Edited 8/09: Major, major, technical additions

Chapter 11 – Upgrades, Modifications, Parts Information, Resources

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Introduction

This chapter is intended to pass on information learned by the reflector group and others to aid owners in determining what direction they need or want to go.

Some owners wish to keep their receivers just as they came from the factory, which is fine but not necessarily prudent, as time has shown that some parts can become real killers. They will probably want to skip this chapter.

Others wish to include improvements to solve some of the "quirks" that were in the receivers when manufactured or replace certain "killer" or less than desired quality parts. Or to replace original parts that have reached the end of there practical life after 50+ years.

Then there is a group which wishes "wring out" every bit of performance from their radios by taking advantage of the last 50 years of component improvements. Or using components that have been invented since the radios rolled off the production floor. Their mantra is "There's never been a tube radio made that I can't make better".

Special mention must be given to Dallas Lankford who well could be the Nicola Tesla [The true inventor of radio] of boat anchor radio experimenting and upgrading. With a lifetime commitment to search for improvements and sharing them with others Dallas has written numerous articles not only posted on his web site but for the NRC, and the HSN as well. His combination of technical expertise combined with a down home writing really draws one to his articles. He maintains a "solid gold" web site at www.Kongsfjord.no

To these people I say "Rejoice Pilgrim, heat up them there soldering irons, get out the good parts catalogs and let 'er rip, this chapter is for you!" No matter how fine a new R390A was when it was brand new, information in this chapter will make you "A" far better than any when new. Go for it!

This data has been added on a "ad hoc" basis and the order of listing has no relevance to the importance of the information. First time owners/readers are advised to check the table of contents carefully as the are certain parts that pose a risk to the owner or are likely, if not replaced before operating the receiver, can cause serious (read: expensive to fix) damage.

Some material was originally posted in the "Pearls" which is graciously maintained by W. Li.

Rational for receiver decisions beyond the scope of this manual may be found at the R390A Pearls site.

This chapter and chapter 9 will probably have more pages than the chapters 1 thru 8

This chapter because of its nature will never truly be done, additions, corrections and new material keeps adding to the chapter. It will be edited and revised as time is available then posted on the usual sites.

R-390/R-390A AGC - Ending The Moment Of Silence: The Way They Should Have Done It

By David Wise

Edited by Perry Sandeen Digital figures courtesy of DW Holtman

This easy front-panel-only wiring change adds a terminal strip and a couple of caps and resistors, and replaces the AGC switch.

- No more silence or blasting
- No side effects
- Compatible with all IF decks, no deck changes needed
- Can be done on the R-390 but not quite as neat mechanically

PRINCIPLES

Collins designed three AGC rates into the R-390, and carried the circuit forward to the R-390A with no change. FAST is a couple of 0.1uF caps, for a time constant of about 0.015 second. MED is a 2uF cap, which gives you about 0.5 second. SLOW multiplies the 2uF cap by about 15* by stringing it across a triode, from plate to grid. Any time the grid moves, the plate moves in the opposite direction about 14 times as much. With 15 times as much charge moving through the cap, to the AGC bus it looks 15 times as big.

* Please understand that due to the nonlinear transfer characteristic of V506, this number is approximate.

This is the next-cheapest possible way to get the SLOW time constant. Since the R-390 IF and RF amplifiers do not load the AGC bus, they could have simply switched in a big series resistor. Either grid current was a concern, the Carrier meter created a compelling argument for the time constant tube, or they anticipated the R-390A's RF deck voltage divider, R201 and R234.

ORIGINAL DESIGN (Figure 1)

Refer to Figure 1, an IF deck schematic excerpt from the Y2K Manual. The cap, C551, has one end connected to V506A's grid, which is also the AGC bus. The other end goes out to S107's rotor on the front panel. There, the MED position grounds it while the SLOW position routes it back to V506A's plate. The plate goes to B+ 205V through R549 (82k).

When the grid is at 0V (no signal), V506A conducts about 2mA of plate current, pulling the plate down to about 27V. At -20V AGC, the tube is cut off and the plate is at +205.

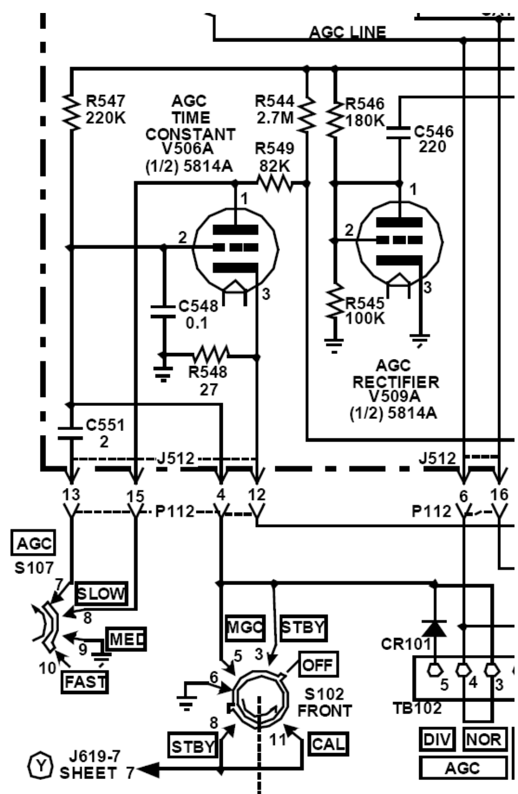


Figure 1

THE PROBLEM

The grid end of C551 is on the AGC bus. At MEDium speed, the other end is grounded. In SLOW, however, the other end is anywhere from +27 to +205. Obviously the cap has to charge up to get there, by dumping [conventional] current onto the AGC bus, which forces it positive and causes a huge signal overload. I call this the "Blast".

If not for two ameliorating factors*, it would last 15 times as long as the MOS because of the multiplication effect.

* First, until V506A gets out of saturation, its grid is driven positive, draws current, and charges the cap more quickly. Second, R545 is bypassed by the AGC rectifier. (IF decks with the Lankford AGC mod will be even faster, limited by the rectifier's effective impedance of around 7k.)

Now switch back to MED. The non-grid end of C551 instantly goes from a large positive voltage (+27 to +205) to ground, which forces the AGC bus negative by the same amount. We're talking up to -200V or ten times max. Every tube cuts off hard until this charge drains away. This is the well-known "Moment of Silence".

These effects are intrinsic to the dual use of C551. In MED, it has X volts, while SLOW gives it 10*X. While switching circuitry could force the cap to the correct charge level as the switch moved between detents, it would be quite complex and utterly out of the question.

IDEA THE FIRST

The only feasible solution is to use separate caps for MED and SLOW.

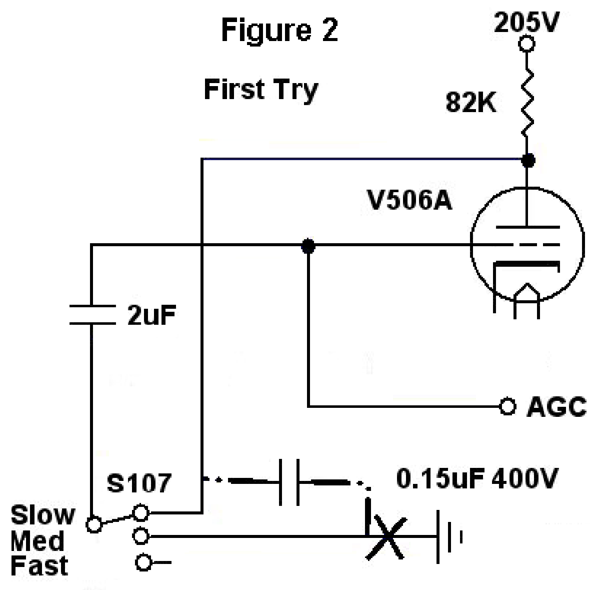
The brute-force approach is to add a second 2uF cap, and use one for MED and the other for SLOW. However, C551 is a big, expensive cap, and Collins would not have sprung for a second one, even on a Government contract.

IDEA THE SECOND

And rightly so, it's not necessary. We could use V506A for MED as well as SLOW, multiplying a small cap to get the old 2uF time constant.

Figure 2

First Try



My first try was beautifully simple. In MED, instead of grounding C551, hook a small cap in series, which requires nothing more than that you remove the ground wire from S107, and substitute a cap. The other end goes to the SLOW contact, and ground is not used. In SLOW, there is no change. In MED, we go from plate, through a small cap, then through C551, to grid. The small cap is sized to give medium speed when closed-loop. Since it's much smaller than C551, the latter won't affect the total much, so we just divide 2uF by 15 and round to the nearest standard value: 0.15uF. I scoped AGC attack and decay during step changes of signal level, and they are indistinguishable from the stock arrangement. This mod is documented in Figure 2.

I should mention that when you switch to MED, you may get a small dip or surge, a Blast or MOS in miniature, at most 7% (1/15) of the real thing. This is the only thing that went right.

NOT SO FAST

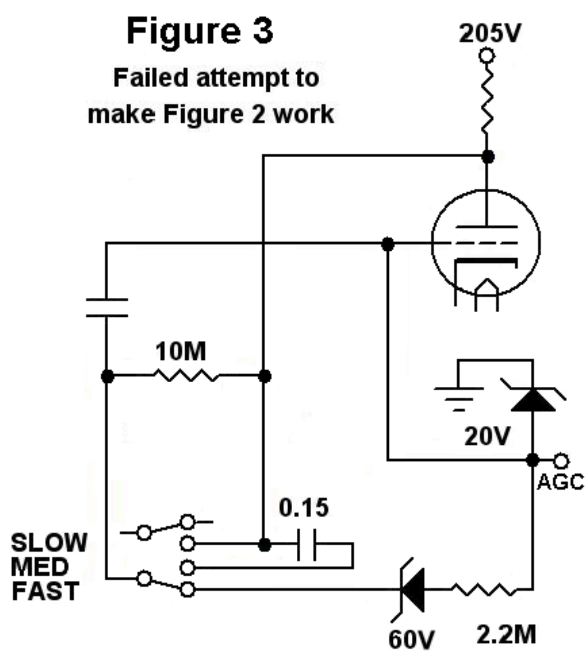
This works as advertised - until you meet the real world. There is a huge bug lurking. Let's say you've been listening to BBC at +60dB. Your "favourite programme" ends, and you go hunting, switching to FAST so as not to miss anything while you scan. Down in the mud you hear something, but it's fluttery. You switch back to SLOW to smooth it out --- and all goes silent. It's the Moment all over again, but it goes on and on. Finally, after maybe ten seconds you can hear again, and your quarry's gone off the air.

What happened? When you were on BBC, C551 was sitting at about -10 on the grid and about +100 on the plate. When you went FAST, it was disconnected. Since you have conscientiously installed a nice new cap, it holds its charge. Your new signal is very weak. You go back to SLOW, and Pow! the stored charge slams the AGC bus right back to -10. Not only that, you're on the SLOW time constant, so it takes 15 times as long to get back to equilibrium.

It works the other way too. Let's say, instead of going to a program you don't like, BBC went off the air. You sit mulling over what you've heard, until the static comes up. You go to FAST, hit another good powerhouse, and go back to SLOW so the Debussy doesn't get clobbered by the fades. *Blast* goes the audio as you endure seconds of overload. It's shorter than usual (C551 starts at +27V instead of 0), but still -- we were trying to eradicate this nonsense, weren't we? We've just made it better some ways, worse others, and unpredictable to boot.

IT'S SUPPOSED TO BE AUTOMATIC BUT ACTUALLY YOU HAVE TO PRESS THIS BUTTON

I tried a couple of embellishments on this mod. I was reluctant to give it up outright, because as far as I can see it is very much the simplest possible mod. The problem is charge storage on C551. First I tried to gradually leak it off to the right value with a bleed resistor. This was doomed, because if you make the R-C551 time constant less than about 20 seconds (and you must, if it's to be useful), there is a noticeable AGC offset until the cap reaches equilibrium.



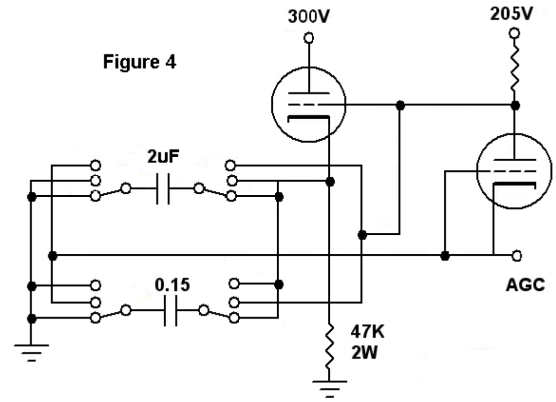
Then I added a zener diode to limit the stored charge (connected to the FAST terminal so it wouldn't disturb MED or SLOW operation), but it only works in certain ranges of signal level and sometimes only if you switch speeds at certain narrowly-defined intervals, and also was getting to be a kluge, see Figure 3. I even entertained a "Dump AGC" button. I was out of the sweet spot, to put it mildly.

EXODUS

At this point I concluded that I had exhausted the possibilities of the existing C551. It's that darn grid connection in the IF deck. As long as C551 is shackled to the AGC bus, you can't manage its charge without also distorting the AGC. I mentally threw out all constraints and started sketching, just to see where it would take me.

A NEW HOPE

The first sketch is reproduced in Figure 4. It uses a 4-pole switch for maximum flexibility. The big idea here is that when a cap is not being used for AGC, we hold it at the same potential it would have if it were.

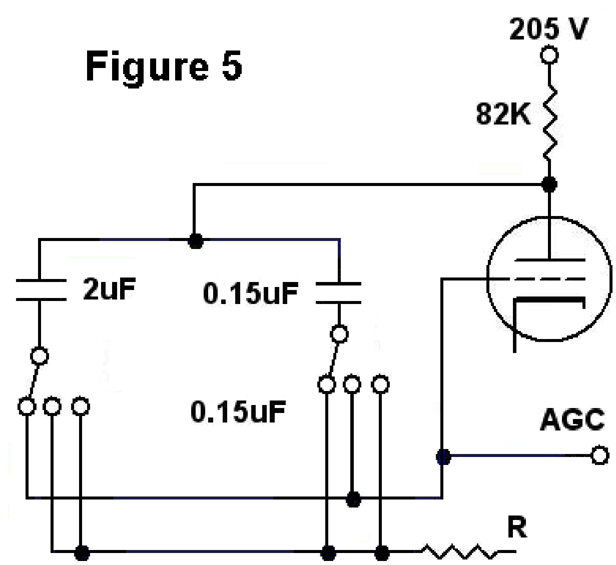


That way, when we switch it in, there's no bump. A new cathode follower (don't panic, this is a gedankenexperiment) isolates the unused cap from the V506A plate to eliminate all chance of interaction.

Even so, we only partly achieve the ideal. To go all the way, we'd need a second cathode follower plus a regulated negative supply, to isolate the grid end of the caps from the AGC bus. As it is, we get a compromise. The cathode follower output is maybe five volts above its grid, so when the AGC is around five volts, we're perfect. We're never more than 15 volts off, which causes a short, mild, fairly innocuous correction because it's divided by 15.

THE REAL DEAL

Of course, adding a cathode follower is out of the question. In fact, any kind of follower, because we don't have a supply higher than +205. Yes, there is one on the audio deck, but I wasn't about to route it to where I would use it. Too much of a kluge, again. I took another leap of faith and erased it. Since the high ends of the caps were now permanently connected to V506A's plate, I could throw out two switch poles. Getting warm, but what about interaction?



The worst case is MED, where my 0.15uF cap is active and C551 is on standby. Picture a step change in signal. At V506, the grid and plate voltages, formerly steady, would abruptly start to rise or fall. At medium speed, the design time constant is about 0.5 second. Hang a 2uF cap off the plate, it will try to hold the plate steady, which reduces the feedback effect and therefore the effective capacitance. Add capacitor, it speeds up: not exactly intuitive.

Could I use a resistor to isolate the standby cap, and still get what I wanted? I wired it up as in Figure 5. As expected, R=0 threw the medium time constant way off. As I increased it the TC came back to normal. At 470k, I couldn't see any effect, so I sat back and took stock. The V506A plate load is 82k, but the impedance seen by C551 is 82k in parallel with the tube's plate resistance, which is only about 20k. The 470k isolation resistor is more than 20 times that, confirming my observation of no significant interference. What about the standby cap? If it's 2uF, it's on about a one-second time constant, in other words, if the signal holds constant for five seconds, we're right on. That's perfectly reasonable. Any time C551 is out of equilibrium, the charge or discharge current offsets the plate voltage, but the effect on the grid is tiny.

The plate offset was small (15/485), and it's divided further by the tube gain. The 0.5-second MED time constant cuts it even further. By George, I think we've got it!

REVIEW

In SLOW, we have C551 from plate to grid same as usual, plus a 470k - 0.15uF series combo to ground. The latter keeps up handily without causing trouble.

In MED, we have the 0.15 from plate to grid, plus a 470k - 2uF series combo to ground. The latter is an acceptable five seconds behind, and adds less than 0.3% AGC offset which decays over a period of about five seconds, in other words, you can't see it outside the lab. Heck, I couldn't see it *inside* the lab, I'm predicting it mathematically. It also affects the meter ballistics slightly, eyeballed at about 5% overshoot on a step change. Again, it's undetectable against a real-world signal.

In FAST, both caps are on standby. V506A is open-loop, so this only affects the meter ballistics. When we go back to MED, the 0.15uF cap goes active, highlighting a small problem.
It was directly in parallel with C551 and therefore five seconds behind. If we rearrange the 470k resistor slightly, FAST will connect the 0.15uF cap straight to ground. This also allows you to use one of the switch terminals as a tie point. The incremental ballistic effect is small and in my opinion quite worth the convenience. You could also use two separate resistors, but you need a tie point.

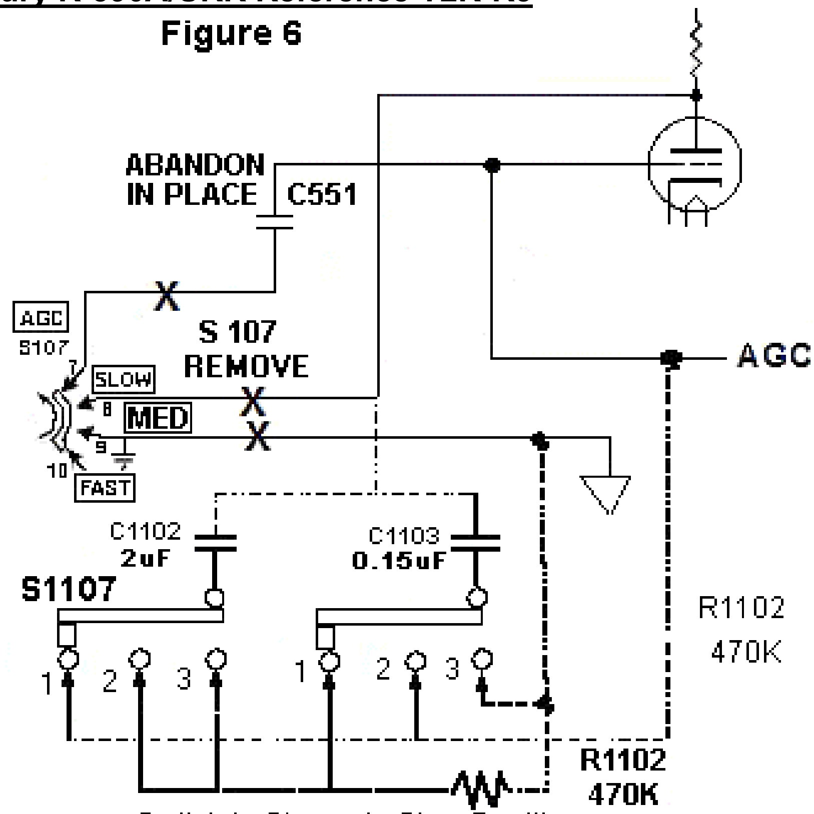
On switching FAST-MED, MED-SLOW, and SLOW-MED, the cap being brought online has a potential equal to V506A's plate voltage. If there is enough signal to generate AGC voltage, it is not part of the cap's potential, and the AGC bus will experience a positive step. It's very small, because the error is inside the feedback loop and gets divided by the tube gain. For example, with a 16V error, the grid (i.e., the AGC bus) goes up 1V and the plate goes down 15.

ONE LAST THING

This is how Collins should have done it. The cost delta would have been: one resistor, 2-pole vs 1-pole switch and one additional 0.15uF 400V Vitamin Q cap

I am not a genius. All I can figure is, the engineer assigned this task was mediocre, inexperienced, unimaginative, or just pressed for time.

Figure 6



Switch Is Shown In Slow Position

I changed S107 on the front panel. I was loath to also modify the IF deck, but the only way to avoid it is to abandon C551 and add my own 2uF cap on the front panel.

In Figure 6 I call it C1102 (C551, version 2). Provided we can find a place for it, we have a viable mod.

FITTING IT IN

Where to put our new caps? Since we're already replacing S107, the open space between it and the ANT TRIM shaft is a prime candidate. How much space do we need? The DigiKey catalog lists six different 2.0 or 2.2/400 caps. They're all less than 2" x 1-1/4" x 1-1/4".

If we elevate a cap this size to where it doesn't quite graze the tallest nearby gear, it squeezes in under the top cover (if used). Let's mount it on a terminal strip. Then C1103 can go on the back side and hang upside-down in front of the gear. To avoid gluing or drilling the panel, we can mount the terminal strip to a flat sheet-metal bracket anchored under the terminal board mounting bracket.

PARTS LIST

Piece of 22ga sheet steel, 1" by 2-1/4".

Terminal strip, Radio Shack #274-688 (5-lug) or similar

470k resistor

0.15uF/400V capacitor, NTE MLR154K630 or similar*

2.0uF/400V capacitor, NTE MLR205K400 or similar**

2-pole 3-position rotary switch

A couple feet of wire

* It only needs to be 400V, but with NTE, the 630V model has more convenient lead spacing.
** 2.2uF is fine too; there are six available from DigiKey, they're all small enough, and several are cheaper.

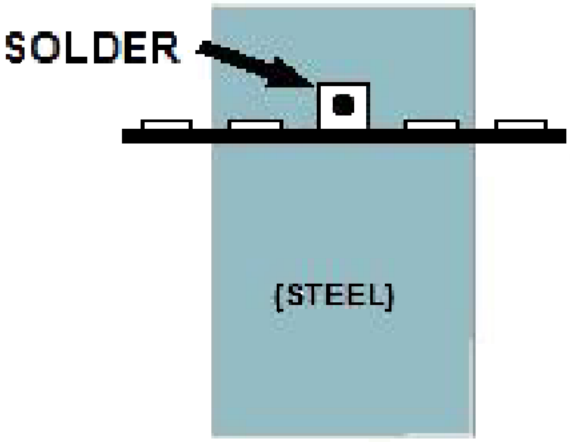
TOOLS

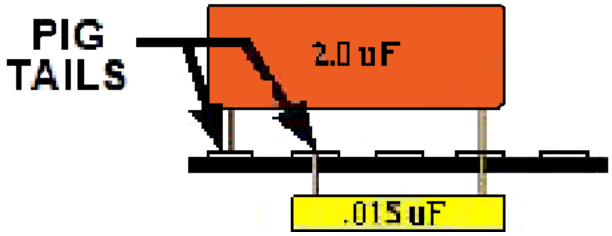
Vise & hacksaw	Sandpaper	Drill
#1 Philips	#2 Philips	Dikes, stripper, soldering iron

ASSEMBLY

Take your sheet metal, drill a 1/8" hole 1/2" from one end, and sand the other half bright. This is our mounting bracket.

Solder the terminal strip mounting lug to the bracket, exactly 1-1/2" from the hole, with the strip facing away from it.





Solder C1102 (2uF) on the "normal" side, and C1103 (0.15uF) on the back. Lug 1 is C1102, 2 is C1103, 3 is unused, 4 is common to both caps. Also wire 4 to 5.

Solder 4" pigtails to lugs 1 and 2. They will go to the A and B rotors on S1107. The wire from the S107 SLOW contact (V506A plate, white with green and orange traces in my radio) will go to lug 5.

Drop the front panel and solder a 6" pigtail to the MGC/STBY contacts on the S102 FUNCTION switch (front wafer, two adjacent contacts wired together). Route this wire, AGC, up near S107 and temporarily tape it down. Replace the front panel.

Note: I didn't drop the front panel; the red wire that you see going off to the right in the assembly photo snakes around to the back panel and connects to the AGC terminal there. Someday I'll go back and do it right.

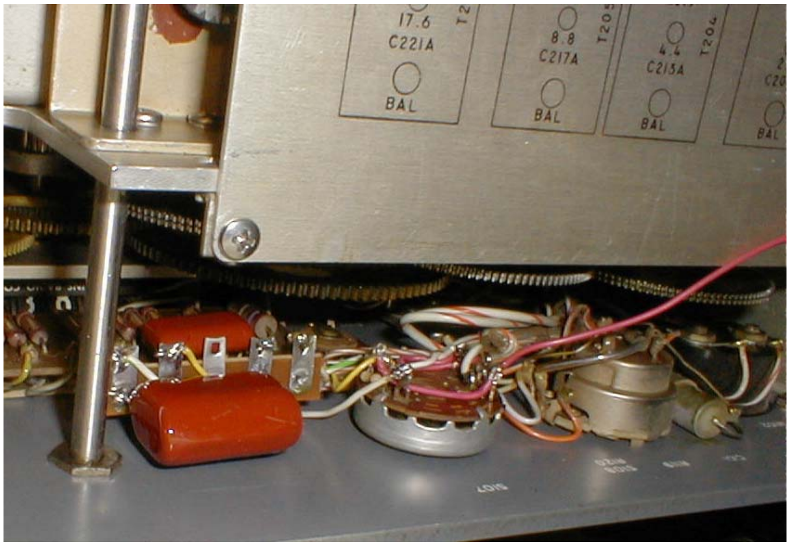
Remove the nameplate, exposing the Terminal Board screws. Loosen the left-hand one, remove the right, slide our bracket under the TB bracket, and replace the screw, through our bracket. Tighten them up, and replace the nameplate.

Disconnect the wires on S107.

The one on the rotor is C551. Tape it off, you won't use it.

The one on the SLOW contact is Plate, solder it to terminal strip lug 5.

The MED contact is Ground, it will go to S1107. Remove and archive S107.



With S1107 on the bench, connect together contacts A1 and B2. Connect together A2, A3, and B1. Solder R1107 (470k) between B1 and B3.

With S1107 behind the front panel, solder the terminal strip #1 pigtail to the A rotor, the #2 pigtail to the B rotor, the AGC pigtail to A1/B2, and the Ground wire to B3. Slip it into the mounting hole, bolt it down, and you're done.

THE R-390

Electrically, the mod can be applied with no value changes. Since the time constant tube has a smaller plate load (56k vs 82k), the meter overshoot will be slightly less. The main issue is where to fit the terminal strip. Not having an R-390, I can't try it myself. It can't go to the left of the AGC switch, the terminal board is there. It would be convenient if it could go below the TB, between it and the frequency readout. There are probably other nooks it could be fit into, although the wires would be longer.

Parts and Services

Al Parker	Filter Caps C603 & C606 "restuff" the filter caps, for R-390A's. Present pricing as of 7/-7 is $50 for the pair (exchange), plus $2.50 return shipping in the USA. The cans are knurled shut, not epoxy or glue sealed. I'll update my webpage on the process soon (when I get a round tuit) it's at: boatanchors.org/filtercap.htm
Howard Mills	Front Panels
Hank Arny	Mostly metal parts. Front panels, Knobs, Partial Chassis
boltdepot.com	/machine-screws.aspx
jamminpower.com	R-390 maintenance manuals for free, if you are willing to wait for them - they are large.
jamminpower.com	R 390A modules, junker modules, SP 600 parts. Andy is not in the parts business per se. He sells parts to help people out.
Fair Radio	R-390A Wide assortment of parts. Stock varies.
Mc Master-Carr	Every conceivable screw one could ever use-even in stainless steel.
Mouser	Electronic parts. Takes small orders. Great selection
Dan's Small Parts	Internet only. Vast assortment of small electronic parts. Frequent sales. A+ site
Antique Radio Supply	They usually have a choice from the Chinese copies up to NOS JAN tubes--
Krize Electronics	Tubes Frank Krize. 1373 Chapel Hill Rd. Natchez MS 39120 Telephone: 601-442-0973 Email: k5svc@NATCHEZ.net
http://www.r-390a.net/ R390	web page by Al Tirevold
http://www.r390a.com/	web page by Chuck Ripple. He also rehabs R390A's. Has some parts
Miltronix.com	Rick Mash re-builds R-390A'a as a full time living as well as doing repairs. Considered the premier re-builder by many. Sometimes has radios for sale.
Triode electronics	Tubes
HSN Newsletter	THE granddaddy pioneer predecessor to the R-390 list. Pure gold for R-390A, SP-600's & R-388's radios plus lots of useful hints and procedures..
Kongsford.com	Want the maximum stable performance from your R-390A? Dallas Langford's site will provide you with about all the proven improvements possible. A++++++
Antique Electric Radio	NOS, Chinese, east European tubes plus BA related parts.
Just Radios	Canadian source for economical capacitors and resistors. Well recommended by users.
Surplus Sales of Nebraska	Expensive source for very hard to find R 390A and other parts. Stock will vary.
Jan Skirrow	Meter replacement circuits
Leed's Electronics	Carrier level meter substitute from in New York
Rich Baldwin KD6VK	NOS silver mica caps in every shape and size Can ship small quantities 8105 Crichton Lane Durham, NC 27713

Painting R-390A Front Panels

An EXCELLENT paint resource can be found here: http://www.r-390a.net/faq-paint.htm

Front Panel

Part 1:

Remove all hardware from the front panel

Cover back of front panel with tape (to protect any paint or lettering on the back)

Spray Aircraft Paint Remover over the front surface

After a few minutes, begin removing paint with a plastic scraper and paper towels

Use a brass bristle brush to remove paint from engraved lettering

After all paint is removed, sand with medium sandpaper (150 grit)

Finish sand with fine sandpaper (400 or 600 grit)

DO NOT use steel wool

Remove tape from rear surface of the front panel

Acid-etch aluminum with Alumiprep

Prop panels up into a semi-vertical position to spray

Overspray "lightly" with a zinc oxide or zinc chromate based primer, found mostly in marine or aviation supply houses

Sand lightly with 1500 cloth

Part 2:

2A: Spray Can Method

Spray lightly with two coats of Dark Machine Gray (Rustoleum #7857), watching the sheen during the second coat until evenly finished and allowing 30 minutes to 1 hour between coats. PPG #55-307, "Battleship Gray", and Plastikote # 1105, "Medium Gray" are also recommended as good matches. For black panels, use Rustoleum Semi-gloss black.

Insert panel into an oven set at 150 degrees F immediately after applying the second coat

Let bake for approximately two hours

Remove and let cure for about 48 hours

2B: Paint Sprayer Method (MUCH better!)

Go by your favorite paint store, and get them to mix your favorite shade of gray (or color match your an existing panel), using oil-based semi-gloss (or satin) base. I have been pleased with Benjamin Moore Satin Impervo. Their best color for the panel is called a Smoke Gray.

After shaking well, mix your paint with thinner (mineral spirits) by 30% to 50%! Yes, I KNOW this looks thin. Do it anyway.

Set your sprayer on the finest, lightest mist you can. Test with a piece of white paper. I use about 40 psi pressure and a very fine mist.

Spray very light coats, about 5 minutes apart. It should be the 3rd or 4th coat before you get complete coverage of the underlying primer. Do NOT spray on heavily, as it will run easily since you thinned it so much.

Spray about 8 to 10 coats total. This has taken you about an hour.

You can speed drying by Inserting panel into an oven set at 150 degrees F immediately after applying the last coat, one to two hours

After fully dry, bake at about 200 degrees for at least two hours. The paint may begin to "stink" at this point, which is normal

Remove and let cure for about 48 hours

Part 3:

Final Preparation

Polish the panel with rubbing compound (liquid type for removing swirl marks during automotive refinishing)

Warm the panel in the oven prior to lettering with a paint stick (about 120 degrees for 2 or 3 minutes)

Apply paint to panel lettering by smearing paint lightly and gently over the engraved letters

Remove excess paint stick with a diaper or soft cotton cloth about 1 to 2 minutes after application

Reapply paint as necessary if cloth creates any voids

Rub entire panel briskly and lightly to polish the panel and remove any "chalky" looking patches around the lettering.

Knobs

Knobs and meter covers and dial covers are done similarly, with a black (gloss or semi-gloss) paint. Rustoleum #7777 "Satin Black", Krylon #1613, "Semi-Flat Black", and Krylon Epoxy #EP705 "Gloss Black" are possible choices.

For a nice trick to painting the knobs, take a look at Chuck Rippel's R-390A video.

Resistor and Capacitor Tally List

Does not include all variable cap assemblies or the 1% carbon film resistors

Tally	Capacitor, Mica
4	Capacitor, Mica 5 pF., ± 0.5 pF., 300 Vdcw.., CM15D050K
5	Capacitor, Mica 12 pF., ± 5%, 500 Vdcw.., CM15D120J.
4	Capacitor, Mica 15 pF., ± 5%, 500 Vdcw.., CM15D150J
2	Capacitor, Mica 18 pF., ± 5%, 500 Vdcw.., CM15D180J.
1	Capacitor, Mica 20 pF., ± 5%, 300 Vdcw.., CM15D200J
1	Capacitor, Mica 24 pF., ± 5%, 500 Vdcw., no. CM15D240J.
4	Capacitor, Mica 33 pF., ± 2%, 500 Vdcw.., CM15D330C
2	Capacitor, Mica 39 pF., ± 2%, 500 Vdcw. CM15D390J
4	Capacitor, Mica 47 pF., ± 2%, 500 Vdcw.., CM15D470G
2	Capacitor, Mica 51 pF., ± 2%, 500 Vdcw.., CM15D510G.
1	Capacitor, Mica 56 pF., ± 2%, 500 Vdcw., CM15D560G.
10	Capacitor, Mica 68 pF., ± 2%, 500 Vdcw.., CM15D680G
1	Capacitor, Mica 75 pF., ± 2%, 500 Vdcw.., CM15D750G..
1	Capacitor, Mica 82 pF., ± 2%, 500 Vdcw., CM15D820G.
10	Capacitor, Mica 100 pF., ± 2%, 500 Vdcw.., CM15D1016.
8	Capacitor, Mica 110 pF., ± 2%, 500 Vdcw.; CM15D111G.
5	Capacitor, Mica 120 pF., ± 2%, 500 Vdcw.., CM15D121G.
5	Capacitor, Mica 150 pF., ± 2%, 500 Vdcw.., CM15D151G.
2	Capacitor, Mica 160 pF., ± 2%, 500 Vdcw.
1	Capacitor, Mica 200 pF., ± 1%, 500 Vdcw..
4	Capacitor, Mica 220 pF., ± 2%, 500 Vdcw.., CM15D221G.
1	Capacitor, Mica: 270 pF., ± 2%, 300 Vdcw.., CM15D271G
2	Capacitor, Mica 300 pF., ± 2%, 500 Vdcw.., CM15D301G.
1	Capacitor, Mica 330 pF., ± 2%, 500 Vdcw..,
2	Capacitor, Mica 470 pF., ± 2%, 300 Vdcw.., CM15D471G.
1	Capacitor, Mica 510 pF., ± 2%, 300 Vdcw., CM15D511G.
2	Capacitor, Mica 1000 pF., ± 2%, 500 Vdcw.., CM30D102G.
1	Capacitor, Mica 1000 pF., ± 2%, 500 Vdcw.., CM30D102G
2	Capacitor, Mica 1500 pF., ± 10%, 300 Vdcw.
4	Capacitor, Mica 1800 pF., ± 2%, 500 Vdcw.., CM15D181G.
5	Capacitor, Mica 2400 pF., ± 2%, 300 Vdcw.., VCM20E242G.

	Capacitor, Ceramic
1	Capacitor, Ceramic: 0.75 pF., ± 0.25 pF., 500 Vdcw.., CC20CKR75C.
1	Capacitor, Ceramic: 0.5 pF., ± 0.25 pF., 500 Vdcw.., CC206K0R5C.
2	Capacitor, Ceramic: 1 pF., ± 0.25 pF., 500 Vdcw. CX20CK010C.
2	Capacitor, Ceramic: 1.5 pF., ± 0.25%, 500 Vdcw.., CC20CK1R5C.
2	Capacitor, Ceramic: 2.0 pF., ± 0.25%, 500 Vdcw..,
9???	Capacitor, Ceramic: 4 pF., ± 0.25 pF., 500 Vdcw
1	Capacitor, Ceramic: 8 pF., ± 0.25 pF., 500 Vdcw.., CC20CH080C.
1	Capacitor, Ceramic: 12 pF., ± 5%, 500 Vdcw., CC20UJ120J.
1	Capacitor, Ceramic: 47 pF., ± 5%, 500 Vdcw.., CC20UJ470J.
2	Capacitor, Ceramic: 5000 pF., ± 15%, 1000 Vdcw

Capacitor, Ceramic, Variable

1	Capacitor, Variable, Ceramic: 1.5 pF min, 7 to 10.5 pF max, 350 Vdcw.;
1	Capacitor, Variable: 3 to 12 pF.,
4	Capacitor, Variable, Ceramic: 3-18 pF., 350 Vdcw.
1	Capacitor, Variable, Air: 3.2 pF. to 60.7 pF., 850 Vac
8	Capacitor, Variable, Ceramic: 5 to 37.5 pF., 350 Vdcw.
8	Capacitor, Variable, Ceramic: 8 to 50 pF., 350 Vdcw..
6	Capacitor, Variable, Ceramic: 8-75 pF., 350 Vdcw.
1	Capacitor, Variable, Glass Dielectric; 1.5 to 8 pF., Corning part no. 692063

Capacitor, Paper

12	Capacitor, Paper: 0.1 uF, ± 10%, 200 Vdcw
7	Capacitor, Paper: 0.033 uF, ± 20%, 300 Vdcw.., (56289) part no. 96P33303S4.
1	Capacitor, Paper: 220,000 pF., ±20%, 100 Vdcw.., (56289) part no. 96P22401S13
12	Capacitor, Paper: 10,000 pF., ± 20%, 300 Vdcw. (56289) part no. 96P1030354.
1	Capacitor, Paper: 2 uF., ± 10%, 500 Vdcw. (53021) type no. 62A.
1	Capacitor, Paper: 47,000 pF., ± 20%, 100 Vdcw. (56289) part no. 186P4730155.
1	Capacitor, Paper: 0.1 uF, ± 20%, 100 Vdcw.; (56289) part no. 96P10401S4.
2	Capacitor, Paper: .033 uF, ± 20%, 300 Vdcw.., (56289) part no. 96P33303S4

Capacitor, Electrolytic

1	Capacitor, Electrolytic: 8 µF, 30 Vdcw., (21520) type no. PP8B30A2.
2	Capacitor, Electrolytic: 2 sections, 45 µF 300 Vdcw., CE52C450N.
1	Capacitor, Electrolytic: 3 section, 30 µF., 300 Vdcw., CE53C300N.
1	Capacitor, Electrolytic: 50 µ*F,* 50 Vdcw.., CE64C500G.

