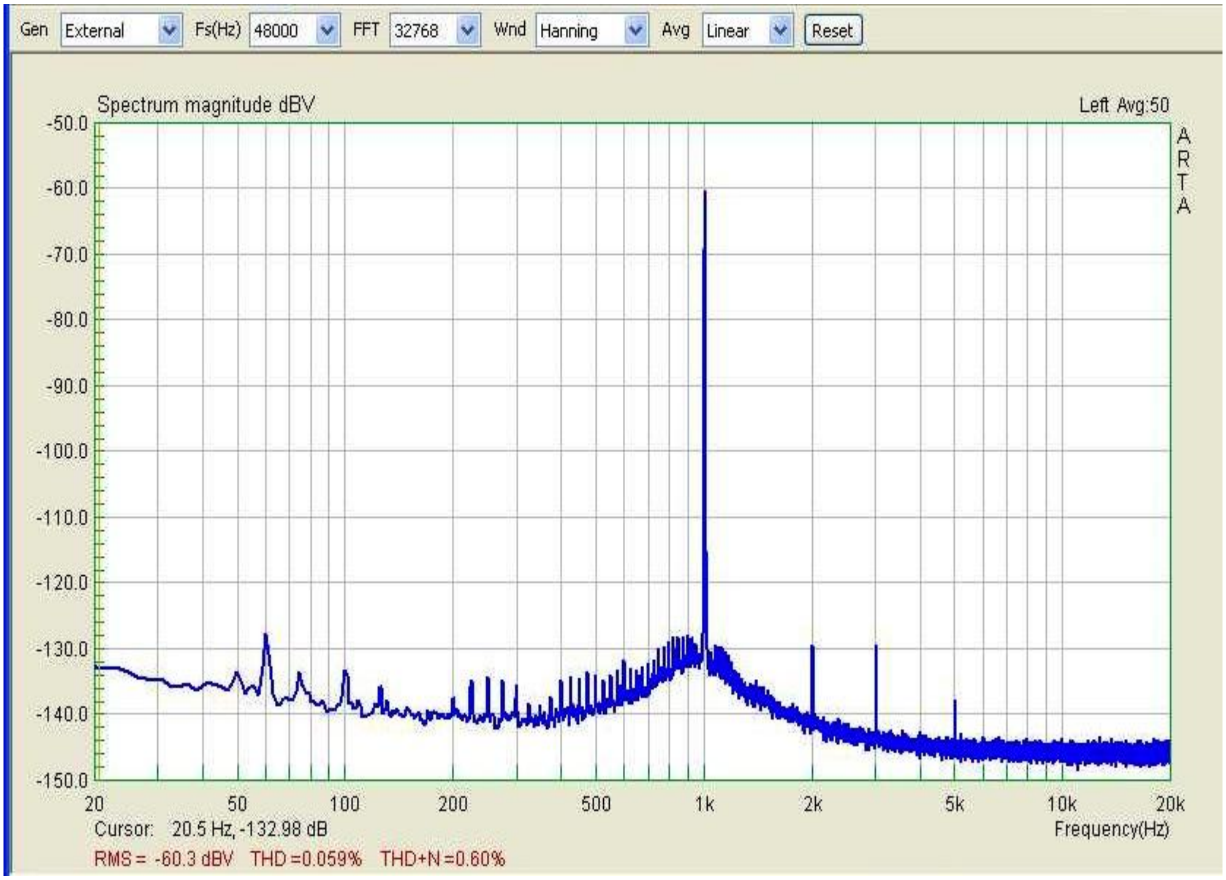
The distortion cancelling pot should only be used if you have a test setup than can see the level of the 2nd H. -- otherwise, make R50 and R51 as equal in value as you can.

IG-339B wrap-up --
After fiddling with resistors, JFETs, and the feedback pot, the overall performance of the B version has distortion under 0.0005% from under 100Hz to 10kHz, and I also think that lower frequency distortion will be well under 0.001%. This unit has 100kHz THD of 0.015% as measured by my HP 339A, while the HP 339A measures itself at 0.012%. You can decide whether keeping the caps from the IG-18 (and padding them for consistent value, if needed) and then adding two more decks to the range switch is overall easier than just building the A version as I and Steve did.

I think that the overall performance of the B version is amazingly good. But what about the questions of the importance of the large tuning cap ratio and the signal level on the JFET's drain? I think the importance of the large tuning cap ratio is a non-issue -- I just can't see that it matters at all. But I do think that the slightly higher levels of distortion in the B version compared to the A version is due to the AC signal level increase on the JFET's drain, especially the relatively higher 2nd H. where we might expect the 3rd to be the highest.

Update 8-11-2012 -- Simple shielding for the power transformer
As noted in the page for the BIG-18, IG-18 #2, using simple shielding for the power transformer greatly reduces line noise artifacts in the oscillator's output. So I decided to try the mod with this unit. The Heath power transformer mounts fairly easily in a 3-1/2" x 3-1/2" x 2-3/8" electrical junction box with a cover plate. I put 1/2" long 6-32 threaded aluminum spacers over the transformer's mounting screws, drilled mounting holes in the back of the junction box and used four 6-32 x 1/2" screws to hold the transformer in the box. I added wires for the primary connections and extended the secondary lead lengths to reach the PC board. I drilled holes through the back of the IG-18 chassis in order to use the junction box's cover screws to hold the junction box to the back panel of the IG-18. The junction box cover neatly sandwiches between the box and the IG-18's back panel. A 3/8" hole through the box cover and the back panel lets the wires through to the inside.

This is pretty crude magnetic shielding by any measure, but the results are very, very good. The spectrum below shows that the overall level of line-noise artifacts is really low -- most of the spikes have been lowered by 20dB or more, compared to those in the 100Hz distortion spectrum shown at the top of the page. Such low noise, combined with any high-pass filtering in the analysis system designed to reduce the effects of hum, will just completely remove line noise from THD calculations at 1kHz and above. The spectrum has a log frequency scale to better show the line noise:



I think this is a remarkable plot (and the 1kHz distortion performance is pretty remarkable too). It might be just as well to use a "wall-wart" supply and feed pre-regulated positive and negative DC voltages to the board's DC input pads. But this was a pretty easy and inexpensive mod to make and the results justify the mechanical work needed to make it happen.

But why is the distortion lower too? Two reasons -- 1) It may be because I replaced four of the five original Heath mylar tuning caps with polypropylene and silver-mica caps -- all except for the 5uF cap for the X1 frequency range; I don't have a big polypropylene for that yet. The quality of the caps really does matter, it seems.

And 2) I inserted a 47 ohm resistor between U1's output on pin 6 and all the rest of the circuitry. I had to cut a board trace to do this but that was easy enough. This isolating resistor gives the opamp a little easier time when driving the tuning circuit's capacitive load. My friend David Barber has suggested that this resistor could be as large as 100 or 200 ohms for even better results.