1/2 Watt fixed ± 10%,

6	27 ohm	13	2K2 ohm	1	100 K ohm
1	33 ohm	2	3K9 ohm	1	120 K ohm
3	47 ohm	1	6K8 ohm	2	150 K ohm
2	56 ohm	1	8K2 ohm	2	220 K ohm
2	100 ohm	2	10K ohm	1	270 K ohm
1	150 ohm	7	22K ohm	1	330 K ohm
1	220 ohm	4	27K ohm	1	390 K ohm
1	270 ohm	3	33K ohm	17	470 K ohm
1	390 ohm	1	39K ohm	2	680 K ohm,
3	680 ohm	3	47K ohm	1	820 K ohm
1	820 ohm	4	56K ohm	4	1 meg.
3	1K ohm	1	68 K ohm	1	1.5 meg
2	1K2 ohm	3	82 K ohm	2	2.7 meg

One Watt Fixed

1	22 ohm, ± 10%, 1 W,	1	39,000 ohm, ± 10%, 1W,
4	560 ohm, ± 10%, 1 W	1	56, 000 ohm, ± 10%, 1 W,
2	2200 ohm, ± 10%, 1 W,	1	82,000 ohm, ± 10%, 1 W,
1	5600 ohm, ± 10%, 1 W,	1	150,000 ohm, ± 10%, 1 W,
1	22,000 ohm, ± 10%, 1 W,	2	180,000 ohm, ± 10%, 1 W,
1	27,000 ohm, ± 10%, 1 W,	1	220,000 ohm, ± 10%, 1 W,

1 22,000 ohm, ± 10%, 2 W,

Tally	**High wattage**
1	Resistor, Fixed, Wire Wound: 4 ohm, ± 5%, 8 W,
1	Resistor, Variable: 100 ohm, ± 20%, 1 W,
1	Resistor, Fixed, Wire Wound: 800 ohm, ± 5%, 8 W.
1	Resistor, Wire Wound: 1200 ohm, ± 5%, 10 W
1	Resistor, Wire Wound: 1600 ohm, ± 5%, 10 W.
2	Resistor, Variable: 2500 ohm, ± 20%, 2 W.
1	Resistor, Variable: 5000 ohm, ± 20%, 2 W,
1	Resistor, Variable: 10,000 ohm, ± 20%, 1 W
1	Resistor, Variable: 500,000 ohm, ± 20%, 2W.

Nolan's R-390A/URR Master Capacitor List

Revision 0.4a BETA (01/13/02) nlee@acadiacom.net

This list is still "BETA". As a result has NOT been checked as closely for errors as usual. Please drop me a message if I left anything out or if you spot an error and I'll correct it and re-post. Thanks, nolan

-----Main Chassis (front and rear panel, etc.)-----

C101 0.22mf 100 WVDC 20% paper
C102 5000pf 1000 WVDC 15% ceramic
C103 50mf 50 WVDC ??% electrolytic industry # CE64C500G (1)
C104 through C107 0.068mf ??? WV?C ??% paper (inside of line filter)

(1) This is the bathtub style capacitor mounted below the line filter on the rear panel. It is NOT an oil filled paper cap even though it looks like one. Watch the polarity.

-----RF Amplifier Sub-chassis-----

C201A 3-12pf 350 WVDC NPO ceramic trimmer (inside of T201)
C201B 8-50pf 350 WVDC N750 ceramic trimmer (inside of T201)
C202 7pf 500 WVDC .25pf ceramic (inside of T201)
C203 330pf 500 WVDC 2% mica (inside of T201)
C204 120pf 500 WVDC 2% mica (inside of T201)
C205A 3-12pf 350 WVDC NPO ceramic trimmer (inside of T202)
C205B 8-50pf 350 WVDC N750 ceramic trimmer (inside of T202)
C206 7pf 500 WVDC .25pf ceramic (inside of T202)
C207 120pf 500 WVDC 2% mica (inside of T202)
C208 75pf 500 WVDC 2% mica (inside of T202)
C209A 3-12pf 350 WVDC NPO ceramic trimmer (inside of T203)
C209B 8-50pf 350 WVDC N750 ceramic trimmer (inside of T203)
C210 7pf 500 WVDC .25pf ceramic (inside of T203)
C211 36pf 500 WVDC 2% mica (inside of T203)
C212 39pf 500 WVDC 2% mica (inside of T203)
C213A 3-12pf 350 WVDC NPO ceramic trimmer (inside of T204)
C213B 5-25pf 350 WVDC NPO ceramic trimmer (inside of T204)
C214 7pf 500 WVDC .25pf ceramic (inside of T204)
C215 100pf 500 WVDC 2% mica (inside of T204)
C216 24pf 500 WVDC 5% mica (inside of T204)
C217A 3-12pf 350 WVDC NPO ceramic trimmer (inside of T205)
C217B 3-12pf 350 WVDC NPO ceramic trimmer (inside of T205)
C218 7pf 500 WVDC .25pf ceramic (inside of T205)
C219 5pf 300 WVDC .5pf mica (inside of T205)
C220 12pf 500 WVDC 5% mica (inside of T205)
C221A 3-12pf 350 WVDC NPO ceramic trimmer (inside of T206)
C221B 1.5-7pf 350 WVDC NPO ceramic trimmer (inside of T206)
C222 7pf 500 WVDC .25pf ceramic (inside of T206)
C223 18pf 500 WVDC 5% mica (inside of T206)
C224 5pf 300 WVDC .5pf mica (inside of T206)

C225A 7-80pf 800 WVAC 4pf variable (front half of antenna trimmer cap)
C225B 6-26pf 800 WVAC 2pf variable (rear half of antenna trimmer cap)
C226 5000pf 1000 WVDC 15% ceramic
C227 0.047mf 100 WVDC 20% paper (4)(z)
C228 1pf 500 WVDC .25% ceramic
C229 5000pf 1000 WVDC 15% ceramic
C230-1 8-50pf 350 WVDC N750 ceramic trimmer (inside of Z201-1)(z)
C230-2 8-50pf 350 WVDC N750 ceramic trimmer (inside of Z201-2)(z)
C231-1 160pf 500 WVDC 2% mica (inside of Z201-1)(z)
C231-2 160pf 500 WVDC 2% mica (inside of Z201-2)(z)
C232-1 2400pf 300 WVDC 2% mica (inside of Z201-1)(z)
C232-2 2400pf 300 WVDC 2% mica (inside of Z201-2)(z)
C233-1 8-50pf 350 WVDC N750 ceramic trimmer (inside of Z202-1)
C233-2 8-50pf 350 WVDC N750 ceramic trimmer (inside of Z202-2)
C234-1 1800pf 500 WVDC 2% mica (inside of Z202-1)
C234-2 1800pf 500 WVDC 2% mica (inside of Z202-2)
C235-1 2400pf 300 WVDC 2% mica (inside of Z202-1)
C235-2 2400pf 300 WVDC 2% mica (inside of Z202-2)
C236-1 8-50pf 350 WVDC N750 ceramic trimmer (inside of Z203-1)
C236-2 8-50pf 350 WVDC N750 ceramic trimmer (inside of Z203-2)
C237-1 120pf 500 WVDC 2% mica (inside of Z203-1)
C237-2 120pf 500 WVDC 2% mica (inside of Z203-2)
C238-1 1500pf 300 WVDC 10% mica (inside of Z203-1)
C238-1 1500pf 300 WVDC 10% mica (inside of Z203-2)
C239-1 5-25pf 350 WVDC NPO ceramic trimmer (inside of Z204-1)
C239-2 5-25pf 350 WVDC NPO ceramic trimmer (inside of Z204-2)
C240-1 68pf 500 WVDC 2% mica (inside of Z204-1)
C240-2 68pf 500 WVDC 2% mica (inside of Z204-2)
C241-1 470pf 300 WVDC 2% mica (inside of Z204-1)
C241-2 470pf 300 WVDC 2% mica (inside of Z204-2)
C242-1 3-12pf 350 WVDC NPO ceramic trimmer (inside of Z205-1)
C242-2 3-12pf 350 WVDC NPO ceramic trimmer (inside of Z205-2)
C243-1 68pf 500 WVDC 2% mica (inside of Z205-1)
C243-2 68pf 500 WVDC 2% mica (inside of Z205-2)
C244-1 1800pf 500 WVDC 2% mica (inside of Z205-1)
C244-2 1800pf 500 WVDC 2% mica (inside of Z205-2)
C245-1 3-12pf 350 WVDC NPO ceramic trimmer (inside of Z206-1)
C245-2 3-12pf 350 WVDC NPO ceramic trimmer (inside of Z206-2)
C246-1 47pf 500 WVDC 2% mica (inside of Z206-1)
C246-2 47pf 500 WVDC 2% mica (inside of Z206-2)
C247-1 33pf 500 WVDC 2% mica (inside of Z206-1)
C247-2 33pf 500 WVDC 2% mica (inside of Z206-2)
C248 5000pf 1000 WVDC 15% ceramic
C249 .5pf 500 WVDC .25pf ceramic
C250 .75pf 500 WVDC .25pf ceramic
C251 1pf 500 WVDC .25% ceramic
C252 2pf 500 WVDC .25% ceramic
C253 4pf 500 WVDC .25pf ceramic
C254 8pf 500 WVDC .25pf ceramic

C255 33pf 500 WVDC 2% mica
C256 0.1mf 200 WVDC 10% paper (5)(z)
C257 47pf 500 WVDC 5% ceramic (z)
C273, C274 5000pf 1000 WVDC 15% ceramic
C275 0.033mf 300 WVDC 20% paper (z)
C276 15pf 500 WVDC 5% mica
C277 5000pf 1000 WVDC 15% ceramic
C278 15pf 500 WVDC 5% mica
C279 15pf 500 WVDC 5% mica
C280 5000pf 1000 WVDC 15% ceramic
C281 1.5pf 500 WVDC .25% ceramic
C282 1.5pf 500 WVDC .25% ceramic
C283-1 5-25pf 350 WVDC NPO ceramic trimmer (inside of Z213-1)
C283-2 5-25pf 350 WVDC NPO ceramic trimmer (inside of Z213-2)
C283-3 5-25pf 350 WVDC NPO ceramic trimmer (inside of Z213-3)
C284, C285,5000pf 1000 WVDC 15% ceramic
C286 100pf 500 WVDC 2% mica
C287, C288 5000pf 1000 WVDC 15% ceramic
C289 2pf 500 WVDC .25% ceramic
C290 2pf 500 WVDC .25% ceramic
C291-1 5-25pf 350 WVDC NPO ceramic trimmer (inside of Z216-1)
C291-2 5-25pf 350 WVDC NPO ceramic trimmer (inside of Z216-2)
C291-3 5-25pf 350 WVDC NPO ceramic trimmer (inside of Z216-3)
C292-1 100pf 500 WVDC 2% mica (inside of Z216-1)
C292-2 100pf 500 WVDC 2% mica (inside of Z216-2)
C292-3 100pf 500 WVDC 2% mica (inside of Z216-3)
C297, C298 5000pf 1000 WVDC 15% ceramic
C299 270pf 300 WVDC 2% mica (inside of T208)
C300 through C308 5000pf 1000 WVDC 15% ceramic
C309 0.1mf 200 WVDC 10% paper
C310 3-60pf 850 WVAC ??? variable (calibration trimmer)
C311 1000pf 500 WVDC 2% mica
C312 150pf 500 WVDC 2% mica
C313 220pf 500 WVDC 2% mica
C314 220pf 500 WVDC 2% mica
C315 220pf 500 WVDC 2% mica
C316 15pf 500 WVDC 5% mica
C317 5000pf 1000 WVDC 15% ceramic
C318 51pf 500 WVDC 2% mica
C319 5000pf 1000 WVDC 15% ceramic
C320 1000pf 500 WVDC 2% mica
C321 12pf 500 WVDC 5% mica
C322 12pf 500 WVDC 5% mica
C323 5pf 300 WVDC .5pf mica
C324 20pf 300 WVDC 5% mica
C325 200pf 500 WVDC 1% mica
C326 5000pf 1000 WVDC 15% ceramic
C327 100pf 500 WVDC 2% mica
C328 5000pf 1000 WVDC 15% ceramic

C329 68pf 500 WVDC 2% mica
C330-1 300pf 500 WVDC 2% mica (inside of Z201-1)
C330-2 300pf 500 WVDC 2% mica (inside of Z201-2)
C331-1 68pf 500 WVDC 2% mica (inside of Z202-1)
C331-2 68pf 500 WVDC 2% mica (inside of Z202-2)
C334 51pf 500 WVDC 2% mica

(z) actual value depends on mod level, I'll add details when I have time

(4) This is the stud mounted oil filled capacitor located on the top side of the chassis next to V201, the 6DC6.

(5) Not used on early receivers. C256 is used to silence hash/static from HR202 Crystal Oven. Less expensive CDE polyester should work fine at this location.

-----Crystal Oscillator Sub-chassis-----

C401 5000pf 1000 WVDC 15% ceramic
C402 5pf 300 WVDC .5pf mica
C403 4pf 500 WVDC .25pf ceramic
C404 4pf 500 WVDC .25pf ceramic
C406, C407 5000pf 1000 WVDC 15% ceramic
C408 12pf 500 WVDC 5% ceramic
C409 150pf 500 WVDC 2% mica
C410 through C415 5000pf 1000 WVDC 15% ceramic
C417 150pf 500 WVDC 2% mica
C418 120pf 500 WVDC 2% mica
C419 100pf 500 WVDC 2% mica
C420 82pf 500 WVDC 2% mica
C421 68pf 500 WVDC 2% mica
C422 56pf 500 WVDC 2% mica
C423 47pf 500 WVDC 2% mica
C424 39pf 500 WVDC 2% mica
C425 33pf 500 WVDC 2% mica
C426 18pf 500 WVDC 5% mica
C427 12pf 500 WVDC 5% mica
C428 5pf 300 WVDC .5pf mica
C429A 8-50pf 350 WVDC NPO ceramic trimmer for 9 MHz
C429B 8-50pf 350 WVDC NPO ceramic trimmer for 8 MHz
C429C 5-25pf 350 WVDC NPO ceramic trimmer for 15 MHz
C429D 5-25pf 350 WVDC NPO ceramic trimmer for 14 MHz
C429E 5-25pf 350 WVDC NPO ceramic trimmer for 4&21 MHz
C429F 5-25pf 350 WVDC NPO ceramic trimmer for 3&20 MHz
C429G 5-25pf 350 WVDC NPO ceramic trimmer for 27 MHz
C429H 5-25pf 350 WVDC NPO ceramic trimmer for 26 MHz
C430A 8-50pf 350 WVDC NPO ceramic trimmer for 11 MHz
C430B 8-50pf 350 WVDC NPO ceramic trimmer for 10 MHz
C430C 5-25pf 350 WVDC NPO ceramic trimmer for 0&17 MHz
C430D 5-25pf 350 WVDC NPO ceramic trimmer for 16 MHz

C430E 5-25pf 350 WVDC NPO ceramic trimmer for 6&23 MHz
C430F 5-25pf 350 WVDC NPO ceramic trimmer for 5&22 MHz
C430G 3-12pf 350 WVDC NPO ceramic trimmer for 29 MHz
C430H 5-25pf 350 WVDC NPO ceramic trimmer for 28 MHz
C431A 5-25pf 350 WVDC NPO ceramic trimmer for 13 MHz
C431B 5-25pf 350 WVDC NPO ceramic trimmer for 12 MHz
C431C 5-25pf 350 WVDC NPO ceramic trimmer for 2&19 MHz
C431D 5-25pf 350 WVDC NPO ceramic trimmer for 1&18 MHz
C431E 5-25pf 350 WVDC NPO ceramic trimmer for 25 MHz
C431F 5-25pf 350 WVDC NPO ceramic trimmer for 7&24 MHz
C431G 3-12pf 350 WVDC NPO ceramic trimmer for 31 MHz
C431H 3-12pf 350 WVDC NPO ceramic trimmer for 30 MHz

-----IF Amplifier Subchassis-----

C501, C502 5000pf 1000 WVDC 15% ceramic
C503 100pf 500 WVDC 2% mica
C504, C505 0.1mf 200 WVDC 10% paper
C506 5000pf 1000 WVDC 15% ceramic
C507 110pf 500 WVDC 2% mica for 16 KHz filter (x)
C508 110pf 500 WVDC 2% mica for 8 KHz filter (x)
C509 110pf 500 WVDC 2% mica for 4 KHz filter (x)
C510 110pf 500 WVDC 2% mica for 2 KHz filter (x)
C511, C512 5000pf 1000 WVDC 15% ceramic
C513 110pf 500 WVDC 2% mica for 2 KHz filter (x)
C514 110pf 500 WVDC 2% mica for 4 KHz filter (x)
C515 110pf 500 WVDC 2% mica for 8 KHz filter (x)
C516 110pf 500 WVDC 2% mica for 16 KHz filter (x)
C517 0.1mf 200 WVDC 10% paper
C518, C519 5000pf 1000 WVDC 15% ceramic
C520 3-12pf 350 WVDC NPO ceramic trimmer (inside of Z501)
C521 0.1mf 200 WVDC 10% paper
C522, C523 5000pf 1000 WVDC 15% ceramic
C524 75pf ??? WVDC ??% ceramic (inside of Z501)
C525 7-10.5pf 350 WVDC NPO ceramic trimmer (BFO neutralization)
C526 100pf 500 WVDC 2% mica
C527 5pf 300 WVDC .5pf mica
C528 0.1mf 200 WVDC 10% paper
C529 0.033mf 300 WVDC 20% paper
C530 150pf 500 WVDC 2% mica
C531 0.1mf 200 WVDC 10% paper
C532 100pf 500 WVDC 2% mica
C533, C534 0.033mf 300 WVDC 20% paper
C535 12pf 500 WVDC 5% mica
C536 0.1mf 100 WVDC 20% paper
C537 1800pf 500 WVDC 2% mica
C538 0.1mf 200 WVDC 10% paper
C539 1000pf 500 WVDC 2% mica
C540 5000pf 1000 WVDC 15% ceramic

C541 0.033mf 300 WVDC 20% paper
C542 47pf 500 WVDC 2% mica
C543 0.1mf 200 WVDC 10% paper
C544 5000pf 1000 WVDC 15% ceramic
C545 0.033mf 300 WVDC 20% paper
C546 220pf 500 WVDC 2% mica
C547, C5480.1mf 200 WVDC 10% paper
C549 0.01mf 300 WVDC 20% paper
C551 2.0mf 500 WVDC 10% oil filled paper (2)
C552 5000pf 1000 WVDC 15% ceramic
C553 0.01mf 300 WVDC 20% paper (3)
C554 1600pf 100 WVDC 1% mica (inside of Z502 - BFO)
C555 50pf ??? WV?C ??% ??? (inside of Z502 - BFO)
C556 50pf ??? WV?C ??% ??? (inside of Z502 - BFO)
C557 ???? ??? WV?C ??% ??? (inside of T501)
C558 ???? ??? WV?C ??% ??? (inside of T501)
C559 ???? ??? WV?C ??% ??? (inside of T502)
C560 ???? ??? WV?C ??% ??? (inside of T502)
C561 ???? ??? WV?C ??% ??? (inside of T503)
C562 ???? ??? WV?C ??% ??? (inside of T503)
C563 ???? ??? WV?C ??% ??? (inside of Z501)
C564 8-50pf 350 WVDC N750 ceramic trimmer for 16 KHz filter (x)
C565 8-50pf 350 WVDC N750 ceramic trimmer for 8 KHz filter (x)
C566 8-50pf 350 WVDC N750 ceramic trimmer for 4 KHz filter (x)
C567 8-50pf 350 WVDC N750 ceramic trimmer for 2 KHz filter (x)
C568 8-50pf 350 WVDC N750 ceramic trimmer for 2 KHz filter (x)
C569 8-50pf 350 WVDC N750 ceramic trimmer for 4 KHz filter (x)
C570 8-50pf 350 WVDC N750 ceramic trimmer for 8 KHz filter (x)
C571 8-50pf 350 WVDC N750 ceramic trimmer for 16 KHz filter (x)

(x) value depends on mod level of IF deck. I'll add details later...
(2) Oil filled metal can paper capacitor mounted to top side of chassis next to chassis harness connector.

(3) Blocking cap for mechanical filters. Recommend working voltage of 600VDC or higher to help decrease the chance of frying the mechanical filters due to failure of C553. This is not a good location to use a cheap replacement capacitor.

-----Audio Frequency Amplifier Subchassis-----

C601 0.01 300 WVDC 20% paper
C602 0.033 300 WVDC 20% paper
C603 3 X 30uf 300 WVDC ??% Electrolytic (Industry # CE53C300N)
C604 0.01 300 WVDC 20% paper (6)
C605 0.01 300 WVDC 20% paper (6)
C606 2X45uf 300 WVDC ??% Electrolytic (Industry # CE52C450N)
C607, C6080.01 300 WVDC 20% paper
C609 8uf 30 WVDC ??% Tantalum Electrolytic (7)
C611 5000pf 1000 WVDC 15% ceramic

C612 68pf 500 WVDC 2% mica

(6) For improved audio performance, you can use .022uf caps at this location.

(7) Also known as "the capacitor that rots off". Leaks sulfuric acid when the seals fail. Watch the polarity when replacing.

-----PTO (VFO) Sub-chassis-----

C701 370pf 500 WVDC 1% ???? (inside of PTO)
C702 10pf ??? WV?C ??% ???? (inside of PTO)
C703 10pf ??? WV?C ??% ???? (inside of PTO)
C704 15pf 500 WVDC 2% ???? (inside of PTO
C705 5000pf 1000 WVDC 10% ceramic (y)
C706 1.5-8pf ??? WV?C ??% glass/silver piston trimmer (inside of Z702)
C707, C7085000pf 1000 WVDC 10% ceramic (y)

C709 510pf 300 WVDC 2% mica (inside of Z702)
C710 through C714 5000pf 1000 WVDC 15% ceramic

(y) C705, C707, C708, C712, C713, and C714 are all constructed into one unit. Some are epoxied together, some are in plastic holders, etc.

Synchronous Detection of DSB and ISB Signals

By Peter C. McNulty, WA1SOV May 2, 1999

Background

A Synchronous Detector designed with the use of an LM311 and an Analog Devices AD607. This is a very basic design. It functions as a product detector and when JMP2 is closed and JMP 1 is open, it functions fully synchronously locking to the 455 KHz I.F. provided at the input. The 455 KHz input should be pre-filtered to remove unnecessary noise from the signal so as to improve locking performance. This circuit has been tested with a Hammarlund SP-600, and it functions quite well with the I.F. out of this receiver connected to J1 of this circuit. Most of the testing has been done with the SP-600 in its 8 KHz position however, any of the passband selections work fine. The narrower the bandwidth selection the better rejection there is against unwanted adjacent channel interference. Additional I.F. selectivity can be had by filtering just the signal provided to U1 and not the I.F. sent to U2. This is not shown below but can easily be implemented. Further improvement could be had, by regenerating the carrier with a Phase Locked Loop such as the NE564. This is necessary for use on signals with marginal carrier signal or suppressed carrier. The loop filter designed for use with such a PLL would provide much greater rejection of unwanted adjacent channel noise. This is an experimental breadboard and is by no means a polished performer. It does afford the experimenter the opportunity to play with a configuration that can provide synchronous detection. Have Fun!

Theory of Operation

The signal applied to J1 is between 1 to 2 volts, rms and is hard limited by U1, LM311. The limiting is necessary to remove amplitude variations of the signal due to the modulation components present. It is only the carrier we want the Phase Locked Loop in U2 to lock-up to. U1 is configured as a zero crossing detector. R1 & R2 bias the input signal to half of VCC. R3 & R4 provide the reference voltage to the comparator, which will determine the switching point of the comparator. The output of U1 is pulled up to VCC so it provides an output which swings between VCC and ground (+4.5v). It is not recommended to apply greater than 4.5 volts to this circuit because of limitations in U2. Nominally the AD607 runs at as little as 3 volts and under no circumstances wants to run above 5 volts. 4.5 volts is fairly standard and can be gotten from a wall transformer or 3 AA batteries. The output signal of U1 is sent through a trim pot which will limit the amplitude of the PLL reference signal to about 0.5 volts at the FDIN input at pin 1of U2. This signal is also biased at one half of VCC with R10 & R11. The selection of C3 and these resistors is such to minimize any phase distortion caused by the input coupling of these components.

The input signal from J1 is also sent through a resistor network, which conditions the signal for suitability at the input of the product detector in the AD607. These detectors are basically doubly balanced mixers and the local oscillator differs for the two detectors by 90 degrees. The PLL in the AD607 basically takes the reference signal that was limited in U1 and provides a new signal, which is in phase and in quadrature to that reference signal for application to the product detectors as shown in the U2 block. The PLL also has a loop filter, which is optimized for 455 KHz, however I did not notice much difference in performance when this time constant was lowered by an order of magnitude. So I wouldn't be too concerned about this when running I.F.s below 2 MHz.

The detected baseband signal is present at pin 18 of U2 (IOUT), and Pin 17 (QOUT) along with other mixing components. These other components should be filtered and R13, C9, R14, & C8 are selected to filter the outputs with a minimum of phase distortion. If you are going to only utilize this detector as a DSB Detector, then the Quadrature output available is not necessary, and it need not be used for anything. Just pass the In-phase output to an audio amplifier auxiliary input for amplification.

Independent Sideband Synchronous Detection

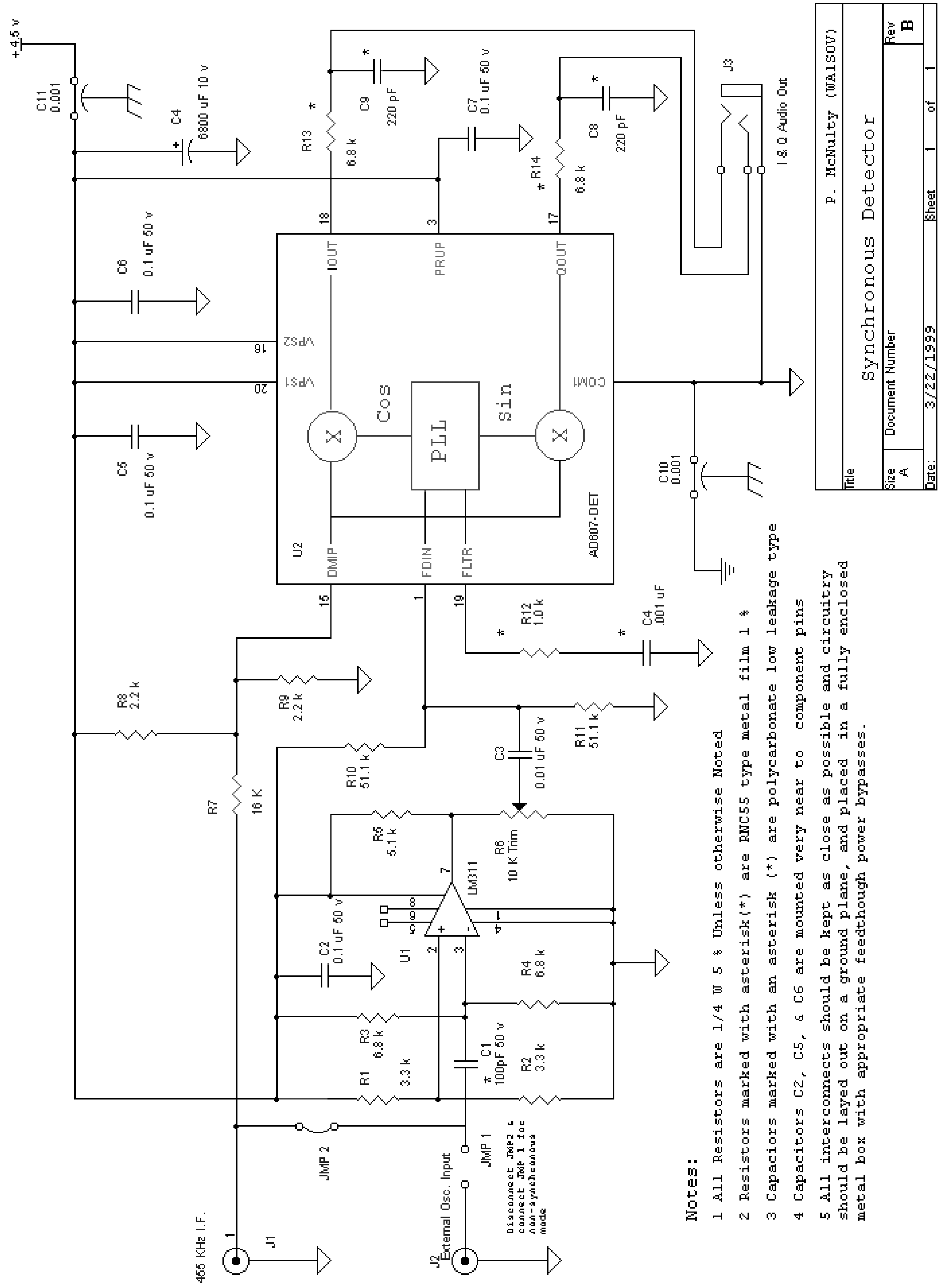
This circuit has been used to provide a source for detection of independent, upper, and lower sideband signals. Unless there is a carrier present with these signals, it cannot operate in the synchronous mode as configured. An independent sideband (ISB) signal can be detected with this detector by utilizing both outputs (IOUT, & QOUT) and passing them to a phase demodulation network. The principle is very much like phasing sideband modulation, but in reverse. The ISB performance of the detector when locked to a carrier was somewhat disappointing. For some reason that I have not discovered yet, the detector will not demodulate ISB when locked to the incoming carrier. It will however demodulate ISB perfectly when a non-phase locked oscillator running near 455 KHz is applied to J2 and the jumper at JMP 2 is removed and JMP is shorted. For ISB detection an external audio phasing network is necessary.

For this to happen it is necessary to utilize wideband audio phase shift networks known as Allpass Filters coupled with doubly balanced mixers. This network is much like the DOME networks originally designed in the 50's. An Allpass Filter design (see http://home.worldnet.att.net/~wa1sov/technical/allpass/allpass.html) can be used or a Digital Signal Processor (DSP) implementing a FIR filter can be employed. The ladder was how I have debugged this design. The concept is pretty simple. If you realize that the IOUT signal contains both sidebands of the signal in phase with each other, and that the QOUT signal contains the sum of upper and lower sidebands 90 degrees out of phase with the IOUT, then with additional phase delay the upper and lower sidebands can be demodulated independently. The IOUT channel gets passed through one Allpass filter, which alters the phase of the signal, by +45 degrees. The QOUT signal is passed through another Allpass filter, which alters the phase of this signal, by -45 degrees. This leaves a net difference of 180 degrees between the IOUT and QOUT Phase processed signals. Adding the IOUT and the QOUT phase processed signals yields a cancellation of the lower sideband. However, the upper sideband remains at the output of this summation and at twice its original amplitude. Now the same two signal are also subtracted from each other in another network to obtain the lower sideband. The lower sideband is also at twice its original amplitude. Thus the lower sideband appears on one channel and the upper sideband would appear on the other channel when applied to the input of a Stereo amplifier. In order for this to work with any degree of opposite sideband rejection close attention to phase accuracy and amplitude matching are paramount. I have obtained without any special optimization of these circuits when using a DSP running a 256 point FIR Filter, opposite sideband rejections of 30 dB. Better can be obtained though through rigorous optimization. Stay tuned!

The signal demodulated does not have to be synchronous to detect signals independently. The system can be used with an external function generator, which is not phase coherent (free running) with the signal. It will function as an independent sideband product detector and Upper and Lower Sideband can be discerned well with this mode.

I hope this article has stirred some interest in this type of detection. Additionally, CQUAM AM Stereo can be demodulated with the I and Q outputs of this detector by simply summing to obtain one channel and differencing to obtain the opposite stereo channel. This does not require the Allpass phasing network described above. Jerry, WA2FNQ transmits the old QAM Stereo signal that can be demodulated using this technique. There are more developments in the works so check back often.

Stay Tuned!


P. McNulty (WA1SOV)
Title Synchronous Detector
Size A
Document Number
Rev B
Date: 3/22/1999
Sheet 1 of 1
+4.5 v
C11 0.001
C4 6800 uF 10 v
C6 0.1 uF 50 v
C5 0.1 uF 50 v
R13 * 6.8 k
C9 * 220 pF
C7 0.1 uF 50 v
R14 * 6.8 k
C8 * 220 pF
J3
I & Q Audio Out
IOUT
PRUP
QOUT
VPS2
VPS1
COM
Cos
PLL
Sin
DMIP
FDIN
FLTR
U2
AD607-DET
C10 0.001
R12 * 1.0 k
C4 * .001 uF
R8 2.2 k
R9 2.2 k
R11 51.1 k
R10 51.1 k
C3 0.01 uF 50 v
R7 16 K
R5 5.1 k
R6 10 K Trim
LM311
U1
C2 0.1 uF 50 v
R4 6.8 k
R3 6.8 k
C1 * 100pF 50 v
R1 3.3 k
R2 3.3 k
JMP 2
JMP 1
External Osc. Input
Disconnect JMP2 & connect JMP 1 for non-synchronous mode
455 KHz I.F.
J1
J2
Notes:
1 All Resistors are 1/4 W 5 % Unless otherwise Noted
2 Resistors marked with asterisk(*) are RNC55 type metal film 1 %
3 Capacitors marked with an asterisk (*) are polycarbonate low leakage type
4 Capacitors C2, C5, & C6 are mounted very near to component pins
5 All interconnects should be kept as close as possible and circuitry should be layed out on a ground plane, and placed in a fully enclosed metal box with appropriate feedthrough power bypasses.

Temperature Control Circuits

Here is a simple temperature controller suitable for a variety of purposes; it was designed to heat the copper float out of a toilet for a barometer! The values and components are not critical; the components shown were selected primarily for easy availability. The temperature setting resistor depends on the NTC type thermistor and should be selected to heat the oven to the desired operating temperature. Temporarily connect a current meter in series with the power supply to monitor the oven's performance. The circuit should draw about 500 ma until the set point is reached and then the current should cut back to about 200 ma (depending on the set temperature, oven structure, and insulation quality). A little cycling before settling is fine but if the current cuts on and off repeatedly, then move the thermistor closer to the heater resistor or decrease the 330k resistor. Higher heating current may be achieved by lowering the value of the heater resistors but at some point the power transistor dissipation may become too high.

Higher resistance negative temp-co thermistors may be used by increasing the resistor connecting from the thermistor to the zener by a proportional amount. The thermistor and 270 ohm resistors are mounted on the float (or other metal oven structure) and the other components are mounted in a separate metal box. To simplify construction, connect the oven structure to the negative terminal of the supply (negative ground) and connect the metal circuit box to the positive terminal (positive ground). This wiring allows the resistors and thermistors to be directly soldered to the oven structure and it allows the 2N3055 to be directly connected to the metal case without insulating hardware. Obviously, the case must not touch the oven structure.

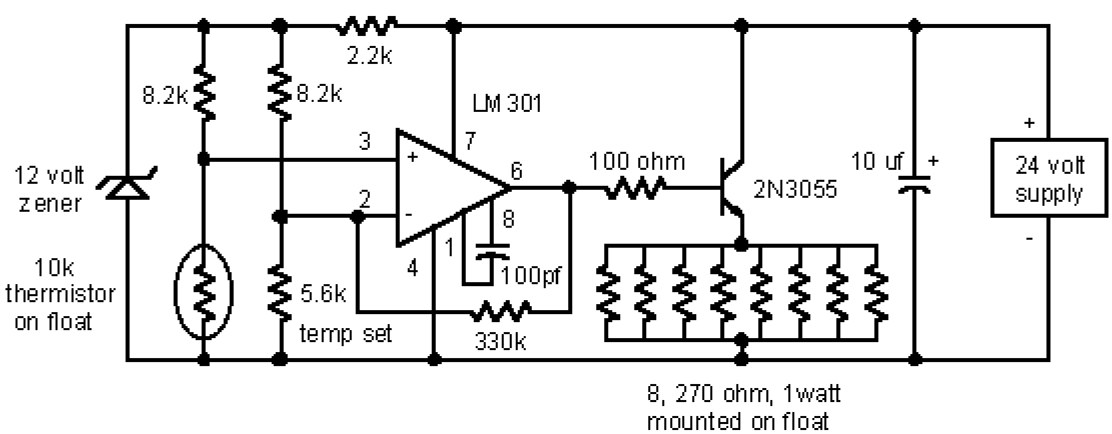
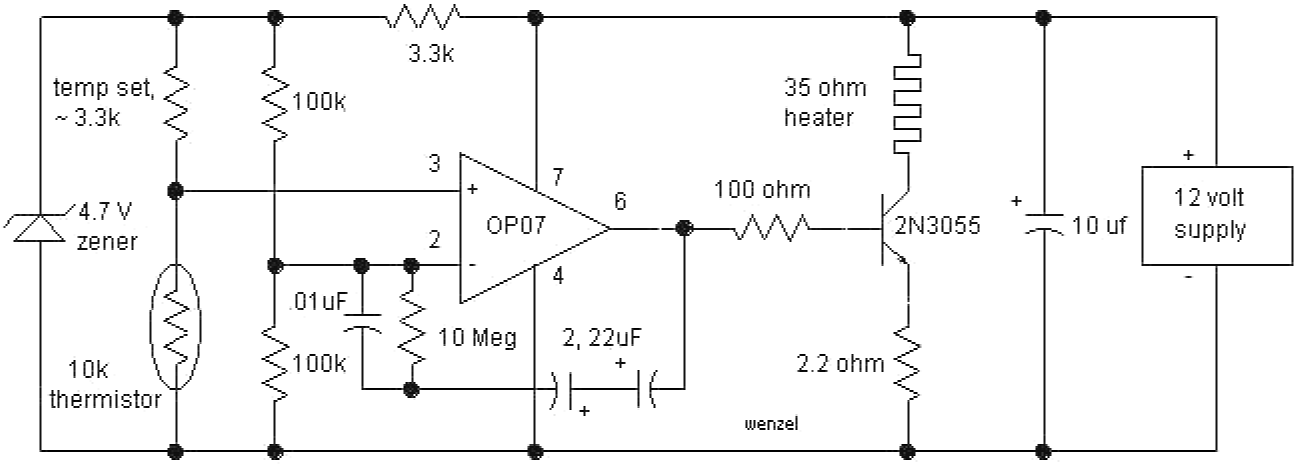


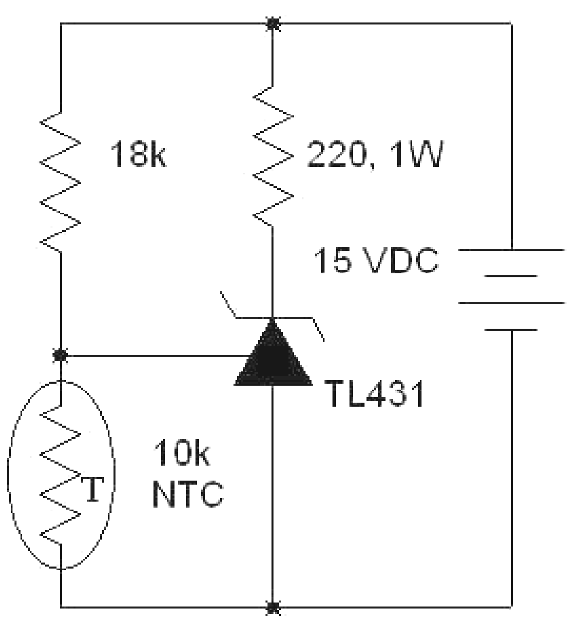
Figure 3: Temperature controller for water barometer.

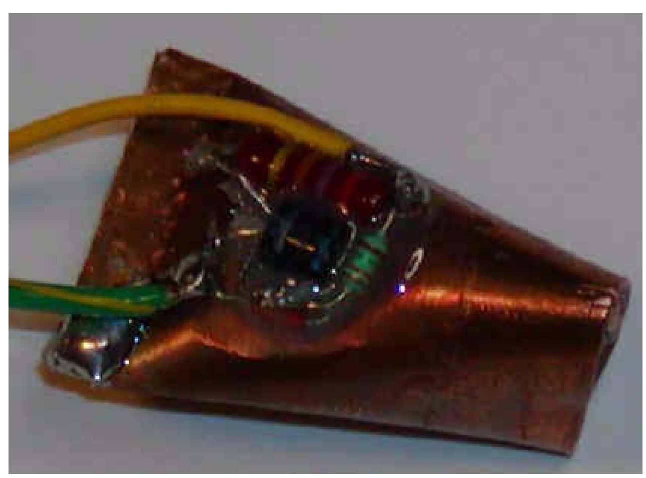
Below is a temperature controller capable of extremely precise control. See http://www.techlib.com/electronics/barometer.html for the oven structure details to make an oven as good as any commercial products with parts from the local plumbing supply house!



The oven controller below is about as simple at they get! It uses a TL431 shunt regulator as the error amplifier and power controller and only three other components. The temperature set by the 18k resistor and the 220 ohm, 1 watt resistor provides the heat, along with the TL431 itself. The oven structure is made by completely flattening one end a 1.25" long piece of 1/2 inch copper tubing with a vise, turning it 90 degrees and flattening the other end, leaving enough of an opening to accommodate the parts to be ovenized. The components are soldered and glued directly to the copper as seen in the photo. The thermistor is a bit difficult to see; it is the orange component that looks similar to a diode buried in the epoxy at the bottom of the photo. The thermal gain of this oven will be about 20 or 30 when it is inside a 2" x 2" x 2" Styrofoam insulator which is sufficient to greatly improve the performance of voltage references, crystal oscillators, etc. To achieve best performance use long lead wires for the circuits in the oven and tuck several inches of wire inside the oven after the circuit to prevent heat loss out the wires.

Warm-up current is about 50 mA and operating current is only about 25 mA with the foam described above. **The operating temperature is about 50° C with a 10k NTC thermistor** with a 3%/°C slope. Other NTC thermistors will work but the 18k resistor must be changed to set the proper operating point.





CAPT. LEE'S PRODUCT DETECTOR FOR THE R-390/ R-390A - REVISITED

by Paolo Viappiani

Editors Note: Because of space limitations the R-390 circuitry and text have been omitted

The simplest way to add a true Product Detector to an R-390A (or to an R-390) is undoubtedly the one described by Capt. Paul H. Lee in his 1968 and 1973 CQ articles ("Modifying The R-390A For SSB", CQ #1/Jan. 1968, pages 55-58; "Using The Surplus R-390 Receiver For SSB", CQ #8/Aug. 1973, pages 43-44 & 82-83).

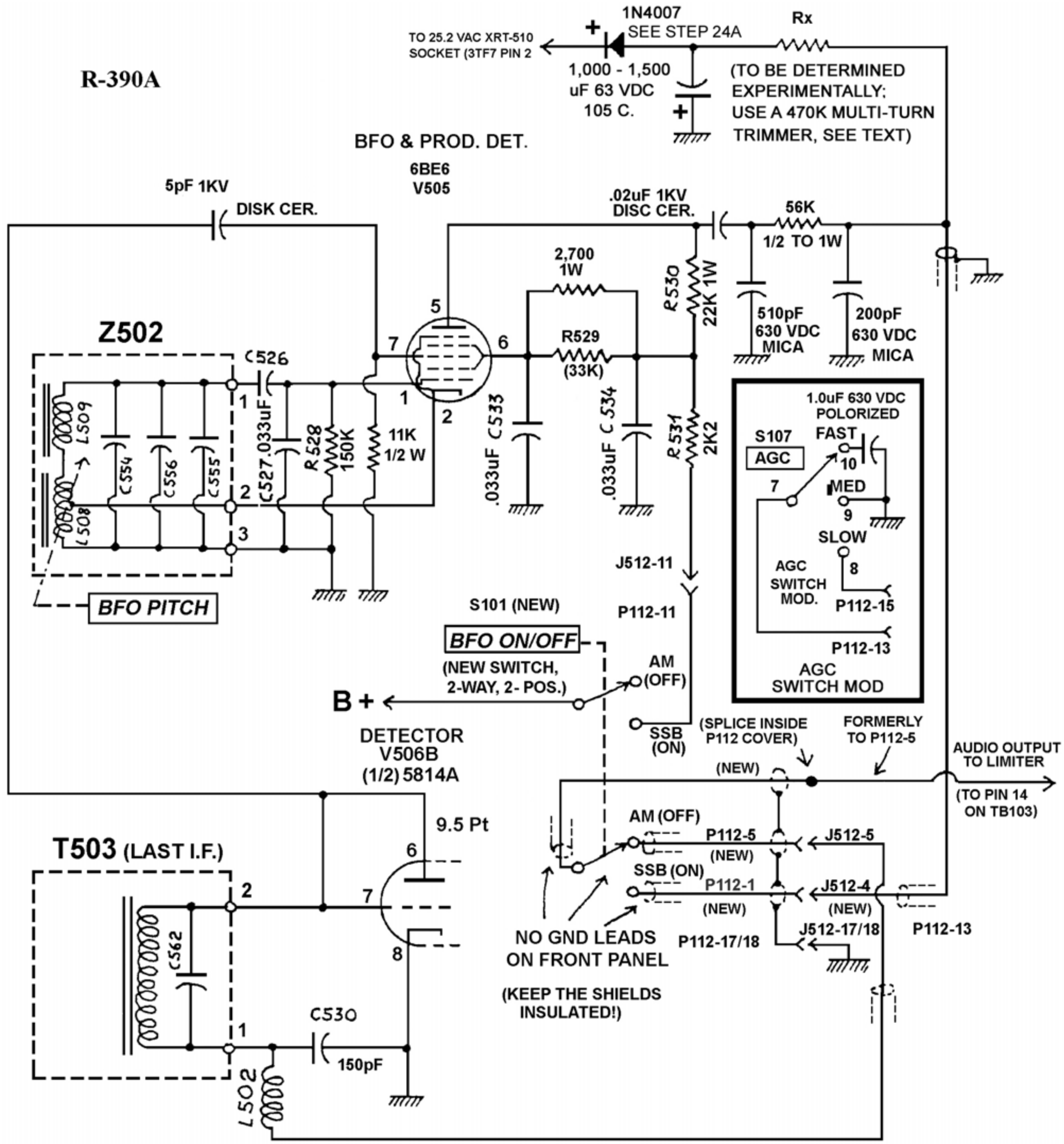
The proposed conversion is quite simple and can be performed in a week-end or so (you have only to make some rewiring and to replace the 5749/6BA6W BFO osc.- with a 6BE6 -osc./mixer-); in addition (unless you have already performed the Dallas Lankford AGC mod described on HSN #27) it might be useful to install the recommended 1.0 mF capacitor between the "Fast" terminal of the AGC switch and a ground lug. The mod works fine both in the R-390 and in the R-390A; but, as Dallas Lankford remarked in HSN #27, the Noise-Limiter is bypassed in the CW/SSB position and a regenerative effect had been reported by Eugene A. Hubbel and others (see: Eugene A. Hubbell, W7DI, "Improving the R-390A Product Detector" on Ham Radio #7/July 1974, pages 12-15). I do not agree with Mr. Hubbell's approach (it requires new holes in the IF subchassis and the addition of expensive components like the Potter & Brumfield relay) and after various trials I succeeded in fixing both the mentioned problems.

I installed the Capt. Lee's Product Detector in a R-390A of mine an EAC unit built in 1967 and had the same mod successfully performed in a R-390 by my friend Rick Mish of Miltronix in Toledo, Ohio. In both the receivers all the regeneration problems have been fixed simply by grounding all the shielded leads of the three involved shielded cables (carrying the audio signal to the new BFO switch) not at the switch side but inside P112 (or P117 in the R-390) only, and the Limiter (and Squelch in the R-390) operation has been restored by applying a small negative bias voltage to the Product Detector Output. As per Mr. Hubbell's idea, the negative bias voltage has been obtained by rectification and filtering of the 25.2 VAC present at the 3TF7 current regulator socket, but with a proper adjustment of that voltage no need was found for also applying a small positive bias to the normally grounded end of the Limiter pot for proper operation (differently from Mr. Hubbell's circuit). So the entire mod becomes simpler and allows using the Limiter both in AM and in SSB.

The following schematic shows the proposed mods for the R-390A. The "Rx" value was found to be 100K; but I think it is better to install a 470K trimmer first and adjust it for the proper bias voltage. After the adjustment, the trimmer can be replaced by a fixed resistor. Please note the restoration of the original 56K/200pF R-C filter proposed by Capt. Lee (and removed by Mr. Hubbell) at the Product Detector output.

In any case I recommend performing the simple modification the Limiter circuit proposed by Bill Kleronomos, KD0HG on ER #70/Feb. 1995 ("Updating That R-390A - An Improved Current Regulator And Noise-Limiter", page 36): in my experience, the addition of a 33K resistor in series to R527 really leads to a Limiter that works effectively and does not noticeably degrade the sound of AM or SSB signals

I think the proposed mods are well worth the effort.



The conversion has to be performed as follows:

1. Disconnect the two control shafts from the I.F. sub-chassis, unplug all plugs from it and remove the sub-chassis from the receiver.

2. Carefully remove the cable clamp and cover from the multi-conductor plug P112 slipping it back along the cable. Remove the wire from pin P112-5 and leave it hanging.

3. Remove the BFO B+ wires from the BFO OFF/ON switch S101. Remove and discard S101, but save the knob.

4. Cut three 20" lengths of single-conductor shielded audio cable (or RG-174 cable); from one end of each piece strip back 1" of outer plastic jacket AND SHIELD BRAID and use the insulating sleeves for protection.

5. Take the new switch, a 2 pole/2 position unit for S101, and connect the inner conductor of the three shielded leads to it, in order to form the audio changeover circuit as shown in Fig. 1.

6. Mount the new switch in the vacant hole on the front panel and replace the knob. Connect the BFO B+ wires to the other pole of the switch (the BFO B+ is to be ON in the BFO "ON" position, which will become the "CW/SSB" position of the new S101).

7. Twist the 3 shielded wires from S101 into a cable and wrap with plastic tape at 3" intervals; cut to the required length to reach plug P112 leaving sufficient slack for clearance over the I.F. sub-chassis and tubes and run the 3 wires through the cable clamp and plug cover.

8. From the free end of each of these 3 wires remove the outer plastic jacket and carefully unravel about 1" of the shield braid, forming a 1" pigtail lead. Twist the 3 pigtails together and solder them, being careful not to melt the plastic insulation on the inner conductors. Slip an insulating sleeve over the common shield lead and over the joint.

9. Connect the inner conductor of the shielded wire from the CW/SSB (BFO ON) audio terminal of S101 to pin P112-1 (a "spare" free pin in P112); use an insulating sleeve for protection (as is done for the other wires on the plug).

10. Connect the inner conductor of the shielded wire from the AM (BFO OFF) audio terminal of S101 to pin P112-5. Use an insulating sleeve as in step 9 above.

11. Slip an insulating sleeve over the free end of the remaining wire, which should be the one connected to the rotary arm of the audio section of the new S101; this is the "audio input" lead. Solder it to the free

12. end of the wire left hanging in step 2 and slip the insulating sleeve down over the bare connection.

13. Carefully solder the "pigtails" lead to pin P112-17 or to P112-18 (both ground terminals in P-112); this one must be the only connection to ground of all the three added audio cables. Pay close attention in order to avoid undesired contacts in P112 (eventually use tape or insulating sleeves as required); carefully replace the cover and cable clamp on P112. Tape the three new wires to the existing cable just outside the clamp.

14. Turning the I.F. sub-chassis over, carefully remove the bellows coupling on the BFO PITCH control shaft; remove the shaft by loosening the panel bearing (in order to clear some working space around the XV505 socket). Remove V505, 6BA6 (BFO tube) and add it to your parts box.

15. Remove the ground (and all wires) from pin 2 of XV505. This may involve shifting several ground leads to other ground tie-points on the sub-chassis.

16. Move the existing lead from XV505 pin 7 to pin 2 (this is the cathode tap on the BFO coil Z502) and in the following steps be sure to leave enough room for replacing the bellows shaft coupling.

17. Connect the 11K 1/2W resistor from XV505 pin 7 to ground.

18. Remove and discard C535.

19. Connect the 2.7K 1W resistor in parallel with the existing screen dropping resistor R529.

20. Connect a 5p SILVER MICA or DISK CERAMIC capacitor between XV505 pin 7 and XV506 pin 6 (and/or 7). This is the I.F. coupling into the injection grid of the 6BE6 Product Detector.

21. With a pair of small metal shears cut a 1/4"V-shaped notch in the lower edge of the interstage partition near the rear of the BFO coil Z502. Cover the edges of this slot with short pieces of plastic tape.

22. Mount the 200 pF and the 510 pF SILVER MICA capacitors on the grounded center post of the XV506 socket, letting them be supported in space by their own ground leads (about 1/4" long).

23. Connect a 56,000 Ohm, 1/2 to 1 W resistor between the free ends of the 200 pF and 500 pF capacitors.

24. Connect a 0.02 uF/1,000 V DISK CERAMIC capacitor from XV505 pin 5 to the 510 uuF end of the Ohm resistor.

24A. Make a bridge rectifier using four 1N4007 diodes and the ripple will be 120Hz instead of 60Hz which will be much easier to filter.

25. Locate XRT510 (the 3TF7 current regulator socket in the corner of the sub-chassis below the BFO tuning shaft) and solder the cathode end of a 1N4007 diode to XRT510 pin 2; let the diode be supported in space by its cathode lead or install an insulated standoff on the side of the I.F. sub-chassis for a more adequate support (this requires drilling a hole in the sub-chassis).

26. Connect a 1,000-1,500 uF, 50 to 63 VDC/105° quality electrolytic capacitor from chassis ground (positive terminal) to the anode lead of the 1N4007 diode. A ground lug under the spade bolt holding the 12 mH RF choke (L502) works fine for the positive (grounded) terminal of the capacitor. Be sure the negative lead of the capacitor is properly insulated from the chassis ground.

27. Add a small insulated standoff to the XV506 screw towards the BFO can (if there are not enough screw threads to mount the insulated standoff securely, you will have to replace the screw with a longer one) and connect its lug to the junction between the 1N4007 diode (anode lead) and the electrolytic capacitor (negative lead). You can use a small piece of insulated wire for the purpose.

28. Place a small 470K, 25-turn trimmer on the outer side of the I.F. sub-chassis (you can use a drop of glue for fixing it temporarily in place) and route two insulated wires connecting the trimmer center lead to the 200 pF end of the 56,000 ohm resistor (see step 22) and one of the outer trimmer leads to the junction between the 1N4007 diode (anode lead) and the electrolytic capacitor (negative lead).

29. Use 12" of RG-174 cable for the CW/SSB audio lead. Remove 1" of the plastic jacket from one end, and make a 1" braid pigtail on this end. Slip a 7/8" insulating sleeve over the pigtail and ground the pigtail to the center ground post of the XV506 socket. Connect the center conductor to the 200 pF end of the 56,000 Ohm resistor.

30. Lay the RG-174 in the V-shaped slot in the interstage partition and tape it in position with a 2" length of plastic tape. Cut the wire to length to reach pin J512-1 of the rear cable socket (the formerly unused pin that mates with pin P112-1 of the cable plug). Strip back 1/2" of the plastic jacket and braid from this end of the shielded wire; connect the center conductor to pin J512-1, using an insulating sleeve for protection.

31. Carefully replace the BFO shaft and bellows coupling removed in step 13; make sure the coupling doesn't accidentally ground any component or wiring.

32. Replace the IF sub-chassis in the receiver (allowing a bit of space in the bottom for the crossing of the two trimmer leads); insert all the plugs previously removed and temporarily reconnect the 2 control shafts and their front panel knobs (make sure they are properly positioned).

33. Plug in a 6BE6 tube in the XV505 socket; turn on the receiver and switch the new S101 to the CW/SSB position (BFO "ON"). With the antenna disconnected and the BANDWIDTH switch in the 1 Kc position, set the BFO PITCH control for the lowest pitch of the hiss. Then, without rotating the shaft, loosen the knob set screw and set the knob pointer to "0".

34. Connect the antenna and enjoy both AM and CW/SSB reception. Switch the Limiter on and with various settings of the LIMITER control carefully adjust the 470 K trimmer for the most satisfactory operation (caution: this adjustment requires several trials and a long time listening).

35. After having adjusted the trimmer properly, disconnect its leads (this requires the I.F. subchassis removal) and measure the resistance with an Ohmmeter. Remove the two wires crossing the bottom of Hollow State Newsletter 40 (Winter 1996/97) the sub-chassis and connect directly the standoff near XV506 and the 200 pF end of the 56,000 Ohm resistor with a fixed resistor of the nearest value to the measured one (1/2 to 1 W rating).

36. Replace the I.F. sub-chassis and enjoy your R-390A!

AGC and Limiter mods*

While you are in there, if you have not yet performed the Chambers/Lankford AGC mod in you R-390A (see HSN #27 page 3) you should consider doing it; in my opinion that AGC mod is the best one ever proposed. Otherwise, you can perform the simpler one proposed by Capt. P. H. Lee.

To do this, you have to simply connect to the unused terminal (#10 in the R-390A, #2 in the R-390) of the AGC switch (FAST position) a 1.0 uF/630 V mylar or poly capacitor with the other lead connected to ground (i.e. to a ground lug placed in the vicinity).

It is not the best possible solution, but it really works and can help in several circumstances. In order to reduce the severe audio distortion created by the Limiter circuit you can also perform the mod proposed by Bill Kleronomos, KD0HG, on ER #70: it is a very simple mod, as you have only to add a 33K 1/2W resistor in series of a 27K 1/2W one (R527 in the R-390A), and I think it is well worth the effort.

***One might want to do the AGC mod proposed by David Wise at the beginning of the chapter.**

High-gain Product Detector

"Fig. 6 shows the circuit as a high-gain product detector. This circuit has good dynamic range and a microvolt input will deliver about a millivolt output. Notice that this circuit is identical to fig. 4 (Mosfet Mixer Stage) except for the details of the output circuit. The output transformer is a regular Radio Shack item. The 500-ohm output is convenient if you go into a low-pass audio filter built from 88-mH toroids, as it provides a good impedance match.

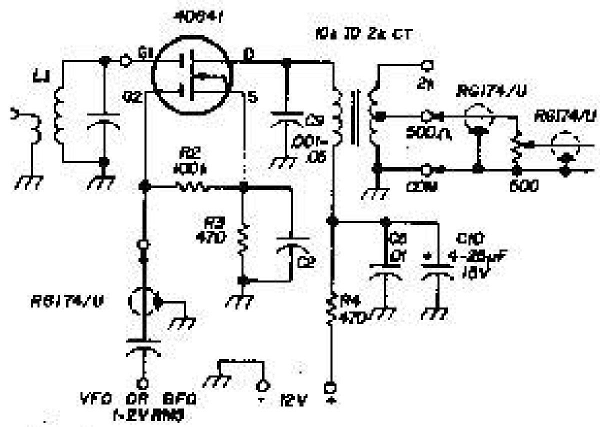


fig. 6. High-gain product detector. Except for the output this circuit is identical to fig. 4.

Notice the volume control hookup. If the volume control is located remotely it is not desirable to ground the low side of the control to the chassis at the remote point.

This can lead to ground loops. Instead, carry the ground with the shield of the miniature coax, and from the control to the next stage on the shield of that same piece of coax. This ground loop problem is very easy to control in a modular setup like this. You can mount the boards on pieces of bakelite or plastic when necessary and have full control of the ground situation without any radical rebuild.

Notice that the output is bypassed for both rf and audio (C5 and C10). Also notice C9. This capacitor can roll off the high-frequency audio components. That little transformer is quite efficient out to 70 khz or more, and there is no need for response above about 2.5 khz. In fact, if you use a .05 mfd capacitor at C9, you will have a definite peak around 1 khz that is helpful for CW work, but still results in useable ssb performance. Again, the bfo voltage is brought in with RG-174/U coax. With a high-gain IF strip, it may be helpful to locate the bfo well away from the strip - or shield it.

...Slightly more gain will result if C2 is paralleled with an audio bypass capacitor,...

The circuit in Fig. 6 can also be used as a synchrodyne detector. Tune L1 to the operating frequency, add an audio filter to the output and use a high-gain audio amplifier. You now have a simple, direct-conversion receiver."

R-390A Upgrading

Panel and Frame

3-prong Power Cord: (optional, recommended if not already installed) Change any two-prong power cords to 3-prong. Insert ground lug from green wire underneath one of the AC terminal cover screws.

Inrush Current Limiter: (optional, highly recommended) Keystone CL80 (or CL90 if you never run the ovens) added between 120 VAC fuse and line filter. This mod limits AC voltage to the radio over a brief few seconds. It also reduces the incoming AC voltage by a couple of volts. Documented on Jan Skirrow's website at http://www.skirrow.org/Boatanchors/ and in HSN issue 48, page 3.

B+ Fuses: (these can be added if not present) Fuse F102 is located in the B+ line between pin 5 of plug P111 and pin 5 of plug P119. Fuse F103 is located in the B+ line leading from pin 2 of plug P119. Chassis wiring has to be unlaced, rerouted, and re-laced to complete this installation. See the wiring schematic in the Y2K Manual on pages 6-83 and 6-84 to determine which changes are necessary.

Ground-fault protection capability: (optional) Insulate the line filter from the chassis using insulating sheet, insulated feed-throughs, two soldering lugs, and 0.01 uF 1000 VDC poly capacitor. For details see HSN issue 43, pages 7 & 8. (I do not add this mod to mine)

PTO

Field Change 7: R702 changed from 56K to 220K. R702 is connected to pin 6 of V701. Documented in Navsea 0967-LP-063-2120. (I do not add this one)

AF Sub-chassis

Production Mod 1: C612, 68 pF, added in parallel with R601. R601 is 680K connected to pin 2 of V601.

WA4HHG (Chuck Rippel) Easy Audio Mod: Replace C601, C604, and C605 with a 0.022 uF 400V orange drop capacitor, and replace R614 with a 560 ohm 2 watt resistor. Documented here and in HSN issue 46, page 6 (adding the C601 capacitor in the HSN version). Note that some prefer even larger values, up to 0.047 uF. Replacement capacitors which are too thick can be mounted either off the edge or under the circuit board.

Kleronomos Real Audio Mod: This is a major (non-reversible) mod that is documented in Electric Radio issue 42. It converts the AF deck to deliver 5 watts of push-pull audio into an 8-ohm speaker.

Production Mod 2: R504 changed from 1000 to 560 ohms. R504 is connected to pin 7 of V501.

SSB Mod: (optional, highly recommended) Uses two 1N4148 diodes and one 47pF ceramic disk (or silver mica) capacitor.

-one diode in parallel with R547, cathode as follows
-----|<-------pin 2 V506A

-one diode in parallel with R546, cathode as follows
----->|-------pin 1 V509A

-one 47 pF cap in parallel with C535

Mods very similar to this one are described be Lankford and Cornelius in issues of Hollow State News.

R-390A Audio Modifications by Chuck Rippel, WA4HHG

Some of you may know that I restore R390A receivers as a hobby. Here is an easy "soft " (reversible) modification which I came up with that really improves the R390A audio by mellowing it out a bit.

The audio tubes, V602A and V603 are coupled with 300V, .01mfd "Vitamin Q" type capacitors. These have been found to leak causing distorted audio. The values are also tend to make the audio amplifier somewhat restrictive. Changing these to .022 (or even .047 or .05 mfd) 400V (min) Sprague Orange drops improves bass response and the increased voltage rating coupled with modern construction helps to keep the leaking problem from re-occurring. Do NOT use disc ceramic capacitors!

Parts needed:

2- .022mfd Sprague Orange drop Capacitors
1- 560 ohm, 2 Watt resistor

Procedure:

Unplug the receiver and remove the bottom cover. Locate the audio sub-chassis (it's on the bottom of the radio and has five tubes on it) . Remove the audio chassis by pulling off the 2 modular plugs and loosening the 4 green captive Phillips screws. Lift the audio chassis onto an appropriate workspace and turn it upside down. There is a 2" wide circuit board which runs 2/3 the long length of the audio chassis. Locate C-604, C-605 and R-614 which are mounted on that pc-board. Remove the two screws holding the board to the audio chassis then fold it with the cables off to the side in order to expose the bottom of the board.

Remove C-604, 0.01 mfd 300v Vitamin Q.

Remove C-605 0.01 mfd 300v Vitamin Q.

Remove R614 560 ohm 1 watt resistor.

Replace C-604 and 605 with the orange drop capacitors. Be sure to first make sure the capacitors do not take up too much vertical space and not allow the audio deck to properly seat in the main chassis. If space is a problem, try mounting the capacitors on the opposite side of the pc board from which they were removed. This is especially true if you use 0.047's or 0.05's or use 600V capacitors.

Replace R614 with the 560 ohm 2 watt resistor.

Replace the PCB, put the audio chassis back in the receiver and re-assemble the radio.

During operation, you will immediately notice that there is a more mellow sound to the receiver. This is due to the increased bass response allowed by the larger value capacitors. If you check the 560 ohm resistor, R614, you will probably see that it has changed value going higher in resistance. I have seen them go to 700-800 ohms. The receiver can be put totally back to "stock" in a matter of minutes by simply reversing the process.

2. Relocate the 3 power resistors from the underside and remount by the squelch plate

IF Sub-chassis

IF Deck

Increase the BFO injection for R-390A diode detectors by paralleling a 47 pF 500 volt silver mica across C535 12 pF. K50P

Production Mod 2: C507 and C517 changed from selected value to 51 pF. C508 through C510 and C513 through C515 changed from selected values to 82pF. Trimmer capacitors C564 through C571 added. See schematic and a modified IF deck for details and correct layout.

Production Mod 2: R504 changed from 1000 to 560 ohms. R504 is connected to pin 7 of V501.

SSB Mod: (optional, highly recommended) Uses two 1N4148 diodes and one 47pF ceramic disk (or silver mica) capacitor.

-one diode in parallel with R547, cathode as follows
-----|<-------pin 2 V506A

-one diode in parallel with R546, cathode as follows
----->|-------pin 1 V509A

-one 47 pF cap in parallel with C535

Mods very similar to this one are described be Lankford and Cornelius in issues of Hollow State News.

3TF7 Substitution Mod: (optional, recommended) Add jumpers on RT510 between pin 7 and pin 5, and between pin 2 and pin 4. This allows you to later substitute a 12BH7A tube in place of your 3TF7 if (when) it ever fails. (HSN issue 10, pages 1&2 or HSN reprints, page 1)

RF Amplifier Sub-chassis

Production Mod 2: Pin 7, V201, connected to ground instead of pin 2.

Production Mod 2: C275 changed from 5000 pF to 3300 pF. C275 is between S208 rear, pin 7 and ground. Note: *The consensus of opinion is that C275 should be* 0.033 uF. *This value was consistently used in manufacturing contracts during the 1960's.*

Production Mod 3: Series network of C256, 0.1 uF, and R235, 47 ohms, inserted between terminal 1 of HR202 and ground.

Production Mod 4: C257, 47pF, added in parallel with C227. C227 is located betweenV201, pin 2 and ground.

Field Change 7: R210 changed from 56K to 220K. R210 is connected to pin 6 of V207. Finger stock added to shield RF chassis if available. Documented in Navsea 0967-LP-063-2120. (I do not add this one)

Power Supply Sub-chassis

Solid State: (optional, only if 26Z5W tubes not available) Cover tube socket with labels regarding solid state mod. Use 3A, 1KV rectifier diodes. One diode between pins 1 and 4 of socket XV801, connecting the cathode lead to pin 4. Another diode similarly installed on XV802. Documented in Navships 0967-063-2110.

Many suggest adding a 220 ohm dropping resistor to lower the B+ voltage to 240 VDC. (Also see HSN reprints, page 5) Some versions add the dropping resistor to the Power Supply Chassis with the 220 ohm 10 watt resistor between T801-6 and ground (http://www.qsl.net/wb4tur/milt/index2.htm website and in HSN issue 48, pages 5 & 6). Some add the 220 ohm resistor to the AF deck (Navy EIB 895 and HSN issue 2, page 5 or HSN reprints, page 5). I prefer to add it to the power supply deck so that I can freely swap power supplies between tube and solid state type and the mod travels with it.

Other Mods

Other

Tube shields: Tube shields are needed on V201, V206, V505, and V701 (HSN issue 5, page 3 or HSN reprints, page 3). Others are often removed if shiny, but should be kept in place if black (inside and outside). IERC shields are the most highly regarded, with WPM coming in second.

Potential Low-Noise Tube Substitutions:

I don't do the following unless I run out of tubes. I tried all of these in my Collins R-390A for several months, but could not really tell any difference. AGC action may have been a bit worse, and sensitivity measurements didn't really change. Resting noise level did decrease with no signal injected, but it was compensated by less gain when signals were applied. Net result: no measurable change in sensitivity. If I had an abundance of the following tubes that would allow me to "tweak" the performance a bit, I might have had better results. Regardless, here are some easy substitutions for you to try.

6BZ6 instead of **6DC6** in V201 and **6AH6** instead of **5749/6BA6W** in V508 (National Radio Club reprint R-57)

6JH6 instead of **5749/6BA6W** in V501, V502, and V503 (Electric Radio issue 26, pages 22+)

6AB4 instead of **6C4**

Inrush Current FAQ

Inrush Current Protection

1. Inrush Current Protection
2. Socratic Exchange: Theory & Protection
3. Voltage Spike Protection

The following series of posts to the BoatAnchors list consider the problem of protecting equipment from the transient high line current that can occur when the power is first turned on.

Unfortunately in the course of saving these and putting them together, the headers were lost, along with the names of some of the posters. Thus, I've generally omitted the identification of the poster. This also leaves me free to correct the spelling and delete the repetitive parts. But, if you recognize your stuff, and would like your name reattached, just ask!

As always, these posts are offered without any guarantees!!

Inrush Current Protection

Post 1: Selecting an Inrush Current Limiter

Some time ago I discovered a neat little device that solves the inrush problem and, as a side benefit, reduces high line voltage. All this for a bit over $2! The device is an Inrush Current Limiter made by Keystone Carbon Co. The beasties look like ceramic disk capacitors with a black vitreous coat.

The limiter is a **Positive Temperature Coefficient Thermistor** which is designed to handle current. When cold (room temp -25C) they exhibit some resistance. As current passes through them and they warm up, this resistance drops by a factor of about 100. The limiters are rated by current handling capability (1.1 to 16 Amps) and cold resistance (0.7 to 120 Ohms). Not all possible combinations of resistance and current are available but at last look there were about 20 different types.

You use the limiter by installing it in series with the line cord (preferably the hot lead) input to your BA. This can be done in a fashion that is totally esthetically pleasing (read "out of sight") and completely reversible.

IMPORTANT: Since the device is a resistor (and a HOT one at that) you must mount it away from heat sensitive components. I have mounted them under chassis without trouble but keep 'em away from just about everything. Don't attempt to heat sink it - that ruins the operation!

Pick the right value by first measuring the steady state current of your BA. That is, after it is fully warmed up and all accessories are turned on. While you're at it , also read your line voltage. Pick a unit that has a MAX steady state current of 120 - 130% greater than the current you measured and has the HIGHEST no-load or cold resistance.

Example: You measure 2.5 Amps (a moderately hungry BA!) and the line voltage is 123V. The KC008L is rated for 3.0 Amps with a cold resistance of 47 Ohms – a nice fit.

Benefits: A BA drawing 2.5 amps probably has a transformer with a primary DC resistance of about 3 Ohms. Inrush, at the peak of the AC sine wave, could be as high as 40 Amps but probably not less than about 20 Amps. With the limiter installed, the inrush will not exceed about 2.6 Amps at 123 line Volts. After the limiter warms up it will have about a .49 Ohm resistance (actually a bit higher because we're not drawing the full 3 Amps.). This means that the line voltage across the transformer will be about 122 Volts (also a bit lower because of the higher resistance). This example came from real life and my actual results showed that the line voltage was reduced to 118 Volts (the BA was rated for 117) which means that the limiter was adding about 2 Ohms.

Negatives: If your area suffers from brownouts, the limiter will exaggerate the effect. If voltage drops, current drops. The limiter will cool a bit, its resistance will rise, and the voltage your BA sees will drop more than the line voltage. This is a very minor problem for me but I feel bound to mention it.

Post 2: Experience with a 51S-1 Receiver

In the past few years, a new kind of thermistors has become available for limiting start-up surge currents in electronic instruments. They differ from conventional thermistors in having a negative temperature coefficient (resistance decreases with increasing temperature), and this property gives them a useful self-regulating characteristic. Placed in the ac line of an instrument, they initially have a high resistance, which limits the inrush current through the instrument. Upon application of power, the current through the thermistor causes selfheating, which lowers the device's resistance. At some point the resistance stabilizes to a value that depends on the equilibrium temperature of the device. The equilibrium temperature is determined by the steady-state current drain of the instrument and the ambient air temperature surrounding the thermistor. Current-inrush thermistors are inexpensive and provide an effective way to protect power supply components in vacuum tube receivers, particularly those that use solid-state rectifiers. Note that you should not use current-inrush thermistors to protect transmitters or amplifiers; they are only suitable for instruments that draw a relatively constant current from the line. (See later post) Here are the details for protecting a typical boatanchor receiver, in this case a Collins 51S-1.

The steady-state current drain for my 51S-1 is about 0.8 Amps at 120VAC. To measure the inrush current, I temporarily removed the 1.5 ampere slow-blow fuse and jumpered a 1 ohm resistor across the fuse terminals. By measuring the voltage developed across the resistor with a scope, I determined the peak inrush current to be slightly more than 7 amperes! The equivalent load resistance presented by the 51S-1 at turn-on is thus (120 VAC/7 Amperes) = 17.1 ohms. As the filter capacitors charge and the tube filaments warm up, this load resistance increases to a steady-state value of (120 VAC/0.8 Amperes) = 150 ohms. A 7 ampere inrush current is very hard on the power switch, and isn't so great on the power transformer, rectifier diodes, and filter capacitors.

The most suitable inrush thermistor I could find was Digikey ………………. (1-800-DIGIKEY) part number KC014L-ND, at a price of $2.13. This thermistor is specified at 50 ohms resistance at room temperature (54 ohms measured), and dropping to 0.89 ohms at 1.1A load. I measured the resistance at 1.1 ohms at the current drain of the 51S-1. To install the thermistor, I clipped the wire to the fuse socket of the 51S-1 and relocated it to an unused lug on a nearby turret. I then soldered one lead of the thermistor (which physically resembles a small disk capacitor) to the same lug and the other to the recently vacated lug on the fuse socket. I used a bit of Teflon tubing on the leads, and kept the leads long so I could suspend the thermistor in free space away from other components. The thermistor dissipates about a watt of heat and runs rather hot.

After installing the thermistor, I replaced the fuse with a 1.5 Amp fast-blow type. I then re-measured the peak inrush current and found it now to be only about 1.8 Amperes, which is consistent with the theoretically expected value of 120 VAC/(54ohms+17ohms) =1.69Amperes. The peak inrush current is now only slightly greater than the steady-state current drain and should thus pose no problem for any of the power supply components. Note that this particular thermistor is appropriate for almost any boatanchor receiver that draws 75-150 Watts from the power line.

Concern is often voiced about a related turn-on problem (actually, a turn-OFF problem), namely the inductive voltage spike caused by the power transformer inductance when the power is switched off. This spike is reputed to cause sparking and welding of contacts in hard-to-replace power switches, particularly in rigs like the KWM-2 and S-line. I checked on this problem with my 51S-1, but measuring the peak voltage developed across the power switch when the rig was shut off. (My Fluke 87 DMM has a peak-reading feature which can capture voltage transients as short as 1 msec.) To my surprise, I found that the inductive voltage kick was only about 5 volts higher than the line voltage, and was no cause for alarm. I had thought about using an MOV surge suppressor across the switch contacts, but decided it wasn't necessary. This is not to say, of course, that the problem isn't greater in some other rigs, but 51S-1 owners need not worry.

Post 3: Inrush Protection for Transmitters

Comment:
>>Note that you should not use current-inrush thermistors to protect transmitters or amplifiers; they are only suitable for instruments that draw a relatively constant current.

Response:
Au contraire. Inrush current limiters work nicely in transmitters and transceivers and probably in amplifiers as well, although I've not tried that. The only stipulation is that the device must be selected to allow the maximum current needed by the transmitter. The resistance of the thermistor after the initial surge is very small, a fraction of an ohm, and less than the resistance provided by the typical AC mains. Consequently, its effect upon the load regulation of the transmitter is negligible. I used an inrush limiter a while back in an Eico 753/751 transceiver supply with excellent results. Prior to using the inrush current limiter, the power on surge produced an unnervingly loud KWUMMP! After installing the inrush current limiter, powering up the unit produced no audible effects at all. I don't happen to remember the voltage drop across the inrush limiter when just the receiver was operating, but I did measure it and found it be negligible; on the order of only a volt or two.

And Someone Else Added:
Of course, on larger transmitters one has to use thermistors on each element. Generally the current draw is too large to protect the entire transmitter. The filaments transformers, the plate transformer, low voltage transformers should all be individually "thermistorized". In mine, I find a volt or two drop at the thermistors is just what the doctor ordered as the line is slightly high.

Post 4: Mounting Caution
Don't solder them into your circuit unless you want trouble. They do get hot in operation and repeated heating and cooling of a solder joint will cause it to crystallize and eventually fail. This was a common failure in televisions with thermistors used in the degaussing circuits, and even with some of those cement block power resistors on circuit boards.

Put in a small screw terminal strip to mount the ICL. Crimp terminal lugs on the ICL and then attach it to the strip with the screws. In the long run, this will save lots of grief and it also makes installing and insulating the ICL a snap.

Voltage Spike Protection
Post 1: MOVs

Turning off a rig can cause a big voltage spike across the transformer primary and the AC line. Usually it just burns out or welds your switch, as R-390A users often learn.

A back issue of The Collins Journal suggested getting 240-volt MOVs and wiring them across your primaries to absorb the transient. Note that if the MOV fails (shorted) it will suck lots of current, but you have a fuse in the line, right? These will protect your switch, and apparently your transformers could use it too. I doubt the big toggle snappers in a Viking need it as much as the wimpy switches in an R-390A or KWM-2, but your transformers may last longer this way. And you'll get protection from nasty things that come in through your power line, and your gear won't put glitches back out there when you turn it off.

Post 2: Selecting MOVs
Query:
>> There have been a number of posts touting the use of varistors to protect against voltage surges. Question is: How to decide what specs when buying these little doo-dahs?

Answer:
My background is in Mechanical Engineering, so take what I am about to say with a grain of salt. When I have picked MOVs (metal oxide varistors) in the past, say to protect stuff on the AC line against spikes, there are two things I have been concerned about. First is the clamping voltage. These little do-dads work by turning from a non-conductor to a conductor at the clamping voltage. The other rating is the amount of current they can handle. Usually this is broken into two numbers, a surge number with a time (like 7000 amps for a microsecond) and a steady state value if I remember correctly. So when I picked one to make into a AC surge suppresser, I picked a clamping voltage of about 150 volts with the highest current capacity I could afford.

Socratic Exchange: Theory & PROTECTION
Post 1: Theory
On Fri, 23 Aug 1996, Jan Skirrow, VE7DJX, asked me some excellent questions about thermistors, varistors, and such. I hope he does not mind me posting his questions or my reply to the group. Thermistors are not often seen in boatanchors (or in a lot of modern semiconductor stuff for that matter). I know a little about them because of their use in temperature measurement and instrumentation.

>>First, I conclude that NTC thermistors would be placed in series with, for example, a transformer and would thus limit in-rush current because their resistance is inversely related to temperature, which would rapidly increase on start-up.

Exactly. They are particularly beneficial with power supplies having capacitor input filters.

Look at the special devices sold as Inrush Current Limiters, not conventional thermistors. Keystone is probably the most common NTC Inrush Limiter manufacturer.

The typical resistance ratios of common NTC thermistors (for other than Inrush Current Limiting operations) is generally between 5 to 10 for 0 C to 50 C temperature changes. Plugging these numbers into the typical resistance relationship

R = Ro * exp(B/T) R, Ro in ohms, T in Kelvin gives a Beta in the approximate range of 2800 to 4000. Using a value of 3400 as an average gives an Ro value of 0.0011 ohms (the resistance at absolute zero). So at 50 C, the resistance should be around 41 ohms (and at 0 C, the resistance is 284 ohms and the ratio is: {ta-da...} 6.9).

In true Inrush Current Limiters, the Beta value is MUCH higher. If, for example, B is 10,000, the 0 to 50 C ratio is 290. This is such that a few ohms cold becomes very low resistance when hot. I don’t really know what the Beta number is for these devices but I might be able to estimate it from the specs knowing the dissipation of the hot device and estimating some heat transfer conditions. It is not necessary to know it for picking an Inrush Current Limiter for your operation.

In any event, a typical Inrush Current Limiter might have the following specifications (actually those for a Keystone CL-110):

Resistance at 25 C: 10 ohms +/- 25%

Maximum Steady State Current: 3.2 amps

Approx. resistance @ maximum steady state current: 0.18 ohms

>>I assume your reference to older metal oxide devices doesn’t refer to metal oxide varistors - which seem to be a transient suppressor that functions by clamping the voltage across itself to some fixed level.

NTC thermistors are generally made from oxides of manganese, nickel, cobalt, copper and iron. Metal oxide varistors for transient voltage suppression are generally variations on zinc oxides. Older thyristors were generally silicon carbide.

It is interesting that while quite different in operation, the thermistors and varistors obey similar exponential relationships. The simple thermistor relationship is shown above. The current through a varistor follows a similar one:

I = Io * exp(a*V) I, Io in amps, V in volts

If you look at more exact relationships with both temperature and voltage dependency included, the equations start looking VERY much alike.

Basically a varistor draws very little current at low voltages, but as the voltage increases, the current increases very rapidly.

>>So these would be used by placing them across (for example) switch or relay contacts that switch an inductive load, and would prevent the voltage across the contacts from going too high due to transients, thus arcing and damaging the contacts.

That is one use, although in snubbing an inductive load, the presence of a diode in a DC circuit or a varistor is an AC circuit will slow down the response of the relay. You really need something that will absorb the energy stored in the magnetic field.

The more common use of a varistor is across the AC line as a transient suppressor. The voltage rating is chosen such that the device does not conduct much at normal voltages, but conducts heavily during a voltage transient.

>>So, comprehensive protection for, say an R-390A, would be an NTC thermistor in series with the power transformer and a varistor that clamped at something over normal line voltage (perhaps 150v rms?) across the troublesome main power switch.

Sort of! An Inrush Current Limiter in series with the transformer primary would reduce the current surge during turn-on. A varistor across the main power switch might help a LITTLE but what you really need here is a snubber network of a resistor in series with a small capacitor. Typical values might be 10 to 100 ohms in series with a 0.01 to 0.05 uF capacitor (rated at 1 KV minimum). A better approach would be to use a better switch!

A 130 volt varistor, like a V130LA20, would be a good choice to add after the power filter network across the line. It would protect against line voltage transients. However, it won't protect the filter here. You should probably use a proper transient protected multiple outlet strip to power the radio anyway. The best ones will have 3 varistors inside. One from line to neutral, and one each from line and neutral to ground.

Inrush Current Limiters and Transient Voltage Suppressors are quite inexpensive today. Small and unobtrusive, they can often be tucked inside your Boatanchor giving you some added protection.

Post 2: Additional Comments
>>I seem to remember horror stories about some so-called transient protected outlets that worked once, and then provided no protection as the varistors went south. All of my outlets are so protected, and I hope they all work! This is important too. In transient suppression, you want to have as much impedance between the source of the transient and the device you want to protect as you can get. Thus for best protection, a staged approach is a good one. At the service entrance to your house, you should have one of the lightning arrestor/transient suppressor blocks made for this purpose. These cost $15 to $30 at a commercial electrical supply house. The only problem is that with installation at the service entrance, you usually have to pull your power meter. Between the service entrance and the wall outlet, your house wiring provides some distributed capacitance and inductance. A 3-MOV protector at the outlet is a good idea here. Checking them is a problem as there is no simple way to do this. If your circuit breaker or fuse blows upstream of the protector for no apparent reason or during a thunderstorm, you can probably assume the protector "went south" and needs to be replaced.

Finally at your equipment, its line cord and RFI filters provide even more impedance. A transient protection MOV inside the rig provides the final stage of protection. It can be smaller in its ratings since the earlier protectors should have already taken most of the energy away from the transient.

Nothing protects against a direct-hit of lightning though. But I would still rather have a few MOVs explode, and maybe a line-filter or two, than the entire rig to replace!

-Subject: [R-390] In-rush current limiting

For those of you who want the ultimate protection for your radio don't forget that in-rush current limiters can be used in series. If one fails, the other will continue to limit the initial startup current spike. Antherm (another NTC thermistor manufacturer) recommends this for the ultimate in protection. For those of you who do not have the Mouser P/N for the Keystone CL80 it is 527-CL80.

I actually use two CL90 (527-CL90) in series. The CL-90 is rated at 2 amps at 120 ohms. The initial resistance is almost three times that of the CL-80. It is cutting the current a little close but I have measured the current and it is just under 2 amps and I have disabled the oven switch.. There is sufficient built-in margin (30%) in these devices to warrant using the CL-90 in lieu of the CL-80 in my humble opinion. When you consider that the stock solution was a bare piece of wire, there is nothing to risk.

From: km1h@juno.com Date: Wed, 12 Apr 2000 00:35:05 -0400
Subject: Re: [R-390] C553 adventure

For those without Variacs or not wanting such a beast in their living room check out www.harbach.com Allen offers a step start PC board that is just the ticket to inrush current. Although designed for Heath linear amplifiers it is 100% adaptable to any 120 or 240VAC application. Its real small, real cheap, and will fit inside just about any BA. In a 390xx I wonder if inrush current is even a problem? Depends a lot on the xfmr and the wiring. Be interesting to actually measure it.

From: "Charles A. Taylor" <calltaylor@prodigy.net> Date: Thu, 13 Apr 2000 12:20:40 -0400
Subject: Re: [R-390] RE: In-Rush Current and R-390A's

wrote: >Just let me know who your electric company is so I can buy some stock. Carolina Power & Light (CPL). Superb quality power at the outlet ...115 VAC +/- 0.5 V at the transformer, but the stock dividends aren't that hot.

>IMO the 390A is no more prone to on -off failure than any other quality radio such as a 75A4, S-Line, etc.

Tube radios do suffer the effects of current inrush into the filaments/heaters. We've probably all heard cases of radios that were left on for years and played without failure. Buddy at the workplace enlisted in the Navy from Memphis and got out after 16 years. His g-mother had an old RCA or some- such that she left on. It was on when he left for boot camp, and was still on when he came back to live in Memphis. The tubes were still lift, it still played. The tubes' transconductances weren't too hot, though.

Point is, ballasts are expensive and some tubes, too. Some aren't manufactured anymore. Balanced against $0.06/7 per kilowatt-hour, it's not a hard decision to make.

I usually operate my R-390A on my days off (Saturday and Sunday only one weekend a month). Variac it up on the evening of my last work- day, and leave it energized until the morning of my workday.

The variac-in' and leavin' DOES in fact make all the difference in tube life. I've fiddle/fooled/fixed/played with R-390As since Navy tech school years, and saw a bunch of them continuously energized aboard ship and shore stations. They stay live a long time if not turned on and off frequently. Same here...mine hasn't need a new tube or filter electrolytic in many, many days.

From: Roy Morgan <roy.morgan@nist.gov> Date: Thu, 13 Apr 2000 13:12:25 -0400
Subject: Re: [R-390] RE: In-Rush Current and R-390A's

<Economic Analysis Mode ON>
Current AES price for a 3TF7is $36.45.
Assume one ballast tube failure due to on/off cycling.
Assume no other tube or component wear.
Estimate R-390A power consumption at 150 watts (ovens off).
$36.45 worth of power at $0.07 per kilowatt-hour is 520 kilowatt hours.
520 kilowatt hours at 150 watts is 3471 hours.
3471 hours is 144 days or about 6 months.

Conclusion: If ballast tube failures occur less often than every 6 months due to on/off cycling, you should spend your money on tubes not on electricity.
</Economic Analysis Mode OFF>

From: "Ben Hall" <kd5byb@wt.net> Date: Wed, 21 Jun 2000 09:35:39 +0600
Subject: Re: [R-390] Filament Voltages

Well, I haven't tried inrush limiters in place of ballast tubes, but I have installed them in a lot of my hollow-state gear with good success.

I use one by Keystone Carbon, bought thru Mouser Electronics. (I believe it is a CL90) You want one with adequate current rating for your application, with the highest initial resistance you can get. When placed in my R-725, when I hit the power switch, the limiter dropped about 50 volts AC, then ramped up to dropping about 5-6 volts AC when fully warmed in about 15 seconds.

Definitely takes out the turn on "twangggg" that some transformers make. Cannot comment on increased tube life as I've done no scientific testing on that, but it cannot hurt...

From: "Gene G. Beckwith" <jtone@sssnet.com> Date: Wed, 21 Jun 2000 18:08:20 -0700
Subject: Re: [R-390] Filament Voltages

In addition to the current limiter technique, many of us have/are using Variacs for soft starts...Am sure this has all been said before, but just in case you missed it...its another approach...

Btw, searching the fest for Variacs large enough to handle my DX100's and Viking Valients...the transformer 'twange' in my Vallient II is so loud, I've shut it down until I find a Variac with enough current capacity to handle it....if none show up this season, planning to build an outboard step start, that drops out a series resistor in the AC line...have done that with great success in a recently re-herbed Heath "Warrior" linear (that's the own with 4 811's) I have it designed for about three seconds of warm up time for the filaments before full line voltage is applied... (uses a 50 watt wire wound in series with AC line that is shorted out with dc relay).

If any one needs details, most recent ARRL handbooks show several examples ...

From: "Dr. Gerald N. Johnson" <geraldj@ames.net> Date: Wed, 21 Jun 2000 19:36:53 -0500
Subject: Re: [R-390] Filament Voltages

Actually, I think the **series resistor, whether step start or thermistor, does more to limit turn on surges** than cranking up a variac... And the thermistor type as used in large tubed TV sets used to be not hard to find.

From: "Wayne Rothermich" <rother@impulse.net> Date: Wed, 21 Jun 2000 21:56:10 -0700
Subject: Re: [R-390] Filament Voltages

May I offer a somewhat different perspective on tube life and inrush current? In the waning days of tube-type radios and TVs, I worked summers for a local repair shop. Over a span of several years I must have replaced thousands of defective tubes. The odd thing is that very few of these failed tubes had open filaments. Almost all of them failed due to low cathode emission, and most of the rest had what the tube tester called a "short". My experience with Tek 500 series vacuum tube scopes has been similar.

The few open filaments that I saw were mostly in radios that had all the filaments connected in series. Because the resistance of the filaments increases with temperature, a tube that warms up faster than the rest (in a series string) can have more than its rated voltage across its filament until the rest warm up. It could be that this is what caused these particular filament burnouts. I believe some tube types specifically designed for series filament operation had controlled warm-up times in order to minimize this problem.

If the major cause of tube failure is indeed deterioration of the oxide cathode, and not filament burnout, I wonder if the filament startup surge is really affecting tube life to any significant degree. (Admittedly, the R-390A does have a series string of two tubes, but these are also in series with the 3TF7, which acts as a current regulator. This regulation should minimize the effect of any warm-up time differences between the two tubes. All the rest of the filaments in the R- 390A are wired in parallel groups.)

The thoriated tungsten filaments in transmitting tubes are different, and the surge may well shorten the life of these tubes. Pilot lamps certainly play by different rules. These observations only apply to oxide-coated, indirectly heated cathodes in small receiving tubes. To borrow a great line from one of our sages, for other kinds of filaments, "your mileage may vary".

Inrush current may be hard on filter capacitors and rectifier diodes, but my experience suggests that it's not a leading cause of the depletion of the world supply of NOS 6BA6s. Has anyone seen a significant proportion of tube failures in parallel filament radio receivers due to filament burnout?

From: Gene Beckwith <jtone@sssnet.com> Date: Wed, 10 Jan 2001 23:01:49 -0500
Subject: Re: [R-390] Inrush Current Limiters

Even though I'm rapidly getting my whole station up to speed with Variac slow start lines, I'm adding Inrush current limiters to all R-390X rebuilds as part of the standard referb menu.

These little gems run pretty hot, by design, so just a reminder to those who haven't used them, mount them so they stand away from other wiring down there under the chassis...with idea to allow as much air circulation as possible...it's an easy and inexpensive modification/up-grade.

Also adding high voltage in rush protection to some of my older linears. Latest in progress slow start project is to do my Viking 500 with an 18 amp Variac...plan to look at measuring 4-400 filament line and bring out for metering...

Note the T-368 is set up this way and allows for filament voltage adjust with built in panel meter...its just good operating practice for our increasingly scarce replacement components.

From: G4GJL@aol.com Date: Thu, 8 Feb 2001 18:14:37 EST
Subject: [R-390] Inrush limitation. Posted on behalf of Wolfgang DD8BD

Hi all, here my little contribution about my recent efforts :
Inserting a current inrush limiter into the R-390A/URR using the NTC S237 - a long story for a short procedure from a newbie for newbies

Remark
- ------
The used Siemens NTC S237 was found in the CONRAD ELECTRONIC catalogue as the most suiting type. There may be other and more suiting types available at other dealers for electronic parts. Also read Jan Skirrows et aliter articles on this subject at first. If you can help with some improvements please post a notice!
Parts:
- ------
- - NTC resistor e.g. Siemens type S237 [22 ohm at 25□‡C / 4 amperes (crystal oven switched off)]
- - Teflon insulation from surplus wires
- - soldering tin
Tools:
- ------
- - suiting pliers for cutting and bending wires
- - soldering iron (I use a 25 watt soldering iron)
- - suiting screwdrivers for opening the lower deck
- - a pair of metallic tweezers
- - 1.5mm driller or nail with similar round shaft
- - desoldering device(s)
- - a hot cup of coffee
Instructions
- ------------

Where to insert the inrush current limiter:

The NTC resistor is inserted between the centred soldering ear of the fuse holder [F101] and the feed through of the line filter [FL101] . Normally in the unmodified condition these are connected by a wire.

Procedure
- ---------
Get the power supply plug out of the mains connection.
At first you carry your R-390A on your work bench, turn it upside down and open the bottom cover to get access to the fuse holder and the line filter. Put the 3 amp fuse out of the fuse holder.

You cut the wire off at the fuse holder and de-solder to get the ear of the back and centred fuse holder connection open (this is the moveable contact what gets pressed out when inserting the fuse).

You cut the wire also at the feed through of the line filter, but let remain 1 to 1 1/2 cm of the wire at the soldering connection of the feed through. Now take a sharp little knife and a pair of tweezers. Hold the wire at the feed through with the pincette to save the sensible feed through at FL101 from any burden or mechanical drag by working on the wire. Cut off the insulation along the remaining piece of wire.

Try to twist this wire carefully a little bit if there is no solid inner conductor.

Now drink a sup of coffee.
Take some surplus Teflon insulated wire and pull out the inner wire. You need two pieces of this insulation suiting to the length of the NTC wires . Now slide the two wires of the NTC resistor each by one into the Teflon insulation so that there remains enough wire free from insulation for the soldering connections.

You take the 1,5mm drill and wrap about two to three turns of one of the wires of the NTC around it in order to get some little spiral or coil. One wire (that without spiral) of the NTC is fed into the soldering ear of the fuse holder and wrapped as far you can. Now it gets soldered.

Take the spiralled wire of the NTC and insert carefully the now not insulated short wire from the feed through of the line filter so that few millimeters look out of the spiral at the remaining wire from the NTC . Bend or deflect this bit of wire to nearly 180 degrees so that the spiral won't glide off from the wire. You also may squeeze the connection. All this is done for the safety of these connections.

Fetch a suiting metallic pair of tweezers and put it between the housing of the line filter and the soldering eye (with the short remaining piece of wire) of the feed through to avoid heat what could stress the feed through and the soldering points.

Now solder this last connection. Bend the NTC wires to attain greatest distance of the NTC from all other parts around inside the receiver. Put the bottom cover back to the receiver and screw it tight. Remove the tools, straighten up your shack and take the cup back in the kitchen.

If you did not forget the pliers or the pair of tweezers inside the R-390A you should be ready and can switch on the gear reading the digital frequency readout reversed (unless you turn the receiver back in normal position, <grin>).

If you can't receive any signal you forgot to reinsert the 3amp fuse into the fuse holder, hihi!

My thanks to Jan Skirrow and other R-390 enthusiasts , whose interesting articles motivated me to use a "current inrush limiter" for the R-390A and other gear.

From: David Wise <David_Wise@phoenix.com> Date: Fri, 14 Sep 2001 15:38:35 -0700
Subject: RE: [R-390] R-390A Restoration Bulletin 6: reassembly

Bill Hawkins asked me where I put the ICL.(inrush current limiters) I put mine in the most obvious place, namely, I soldered it in place of the short jumper from FL101 to F101. I kept the leads at their full length, which suspends the limiter in mid-air at the rear of the center underside compartment behind the PTO. There are no other heat sources nearby, so all it does is warm up the PTO slightly, surely not a problem with the ovens off. (Of course I have them off!) The long leads keep the solder joints cool at the expense of shock and vibration resistance. Since I'll be turning my set on and off a lot, I consider it de riguer, not just a good tradeoff.

I'll measure the average drop across the ICL. I hope it's about 5V, because my line voltage is a very consistent 120.0 . If the ICL is running too cool (a possibility with a CL080 and no ovens), maybe I could wrap it in a little ball of fiberglass insulation. The trick would be finding something to bind it up in.

From: Scott Bauer <odyslim@comcast.net> Date: Sat, 21 Dec 2002 19:40:46 -0500
Subject: [R-390] in rush current limiters

I have seen the mod for a soft start for the 390-A and thought about doing it. While browsing on ebay, I noticed an external in rush current limiter that can be used. I wonder if anybody has used one of these before. It can be seen on Ebay, item # 1945055954. I am thinking of buying this item and wonder if anybody might have some input or experience with such an item.

From: "Scott Seickel" <polaraligned@earthlink.net> Date: Sat, 21 Dec 2002 20:33:07 -0500
Subject: Re: [R-390] in rush current limiters

Well Scott, that e-bay item would be the expensive way to do it. I purchased an excellent rebuild kit from Walter Wilson and it included the Keystone CL-80 inrush current limiter. You just solder it in-line with your AC power supply. If you do not need a complete rebuild, you can purchase the Keystone Current limiter from Mouser electronics for $2.17.

Date: Sat, 21 Dec 2002 22:11:32 -0600 (CST) From: "Jim Shorney" <jshorney@inebraska.com>
Subject: Re: [R-390] in rush current limiters

Man, this is brilliant. This guy on eBay takes a couple of current limiters like you describe and a PTC for fusing, mounts it all in an electrical box with an outlet, writes some flowery text about it, and sells it for 30 bucks. I gotta order some of those from Mouser and get in on this deal...

From: Tom Norris <cthulhu@fhtagn.org> Date: Sat, 21 Dec 2002 22:31:14 -0600
Subject: Re: [R-390] in rush current limiters

The ebay solution is *convenient* though. It is also portable. Could use it with several receivers. On the subject of convenience versus price, does anyone have a good schematic for a solid state ballast similar to what Chuck Rippel sells for $50. Yes, converse to what I say above about being convenient, I have several 390A's that I would like to fit with some sort of similar item. I may end up just buying them from Chuck, his are stable and RF-quiet.

From: "Scott, Barry (Clyde B)" <cbscott@ingr.com> Date: Mon, 30 Dec 2002 09:56:08 -0600
Subject: RE: [R-390] in rush current limiters

I did this myself a while back. I used a mini-box, single AC receptacle, fuse, CL- 80, and a "computer-style" socket. I have it powering an old RCA upright "curtain-burner" I rebuilt. I didn't want to modify the chassis so I just ran a new 3-wire cord out to this box. Works fine.

Date: Thu, 2 Jan 2003 00:54:12 EST From: ToddRoberts2001@aol.com
Subject: Re: [R-390] in rush current limiters

>Let's see - the item on eBay - take a dual outlet (69cents) a plastic outlet box (69cents) a small 3-wire line cord (99cents) two current inrush limiters $2, a small circuit breaker (99cents) - put the thing together for $5-$6 bucks and sell it for $30 bucks on eBay. Sounds like a nice profit margin! One thing I am not fully sure about current inrush limiters is - don't they run "hot" in normal use? And do they fully return to zero resistance when they are at operating temperature? - Todd Roberts WD4NGG.

From: "Walter Wilson" <wewilson@knology.net> Date: Thu, 2 Jan 2003 06:33:58 -0500
Subject: Re: [R-390] in rush current limiters

Todd, Inrush current limiters do indeed get and stay hot when in use. Also, they must be properly sized for the equipment. That's one of the problems with the ePay version: depending on the current limiter selected, it is optimally sized for a certain load. These things also have max current ratings.

For instance, the CL-80 (commonly used in the R-390A), is rated at 47 ohms when cold, max current of 3 amps, and resistance drops to about 0.5 ohms at max current. If you were really pulling the max 3 amps through the limiter, you would expect a voltage drop of about 6 volts (current/resistance). Used in the R-390A with the ovens OFF, the drop across a CL-80 current limiter is between 2.5 and 3 volts. So if incoming line power is 120 VAC, the radio will only see 117 VAC after the 3 volt drop across the limiter. This is not a bad thing, since the receiver was designed to run on 115 VAC. But it is apparent from these observations that a "one size fits all" solution is not really valid here. If you have one of these ePay solutions, just don't go plugging your power bar into it to feed the whole station.

From: Heinz und Hannelore Breuer <hbreuer@debitel.net> Date: Thu, 02 Jan 2003 13:15:47 +0100
Subject: Re: [R-390] in rush current limiters

Not exactly! Last time I checked Ohm's law was still $U = I * R$ not $U = I/R$

From dallas at bayou.com Fri Sep 24 09:36:01 2004
Subject: [R-390] 3TF7 Substitutes

The standard 3TF7 substitutes, (1) using a 42 or 43 ohm 10 watt resistor in place of the 3TF7, and (2) using an appropriate tube, like a 12BY7A, with a 12.6 volt filament in place of the 3TF7 are both acceptable substitutes. Using a 10 MHz rubidium standard I determined, somewhat to my surprise, that the power resistor is generally a more stable substitute than a 12BY7A. Recently while examining the long term frequency stability of one of my R-390A's with the BFO turned ON (for SSB, ECSS, or CW) using a rubidium standard, I found that the 3TF7 does not do a very good job of stabilizing the BFO and PTO frequencies when the AC line is varied. A change of only 2 or 3 VAC in the line voltage (I used a VARIAC to vary the AC input voltage to the R-390A) causes a substantial (4 or 5 or 6 Hz or more) departure from zero beat. Next, I removed the 3TF7, inserted a 9 pin tube test extender into the 3TF7 socket, and powered the BFO and PTO filaments with an external regulated 12 VDC supply (12 VDC was found to give almost exactly 300 mA filament current). With this arrangement, no change in zero beat was observed as the AC line voltage was varied from 120 VAC nominal down to 100 VAC and back up to 120 VAC. WOW. Whoever designed the original BFO and PTO filament stabilization circuit was on the right track. They just used the wrong method to stabilize it. Current regulation is the wrong approach; voltage regulation is the correct approach.

Rather than rewire the 3TF7 socket, I opted to make the mod "plug-in" in so far as it was possible. I cut the metal flange off a miniature 9 pin ceramic tube socket, pushed 9 pieces of #18 tinned solid copper wire in each receptacle, soldered them, cut off the ends to the appropriate length for a 9 pin tube, deburred and polished the tips, drilled out the cylindrical center piece of metal and removed it.

I ground off most of the head of a 6-32 brass screw of the appropriate length and attached an inch long (or somewhat longer) insulated spacer, and mounted an insulated standoff on the threaded end. This provided me with a home made tube socket extender on which I could build most of a 12 VDC regulator.

There is a nut on the front of the IF deck where I added a ground lug. I ran a diode from the #2 pin lug of the adapter to the standoff, and a 1000 mF 50 volt electrolytic from the standoff to the ground lug. The ground tab of a 3 pin 12 volt 1 amp regulator was attached to the RF deck corner nearest the IF deck using one of the green screws that hold the oscillator deck to the RF deck plate; the regulator pins stick up above the top edge of the RF deck plate.

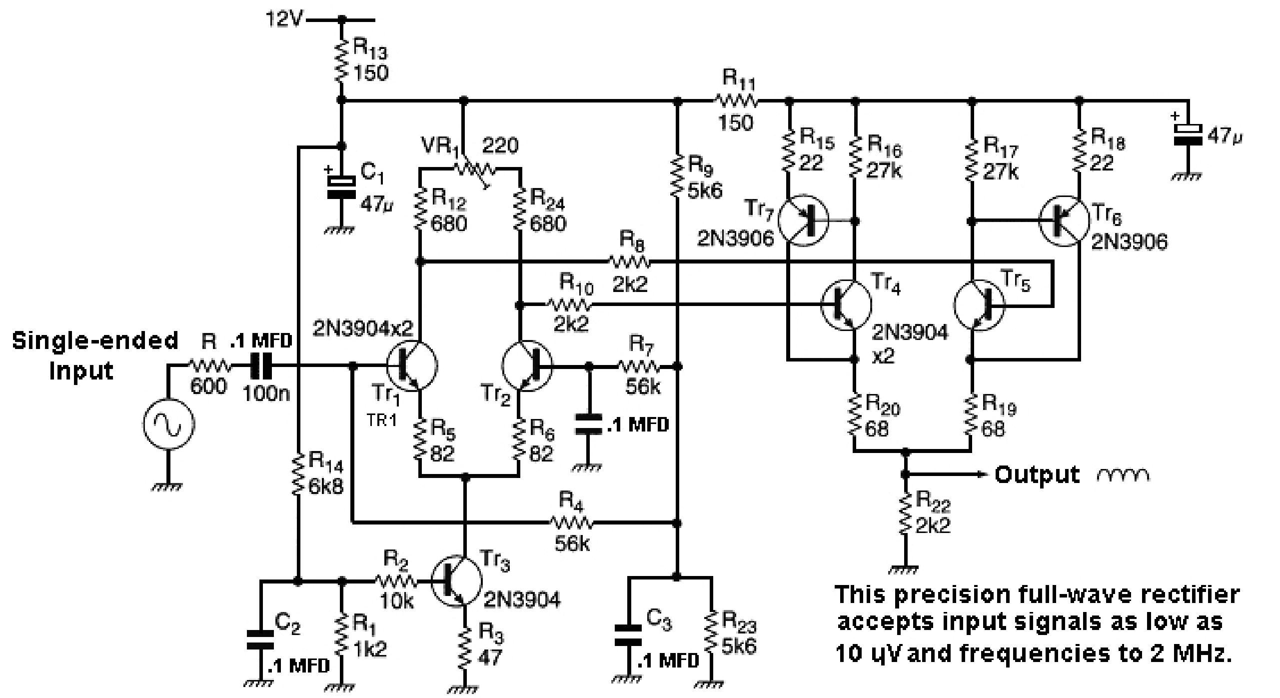
The input and output pins of the regulator were bypassed to the ground pin with 0.1 mF 50 volt capacitors, and the regulator ground pin was wired to the added ground lug on the IF deck. An insulated wire from the standoff to the regulator input pin and an insulated wire from the regulator output pin to lug 7 of the plug-in adapter completed the modification. Well almost... three (3) complete wraps around the lugs of the home made adapter with Scotch Glass Cloth Electrical Tape and heat shrink tubing on the standoff protected the plug-in adapter from shorts.

This mod is not 100% plug-in because to remove it you have to (1) remove the nut on the front of the IF deck to remove the ground lug, and (2) remove the green screw on the RF deck to remove the 3 pin regulator.

I have now had the mod running continuously for about 48 hours. No problems were expected and no problems have been experienced. Not only does this mod give you improved frequency stability for ECSS, SSB, and CW, it should also provide a permanent solution for the 3TF7 replacement problem. There is still some very slow frequency drift, as much as 1 Hz per hour, sometimes more. I currently do not know the cause of this drift.

> I have already spotted a few typos in my posting, which I believe will be > obvious to those who read it. > > I meant to include that I used a 1N4003 diode, 200 PIV and 1 amp. I presume > a 100 PIV 1 amp diode would be fine. I just happened to have 1N4003's on > hand. > > Also, be sure to mount the 3 pin regulator to the RF deck plate (unless you > want to drill a hole in the IF deck and mount it there). I don't know how > large a heat sink the regulator actually needs, but the RF deck front plate > is surely much more than enough.
>
> Best regards, > Dallas

Precision AM Rectifier Circuit



Here is a circuit that can rectify alternating signals. Such a circuit has quite a few different names and can be used in a variety of applications.

In a wireless, the circuit would be called an AM full envelope detector. This circuit would be equally at home positioned in the front end of an RMS detector and is then referred to as an absolute circuit. I will refer to the circuit as a full wave rectifier, or FWR, from now on,

Using the components shown in the diagram, this FWR will rectify up to about 2 MHz, with input signal, as low as 10μV. I originally used it for full envelope detection in a 455 KHz IF chain. It maybe possible to squeeze more speed out the circuit by reducing R_8/R_{10}.

The input can he driven differentially or from a single-ended source. Assuming a single-ended input then the input signal is split into two equal parts 180° out of phase with one another,

Splitting is performed by the long-tailed pair Tr_1 and Tr_2. The two resulting signals drive two emitter followers.Tr_4 and Tr_5. connected in conjunction with Tr_7 and Tr_6, which serve as local current feedback pairs. This action ensures super linearity right down to the micro-volt region .i,e, constant bandwidth with wide dynamic range,

Transistors Tr_4 and Tr_5 conduct alternately and the summation of the two emitter currents results in full wave rectification across R_{22}.

The DC static offset voltage will be in the region of 6V or about $V_{cc}/2$, It can easily be removed by an op-amp, leaving just the rectified signal for further processing. The gain produced by long-tailed pair Tr_1 and Tr_2 is about seven. This, suited my needs at the time. Obviously though. this gain can be changed to suit other applications.

Adjust V1 to balance the rectification action with the aid of an oscilloscope. The whole circuit runs from a single 12Vsupply, Most importantly, the circuit is incredibly cheap.

Darren Heywood Buckley Flintshire

Parts List

4	2N3904	2	22R	2	150R	3	2K2	2	27K
2	2N3906	1	47R	1	600R	1	5K6	2	5K6
2	47 Mfd	2	68R	2	680R	1	6K8	1	220R 10 Turn
4	.1 Mfd	2	82R	1	1K2	1	10K		

Adapted From ELECTRONICS WORLD May 2004

Intrigued by Darren Heywood's precision full wave rectifier circuit described in the January 2003 issue of Circuit Ideas *(EW page 36),* Alan Bate has come up with some different ideas

As this kind of circuit is usually implemented by linearizing a diode rectifier forward characteristic within a negative feedback loop, Figure 1 shows the classic full wave circuit using this approach. The problem with such a circuit is the performance of amplifier 1 which not only needs to have wide small signal bandwidth and but also high slew rate to rapidly switch through the open loop transition when the diodes are shut off. See **Figure 2** of a SPICE simulation at 10kHz, using general purpose small signal Schottky diodes and OP37's. Note the fast transition required of the op amp in the diode drive waveform. However, new op amps like the Analogue Devices *AD3038/39,* with a small signal bandwidth of 350MHz, and slew rate of 425V *rms* would now turn this approach into a 'humdinger of a circuit' !

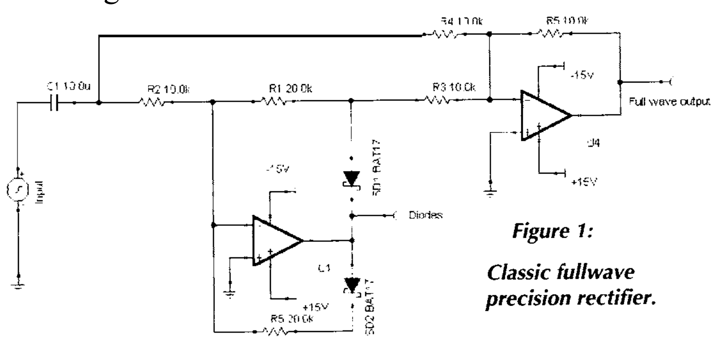


Figure 1:

Classic fullwave precision rectifier.

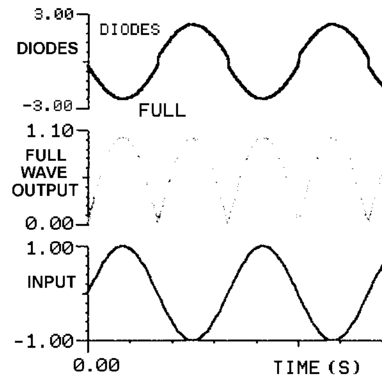


FIG. 2
CLASSIC FULL WAVE
RECTIFIER OUTPUT

Mr. Heywood's approach to linearize the rectifier function is novel as it potentially offers wide bandwidth. However, his circuit suffers from a few limitations, which can easily be put right, greatly improving linearity at low signal level and bandwidth. His sensitivity and bandwidth claims of 20uV and 2Mhz are also worth investigating.

Rectifier out-put stage

The differential output rectifier stage, uses local feedback pairs in an attempt match the "rectifier gain" for each quadrant. There are three shortcomings with this circuit, which showed up in SPICE simulations.

Consider one compound pair, Tr4, Tr7. Degenerating the gain in both transistors is not necessary as there is 100% overall feedback applied to dilute the non-linearities in both transistors. The resistor R15 only suppresses precious loop gain and decreases the current transfer efficiency around the loop by raising the ac input impedance of the second transistor Tr7.

Even if the two compound pairs had infinite loop gain, we would still be left with a large dead-band around the null point. This is due to the comparatively large resistance value chosen for the emitter degeneration resistors RI9 and R20. At the quiescent bias condition, the resistors equally share the output current.

However, at either signal extremity all the output current will pass through one of the emitter resistors while the other compound pair is virtually turned off. This variation in current shifts the turn on threshold at each input, due almost entirely to the volt drop across R19 and R20.

Consider the **Figure 3** Model of the output stage DC requirements. Assuming a typical beta of 100, Vbe for silicon of 0.6 Volt, and using the rule of thumb for transistor intrinsic emitter resistance re = (26 *lie)* (based on Schottky's diode law at room temperature where the emitter current Ie is in mA)

Examining the left hand arm of the output stage Tr4 and Tr7 and letting output current = Iout (mA).

Let transistor current gain = 100, compound pair current gain = Ai = approximately Beta Tr4 x Beta Tr7 x say 50% current transfer at Tr7 base, then Ai = 100 x 100 x 0.5 = 5,000. Max input voltage (seen from the preceding stage) with Tr4 on and Tr5 fully off:
Voltage input for full output = Iout (R101 Ai + R20 + R22) + re/Ie (Tr4)+Vbe4. **(Eq 1)**

At the mid point the two arms of the output stage equally share the output current, so assuming the compound pair gain remains constant with Iout and their Vbes (Tr4, Tr5) are equal;

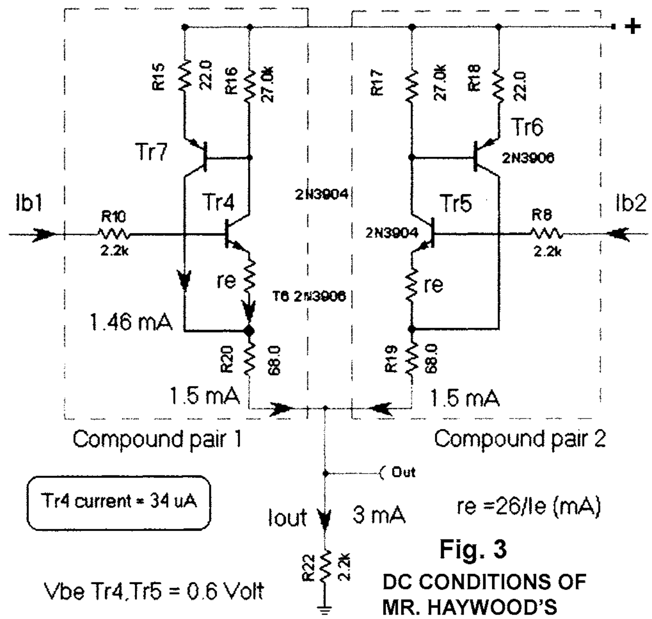


Fig. 3
DC CONDITIONS OF MR. HAYWOOD'S OUTPUT STAGE.

Voltage input at the null point = (Iout/2) (R10/ Ai + R20) + R22. Iout + re/Ie (Tr4) + Vbe4. (Eq 2)

Now, assume the current in Tr4 is «Tr7, i.e. the bulk of the output current passes through Tr7, not through the re of Tr4. Running a DC SPICE analysis showed Ie of Tr7 = 1.46 mA and Ie of TR4 to be only 34.3µA. Hence, we can ignore the volt drop across re. Then: Change in Vin = (Equation 1) -(Equation 2).

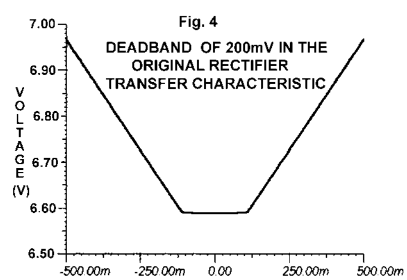
Delta Vin = I(R10/Ai + R20). Iout/2 As the compound pair current gain Ai, is very large, we can ignore the volt drop across R10, simplifying the change in Vin to: Delta Vin = R20.(Iout/2)

For an output current of 3mA which is what Mr. Heywood's circuit runs at:

Rectifier stage dead-band, Delta Vin = 1.5(mA) x 68 Ohms = $\pm$102mV

See **Figure** 4 of the SPICE simulation of the output stage differential DC transfer function about the DC null or steady state showing the dead-band discussed above.

This causes the output circuit to be 'blind' to the middle of the incoming signal, significantly reducing the dynamic range and progressively increasing low level distortion.



This effect when referred to the input is diluted only a little by the modest voltage gain (7: 1) of the preceding stage.

Looking at the compound pairs, the collector loads R 16 and R 17 are unnecessarily large. Here there are conflicting requirements as R 16, R 17 should be:

1. Lowest possible resistance, in order to rapidly remove base charge from Tr7, Tr6 for good high frequency performance.

2. Highest possible value, to efficiently transfer the current signal from Tr4 into Tr7 and achieve the highest current loop gain in the compound pair. R15 and R18 are not a good idea as they not only suppress gain in the wrong place but also raise the input impedance of Tr6, Tr7 and further reduce the efficient transfer of signal current from Tr4, Tr5.

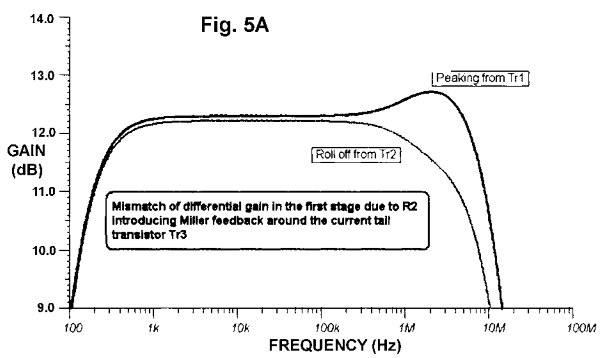
The third improvement would be to ac couple the long tail pair preamplifier to the output stage.

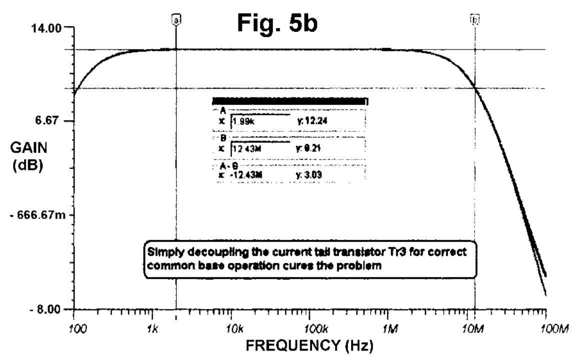
This blocks any unwanted DC imbalance in the first stage, which any way is ac coupled on its input so there is no point in preserving DC coupling with all the associated temperature drift problems.

The bias requirements of the output stage can then be optimized separately.

Differential pre-amp stage

Considering the long tail pair input stage the current tail transistor Tr3 should be operated as a constant current device for efficient current signal transfer at the emitters of Tr1, Tr2. This means operating Tr3 in grounded base to give the highest ac output impedance (around 1MΩ at low frequency for common base). This requires voltage DC bias of the base and de-coupling it to ac right at the base with a (low inductance) ceramic capacitor. R2 defeats this, allowing a 'soggy' operation somewhere between common emitter and common base, giving a less efficient constant current sink and worse allows Miller feedback around Tr3 which adds a very undesirable capacitive lag across the current tail. This causes the high frequency gain to peak through Tr1 as it provides ac bypass to Tr1 emitter and to roll off at Tr2's frequency response by shunting the signal path to Tr2 emitter! This in turn, gives an increasing imbalance in the differential ac gain with increasing frequency. See **Figure 5a** and **Figure 5b** SPICE simulations demonstrating this effect. A minimal change is to decouple Tr3's base. The high frequency bandwidth of the circuit is also limited by the unnecessary resistors R8, R10, which with 'circuit strays' and the input capacitance of the following stage form low pass RC filters.



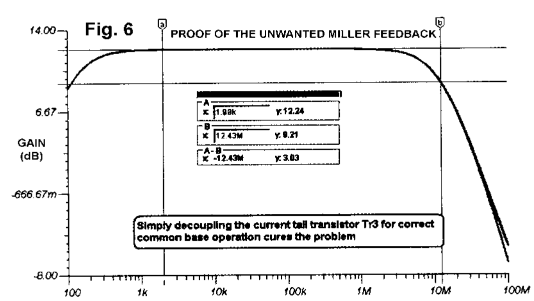


There is also a second Miller feedback effect from the base collector capacitance of Tr1. Miller feedback is minimal at Tr2 due to its de-coupled base and common base operation (with the long tail pair driven single ended). This second Miller effect can easily be demonstrated by driving the input from a low impedance generator rather than the 600 Ohm source used by Mr. Heywood. Compare **Figure 6** of the SPICE frequency response simulation of the first stage output, driven by a zero Ohm source with the 600 Ohm source performance of **Figure 5b**. Driving from a low generator source impedance 'absorbs' Tr1 's base collector capacitive current. This effectively minimizes the 'Miller effect' to the internal workings of the transistor. Adding an emitter follower to the input would be a practical way of achieving the same result.

Re-vamped circuit

With the above points in mind, I have re-vamped the circuit for interest sake, optimizing with the aid of SPICE circuit simulation.

See the **Figure 7** plot of rectifier gain versus Tr4, (and Tr5) collector loads R16 (R17) values.



Examining the plot shows a rapidly diminishing return on gain increase with increasing load value. 3.3k Ohms appears about the optimum value, giving near maximum gain while maintaining a much lower resistance value than used in the original circuit and therefore enabling a faster turn off response. See also **Figure 8** of the revamped circuit and **Figure 9**, comparison of linearity with the original circuit at 500kHz.

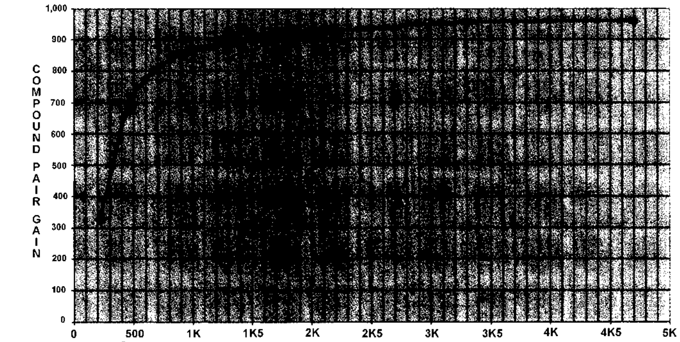


Fig.7 Gain Versus R16 Value.

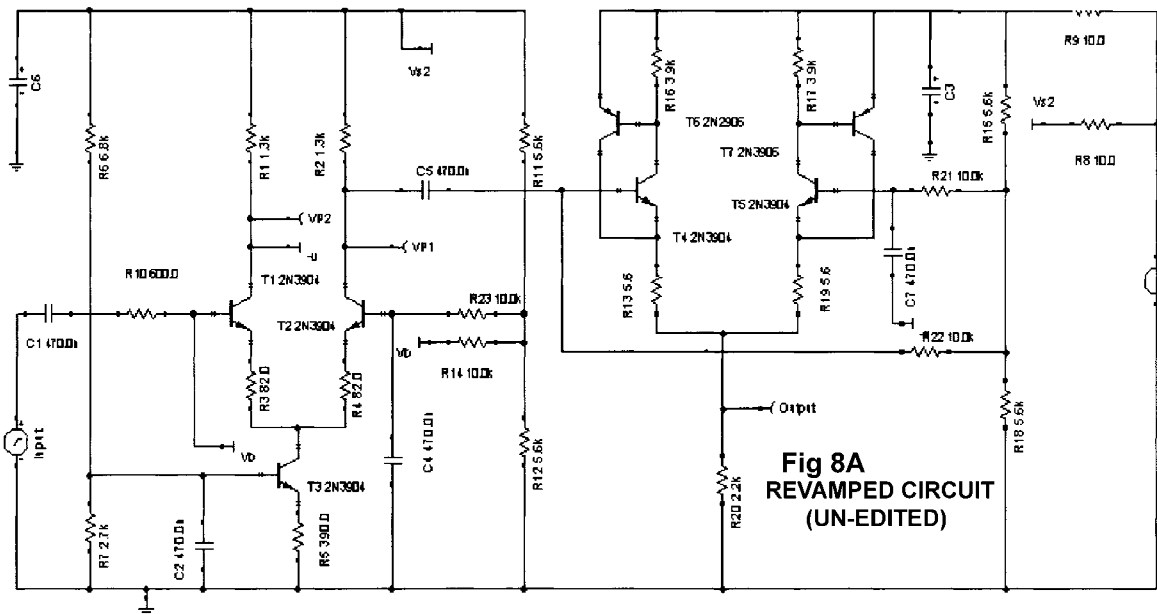


Fig 8A
REVAMPED CIRCUIT
(UN-EDITED)

Further development.

With bandwidth limited by the Miller effect in the first stage and wishing to find a simpler output stage, my thoughts turned to designing a rectifier output stage, which would use the Cascode principle. The 'Cascode' was an amplifier topology invented back in 'valve days' to overcome Miller feedback.

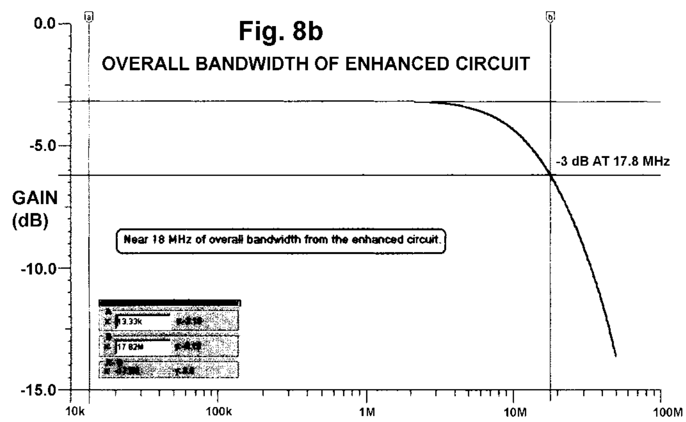


Fig. 8b
OVERALL BANDWIDTH OF ENHANCED CIRCUIT

Voltage generation of the signal at the amplifier's first stage output was avoided and hence there was no internal feedback through the anode to grid capacitance (or base collector in transistor talk). The Cascode was used universally as the front end RF amplifier in valve VHF/UHF TV tuners to achieve high gain at RF. See **Figure 11**.

The first stage of triode amplification was fed directly into the low ac impedance of the cathode port of a second triode valve, operating in common grid (common base, in transistor talk). The first valve gave no voltage amplification but very high power gain due to the amplified signal current from the valves mutual or transconductance, gm.

The second stage had no Miller limitation due to the grounded control grid, preventing any anode to grid Miller capacitive feedback.

Fig. 9a ORIGINAL CIRCUIT LINEARITY AT 500 kHZ.

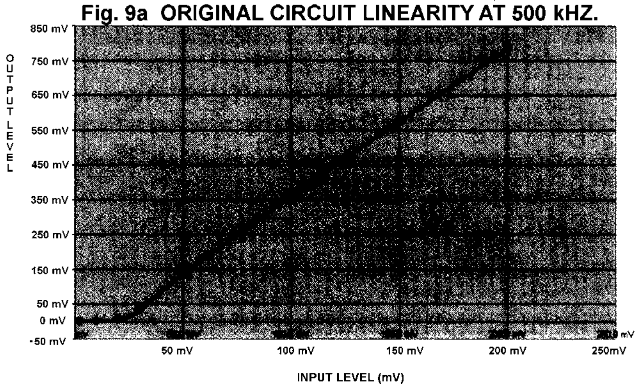
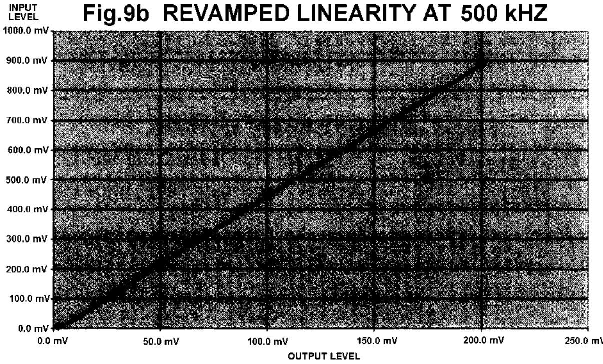


Fig.9b REVAMPED LINEARITY AT 500 kHZ



The common grid topology has high output impedance (analogous to transistor common base), allowing high voltage amplification to be achieved. The cascode principle adapts readily to transistor circuits with the advantage of folding the circuit, avoiding build up of DC levels by using complimentary PNP devices for the second stage.

They didn't have the advantage of a 'PNP valve', hence the valve circuit stacked up in its DC requirements. **Figure 12** shows my use of this technique with the output stage configured in differential common base.

Fig.10 REDUCED DEAD-BAND: ENHANCED CIRCUIT

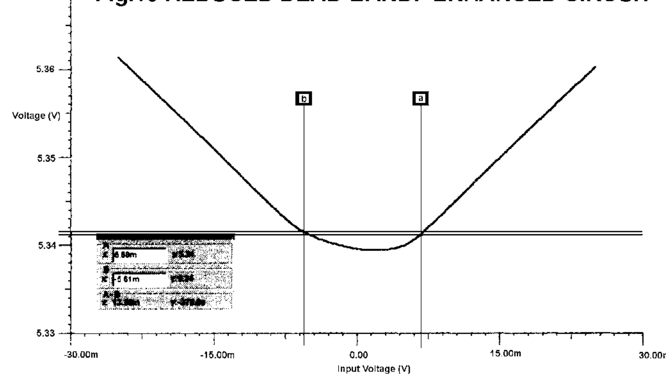
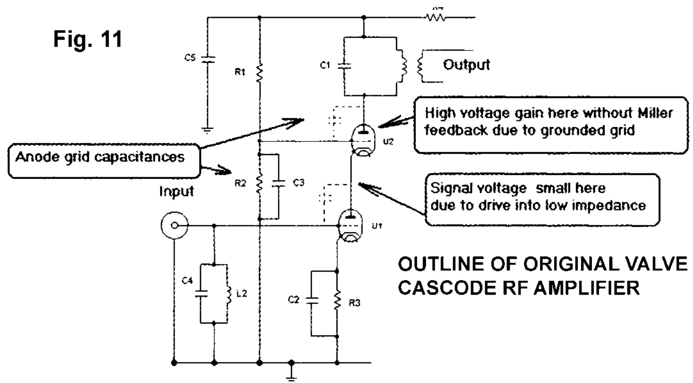


Figure 10, comparison of frequency response with the original circuit.

The low impedance of the PNP emitters clamp the first stage output signal voltage so that the voltage change becomes minimal but near linear signal current is transferred into the output circuit. AC coupling and the high value emitter bias resistors allow each PNP transistor to be accurately biased on the verge of conduction.

Fig. 11



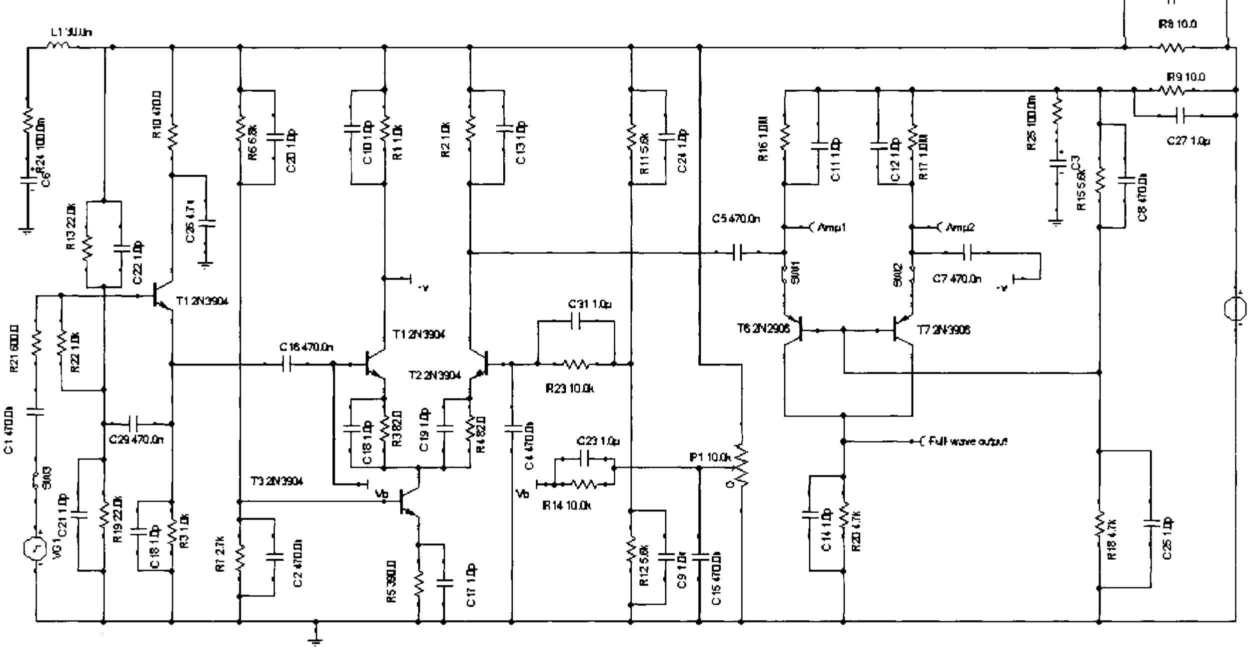
OUTLINE OF ORIGINAL VALVE CASCODE RF AMPLIFIER

As soon as each amplified signal from the first stage swings positive the relevant PNP transistor conducts and the signal current flows on into the output resistive load. Common base operation ensures the PNP transistors will have very closely matched gains. Because hfb = hfe/(hfe + 1) and with hfe variations of 100 minimum to 300 typical for the 2N3904 devices used, the common base gain or emitter efficiency will only vary around 0.9901 to 0.99668.

The common base operation also enables high voltage gain to be achieved due to the inherently high output impedance of around 1M Ohm at low frequency, allowing a high collector load value.

See **Figure 13** of the simulated linearity at 500kHz showing non- linearity at very low levels (where re will be large and current gain at its lowest) and the simulated drive and output waveforms at 500kHz (**Figure 14**). Note the clamping of the pre-amp output voltage on each positive half cycle where the associated PNP turns on.

Fig. 12 CIRCUIT OF FOLDED CASCODE APPROACH (SPICE DRAWING)



The active part of each signal offers little voltage variation and hence minimal Miller feedback, allowing near full bandwidth to be achieved on the positive half cycles. What happens on the negative cycles is a 'don't care' as the corresponding output device is shut off. This allows some over loading of the first stage to occur on large signals, with the input transistors bottoming on the unwanted negative half cycles.

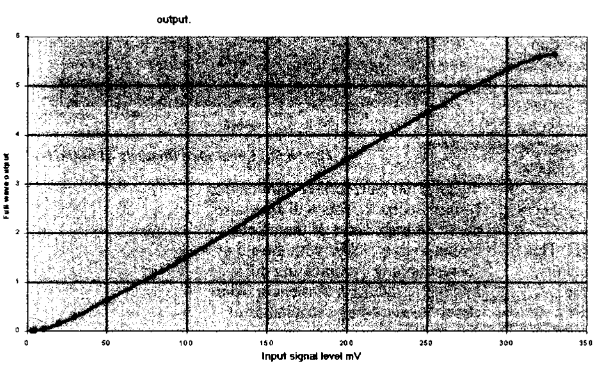
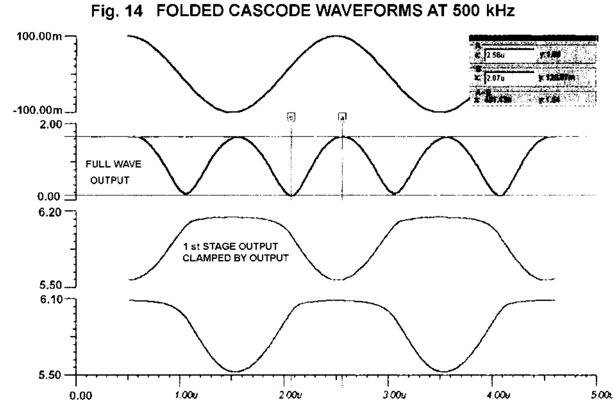


Figure 13: Linearity at 500kHz of the enhanced circuit using the PNP rectifier output.

Fig. 14 FOLDED CASCODE WAVEFORMS AT 500 kHz



This effectively increases the dynamic range at frequencies where the propagation delay of Tr1 and Tr2 can be ignored. Finally a spare NPN transistor (the original circuit uses seven transistors) is used as an input (where re emitter follower to provide a low AC impedance drive, of around 15 Ohms to the long tail pair, further minimising Miller feedback. This will also raise the input impedance of the input circuit to over 50k Ohm with the aid of 'bootstrapping', eliminating the loading of the input bias chain from about 10 kHz. Stray capacitance of 1 pF has been assumed in the simulations across each resistor. Circuit strays have not been included in the simulations, as this will depend on layout.

Word on bread boarding

My own bread board method today is to glue surface mount components using 'super glue' to industrial fibre glass FR4 board with copper on the opposite side to serve as the reference (ground) plane. Unpopulated etched boards can also be used with the green masking serving as an insulating barrier. This has perfectly adequate insulation for low voltage circuitry.

If space is not an issue use 0805 size passive chip components as they are easier to manage than the smaller 0603 size components. Use the smaller 0403 components only if you have bionic eyesight! Wire point to point to the chip ends using the fine Kynar insulated wire available from RS or Farnell components. Glue down the wires to the required layout. Keep wiring as short as possible, especially to inputs of active devices.

I have found bread boarding like this gives a reasonable approximation to a final PCB layout using a ground plane.

Note on Miller feedback

This is the effect that unless prevented, kills bandwidth in any voltage amplifier. Assume we have an amplifier device with a voltage gain Av of 100 and ideal infinite input resistance with zero output impedance. It has unwanted feedback capacitance between the input and output terminals of Cf. The capacitance will cause feedback signal current from input. Usually, this feedback is small say 2pf.

The capacitor current will be proportional to the voltage across the capacitor, which in our case is Vin -(-Av. Vin) or Vin(1 +Av) = 101 Vin.

If the capacitor volt drop has increased 101 times due to the amplifier then the feedback capacitive current will be101 times larger for the same input voltage. The feedback capacitance Cf, now looks 101 times bigger at the amplifier input! The problem occurs when we drive the amplifier input from a real world generator of finite source impedance Rs. With no voltage gain, the capacitance will be the input capacitance (Cin +Cf) which will form an RC lag with the generator source impedance, giving in turn a low pass RC filter. -3 dB bandwidth = 1/ (2π Rs (Cin +Cf)). If we now add gain Av, the bandwidth will shrink accordingly as-3 dB bandwidth = 1/ (2π Rs (Cin + Cf (l + Av)))

The bandwidth has now reduced in proportion to the voltage gain of the amplifier, in our case by one hundred times! This occurs inside any bipolar transistor due to the collector to base capacitance and the base spreading resistance rbb giving a minimum input source impedance even when voltage driven. This is why the common emitter gain of a transistor is often considerably less than its transition or unity gain frequency would imply.

Editors Notes:

1. I have re-drawn circuit Fig. 12, translating the circuit from SPICE to what normal human being's use and titled it Fig. 12A. The screen values on the figures are unreadable on the copy I have. If you can figure them out great.

2. Where the original schematic used lines ending in symbols to signify inter-connections they were removed and replaced with solid lines. After I did that the circuit explanation became more clear.

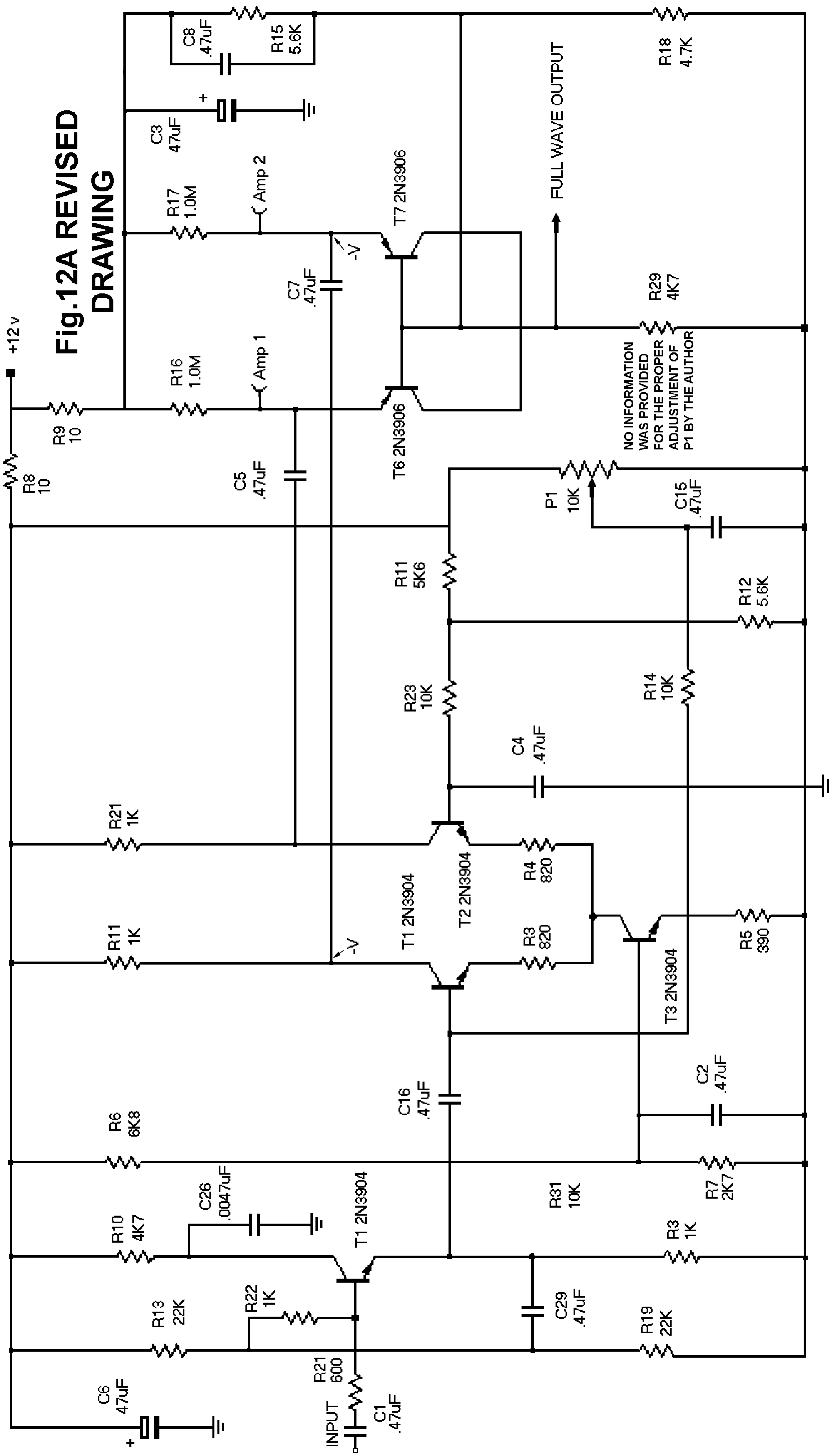
2. I removed the 1pF stray capacitor symbols, the power line inductance and the ESR values to the electrolytic capacitors as they are irrelevant to this circuit.

3. There was nothing in the text of drawings explaining what P1 did or how to adjust it. Reading the text of Darren Heywood's precision full wave rectifier circuit that's included with this addition may provide the answer.

4. I believe the author Alan Bate while having great technical skills has been out in the noonday sun far too long when he doesn't recommend using ordinary discrete components for construction.

5. Be aware that you might have to do a lot of circuit layout changes to get this circuit to operate as one would wish.

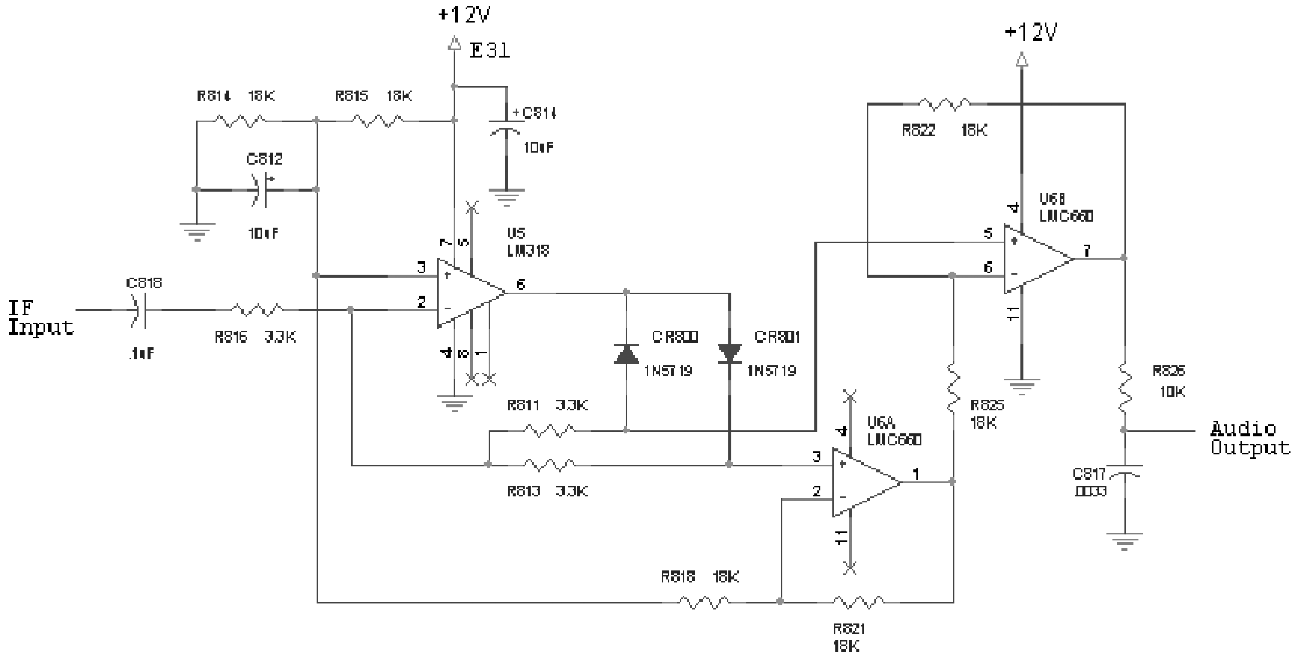
6. Last but not least, I made a great effort not to make errors in my drawing BUT CHECK IT YOURSELF!



An Improved, Precision, Full Wave AM Detector

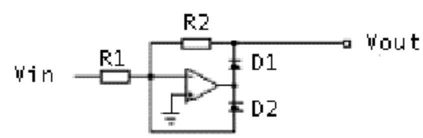
By Rob Schenck, K2CU K2CU@arrl.net

Detector Schematic



Theory of Operation

This circuit is a variation of the classic "Precision Half Wave Rectifier" as described in many op-amp application notes as shown:



Its operation is based upon basic principles of operational amplifier circuits. An ideal op-amp has extremely high input impedance, extremely low output impedance, and extremely high gain. Note the operative word "extremely". The non inverting input (+) of the op-amp is connected to ground. Feedback is provided by the two paths R2+D1 and D2. The op-amp will drive its output pin to a voltage such that the inverting input (-) is at ground voltage. There will be a slight variation (a few millivolts) from 0 volts as a result of what is known as input offset voltage, a property of real op-amps. The inverting input is often called the summing junction, as the sum of the currents in that node will be zero. Since virtually no current flows into the inverting input (-) of the op-amp itself, any current coming in through the input resistor R1 will be met by an equal and opposite current coming through either of the two paths of R2+D1 or D2.

The I.F. input signal will produce an input current of: I in = V in/R1

During the negative half of the input AC waveform, the op-amp output will drive positive such that the current through R2+D1 will be equal and opposite the input current, or:

$$I_{in} = V_{in}/R1 = -I_{D1} = -I_{R2}$$

The voltage developed across R2, and hence the output voltage as the left side of R2 is at ground potential, will be:

$$V = V_{out} = -I_{R2} * R2 = -V_{in}/R1 * R2$$

By setting R1 = R2, the output voltage will be; V out = V in

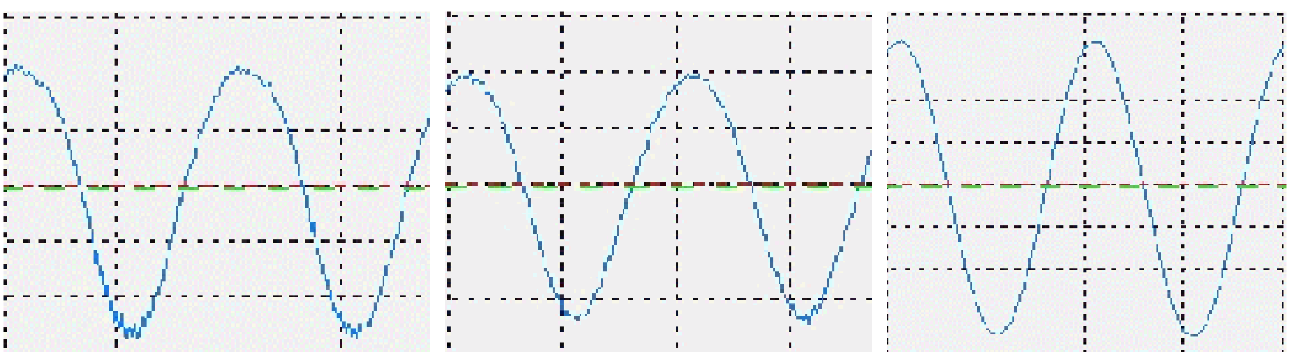
Similarly, the positive going input cycle produces a feedback current through D2. Any non linear voltages developed across the diodes is not relevant as the output is derived from the feedback resistor only.

A full wave detector was desired for a 50 KHz IF application so that the ripple frequency out of the detector would be 100 KHz, and more easily filtered. In the full wave detector version of the circuit, a second feedback resistor is added to the D2 diode path to produce a voltage of the other half of the input waveform. A virtual ground at +6 Volts was created with R814/R815/C812, so that the circuit could operate from a single supply. Again, any non linear voltages developed across the diodes are not relevant. The two rectified half waves are combined in the next two op-amp stages. The two outputs of the rectifier on R811 and R813, are connected directly to the non-inverting inputs of the op-amp stages so that there will be minimal loading to these outputs. U6A provides a gain of two to the signal voltage on R813. U6B provides a gain of two to the voltage on R811, and inverts and adds the output of U6A. The result is a full wave rectified signal at the output of U6B. Simple low pass filtering will remove the carrier energy which is at twice the ripple frequency.

The ability of this circuit to accurately track the input waveform is dependant on the speed of the op-amp and the switching speed of the diodes. For the circuit to track the input, the op-amp must be considerably faster than the input frequency. The LM318 with its 15 MHz bandwidth was selected for this reason and because it is readily available and stable. This circuit was built up and set up for comparison with the stock envelope detector in a Drake R4-C with a 50 KHz I.F. A HP 8640 signal generator was used to provide a 1Khz modulated carrier at 3.885 MHz which was input to the antenna connector. The input level was set to S9 (30 uV). The detectors were compared at modulation levels of 80% and 99+ %. A Link Inc. , PC based oscilloscope was used to record and display the resulting waveforms.

The distortion is visible at the crest of the wave, where the input signal is at the modulation trough in this inverting detector. The distortion at 80% is less visible than at 99+%, but still audible.

Applying the 99+% modulated signal and using the precision rectifier results in the waveform on the right below



R4-C Stock Detector at 99+%   R4-C Stock Detector at 80%   Precision Detector 99+% Modulation

This waveform displays virtually no distortion not only visually, but also audibly. A pure tone is heard coming from the speaker. Actual on air comparative listening tests of 75 meter AM signals at levels from just over the noise to well over S9 were quite conclusive. Low level signals could be more easily heard through the noise, and high level signals we noticeably clearer.

Comments: Purists may want to use LM318 op-amps for all three stages, but I found no significant improvement when doing same at 50 KHz. I wanted to pick parts that were readily available and in particular in DIP packages so that any builder would be able to build up the circuit on a perf board, preferably the type with a ground plane on one side. This circuit is so stable at 50 KHz, that it was actually built and tested using a prototyping board with no ground plane at all.

Operation at 455 KHz will definitely require a ground plane in the construction. At 455 KHz IF frequency and higher, the combining circuitry requires the use of three LM318's or equivalent.

The Math Behind The Detection Process.

There have been several articles written discussing the production of harmonic distortion by diode envelope detectors. Almost all of these start with the exponential diode equation and then examine the cross modulation products derived from the series expansion of the exponential function.

$$e^X = 1 + X + \frac{1}{2} X^2 + \frac{1}{6} X^3 + \frac{1}{24} X^4 + \ldots\ldots$$

The simplistic AM modulated signal is:

$V(t) = \cos(W_c)\,t + m/2 \cos(W_c - W_m)\,t + m/2 \cos(W_c + W_m)\,t$, where m is modulation index = Carrier + LSB + USB

The term of significance in the e^x expansion series is the square term as it provides the cross product of the carrier with each sideband, as in synchronous detection. Unfortunately, it also provides the cross product of the two sidebands which produces second harmonic distortion of the modulating signal.

Consider: $(a + b + c)^2 = a^2 + ab + ac + b^2 + ba + bc + c^2 + ca + cb = a^2 + b^2 + c^2 + 2ab + 2ac + 2bc$

Where; $\cos(W_c)\,t = a$ $\quad$ $m/2 \cos(W_c - W_m)\,t = b$ $\quad$ $m/2 \cos(W_c + W_m)\,t = c$

The three squared terms produce signals at twice the carrier frequency by the trigonometric double angle formulas:

$$\cos^2(W) = \frac{1}{2} (\cos (2W) + 1)$$

The 2ab and 2ac terms multiply each sideband with the carrier:

$$2 (\cos(W_c)t) (m/2\cos(W_c-W_m)t) = m/2 [\cos (2W_c - W_m)t + \cos (W_m)t]$$

$$2 (\cos(W_c)t) (m/2 \cos(W_c+W_m)t) = m/2 [\cos (2W_c + W_m)t + \cos (W_m)t]$$

Which results in double sideband AM at twice the carrier ($2W_c$) plus the desired, demodulated output $m \cos (W_m)t$.

These detectors are often referred to as square law detectors for this reason.

The problem is the 2bc term which is the cross modulation of the two sidebands themselves. It results in:

$$2bc = 2 (m/2 \cos(W_c-W_m)t) (m/2\cos(W_c+W_m)t) = m/4 [\cos(2W_c)t + \cos (2W_m)t]$$

The $m/4 \cos (2W_m)t$ term represents some 25% second harmonic distortion, which is much more than what is usually experienced in the real world.

In the real world, diode detectors work into a load resistor which often has a parallel "filter" capacitor. If the diode were an ideal device, it would only allow current to flow in the load resistor during precisely one half of the input waveform. The resultant voltage developed across the resistor would then be precisely the half wave rectified voltage of the input signal. The reality is that the current/voltage relation in a diode is a non linear, exponential function as has been discussed. This means that the output voltage on that resistor would be the half waveform of the input signal minus the voltage across the diode.

$$V_{out} = V_{in} - V_{diode}$$

It is the voltage across the diode that is the non-linear portion of the output signal, which becomes dominant at low input levels, such as the modulation trough of an AM signal. There are techniques that have been employed to reduce the diode component of the output waveform. One is to drive the detector into a high impedance load resistor. With the current very low, the voltage produced on the diode will be low. This approach is limited by the generally ignored reverse current of the diode. At high impedance levels, the rectifying function of the diode becomes compromised by the reverse leakage current, resulting in other performance problems.

The precision rectifier uses the features of op-amp feedback design to eliminate the diode error from the rectification process, resulting in a "perfect" rectifier. Well, why does a perfect rectifier work as a detector anyway? Most diode detector discussions talk about the square law function as discussed above to explain the demodulation process. The received carrier multiplies against the received sidebands. This is where the error comes in the thinking that it also should produce an objectionable level of second harmonic distortion. Let's go back to the pure half wave rectified signal that the precision detector produces and the diode with the load resistor tries to emulate. What is the process that takes an input signal at some frequency, Wc, and outputs only the positive half of the waveform. It is just as if the signal were multiplied by one during the positive half cycle, and zero during the negative half cycle. The type of waveform to do this would be a square wave, with a values of one or zero, and at the same frequency and phase as the incoming carrier. The mathematical representation of a square wave is what is known as a Fourier Series expansion. For our sequence of one, zero, one, zero... at a frequency of Wc the function is:

$$f(X) = \tfrac{1}{2} + 2/Pi \cos Wc\ t + 2/3Pi \cos 3Wc\ t + 2/5Pi \cos 5Wc\ t +$$

When this is multiplied times the incoming signal, we get a result that is the half wave rectified waveform. Our interest is in the fundamental term, cos Wc t. The higher frequency harmonic terms and the DC term will produce results that are all at or above the incoming IF frequency. So the product term of interest becomes:

$$Vout(t) = \{A \cos(Wc)\ t + \tfrac{1}{2} AB \cos(Wc - Wm)\ t + \tfrac{1}{2} AB \cos(Wc + Wm)\ t\} * 2/Pi \cos Wc\ t$$

The demodulated sideband terms are:

$$Vout(ty) = AB/2Pi * \cos(Wm)\ t + AB/2Pi * \cos(Wm)\ t \text{ , or}$$

$$Vout(ty) = AB/Pi * \cos(Wm)\ t$$

There are no intermodulation terms to contend with. In fact, functionally, this is identical to synchronous detection. The negative half wave signal is derived in a similar fashion. The two signals, when combined in the summing circuit produce an output with twice the ripple frequency to be filtered by an output low pass filter. This is of value when the IF frequency is only 50 KHz, as in the Drake R4-C and others. For receivers with a 455 KHz or 500 KHz IF, replace the two LCMC660 op amps with a pair of LM318's.

R-390/R-390A Carrier Meter Zero Adjust Mod

David Wise September 2007

The R-390A Carrier Zero pot, R523, is very touchy. A popular mod is to replace it with a ten-turn unit. This works, but it doesn't solve the underlying problem, it only covers up the symptom.

I have developed a simple, reversible mod that makes the original pot work better than new. It spreads out the adjustment range and makes it linear. Only the IF deck changes, and you may not have to remove it from the radio. You don't even have to demount the pot.

This mod is very effective on the R-390A. It can be applied to the R-390 with more effort and lower expectations.

PRE-REQUISITES

1. This mod is for the original 17-ohm meter only.

2. If, like mine, your meter reads high (more than 100dB with 100mV of signal on the balanced antenna input), you can do this mod without removing your IF deck. Otherwise you will have to change a resistor inside.

There is another, more involved mod that also allows you to use a nonstandard meter, which I will write up later.

THEORY

The R-390A Carrier Meter circuit places a milli-ammeter between an adjustable voltage and the variable voltage at the cathode of the AGC time constant tube V506A, whose grid is on the AGC bus. At zero signal, V506A is saturated and conducting a plate current of about 2.2mA . This current develops about 60mV across the 27 ohm cathode resistor R548. If the adjustable voltage on the other side of the meter is also 60mV, the meter indicates zero. As the signal level increases, the AGC bus goes negative, reducing V506A's current and hence the voltage on the minus end of the meter, and the meter goes upscale.

The adjustable voltage comes from a variable resistor in the 4th IF V504 cathode. This tube is not controlled by the AGC bus; instead it uses the "cathode bias" technique, where the cathode current goes through 680 ohm R524, pushing the cathode positive with respect to the grid in a way that tends to minimize change. The operating point is about 13mA and 9V*. The designers inserted a small variable resistor at the bottom of R524 and anchored the meter at the junction.

* Says the manual. In my radio, four different 6AK6 tubes yielded from 7.5V to 8.5V, so the average current is 11mA not 13mA. All my calculations below are based on 11mA.

Getting 60mV from 11mA takes 5.5 ohms*. The smallest pot available at the time of the R-390 was the 15 ohm wire-wound unit that they used. As the radios went into service it became apparent that the change as the slider moved from one turn of resistance wire to the next was objectionable, and in the R-390A they replaced it with a carbon pot (Cost Reduction Report, section 4.3.3, page 15). (There is new speculation that low-ohm pots were also regarded as unreliable and their use became discouraged.) The smallest unit available was 100 ohms, and the engineer unimaginatively tacked a 22-ohm fixed resistor, R537, across it to bring it down to approximately the same maximum value as before. This new pot did not suffer from the stepping effect of its predecessor, but the adjustment range is compressed into the last few ohms of R523. Since the nominal setpoint is 7.3 ohms (7.3 in parallel with 22 is 5.5), the pot is normally at about 7% rotation, and it's really touchy. As the pot ages and gets scratchy, it becomes impossible to keep it stable.

Table 1.

V	R
.01	0.9
.02	2.0
.03	3.1
.04	4.4
.05	5.7
.06	7.3 <- nominal
.07	9.0
.08	10.9
.09	13.0
.10	15.5
.20	100

Voltage vs R 523 Ohms
(stock, 11mA, R 537 = 22)

* These values are for the R-390A. The R-390 needs about 70mV from 10mA. This is discussed at the end of the article.

As the pot goes from 0 to 100, the total cathode resistance seen by V504 goes from 680 to 698, essentially no change, so I simplified the calculations to assume constant current.

You can see how nonlinear this is. 35% of the output range is crammed into the first 10% of rotation. At nominal, 1% adjustment equals 10% change.

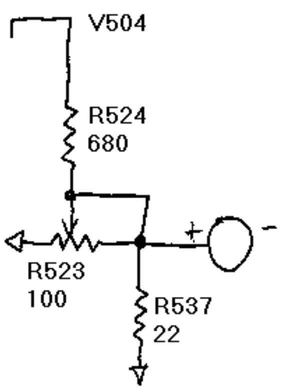


Figure 1
Original

If we are allowed to increase the effective resistance in series with the meter (lowering its full-scale reading), we can improve this dramatically with a small wiring change.

The stock design sets up R523 as a rheostat; the slider and the CW end are connected together. The CCW end is grounded, and R537 goes across. These parallel resistors form the lower leg of a voltage divider. The upper leg is R524 inside the IF deck. See Figure 1.

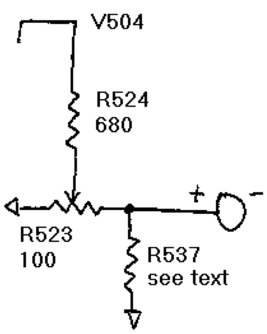


Figure 2
Modified

If we modify this as in Figure 2, we get a current divider, analogous to the voltage divider you're all familiar with. When the pot is CCW, all the current flows straight to ground and the meter reference voltage is zero.

When the pot is CW, almost all the current flows through R537, supplying 100mV. In the middle, some current goes one way, some the other, and we get an intermediate voltage.

Let's derive the expression for reference voltage as a function of pot rotation x. Say I is the current in R524. It splits in R523. I1 goes to the left, and I2 goes to the right and down R523, which we set to arbitrary resistance R.

Then $I2/I = x/((100-x+R)+x) = x/(100+R)$, $I2 = xI/(100+R)$, and $V = I*I2*R$ or $x\ IR/(100+R)$. It's linear!

Table 2.

V	R	
.01	10	
.02	20	
.03	30	
.04	40	
.05	50	
.06	60	<- nominal
.07	70	
.08	80	
.10	100	

Voltage vs R523 Ohms
(modified, 11mA, R 537 = 10)

Notice that 10% rotation yields 10% voltage change, compared to 1% in Table 1. It is ten times easier to adjust. Scratchiness is ten times less obvious, and making replacement unnecessary in most cases.

This is how Collins should have done it. All I can figure is, the engineer assigned this task was new and naive, distracted by other matters, or (the Collins Collectors Association will burn me at the stake) just plain incompetent.

My '54 Motorola was wired like Figure 3. Your mileage may vary.

HOW TO DO IT

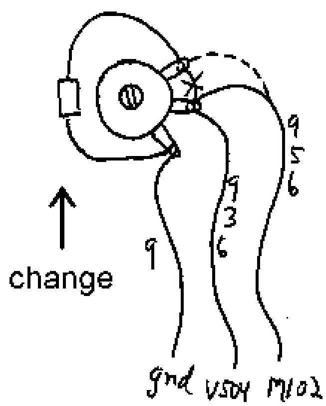


Figure 3
Pictorial
R522/R523

At R523, the slider and CW terminals are jumpered together; then there are two wires and R537. One wire goes to V504, the other to the meter. In my radio, the meter wire is white with green and blue tracers, and the V504 wire is white with orange and blue tracers. Both were on the slider. Remove the jumper and R537, move the meter wire to the CW terminal, and install the new R537 between the CW and CCW pins. That's it! Fire it up and revel in the smooth, easy adjustment. It should zero around 60%; if your 6AK6 is flat you'll have to turn it higher.

The soldering heat may change the old R537. You might as well replace it with a new carbon- or metal-film unit. The value, which can range from 5.6 ohms to 22 ohms, determines the full-scale reading, and depends on many factors.

To approach the original reading, use 5.6 ohms. To meet or exceed the original reading, you will also have to change R549, see the end of the next section. I'm using 10 ohms for R537 and the original R549.

COMMENTARY AND EXPERIMENTS

There are three side effects, one innocuous, one beneficial, and one possibly detrimental. The latter can be neutralized by changing one resistor inside the deck.

First, the cathode resistance seen by V504 increases slightly. At the nominal operating point, the stock design was about 685 ohms. Now it is 705. This is less than 4% change, well within tolerance. I could not detect any change in IF gain.

Second, the meter full-scale reading is now independent of the zero adjust. Third, the full-scale reading may be less than before. It depends on the cathode current of 6AK6 V504. Strong tubes make it worse, because you have to turn down R523 to get a zero. You can increase the full-scale reading by decreasing R537, until you run out of zero-adjust range.

At worse case the stock circuit will show the meter just 4.5 ohms, which is unreachable with the mod as documented so far.

But you may not want the original reading; it may have been too high. Inject 100mV of signal into the balanced antenna connector and note the resulting AGC voltage. At this voltage, you want to read 100dB. I'm using 10 ohms, see Table 3 below.

Note that 10 ohms yields a Thevenin resistance of 9.1 ohms, somewhat larger than the stock value of 5.5 . At 5.5, my meter reads high; this is what enables me to use the simple version of the mod. If your radio needs 5.5 ohms, then you will have to change R549; see the end of this section.

The AGC voltage developed at a given input depends on the 6DC6's gain and its cutoff characteristic. Hot ones require more AGC. So do "wide" ones. I tried five different tubes. If you try this, it's essential that you measure the antenna voltage, because trim and choice of tube affects the input impedance.

The AGC also depends on the cutoff characteristics of 6BA6 IF amps V501, V502, and V503. Wide ones develop more AGC. It also depends on the gain of 6BA6 AGC amp V508. A hot one will develop more AGC.

I played with 6DC6 substitutes. I tried 6DK6, 6CB6, 6DE6, and 6BZ6. As you can see, they're either too sharp or too wide, but they're good enough to keep it glowing until you get the real thing.

Finally, the AGC depends on IF gain. Table 3 was compiled with IF gain set for best S/N ratio. The gain was 600uV, rather than 150, for -7V on the diode load. The Tung-Sol 6DC6 that got -13.3V in Table 3 gets -13.8V when the IF gain is reset to the standard value. In that situation, the perfect R537 would be 11 ohms.

IF gain also affects the 10dB point. At 600uV/-7V, 10dB input does not cause much deflection. At 150, the meter reads above 10dB. Somewhere in between, it's right on. If you don't feel like searching for the S/N sweet spot, this trick will probably get you closer than the stock setting.

The Carrier Meter reading for a given AGC voltage also depends on the no-signal V506A plate current, which in turn depends on B+ and the exact value of the load resistor R549.

TUBE	AGC	R537	Total R
Tung-Sol-13.3	9.6	8.8	
RCA #1	-12.7	8.8	8.1
RCA #2	-12.1	8.2	7.6
RCA #3 '68 JAN	-13.3		
RCA #4 '62 JAN	-13.6	12.2	10.9
6DK6	-8		
6CB6 #1	-9.1		
6CB6 #2	-9.6		
6DE6 #1	-11.1		
6DE6 #2	-12.3		
6BZ6 #1	-14.1		
6BZ6 #2	-15.4		

Table 3.

AGC voltage with 0.1V on balanced antenna input, and R537 setting to get 100dB reading (1450KC, trimmed, IF gain set for best S/N)

Raising B+, or lowering R549, increases the reading. The same 12AU7 that reads 100dB at -13V with an 82K pullup to 205V only needs -9V with a 56K pullup to 180V, the setup that obtains in the R-390. This opens the possibility of tweaking R549 to control the full-scale sensitivity. If you can't get a full-scale reading with my recommended R537 value, lowering R549 will get you there. Of course you have to remove the IF deck to do that, which I was trying to avoid.

THE R-390

TM11-5820-357-35 is inconsistent regarding the 6AK6 final IF (V506) cathode resistor R536. All the schematics show it as 820 ohms, while the voltage/resistance diagram says 700 ohms, 8.2V. One owner reports 900 ohms and 9.4V, so the V/R diagram is probably wrong. In any case, the reference current is less. At the same time, the AGC time constant tube is driven harder (56K from 180V instead of 82K from 205) and therefore runs a higher cathode voltage, about 70mV nominal instead of 57. If you applied my technique to the existing 15-ohm pot, R537 would be 15 ohms, and the maximum reference voltage would be 90mV if R536 is 680, around 75 if R536 is 820. In the latter case, you'd find yourself setting the zero adjust around 95%; kind of close to the edge.

This mod reduces the turn-to-turn resistance jump by a factor of two. Not having any direct experience with the R-390, I can't say if this is enough of an improvement to be worth doing. You could always replace the pot with a 100-ohm carbon like the R-390A. The benefits are not so great and the effort and risk of side effect are larger. Still, it's easy to do and undo; I'd try it.

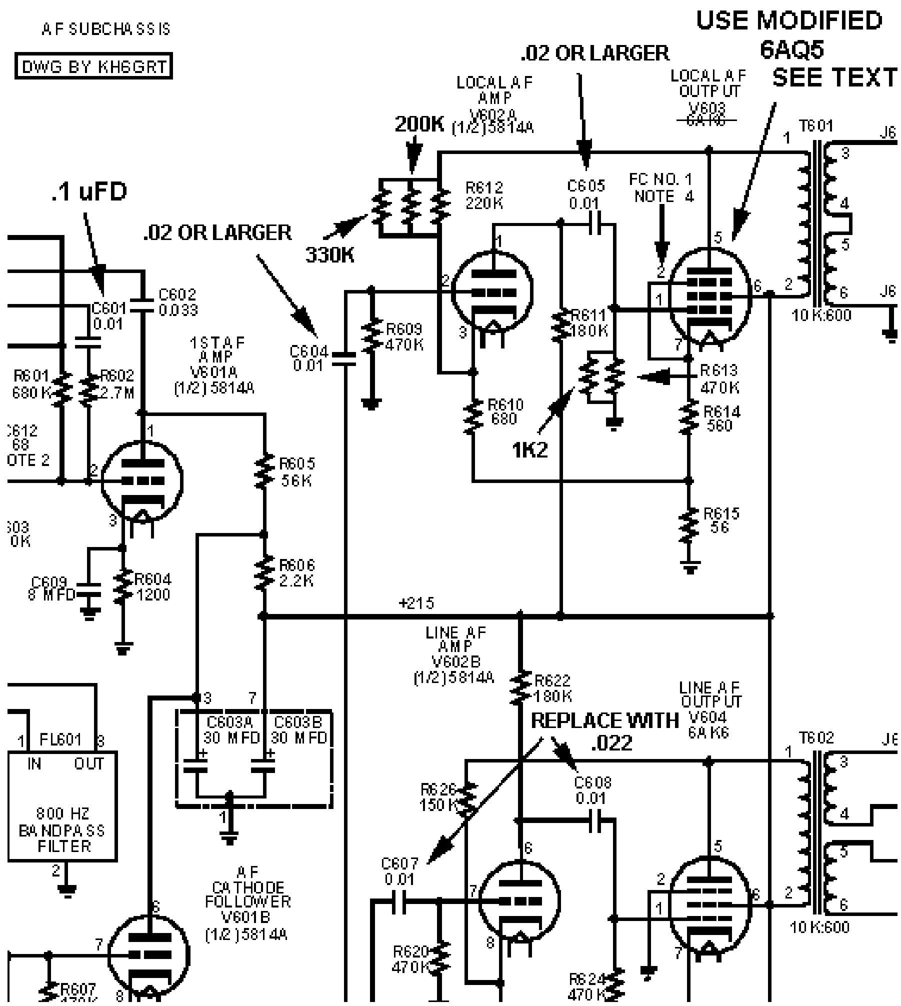
AF LF Fidelity, Distortion, & Power Mods

Dallas Lankford 4/5/04, rev 3.

Recently I implemented several versions of Chuck Ripples LOCAL Audio LF Mod (changing C604 and C605 to 0.02 or larger). It works very well provided you don't go too large which causes motor boating at higher LOCAL GAIN settings.

This change lowers the 6 db point from about 70 Hz to about 30 Hz, I also replace C607 and C608 with 0.022's for similar improvement in the LINE a1udio.

Replacing C602 with a 0.1 reduced the LOCAL to about 30 Hz. I used 630v polyester metIa1ized film capacitors from Antique Electronic Supply, I tried the 6AQ5 triode conversion mod but it was a disaster. Audio power was not increased and distortion was much worse.



For reduced distortion and modest power increase I paralleled 1K2 ohm 1/2 watt across R613 (increases power output) and 330K and 200K 1/2 watt across R612 for more negative feedback (lower power). The result was reduced distortion and increased power (0-9 to 1,0W RMS into 8 ohms real,

The Graham Maynard mod, cutting off pin 7 of a 6AQ5 instead and replace the 6AK6 works very well with my mod having the same low distortion and power output of 1.4 W. RMS max. If you want more power use the LINE output with a high fi.

From dallas at bayou.com Fri Sep 24 08:10:23 2004
Subject: [R-390] Carrier Meter Linearization

TM-11-856A, among others, shows (in Fig. 41) a linear (wrt dB) carrier meter response for the R-390A. As we all know, of course, this is not the case. Carrier meter readings are always 20% to 30% low (20 or 30 dB less than 100 dB when a precision signal generator is used to examine the accuracy of a carrier meter in an R-390A through its 100 dB range). The reasons for the inaccurate carrier meter readings are incorrect values for R524, the R523 + R537 assembly, and R548. After removing R524 and R548 and bringing insulated wires out of the IF deck so that I could vary those values, I determined that (1) R537 (22 ohms nominal) should be removed entirely (to provide more range for meter zero adjustment), (2) R524 (680 ohms nominal) should be replaced by 649 ohms 1% (since R524 is usually higher in value than it should be, you can usually obtain 649 ohms by adding an appropriate resistor in parallel), and (3) R548 (27 ohms nominal) should be replaced by 62 ohm 1%. R523 should probably be replaced by a 10 turn 2 watt high quality 100 ohm wire wound variable resistor to make zero adjustment easier and to improve long term zero set stability (though "zero" will still drift around a bit).

From dallas at bayou.com Fri Sep 24 09:36:01 2004
Subject: [R-390] 3TF7 Substitutes

The standard 3TF7 substitutes, (1) using a 42 or 43 ohm 10 watt resistor in place of the 3TF7, and (2) using an appropriate tube, like a 12BY7A, with a 12.6 volt filament in place of the 3TF7 are both acceptable substitutes. Using a 10 MHz rubidium standard I determined, somewhat to my surprise, that the power resistor is generally a more stable substitute than a 12BY7A. Recently while examining the long term frequency stability of one of my R-390A's with the BFO turned ON (for SSB, ECSS, or CW) using a rubidium standard, I found that the 3TF7 does not do a very good job of stabilizing the BFO and PTO frequencies when the AC line is varied. A change of only 2 or 3 VAC in the line voltage (I used a VARIAC to vary the AC input voltage to the R-390A) causes a substantial (4 or 5 or 6 Hz or more) departure from zero beat. Next, I removed the 3TF7, inserted a 9 pin tube test extender into the 3TF7 socket, and powered the BFO and PTO filaments with an external regulated 12 VDC supply (12 VDC was found to give almost exactly 300 mA filament current). With this arrangement, no change in zero beat was observed as the AC line voltage was varied from 120 VAC nominal down to 100 VAC and back up to 120 VAC. WOW. Whoever designed the original BFO and PTO filament stabilization circuit was on the right track. They just used the wrong method to stabilize it. Current regulation is the wrong approach; voltage regulation is the correct approach.

Rather than rewire the 3TF7 socket, I opted to make the mod "plug-in" in so far as it was possible. I cut the metal flange off a miniature 9 pin ceramic tube socket, pushed 9 pieces of #18 tinned solid copper wire in each receptacle, soldered them, cut off the ends to the appropriate length for a 9 pin tube, deburred and polished the tips, drilled out the cylindrical center piece of metal and removed it.

I ground off most of the head of a 6-32 brass screw of the appropriate length and attached an inch long (or somewhat longer) insulated spacer, and mounted an insulated standoff on the threaded end. This provided me with a home made tube socket extender on which I could build most of a 12 VDC regulator.

There is a nut on the front of the IF deck where I added a ground lug. I ran a diode from the #2 pin lug of the adapter to the standoff, and a 1000 mF 50 volt electrolytic from the standoff to the ground lug. The ground tab of a 3 pin 12 volt 1 amp regulator was attached to the RF deck corner nearest the IF deck using one of the green screws that hold the oscillator deck to the RF deck plate; the regulator pins stick up above the top edge of the RF deck plate.

The input and output pins of the regulator were bypassed to the ground pin with 0.1 mF 50 volt capacitors, and the regulator ground pin was wired to the added ground lug on the IF deck. An insulated wire from the standoff to the regulator input pin and an insulated wire from the regulator output pin to lug 7 of the plug-in adapter completed the modification. Well almost... three (3) complete wraps around the lugs of the home made adapter with Scotch Glass Cloth Electrical Tape and heat shrink tubing on the standoff protected the plug-in adapter from shorts.

This mod is not 100% plug-in because to remove it you have to (1) remove the nut on the front of the IF deck to remove the ground lug, and (2) remove the green screw on the RF deck to remove the 3 pin regulator.

I have now had the mod running continuously for about 48 hours. No problems were expected and no problems have been experienced. Not only does this mod give you improved frequency stability for ECSS, SSB, and CW, it should also provide a permanent solution for the 3TF7 replacement problem. There is still some very slow frequency drift, as much as 1 Hz per hour, sometimes more. I currently do not know the cause of this drift.

> I have already spotted a few typos in my posting, which I believe will be obvious to those who read it. I meant to include that I used a 1N4003 diode, 200 PIV and 1 amp. I presume a 100 PIV 1 amp diode would be fine. I just happened to have 1N4003's on hand. Also, be sure to mount the 3 pin regulator to the RF deck plate (unless you want to drill a hole in the IF deck and mount it there). I don't know how > large a heat sink the regulator actually needs, but the RF deck front plate is surely much more than enough.

Best regards, Dallas

Revised: 3/8/06
Circuit Protection
Home <http://www.skirrow.org/Boatanchors/index.htm> Reference <http://www.skirrow.org/Boatanchors/reference.htm> Articles <http://www.skirrow.org/Boatanchors/how%20to.htm> Links <http://www.skirrow.org/Boatanchors/links.htm>

The following series of posts to the BoatAnchors list consider the problem of protecting equipment from current and voltage transients that can occur when the power is first turned on.

Unfortunately in the course of saving these and putting them together, the headers were lost, along with the names of some of the posters. Thus, I've generally omitted the identification of the poster. This also leaves me free to correct the spelling and delete the repetitive parts. But, if you recognize your stuff, and would like your name reattached, just ask!

As always, these posts are offered without any guarantees!!

Inrush Current Protection

Post 1: Selecting an Inrush Current Limiter

Some time ago I discovered a neat little device that solves the inrush problem and, as a side benefit, reduces high line voltage. All this for a bit over $2! The device is an Inrush Current Limiter made by Keystone Carbon Co. The beasties look like ceramic disk capacitors with a black vitreous coat.

The limiter is a Positive Temperature Coefficient thermistor which is designed to handle current. When cold (room temp -25C) they exhibit some resistance. As current passes through them and they warm up, this resistance drops by a factor of about 100. The limiters are rated by current handling capability (1.1 to 16 Amps) and cold resistance (0.7 to 120 Ohms). Not all possible combinations of resistance and current are available but at last look there were about 20 different types.

You use the limiter by installing it in series with the line cord (preferably the hot lead) input to your BA. This can be done in a fashion that is totally esthetically pleasing (read "out of sight") and completely reversible.

IMPORTANT: Since the device is a resistor (and a HOT one at that) you must mount it away from heat sensitive components. I have mounted them under chassis without trouble but keep 'em away from just about everything. Don't attempt to heat sink it - that ruins the operation!

Pick the right value by first measuring the steady state current of your BA. That is, after it is fully warmed up and all accessories are turned on. While you're at it , also read your line voltage. Pick a unit that has a MAX steady state current of 120 - 130% greater than the current you measured and has the HIGHEST no-load or cold resistance.

Example: You measure 2.5 Amps (a moderately hungry BA!) and the line voltage is 123V. The KC008L is rated for 3.0 Amps with a cold resistance of 47 Ohms - a nice fit.

Benefits: A BA drawing 2.5 amps probably has a transformer with a primary DC resistance of about 3 Ohms. Inrush, at the peak of the AC sine wave, could be as high as 40 Amps but probably not less than about 20 Amps. With the limiter installed, the inrush will not exceed about 2.6 Amps at 123 line Volts. After the limiter warms up it will have about a .49 Ohm resistance (actually a bit higher because we're not drawing the full 3 Amps.). This means that the line voltage across the transformer will be about 122 Volts (also a bit lower because of the higher resistance). This example came from real life and my actual results showed that the line voltage was reduced to 118 Volts (the BA was rated for 117) which means that the limiter was adding about 2 Ohms.

Negatives: If your area suffers from brownouts, the limiter will exaggerate the effect. If voltage drops, current drops. The limiter will cool a bit, its resistance will rise, and the voltage your BA sees will drop more than the line voltage. This is a very minor problem for me but I feel bound to mention it.

Post 2: Experience with a 51S-1 Receiver

In the past few years, a new kind of thermistors has become available for limiting start-up surge currents in electronic instruments. They differ from conventional thermistors in having a negative temperature coefficient (resistance decreases with increasing temperature), and this property gives them a useful self-regulating characteristic. Placed in the ac line of an instrument, they initially have a high resistance, which limits the inrush current through the instrument. Upon application of power, the current through the thermistor causes self-heating, which lowers the device's resistance. At some point the resistance stabilizes to a value that depends on the equilibrium temperature of the device. The equilibrium temperature is determined by the steady-state current drain of the instrument and the ambient air temperature surrounding the thermistor. Current-inrush thermistors are inexpensive and provide an effective way to protect power supply components in vacuum tube receivers, particularly those that use solid-state rectifiers. Note that you should not use current-inrush thermistors to protect transmitters or amplifiers; they are only suitable for instruments that draw a relatively constant current from the line. (See later post) Here are the details for protecting a typical boatanchor receiver, in this case a Collins 51S-1.

The steady-state current drain for my 51S-1 is about 0.8 Amps at 120VAC. To measure the inrush current, I temporarily removed the 1.5 ampere slow-blow fuse and jumpered a 1 ohm resistor across the fuse terminals. By measuring the voltage developed across the resistor with a scope, I determined the peak inrush current to be slightly more than 7 amperes! The equivalent load resistance presented by the 51S-1 at turn-on is thus (120 VAC/7 Amperes) = 17.1 ohms. As the filter capacitors charge and the tube filaments warm up, this load resistance increases to a steady-state value of (120 VAC/0.8 Amperes) = 150 ohms.

A 7 ampere inrush current is very hard on the power switch, and isn't so great on the power transformer, rectifier diodes, and filter capacitors. The most suitable inrush thermistor I could find was Digikey (1-800-DIGIKEY) part number KC014L-ND, at a price of $2.13. This thermistor is specified at 50 ohms resistance at room temperature (54 ohms measured), and dropping to 0.89 ohms at 1.1A load. I measured the resistance at 1.1 ohms at the current drain of the 51S-1. To install the thermistor, I clipped the wire to the fuse socket of the 51S-1 and relocated it to an unused lug on a nearby turret. I then soldered one lead of the thermistor (which physically resembles a small disk capacitor) to the same lug and the other to the recently vacated lug on the fuse socket. I used a bit of Teflon tubing on the leads, and kept the leads long so I could suspend the thermistor in free space away from other components. The thermistor dissipates about a watt of heat and runs rather hot.

After installing the thermistor, I replaced the fuse with a 1.5 Amp fast-blow type. I then remeasured the peak inrush current and found it now to be only about 1.8 Amperes, which is consistent with the theoretically expected value of 120 VAC/(54ohms+17ohms) =1.69Amperes. The peak inrush current is now only slightly greater than the steady-state current drain and should thus pose no problem for any of the power supply components. Note that this particular thermistor is appropriate for almost any boatanchor receiver that draws 75-150 Watts from the power line.

Concern is often voiced about a related turn-on problem (actually, a turn-OFF problem), namely the inductive voltage spike caused by the power transformer inductance when the power is switched off. This spike is reputed to cause sparking and welding of contacts in hard-to-replace power switches, particularly in rigs like the KWM-2 and S-line. I checked on this problem with my 51S-1, but measuring the peak voltage developed across the power switch when the rig was shut off. (My Fluke 87 DMM has a peak-reading feature which can capture voltage transients as short as 1 msec.) To my surprise, I found that the inductive voltage kick was only about 5 volts higher than the line voltage, and was no cause for alarm. I had thought about using an MOV surge suppressor across the switch contacts, but decided it wasn't necessary. This is not to say, of course, that the problem isn't greater in some other rigs, but 51S-1 owners need not worry.

Post 3: Inrush Protection for Transmitters

Comment:

>>Note that you should not use current-inrush thermistors to protect transmitters or amplifiers; they are only suitable for instruments that draw a relatively constant current.

Response:

Au contraire. Inrush current limiters work nicely in transmitters and transceivers and probably in amplifiers as well, although I've not tried that. The only stipulation is that the device must be selected to allow the maximum current needed by the transmitter. The resistance of the thermistor after the initial surge is very small, a fraction of an ohm, and less than the resistance provided by the typical AC mains. Consequently, its effect upon the load regulation of the transmitter is negligible. I used an inrush limiter a while back in an Eico 753/751 transceiver supply with excellent results. Prior to using the inrush current limiter, the power on surge produced an unnervingly loud KWUMMP! After installing the inrush current limiter, powering up the unit produced no audible effects at all. I don't happen to remember the voltage drop across the inrush limiter when just the receiver was operating, but I did measure it and found it be negligible; on the order of only a volt or two.

And Someone Else Added:

Of course, on larger transmitters one has to use thermistors on each element. Generally the current draw is too large to protect the entire transmitter.

The filaments transformers, the plate transformer, low voltage transformers should all be individually "thermistorized".

In mine, I find a volt or two drop at the thermistors is just what the doctor ordered as the line is slightly high.

Post 4: Mounting Caution

Don't solder them into your circuit unless you want trouble. They do get hot in operation and repeated heating and cooling of a solder joint will cause it to crystallize and eventually fail. This was a common failure in televisions with thermistors used in the degaussing circuits, and even with some of those cement-block power resistors on circuit boards.

Put in a small screw terminal strip to mount the ICL. Crimp terminal lugs on the ICL and then attach it to the strip with the screws. In the long run, this will save lots of grief and it also makes installing and insulating the ICL a snap.

Voltage Spike Protection

Post 1: MOVs

Turning off a rig can cause a big voltage spike across the transformer primary and the AC line. Usually it just burns out or welds your switch, as R-390A users often learn.

A back issue of The Collins Journal suggested getting 240-volt MOVs and wiring them across your primaries to absorb the transient. Note that if the MOV fails (shorted) it will suck lots of current, but you have a fuse in the line, right?

These will protect your switch, and apparently your transformers could use it too. I doubt the big toggle snappers in a Viking need it as much as the wimpy switches in an R-390A or KWM-2, but your transformers may last longer this way. And you'll get protection from nasty things that come in through your power line, and your gear won't put glitches back out there when you turn it off.

Post 2: Selecting MOVs

Query:

>> There have been a number of posts touting the use of varistors to protect against voltage surges. Question is: How to decide what specs when buying these little doo-dahs?

Answer:

My background is in Mechanical Engineering, so take what I am about to say with a grain of salt. When I have picked MOVs (metal oxide varistors) in the past, say to protect stuff on the AC line against spikes, there are two things I have been concerned about. First is the clamping voltage. These little do-dads work by turning from a non-conductor to a conductor at the clamping voltage. The other rating is the amount of current they can handle. Usually this is broken into two numbers, a surge number with a time (like 7000 amps for a microsecond) and a steady state value if I remember correctly. So when I picked one to make into a AC surge suppressor, I picked a clamping voltage of about 150 volts with the highest current capacity I could afford.

Theory & Protection

Post 1: Theory

On Fri, 23 Aug 1996, Jan Skirrow, VE7DJX, asked me some excellent questions about thermistors, varistors, and such. I hope he does not mind me posting his questions or my reply to the group. Thermistors are not often seen in boatanchors (or in a lot of modern semiconductor stuff for that matter). I know a little about them because of their use in temperature measurement and instrumentation.

>>First, I conclude that NTC thermistors would be placed in series with, for example, a transformer and would thus limit in-rush current because their resistance is inversely related to temperature, which would rapidly increase on start-up.

Exactly. They are particularly beneficial with power supplies having capacitor input filters.

Look at the special devices sold as Inrush Current Limiters, not conventional thermistors. Keystone is probably the most common NTC Inrush Limiter manufacturer.

The typical resistance ratios of common NTC thermistors (for other than Inrush Current Limiting operations) is generally between 5 to 10 for 0 C to 50 C temperature changes. Plugging these numbers into the typical resistance relationship

R = Ro * exp(B/T) R, Ro in ohms, T in Kelvin

gives a Beta in the approximate range of 2800 to 4000. Using a value of 3400 as an average gives an Ro value of 0.0011 ohms (the resistance at absolute zero). So at 50 C, the resistance should be around 41 ohms (and at 0 C, the resistance is 284 ohms and the ratio is: {ta-da...} 6.9).

In true Inrush Current Limiters, the Beta value is MUCH higher. If, for example, B is 10,000, the 0 to 50 C ratio is 290. This is such that a few ohms cold becomes very low resistance when hot. I don't really know what the Beta number is for these devices but I might be able to estimate it from the specs knowing the dissipation of the hot device and estimating some heat transfer conditions. It is not necessary to know it for picking an Inrush Current Limiter for your operation.

In any event, a typical Inrush Current Limiter might have the following specifications (actually those for a Keystone CL-110):

Resistance at 25 C: 10 ohms +/- 25%

Maximum Steady State Current: 3.2 amps

Approx. resistance @ maximum steady state current: 0.18 ohms

>>I assume your reference to older metal oxide devices doesn't refer to metal oxide varistors - which seem to be a transient suppressor that functions by clamping the voltage across itself to some fixed level.

NTC thermistors are generally made from oxides of manganese, nickel, cobalt, copper and iron. Metal oxide varistors for transient voltage suppression are generally variations on zinc oxides. Older thyristors were generally silicon carbide.

It is interesting that while quite different in operation, the thermistors and varistors obey similar exponential relationships. The simple thermistor relationship is shown above. The current through a varistor follows a similar one:

I = Io * exp(a*V) I, Io in amps, V in volts

If you look at more exact relationships with both temperature and voltage dependency included, the equations start looking VERY much alike.

Basically a varistor draws very little current at low voltages, but as the voltage increases, the current increases very rapidly.

>>So these would be used by placing them across (for example) switch or relay contacts that switch an inductive load, and would prevent the voltage across the contacts from going too high due to transients, thus arcing and damaging the contacts.

That is one use, although in snubbing an inductive load, the presence of a diode in a DC circuit or a varistor is an AC circuit will slow down the response of the relay. You really need something that will absorb the energy stored in the magnetic field.

The more common use of a varistor is across the AC line as a transient suppressor. The voltage rating is chosen such that the device does not conduct much at normal voltages, but conducts heavily during a voltage transient.

>>So, comprehensive protection for, say an R-390A, would be an NTC thermistor in series with the power transformer and a varistor that clamped at something over normal line voltage (perhaps 150v rms?) across the troublesome main power switch.

Sort of! An Inrush Current Limiter in series with the transformer primary would reduce the current surge during turn-on. A varistor across the main power switch might help a LITTLE but what you really need here is a snubber network of a resistor in series with a small capacitor. Typical values might be 10 to 100 ohms in series with a 0.01 to 0.05 uF capacitor (rated at 1 KV minimum). A better approach would be to use a better switch!

A 130 volt varistor, like a V130LA20, would be a good choice to add after the power filter network across the line. It would protect against line voltage transients. However, it won't protect the filter here. You should probably use a proper transient protected multiple outlet strip to power the radio anyway. The best ones will have 3 varistors inside. One from line to neutral, and one each from line and neutral to ground.

Inrush Current Limiters and Transient Voltage Suppressors are quite inexpensive today. Small and unobtrusive, they can often be tucked inside your Boatanchor giving you some added protection.

Post 2: Additional Comments

>>I seem to remember horror stories about some so-called transient protected outlets that worked once, and then provided no protection as the varistors went south. All of my outlets are so protected, and I hope they all work!

This is important too. In transient suppression, you want to have as much impedance between the source of the transient and the device you want to protect as you can get. Thus for best protection, a staged approach is a good one. At the service entrance to your house, you should have one of the lightning arrestor/transient suppressor blocks made for this purpose. These cost $15 to $30 at a commercial electrical supply house. The only problem is that with installation at the service entrance, you usually have to pull your power meter.

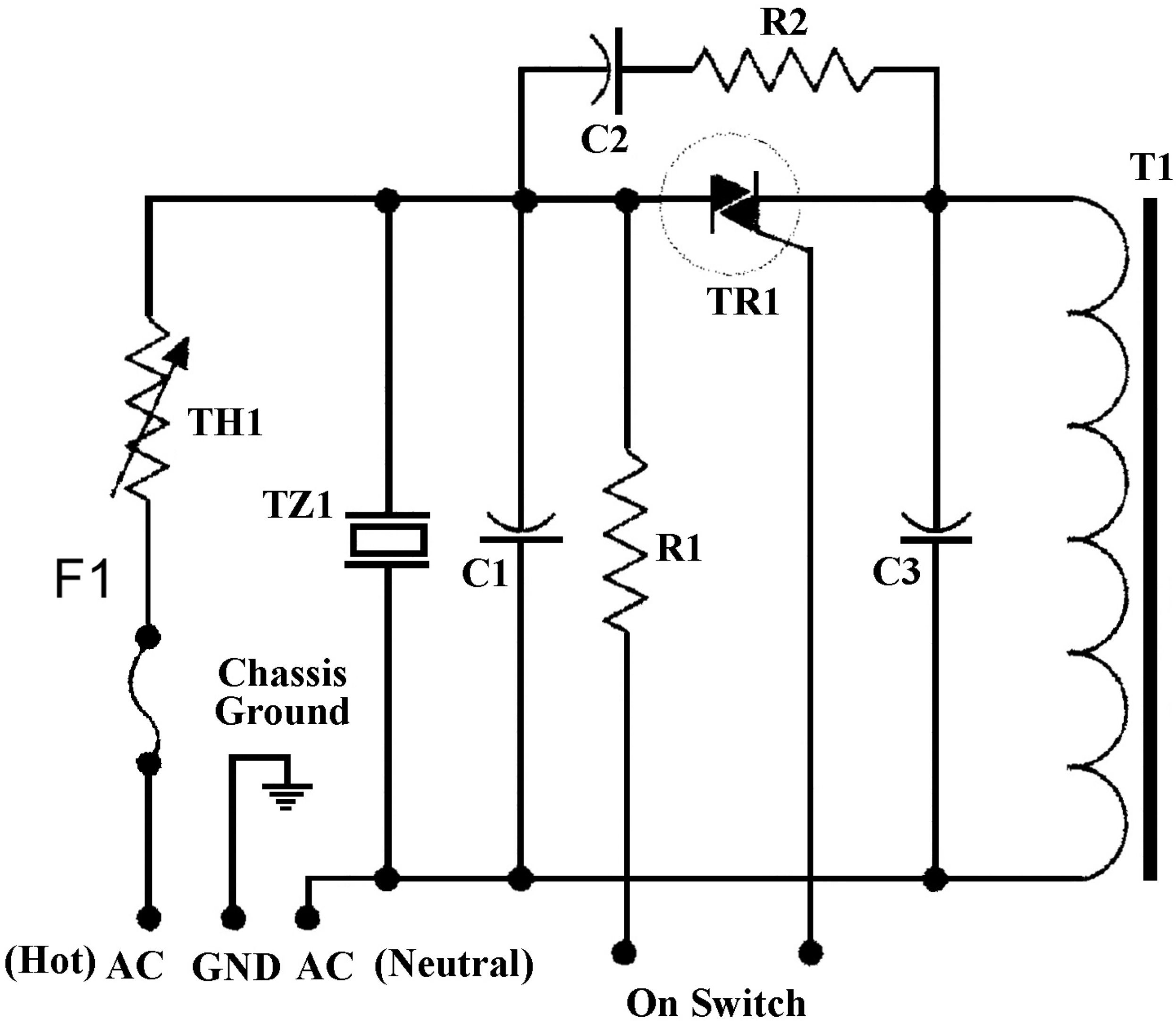
Between the service entrance and the wall outlet, your house wiring provides some distributed capacitance and inductance. A 3-MOV protector at the outlet is a good idea here. Checking them is a problem as there is no simple way to do this. If your circuit breaker or fuse blows upstream of the protector for no apparent reason or during a thunderstorm, you can probably assume the protector "went south" and needs to be replaced.

Finally at your equipment, its line cord and RFI filters provide even more impedance. A transient protection MOV inside the rig provides the final stage of protection. It can be smaller in its ratings since the earlier protectors should have already taken most of the energy away from the transient.

Nothing protects against a direct-hit of lightning though. But I would still rather have a few MOVs explode, and maybe a line-filter or two, than the entire rig to replace!

May 13, 2003

Electronic Power Switch circuit



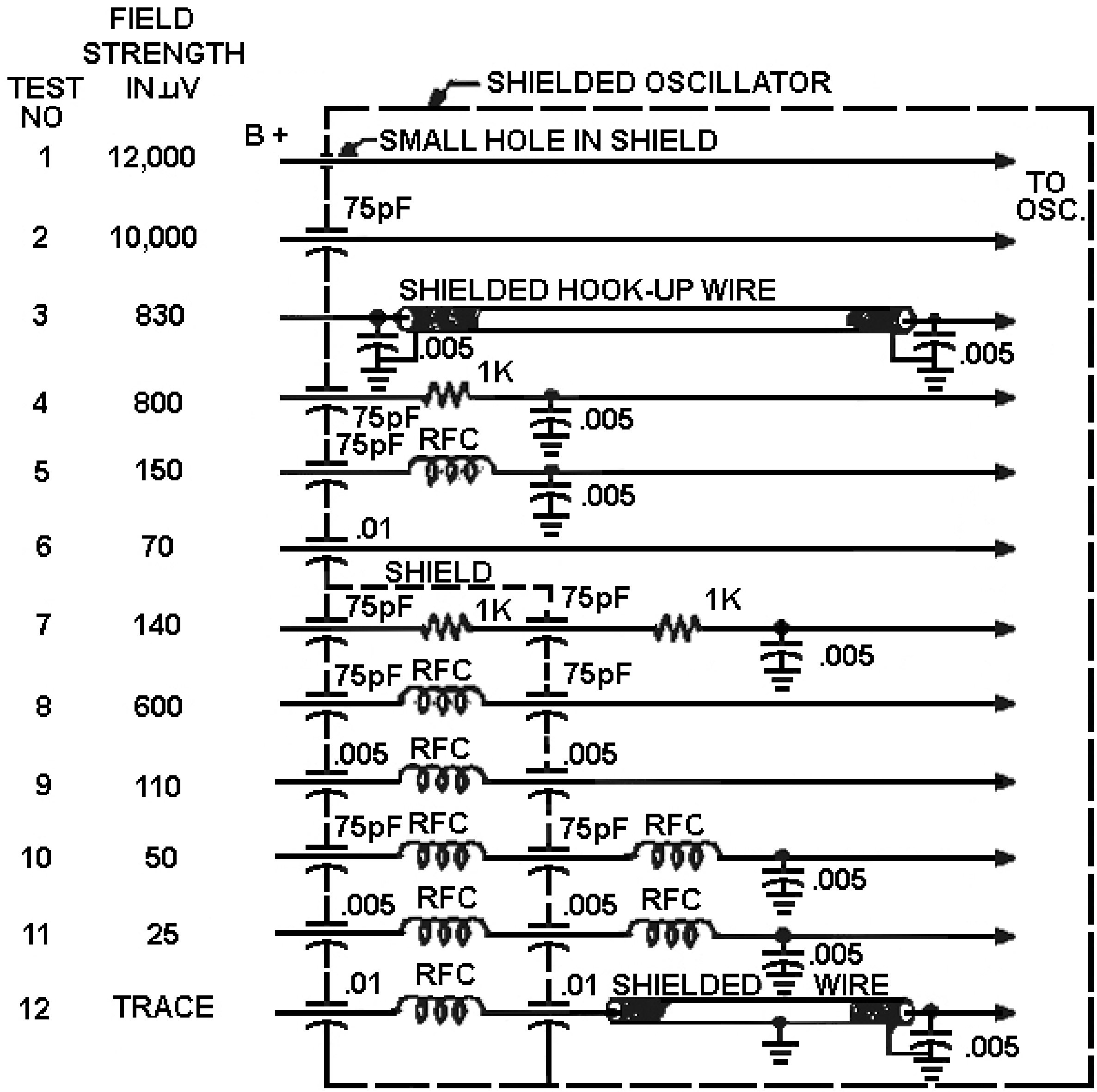
TH 1 Thermistor 2.5 Ω 8A CL 30
TZ 1 MOV 250V – 2,500A Siemens 510K250
C 1 .047µFD. 250 VAC Metallized Film
C 2 .047µFD. 250 VAC Metallized Film
C 3 .047µFD. 250 VAC Metallized Film
F 1 Slow Blow. Value as required by equipment

R 1 75 Ω 2W to 5W
R 2 75 Ω 2W to 5W

T 1 BA power transformer

The triac power on circuit is the same that is used on the Pass/Thagard A75 amplifier.

This circuit is from the 2/06 issue of Audio Express magazine pages 26-31. The author is Julian Evetts.



PARTS USED

1K CARBON RESISTER
RFC - OHMITE Z-50 (7 µHy)
75 uuF CERAMIC FEED - THROUGH
.005 DISK CERAMIC
.01 SPRAGUE HI- PASS
.005 CERAMIC FEED - THROUGH

LEAD LEAKAGE WITH VARIOUS LEAD FILTERING SYSTMS
(COURTESY W1DBM)

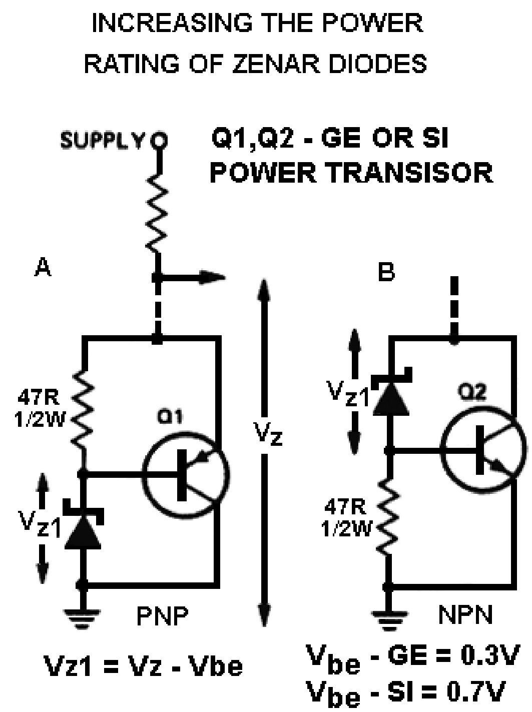
Original drawing courtersy Roy Morgan Digital remanufactioring courtesy Perry Sandeen

From this experiment two conclusions are quickly apparent. First until at least .01µFd of capacitance is used elaborate filter schemes are ineffective. The second is that a much higher choke value than the 7µHy Ohmite Z-50 is needed for the choke to be reasonably effective.

From Radio Handbook, Fourteenth Edition, Editors and Engineers, William Orr, W6SAI, Editor, 1956, Section 15-5, Page 294.

A power transistor can be used to provide a high-powered Zener voltage from a low-wattage Zener. A 400-mW Zener can be used where a 10-W Zener is required or a 1-W Zener can be used where a 50 to 80-W Zener is required by using appropriate transistors for Q1 and Q2 (in the circuits shown).

Where a low rating is required, Q1 would be an ASZ 15 (ge) or a 2N2955 (si), For higher powers, Q1 should be an ASZ18 (ge) or a 2N2955 and Q2 a 2N3055 (si) or an AY8149 (si). The circuit in A has the advantage that the power transistor can be bolted without an insulator to the chassis that serves as a heatsink.



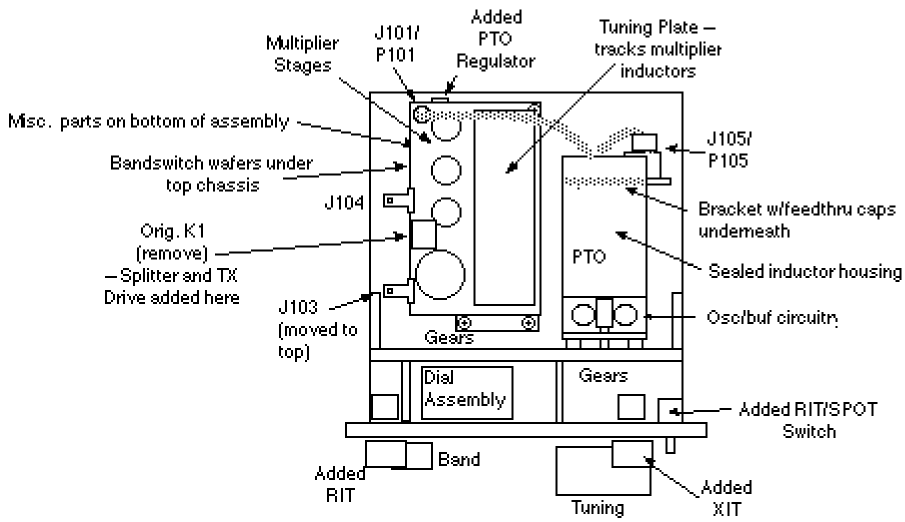
TRANSISTORIZING THE T-368 EXCITER 9/14/95

An experimenter's notes... but not every detail

This circuit is adaptable to the R390A PTO as the frequency range is so similar

This project uses the existing tuned circuits, physical structure, and some parts and wiring of the exciter from the old Collins military T-368 transmitter as the basis of a solid-state VFO/multiplier for most of the HF region of interest to amateurs.

As supplied, it has a 1.5-3 MHz PTO, a mechanical digital readout, and doubler stages to cover 1.5-3, 3-6, 6-12, and 12-24 MHz (they say 12-20, but the dial reads to 24; probably reflects the ratings of the companion PA). It measures about 10" wide by 9" high by 11" deep, so it's not small, and as I received it there was no enclosure for the whole assembly -- no problem for less critical applications like HF, but probably best to add shielding (I haven't yet for HF QRP!). It's built like a tank, and the tuned circuits in the multipliers track the PTO in typically elegant Collins fashion.

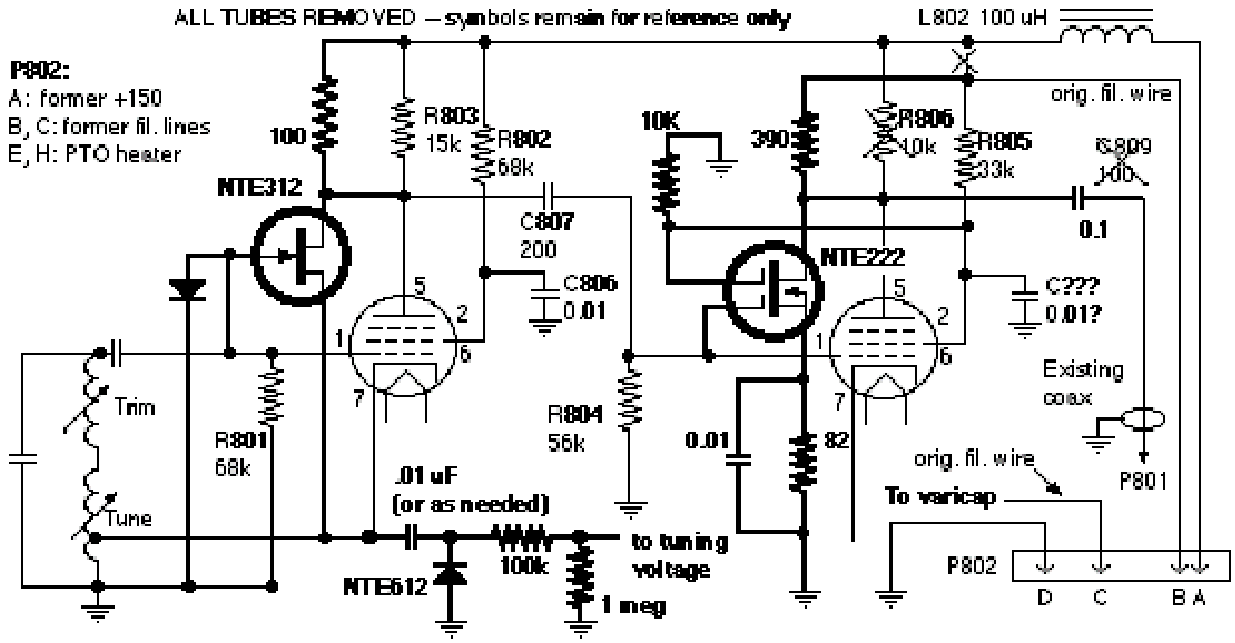


I got it from Fair Radio Sales, a renowned surplus dealer, about 5 years ago without the output tube (the 6000, a rare beast nowadays). It cost $35 then; one of my correspondents informs me that they now run $45 without the output tube. I first used it as an improvement over a Heathkit HG-10 VFO with a DX-60 transmitter, and it was remarkably stable even in its tube form.

Once transistorized, you get an incredibly stable VFO for the home station that covers a wide range, complete with receiver and transmitter offset pots on the front panel. It's not a unit for backpacking, because of its solid, heavy construction. But that weight and bulk is part of what assures mechanical stability. The sheer size of the sealed PTO assembly, plus the fact that the components were designed for the higher RF currents of the tube circuit, means that once you solid state the thing, you have rock-like stability. There's even a thermostat-controlled heater you can hook up around the PTO tuned circuits. True, the unit tunes pretty fast on the higher bands, but the gearing is so good that there is no backlash. The offset pots can help you zero in if your hand isn't too steady.

My sources for circuit concepts and building blocks:

ARRL Handbook	QRP Classics (ARRL)
W1FB's Design Notebook (ARRL)	W1FB's QRP Notebook (ARRL)
Solid-State Design for the Radio Amateur (ARRL)	Peter Anderson, KC1HR, "Transistorizing Surplus VFOs", QST, Feb. 1989, page 45-46 (in *Hints and Kinks)*



T-368 manual pages and schematic relating to exciter. On ordering, I asked only for the pages from the T-368 manual related to the exciter, and Fair Radio quoted me $10 at the time.

THE PTO

Removing the PTO takes some care -- you have to loosen not only the mounting screws, but the hex set screws on the PTO shaft gear. I suggest you set things at 3 MHz at the top end of the band, so you know where to reset things. All the electrical connections unplug, and all these wires are re-used, also the feedthrough capacitors attached to an angled bracket under the PTO. All the circuitry outside the tank circuit is on a small sub-chassis under the tube sockets.

The original approach - transistorize existing circuit

I started with the QST article, which suggests using 2 40673 dual-gate MOSFETs. (The NTE222 is a currently-available substitute for the discontinued 40673 -- but they're about $5 each locally!). Read this for some precautions about turning the PTO too far beyond its range, and damaging the internal slugs. I think it will go more than 50 kHz beyond, so you have some slop for recalibration. Also, the author warns not to open the PTO sealed inductor can if you can avoid it -- the hermetic seal is one key to super-stability.

Having used various VFO circuits, I decided to save a few bucks, and used a plastic VHF JFET for the oscillator (NTE312 or other similar device). I took a 100 ohm resistor from regulated V+ (after the existing RF choke) to drain; hooked drain to the original plate pin; hooked the gate to the original tube grid connection; and hooked the source to the original cathode connection. A 1N914 diode goes from gate to ground (cathode grounded). That author had put the diode from gate 1 to source, and this meant no oscillation for me without some kind of "kick", like temporarily grounding the gate! When the diode went to ground like all other oscillator circuits I've seen, all is well.
The tuning diode circuit provides an adjustment range for RIT/offset. These values gave me a tuning range of 7 kHz at the 3 MHz end of the PTO when I tested it before re-installation.
However, I didn't discover till I put things together that the tuning range at the bottom end (near the 3.5 and 7 MHz ham bands) was substantially lower! At 80 meters, I have barely enough to get an 800 Hz CW offset. At 40 meters, I even have only about +- 2 kHz.

I later put in an NTE618 tuning diode, designed for AM radio tuning service (440 pF at 1 volt!) Of course, I reduced the size of the .01 uF capacitor until the tuning range was reasonable. I think it was about 100 pF, can't remember :-). Tuning range is about +/- 2 kHz at 1.75 MHz, multiplying to +/- 16 kHz at the highest band.. My 10-turn pots make even the widest range quite manageable. The stability is SOMEWHAT degraded by the tuning diode arrangement, but I still got less than 10 Hz drift overnight relative to WWV at 10 MHz.

I used a regulated 6 volt bus for the oscillator (and RIT bias -- see below), coming from a 3-terminal regulator mounted on the back of the multiplier. My regulator was an out-of-tolerance 7805; I'll bet 5 volts from a good 7805 would be fine.

The buffer is almost identical to the QST article above, except for output coupling and operation from a separate power bus. Here the original plate pin serves only as a tie point, the 10K plate resistor and 100 pF coupling capacitor being removed. Handle the MOSFET with all due electrostatic precautions -- wrist strap, etc.

4 wires that enter the PTO circuit from the power connector are rewired as shown; the other two are for the VFO oven heater. You should not find this last option necessary, unless you are operating in the Arctic! At the feedthrough capacitors under the PTO, you have convenient points to solder your connections to the rest of the assembly.

Re-installation of the PTO IS a bit tricky because of the split gear that eliminates backlash. You have to turn the one gear so the springs are compressed, hold them in place with a screwdriver until you get them meshed with the gearing from the dial.

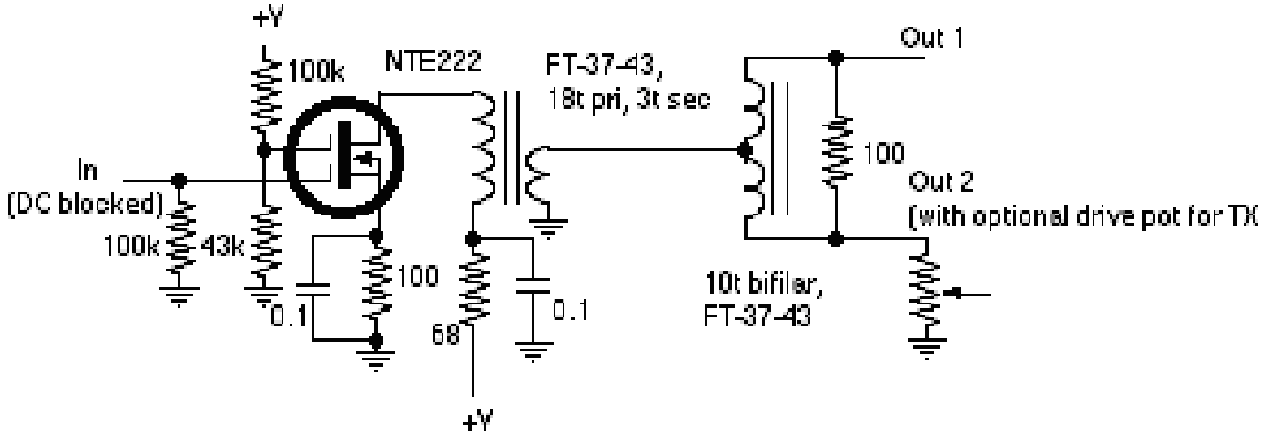
Later...
I just ripped out all the existing parts (in the tube compartment) and built the oscillator/buffer more cleanly, according to the homebrew books.. The tuned circuit components to the left, plus the coupling capacitor, are the ones in the sealed can, and three pins come into the tube compartment for ground, cathode/source, and grid/gate. Don't know exactly why I did this, other than to have less clutter in there.

MULTIPLIER STAGES
I was on my own on the multiplier stages. It was possible to get some results by simply plugging in JFETs in the appropriate places, but little anomalies would creep in from the long wiring, old-style "Vitamin Q" bypass/coupling capacitors, etc. As I went on, I became less and less concerned with historic preservation, and more and more inclined to gut the unnecessary parts.

On top of the chassis, rip out the relay K101, intended to switch between this exciter and an external one for FSK. Use the two BNC connectors on angled brackets -- move both of them to the top, with room for the power splitter /drive level pot.

Obviously JFETs are the best choice, to provide high impedance loads to the tuned circuits that were designed for tube grids. I used the NTE312 again.

FIRST VERSION: I just wired the gate to the grid pin of each stage (with a 47K resistor to ground), the drain to the plate pin, and the source to chassis ground, clipping the existing grid resistors to ground. Actually, the first stage has its gate (and associated resistor) going directly to the BNC connector from the VFO (J101).



The plate supply wire was used, with its decoupling networks (1K series resistor and .01 uF capacitor to ground) . 1K, of course, is way too high for the transistor circuit, so I bridged something in the range of 68 to 100 ohms across it. They are on the bottom of the multiplier assembly -- three Phillips screws are accessible with a long screwdriver. The bandswitch and tuning shafts couple to the front panel with slotted plates to allow some free play, so getting them back in takes some coordination. I got lots of practice by taking it in and out many times!

The first stage works as a buffer on the low band, and a doubler above that. The low band switches in a load resistor instead of the tuned circuit -- R106, 10K (on the bottom again!). This obviously needs to be lower. I plugged in different resistors (in the 3300 ohm range, can't remember) across this resistor, until the output at S101, pin 8 was the same on bands 1 and 2.

The second multiplier stage had a tendency to self-oscillate. I cured it with a series RC network from gate to ground, in parallel with the gate resistor: 2200 ohms and 0.1 microfarad. This obviously loads down the tank circuit enough to reduce the Q. A resistor directly across the appropriate tuned circuit might work as well; the series R-C method was simply used because it was easiest to apply without taking the assembly off a million more times. Another such network went on the input of the third multiplier when I discovered another oscillation on the highest band.

SECOND VERSION: Further reflection led me to dislike the sheer length of the unshielded bandswitch wiring in the original circuit. So, the little terminal board on the bottom was completely ripped out, and replaced with a piece of double-sided PC board. Everything was basically re-wired as cleanly as possible near the new PC board ground plane, with excursions to the bandswitch being as direct as possible. The dog-bone ceramic coupling capacitors between tuned transformer stages were saved and re-used for "historical" reasons -- with some wire and insulating tubing to extend them, they were placed directly between the appropriate tuning cans, between the new PC board and the metal chassis. Also saved were the 15 pF capacitors that had been placed near the tube grids, which are part of the overall tuning capacitance, apparently being placed nearer the tube grids to aid in bypassing?

The final tuned circuits were also used in this second rewiring. The lowest band tuned circuit was used as a tuned load for the first buffer/multiplier on the lowest band, and the output attenuated with resistors to match the output of the other bands. The others were incorporated into the circuit by means of 20 pF coupling capacitors. The slugs do not make much difference in the upper 3 bands, but overall waveform purity seems to be better with these coupled in, so you might as well use them. All of this fed into a broadband step-down transformer, wound on an FT-37-73 core, approximately 15 to 5 turns ratio (have to look again). This yielded about 1 volt peak to peak output.

OUTPUT AMPLIFIER

FIRST VERSION: In the area of the IPA (intermediate power amplifier, another word for "driver") tube, I built a broadband buffer stage based on W7ZOI's buffer in the VFO section of the "high performance communications receiver" project in the ARRL Handbook. Any broadband buffer circuit with a hi-Z input and a high output into 50 ohms would be suitable. The input comes right off S101, section 2, rear, pin 12.

The splitter transformer requires a 25 ohm input. When built separately, the splitter would be preceded by a 2:1 impedance transformer (approx. 4:3 turns ratio autotransformer). With an output transformer already present, just take some turns off the secondary (originally 5 turns).
I got output of about 3V peak to peak into 50 ohms on the lower 3 bands, and 2V p/p on the highest. Pure sine waves are not a concern for me, since diode ring mixers generate harmonics anyway, and even would rather have a square wave. If it is for you, you can put in a low pass filter for each band, using the S101 section 1 rear (rewired) to pick them.

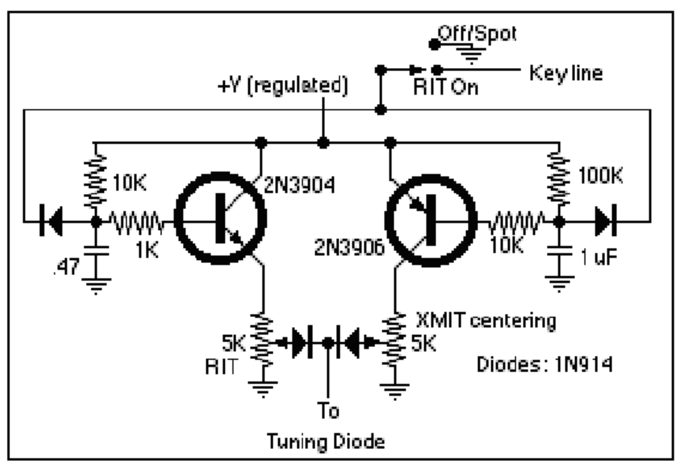
SECOND VERSION: I used a broadband output amplifier based on a 2N3866 (from DeMaw's "W1FB's Design Notebook"), with adjustable gain (pot in place of the emitter resistor). This gave me about 4 volts p-p on the low end, about 3 volts at the upper limit. This is enough to accommodate any combination of splitters, high-level diode mixers, or what have you.

FOR FUTURE EXPERIMENTING: Consider also some CMOS chips as output amps. A recent QST article featured 74HCxxx chips to put out nice clean 5 volt P-P square waves. There is no better driving signal for diode ring mixers.

FINAL ALIGNMENT: Make the best compromise you can between uniformity of output in a given band, and waveform purity (least feedthrough of sub-harmonic signals). This will take some head-scratching and fiddling, but you can do it.

OFFSET TUNING -- TRANSMIT AND RECEIVE

I have two pots (found matching knobs at a local surplus house!) and a switch added to the front panel, and a transistor switching arrangement (built on the pot terminals) to select the correct pot when the transmitter is keyed. I used 2000 ohm 10-turn pots -- you can use probably use anything from 2 to 10K. You can fiddle with resistors on either ends of the pot(s) to limit tuning to the most linear part of the varactor's range if you wish.



The capacitors hold the VFO at the transmit frequency until the CW waveform has decayed to zero. Your mileage may differ as far as the values needed.

Remember that the diodes affect temperature stability -- as the temperature goes up, the voltage drop goes down, reducing the output voltage and lowering the frequency. If you're lucky, this compensates for the temperature characteristics of the varactor diode! For all I know, the transistors behave similarly. Thus you might best place them near the VFO in a small enclosure. I've gotten by with being sloppier, and the thing is still more stable than you can imagine.

The SPDT switch (front panel) attaches the key lines either to the system keying line, or directly to ground. When grounded, you RX and TX on the same frequency for SSB or for spotting a CW signal to zero beat. When in the other position, you have receive offset for a CW beat note, or for RIT in SSB or CW.

John Seboldt K0JD
Milwaukee, WI (but "forever a zero", having moved out of zero-land for the first time in 1999)
k0jd at seboldt dot net

THE AUTHOR:
John Seboldt, K0JD, began hamming as WN0QXG in about 1967. Music, electronics, and ham radio grew side by side in his youth, leading to work in the broadcast industry while studying music at Luther College, Decorah, IA, and The University of Iowa, Iowa City. Church music has been his main field -- he served 15 years in the Twin Cities, and moved to Milwaukee in 1999. (Check out samples of his music work at www.seboldt.net/annunciation). For now, technology has again claimed his working hours: at Time Warner Cable he's a Broadband Technician, having worked in the cable industry since 2001.

From: Garey Barrell <k4oah@mindspring.com> Date: Mon, 29 Jul 2002 14:13:05 -0400
Subject: {Collins} Re: Lankford-Chambers SSB Mod For R-390A

The "Chambers-Lankford AGC/BFO Mod" was published in Issue #27 of the "Hollow State Newsletter" in the Fall of 1991. Apparently it was intended as a compromise modification to allow reasonable CW / SSB reception with the standard diode detector, by greatly increasing the BFO injection level. Recognition was given to the Capt. Paul H Lee Product Detector modification from CQ Magazine, but this mod was deemed unsatisfactory primarily because it required modification of the R390A mainframe. Also the Noise Limiter was bypassed, and there was a regenerative effect in the audio response. Hubbell "cured" with the addition of a small relay in his article in, "Improving the R390A Product Detector" Ham Radio Magazine.

Another article, "Capt Lee's Product Detector for the R-390/R-390A - Revisited" by Paolo Viappiani appeared in Issues #40 and #41 of the same publication. He claimed to have resolved the regeneration problem by proper (single-point) grounding of the shielded cables added and restored the Noise Limiter function by the addition of "a small negative bias voltage to the Product Detector Output."

73, Garey - K4OAH Atlanta

To: <collins@listserve.com> From: "Jack Quinn" <k0hehjdq@mindspring.com>
Subject: {Collins} R390A Lankford SSB Mod

Hi All:
You might want to go to KK4DF web site where he has a list of production Mods, Field changes and other mods including the Lankford SSB mod which is as follows (uses 2 diodes and a cap.)

-one diode in paralled with R547, cathode as follows
---]<--- pin 2 V506A

-one diode in parallel with R546, cathode as follows
--->[--- pin1 V509A

-one 47 pf cap in parallel with C535

I have the whole print out of this mod and discussion from the R390 group if any would like to have a copy.

73's Jack, K0HEH

From: "Drew Papanek" <drewmaster813@hotmail.com>
To: r-390@mailman.qth.net

Hello All,

My vote for SSB reception is the 2 diode (Lankford) modification for its simplicity. This method works quite well. A couple of points:
1. Make sure that AGC capacitors are not leaky, especially the large oil filled unit atop IF deck (This should not be a problem for Bruce as his receiver was done by Rick Mish) 2. When increasing BFO injection capacitance remember to adjust BFO neutralization in AGC IF amp.

The 2 diode modification gives good SSB reception even without altering BFO injection. Done this way and with AGC capacitors verified leak-free (by Dr. Jerry's method) I have not found any change to reception of AM and CW signals strong or weak.

Capt. Paul Lee's 6BE6 product detector modification has been mentioned. The National NC-300 uses a 6BE6 as a product detector giving good results; this should perform satisfactorily in the R-390A (although the NC-300's final IF is 85 KHz vs R-390A at 455 KHz). Lee's R-390A mod calls for alteration of mainframe wiring harness to run shielded audio cables to a multisection BFO switch. This accounts for some of the "invasiveness" of the mod. I would suggest leaving mainframe wiring and BFO switch intact and use a small relay in IF module for audio switching. The relay would be powered by B+ switched to the BFO-turned-product detector when in SSB/CW mode. A quick perusal of Mouser's catalog shows a 24V relay with coil power consumption of 70 mW. A series dropping resistor to operate this relay from the (approx.) 180V B+ line would dissipate only 450 mW. In this fashion modification would be limited to IF module only.

Drew

From: "Jim DiMauro" <jfd@warwick.net> Subject: [R-390] SSB Conversion Photos

Hi All:

For those interested, I shot a couple of photos of the SSB conversion on my Collins R-390A. Here are the links:

http://webusers.warwick.net/~u1016524/r390afront.JPG
http://webusers.warwick.net/~u1016524/r390assbadapt.JPG

You can see on the front panel that the Mode switch positions are "LO-UP-AM-BFO" and the dial lock control is now a "VFO Vernier," which provides kind of a fine tuning of the SSB signal, although in my opinion it's not really necessary, since SSB signals are easily tuned in with the main tuning knob.

The subchassis on the underside of the radio supports a 6U8 and associated circuitry. Near as I can tell from the photographs it looks identical to the rare factory SSB conversion described on r390a.com, but since my radio is a Collins-made unit from the first contract in 1954, it's almost certainly a later addition.

So there you have it.

73, Jim WA2MER

From: Llgpt1@aol.com Date: Sat, 19 Jul 2003 21:19:11 EDT
Subject: Re: [R-390] SSB Conversion Photos

That is basically the same conversion as pictured in Paolo Viappiani's R-390 - R-390A Handbook. The Schematic is printed in it also, schematic is in English, noted as an EAC conversion.

From: "Bernie Nicholson" <vk2abn@batemansbay.com> Date: Thu, 24 Jun 2004 22:14:50 +1000
Subject: [R-390] Product det

I have sent the scanned HAM RADIO article to 35 respondents so far (prod det.pdf) ,

The circuit works well but I haven't tried the noise limiter mod, also I have used a 12AU7 double triode prod det circuit and it works even better. I mounted it in the 3TF7 socket and substituted 12BA6 tubes for VFO & BFO this also works well I used a subminiature relay to switch the audio lines it had a 9K coil and I put 22K in series to HT When u switch off the bfo the circuit reverts to the original for receiving AM , I had a few spare IF modules so I can unmodify it pretty easily, Kind Regards To every one

From: MURPH <rickmurphy1001@earthlink.net> Date: Tue, 14 Jan 2003 18:49:40 -0500

Has anyone done the W3JHR Captain John Lee modification i.e. - Changing the bfo to a product detector using a 6BE6 ? It's suppose to fix all the SSB ailments and the AGC problem is addressed by adding a 1.0 mfd cap from pin 10 to ground on the AGC switch S107. January 68 CQ magazine.

Richard

From: "Walter Wilson" <wewilson@knology.net> Subject: Re: [R-390] R390a - S.S.B.

I've worked on a unit that had that modification done. The unit I worked on had problems with the relay that must be added to the IF deck to make this thing work properly. There are several versions of the Capt. Lee mod, so look for the later ones if you plan to do it. It certainly falls into the category of not-easily-reversible modifications, which I tend to avoid.

Walter Wilson - KK4DF http://r-390a.us

From: "Drew Papanek" <drewmaster813@hotmail.com> Date: Wed, 15 Jan 2003 17:23:43 -0500

Hello all,

I have found the 2 diode (Lankford) modification to be extremely effective for SSB reception on the R-390A. The addition of these diodes is not an alteration to the detector as some might suspect. Instead, diodes are used to shunt 2 AGC time constant determining resistors. The result is a short negative going time constant (fast attack) and the original positive going time constant (slow decay). Fast attack enables AGC to capture signal peaks and slow decay "remembers" (for a short time) those peaks so that gain is appropriate for future peaks. Result is a tolerable signal to BFO ratio at the detector (minimizes distortion). That time constant alteration accounts for most of the mod's effectiveness.

The Lankford modification also increases BFO coupling cap for more injection, and adds capacitance to AGC line. The mod works well without these capacitor changes.

For good AGC action on all modes using medium and slow AGC switch settings verify that the 2 uF oil filled AGC cap is not electrically leaky.

On replacing leaded caps/resistors in those crowded spaces: Trying to undo the connection from a tube socket pin or other terminal often does damage.
Cutting out old component leaving a stub at the terminal and splicing in new component using small copper tubing sleeves was a good suggestion.

Alternatively, the new component's leads can be coiled using a piece of the old component's lead as a form, and these coils slipped over the aforementioned stubs and soldered. This works especially well if new component has smaller diameter leads than old. Such would be the case when using 1/4 watt carbon film or metal film resistors to replace old 1/2 watt units (calculate dissipation to determine suitability of replacement).

Unless they're cooked, leave those 2.2K decoupling resistors alone. As Bob pointed out, the variation in voltage drop across these (due to drifting out of tolerance) is miniscule compared to changes brought about by line voltage variation and normal gm tolerance range of tubes. Of more importance would be screen and cathode resistor values. For IF stages after "mechanism a philharmonic" (that funky fractured translation of "mechanical filter"), resistor values become less critical. This is because that section of the IF chain has much more gain than needed and we reduce it anyway by tweaking IF gain pot for best S+N/N ratio.

Drew

From: "bernie nicholson" <vk2abn@bigpond.net.au> Date: Wed, 29 Jun 2005 10:30:01 +1000
Subject: [R-390] ssb converters

The Collins SSB converter is a BEHEMOTH, it has twin motor driven BFO s driven by phase discriminators, FORESTS of Tubes, and I have seen smaller power transformers in 2KW linears the unit also weighs MORE than a 390 ,BUT they do work.

I have repaired lots of converters for people in OZ and the best and simplest unit is the TECHNICAL Material Corp MSR 591, with this unit and using the 2 KHZ filter in the 390A the Signal to noise ratio is the best I have heard on any receiver EVER.

my personal Bias is towards the IF module conversion as per HAM RADIO Magazine, BUT using a 12AU7 double triode in the Ballast tube hole as a product detector,

I also have removed the 2 KHZ & 4 khz filters from a module and substituted dedicated asymmetrical USB & LSB filters and Crystal controlled the bfo with a 455.00 khz crystal from the 1khz filter can. I use a High resistance relay to switch the Audio line either through the product detector or for AM it reverts to the original circuit using the supply to the BFO, so only the module needs to be modified and the unit can be transferred from one RX to another

[I have 3 receivers the 390a type and quite a few spare modules] my favourite RX is this 390 a with SELECTABLE Sideband , the agc has also been modded to give attack and decay times more compatible with ssb I use the Slow switch position for this ,

Hope this info is of interest Regards to ALL

High-gain Product Detector

"Fig. 6 shows the circuit as a high-gain product detector. This circuit has good dynamic range and a microvolt input will deliver about a millivolt output. Notice that this circuit is identical to fig. 4 (Mosfet Mixer Stage) except for the details of the output circuit. The output transformer is a regular Radio Shack item. The 500-ohm output is convenient if you go into a low-pass audio filter built from 88-mH toroids, as it provides a good impedance match.

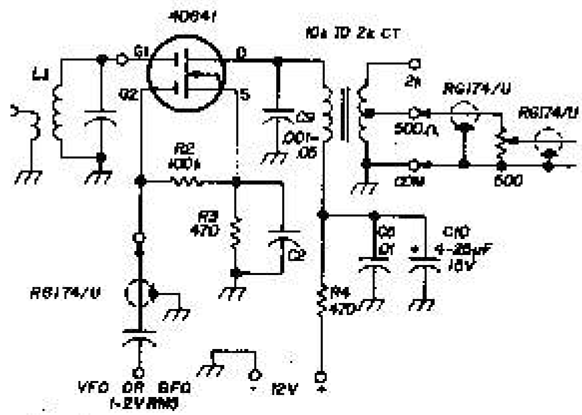


fig. 6. High-gain product detector. Except for the output this circuit is identical to fig. 4.

Notice the volume control hookup. If the volume control is located remotely it is not desirable to ground the low side of the control to the chassis at the remote point.

This can lead to ground loops. Instead, carry the ground with the shield of the miniature coax, and from the control to the next stage on the shield of that same piece of coax. This ground loop problem is very easy to control in a modular setup like this. You can mount the boards on pieces of bakelite or plastic when necessary and have full control of the ground situation without any radical rebuild.

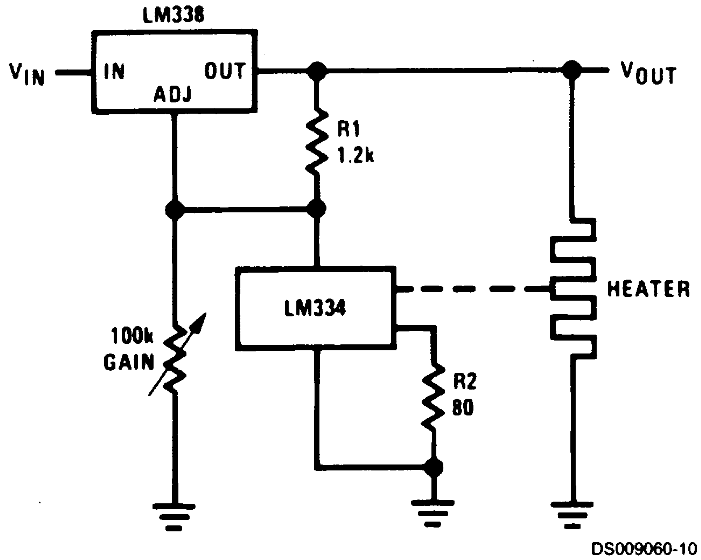
Notice that the output is bypassed for both rf and audio (C5 and C10). Also notice C9. This capacitor can roll off the high-frequency audio components. That little transformer is quite efficient out to 70 khz or more, and there is no need for response above about 2.5 khz. In fact, if you use a .05 mfd capacitor at C9, you will have a definite peak around 1 khz that is helpful for CW work, but still results in useable ssb performance. Again, the bfo voltage is brought in with RG-174/U coax. With a high-gain IF strip, it may be helpful to locate the bfo well away from the strip - or shield it.

...Slightly more gain will result if C2 is paralleled with an audio bypass capacitor,...

The circuit in Fig. 6 can also be used as a synchrodyne detector. Tune L1 to the operating frequency, add an audio filter to the output and use a high-gain audio amplifier. You now have a simple, direct-conversion receiver."

Last Update: 05/12/97 Web Author: David White

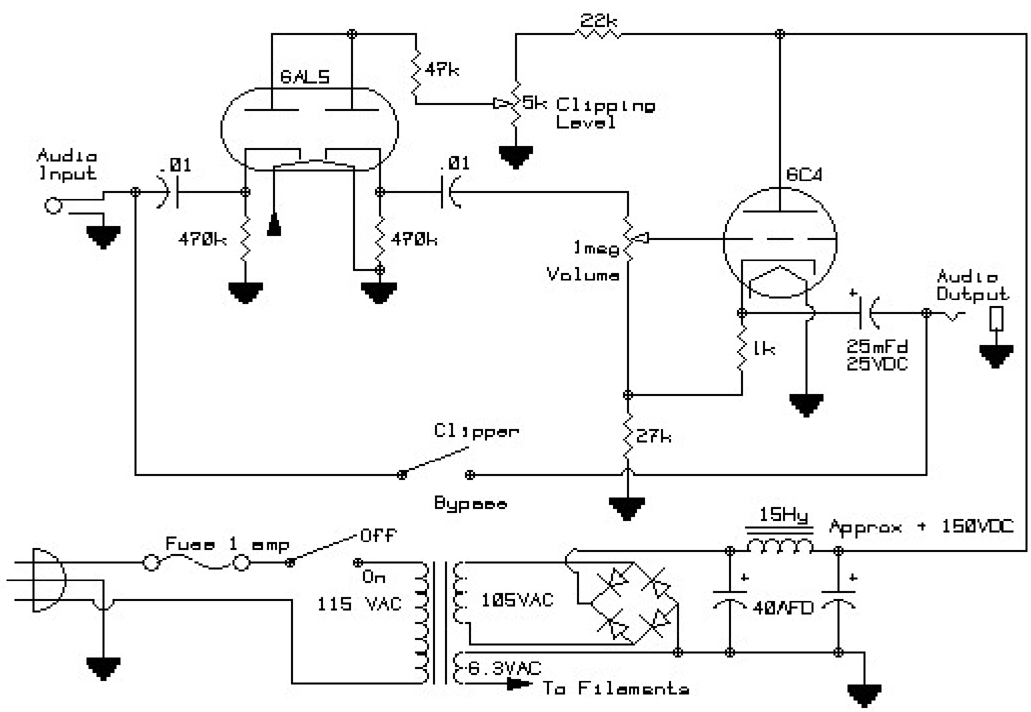
Temperature Controller



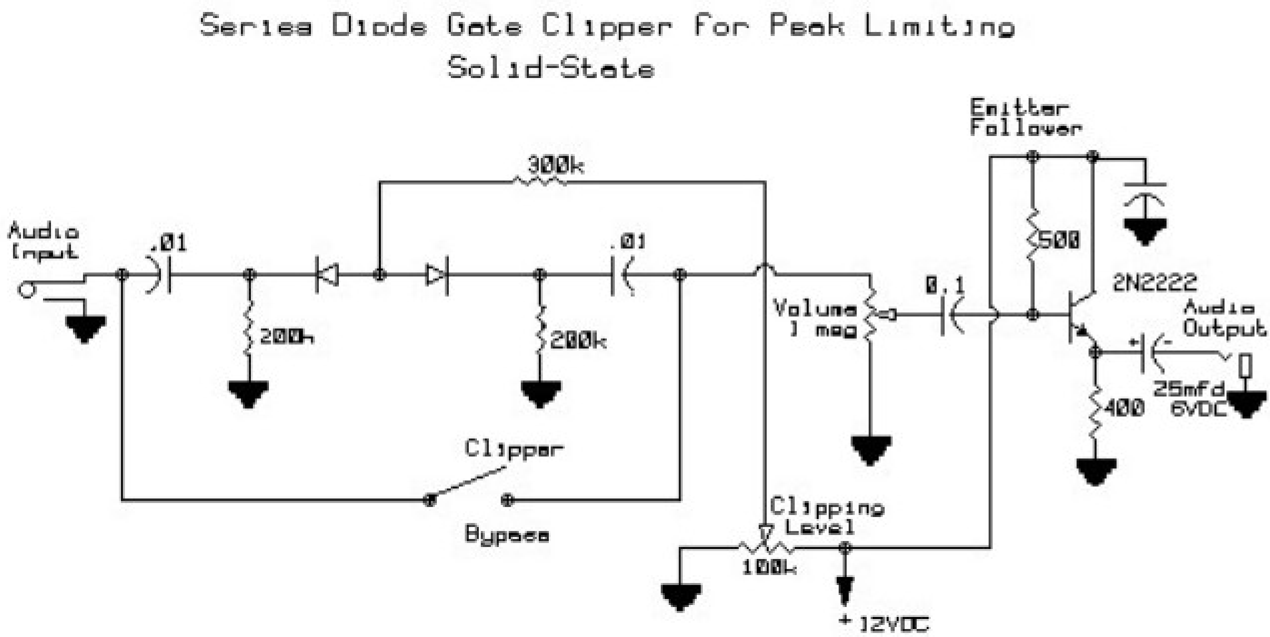
An Excellent Audio Clipper

Here are two versions of the best clipper circuit I could find in a short search of the radio handbooks I have on my, somewhat limited, book shelf. I was trying to find a fairly simple clipper circuit I could use ahead of a good audio filter to help eliminate static crashes, and the key clicks from my own QSK rigs, and to provide some of the advantages of AVC for those receivers lacking one that operates on CW. They are supposed to be very effective and are fully adjustable. One must balance three controls in order to get the best use out of these: the receiver audio gain control, the clipping level, and the clipper audio gain. Although an oscilloscope is not essential, it would make it a bit easier. First you set the clipper audio gain control at about center with clipping level at zero. Then you adjust your receiver audio gain for a comfortable level. Then adjust the clipping level on a signal you find to be at that comfortable level so that there is no apparent change in audio quality. Lastly, adjust the clipper audio gain for your chosen level. What SHOULD happen is that all signals appear to be at the same level, and there should be no apparent distortion. Either clipper has a high input impedance, 470k in the tube version, and 200 k in the SS version. Output impedance in both cases is "low", I think about 1k or less. Any clipper should be followed by a good audio filter of some sort, whether hollow or sand-state. I prefer a filter such as the MFJ or Vectronics sand-state filter because they are so small and they work like gang-busters.

Series Diode Gate Clipper for Peak Limiting



This is the tube version of the Series Diode Gate Clipper. This circuit was taken in part from the "Clipper/Filter" which is shown in the 1961 ARRL Handbook, minus the filter part. The first 470 k resistor shown above, connected to the first cathode of the dual-diode, is shown as a 47 k resistor in the Handbook article. This is a mis-print and should be 470 k. I came to this conclusion after reading a one-and-a-half page article in Bill Orr's ***Radio Handbook***, 19th edition, under ***Special Circuits***, on how the similar solid-state circuit shown below works. Experimentation will tell if my conclusion is correct. Any low or medium mu triode will work well as the cathode-follower. 1/2 of a 6SL7, 12AU7, or 1G6, or a 30 works just fine. If you use a 30, you will want to use a dual-diode with a similar filament requirement and change the power-supply accordingly. Likewise for the 1G6



.

Here is the sand-state version of the clipper shown above.

The emitter-follower is one I borrowed from Doug Demaw's ***Solid-State Design for the Radio Amateur.*** I have not yet tried either circuit, but hope to soon.
I hope you find one of these useful.
Ken Gordon W7EKB

The "Select-o-ject"

The Selectoject is a receiver adjunct that can be used as a sharp amplifier or as a single-frequency rejection filter. The frequency of operation may be set to any point in the audio range by turning a single knob. The degree of selectivity (or depth of the null) is continuously adjustable and is independent of tuning. In phone work, the rejection notch can be used to reduce or eliminate a heterodyne. In C.W. reception, interfering signals may be rejected or, alternatively, the desired signal may be picked out and amplified. The Selectoject may also be operated as a low-distortion variable-frequency audio oscillator suitable for amplifier frequency-response measurements, modulation tests, and the like, by advancing the "selectivity" control far enough in the selective-amplifier condition. The Selectoject is connected in a receiver between the detector and the first audio stage. Its power requirements are 4 ma. at 150 volts and 6.3 volts at 0.6 ampere. For proper operation, the 150 volts should be obtained from across a VR-150 or from a supply with an output capacity of at least 20 mA.

The wiring diagram of the Selectoject is shown in Fig. 5-45.Resistors R2 and R3, and R4 and R5, can be within 10 percent of the nominal value but they should be as close to each other as possible. An ohmmeter is quite satisfactory for doing the matching. One-watt resistors are used because the larger ratings are usually more stable over a long period of time.

If the station receiver has an "accessory socket" on it, the cable of the Selectoject can be made up to match the connections to the socket, and the numbers will not necessarily match those shown in Fig. 5-45. The lead between the second detector and the receiver gain control should be broken and run in shielded leads to the two pins of the socket corresponding to those on the plug marked "A.F. Input" and "A.F. Output." If the receiver has a VR-150 included in it for voltage stabilization there will be no problem in getting the plate voltage - otherwise a suitable voltage divider should be incorporated in the receiver, with a 20- to 40-mfd. electrolytic capacitor connected from the + 150-volt tap to ground.

In operation, overload of the receiver or the Selectoject should be avoided, or all of the possible selectivity may not be realized.

The Selectoject is useful as a means for obtaining much of the performance of a "Q Mulitplier" for a receiver lacking one.

C1 -- 0.01-µf. mica, 400 volts. C2,C3 -- 0.l-µf. paper, 200 volts C4, C8 -- 0.002-µf. paper, 400 volts. C5 -- 0.05-µf. paper, 400 volts. C6 -- 16-µf. 150-volt electrolytic. C7 -- 0.0002-µf, mica. R1 -- 1 megohm, ½ watt. R2, R3 --l000 ohms, 1 watt, matched as closely as possible. (see text). R4, R5 -- 2000 0hms, 1 watt, matched as closely as possible. (see text).	**R6 -- 20,000 ohms. ½ watt. R7 -- 2000 ohms, ½ watt. R8 -- 10,000 ohms, 1 watt. R9 -- 6000 ohms, ½ watt. R10 -- 20,000 ohms. ½ watt. R11 -- 0.5-megohm, ½ watt, potentiometer (selectivity). R12 -- Ganged 5-megohm potentiometers (tuning control). (IRC PQ11-141 with IRC M11-141.) R13 -- 0.12 megohm, ½ watt. S1, S2 -- D.p.d.t. toggle (can be ganged).**

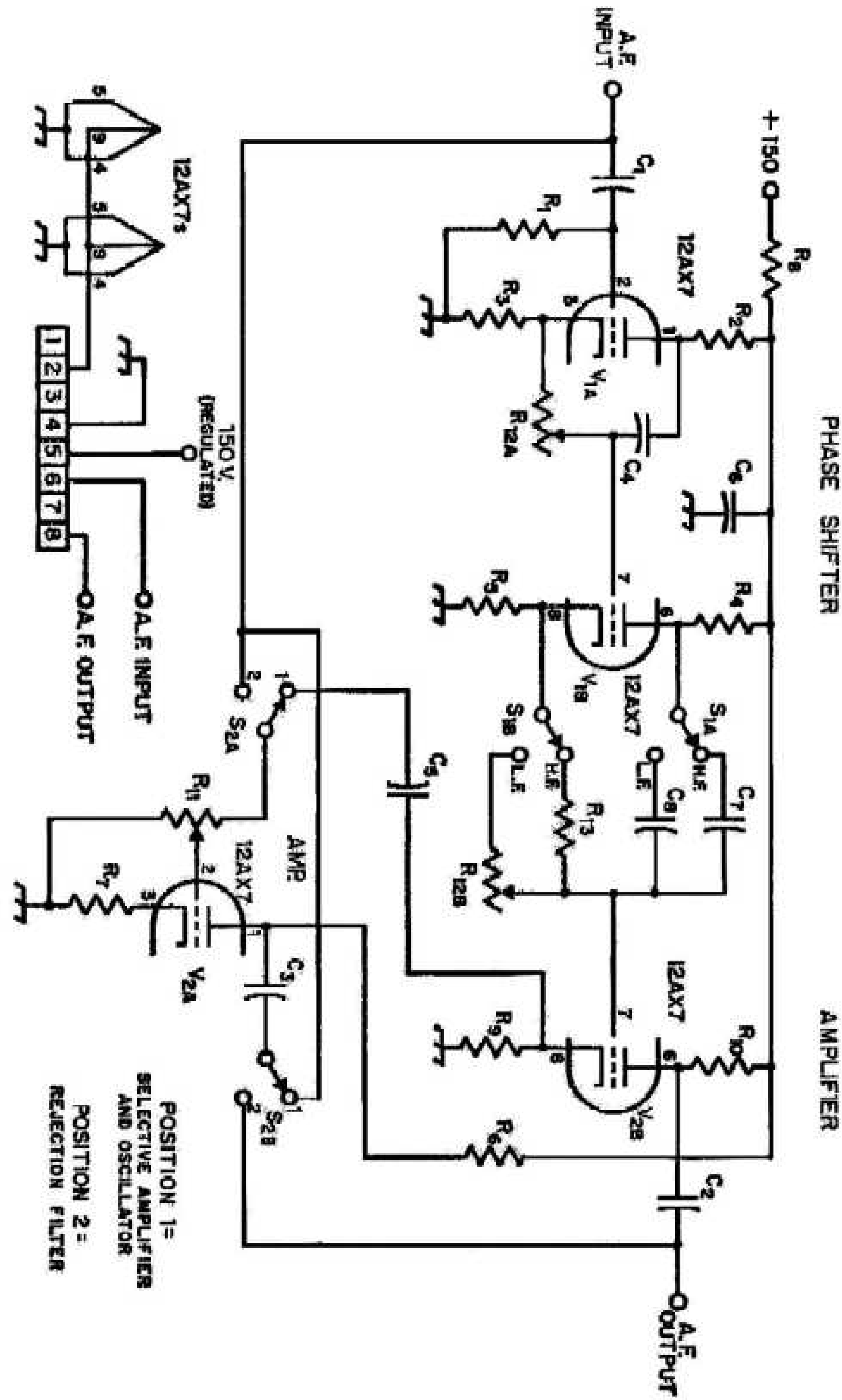


Fig. 5-45 -- Complete schematic of Slectoject using 12AX7 tubes.

A Medium Wave Audio Processor

by Bruce Carter, Texas Instruments Applications

Introduction

AM radio has long been the domain of talk radio and sports broadcasts. Many people seldom change to "the other band" on their radio unless it is to hear their favorite team play, to hear a talk show, or to get the news. Before FM became widely available, AM was the primary radio band used for music. Throughout much of the third world, this is still the case. Vast distances between cities make it impractical to use FM as a broadcast band for small, isolated groups of listeners

Many attempts have been made through the years to revive the AM band. Most recently, the FCC extended the band to 1700 kHz in an attempt to alleviate some of the nighttime clutter. In the late 1970's and early 1980's, the FCC attempted to revive interest in the band by approving several standards for stereo broadcasting. AM stereo never caught on with the public, due to the reluctance on the part of the commission to select a standard, combined with the growing popularity of talk radio and a proliferation of FM music stations.

Ironically - the rebirth of the AM band is happening due to a seeming unrelated event, the passage of the Telecommunications Act of 1996. This act has led to wholesale consolidation of radio station ownership. The large radio corporations program a bland mixture of rock, country, and talk stations in every city, and have disenfranchised many niche format fans. Niche formats are forced off the air, to streaming internet audio, or onto the AM band.

As more and more specialized music ends up on the AM band, there is renewed demand for improved AM receivers. Unfortunately, there are some disadvantages to AM:

High frequency interference is a problem at night. AM frequency response is not limited (unlike FM, which is severely rolled off above 15 kHz). Even though AM stations are allocated on 10 kHz channels, the FCC does not require that the audio frequency response be limited to less than 10 kHz. High frequency response can extend to 20 kHz or even beyond. At night, however, distant stations will be present on adjacent frequencies, and audio from those stations will mix with the high frequency audio, producing noise.

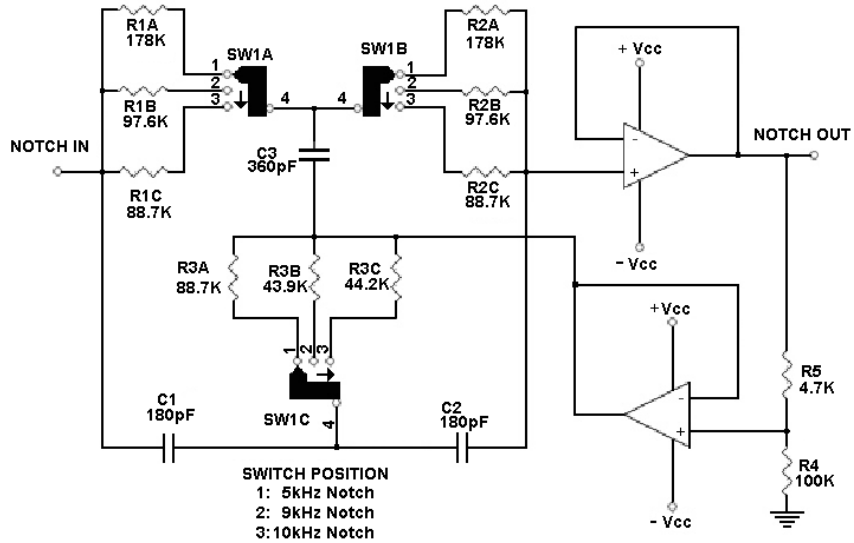
The 10 kHz spacing of AM stations causes another problem. The channel spacing is close enough that the carrier frequencies from adjacent channels are audible. These carriers produce audible tones in the audio - 10 kHz in North America, and 9 kHz in Europe and much of the rest of the world. Because AM propagates much farther at night, the problem becomes worse. In almost every case, there will be a carrier below and above the desired station. Not only will both carriers contribute to a 10 kHz whistle, but they will be at slightly different frequencies. The FCC allows AM stations to deviate (20 Hz from their assigned frequency, so the two carriers will mix together, creating an even more annoying warbling 10 kHz whistle tone.

When the carriers from adjacent stations mentioned above mix together, they create sum and difference frequencies. The sum frequencies are above the range of human hearing, and not of concern. The difference frequencies, however, are a concern. Each individual carrier can be (20 Hz from the assigned frequency. This means that their difference frequency can vary from 0 to 40 Hz (If one is 20 Hz low and the other is 20 Hz high). This difference frequency will show up in the received audio, and may be quite annoying through a subwoofer.

The Motorola C-Quam system has become the worldwide standard for AM / Medium Wave stereo transmission. Unfortunately for stations that broadcast music, C-Quam utilizes a 25 Hz pilot tone in the audio from the phase of the carrier that carries the difference (L-R) modulation. This 25 Hz pilot, unlike the pilot used for FM stereo, is not used in the demultiplexing process. It is only there to identify the station as an AM stereo station (to light the stereo light).

Whistle Filter

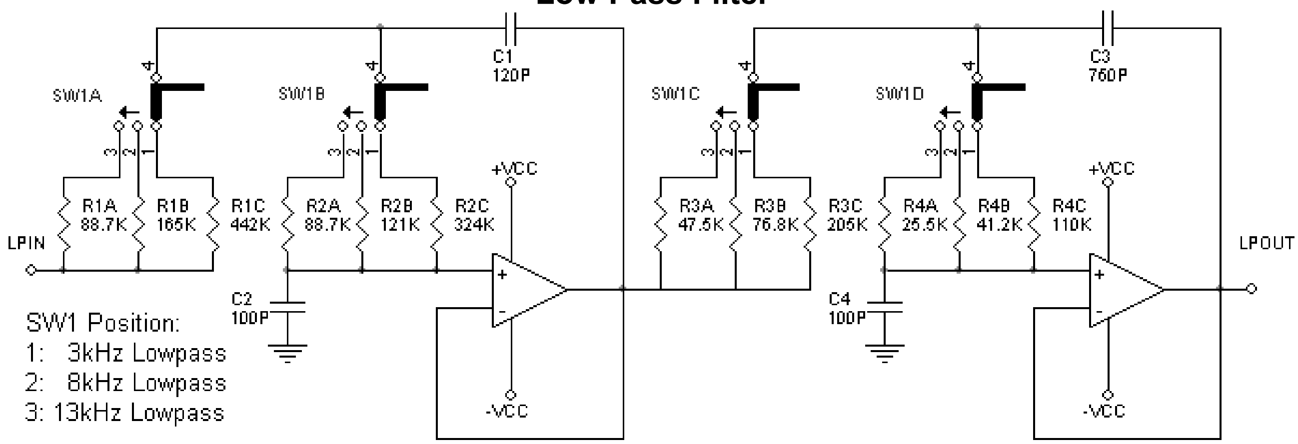
Because it is low in frequency, it may take longer for the pilot detection circuitry to light the light than it does for the decode circuitry to operate - so the listener is likely to hear stereo separation before the light turns on. The stereo pilot will appear in both channels of the received audio. Unfortunately, this stereo pilot will make low bass unlistenable on AM Stereo stations. The pilot will also beat with any carrier difference frequency, creating even more low-frequency tones.



The schematic above is the notch filter section of the AM signal processor. It is extremely difficult to form a high Q notch filter at high audio frequencies. The Twin T notch configuration was chosen because it is unity gain, and relatively easy to tune. All passive components should be 1% except for R11 through R14, which can be 5%. The depth of the notch depends to a large extent on the matching of the components. C3 should be formed from placing two 120 pF in parallel. Care should be taken that these capacitors should be taken from the same batch. Similarly, R1 and R4 should be from the same batch, as should R2 and R5, and R3 and R6.

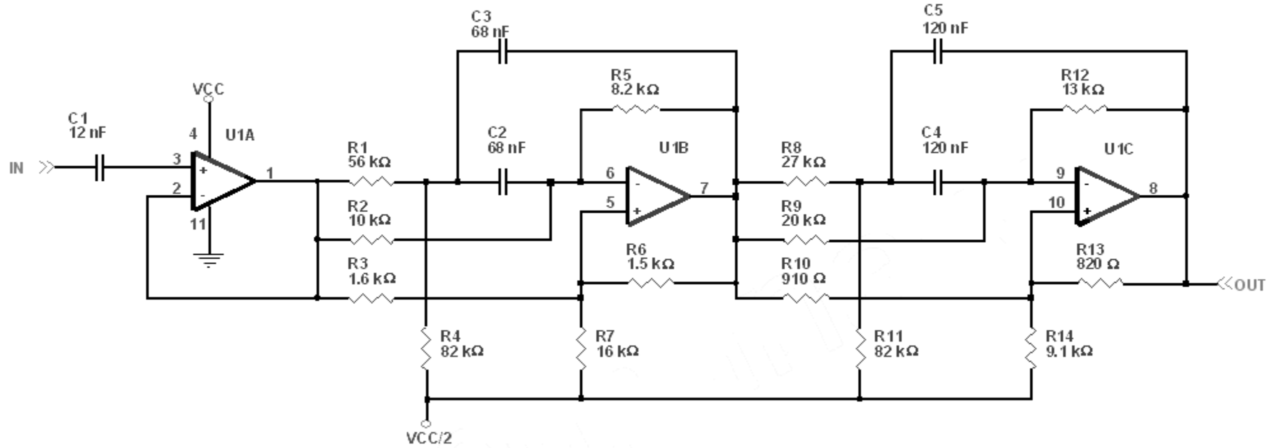
There are areas of the world where both 10kHz and 9kHz stations are receivable. In these areas, virtually any frequency from 1 kHz to 20 kHz is possible. It would be difficult to design a filter to reject all of these frequencies. In the unlikely event that a particular frequency is objectionable, this design can be modified to reject it: the center frequency of the notch is 1/2pR1C1. R2 = R1, C2 = C1. C3 is twice C1, R3 is one half R1. If there are more than one tone in the audio, two or more notch stages can be cascaded to eliminate them. Sections of the filter can be eliminated. A North American listener can eliminate the 9 kHz notch. The 5 kHz notch can be eliminated for listeners that do not care about shortwave.

Low Pass Filter



Each op-amp is configured as a Sallen-Key low pass filter. Stage 1 has a Bessel characteristic, and stage 2 has a Chebyshev characteristic. The combination of the two form a Butterworth filter with a roll-off of 0.5 dB at the indicated frequencies. The three frequencies selected - 13 kHz, 8 kHz, and 3 kHz are selected to match the settings on a Hammarlund SP-600JX receiver - which the author uses for Medium wave and shortwave listening.

Rumble (High Pass) Filter

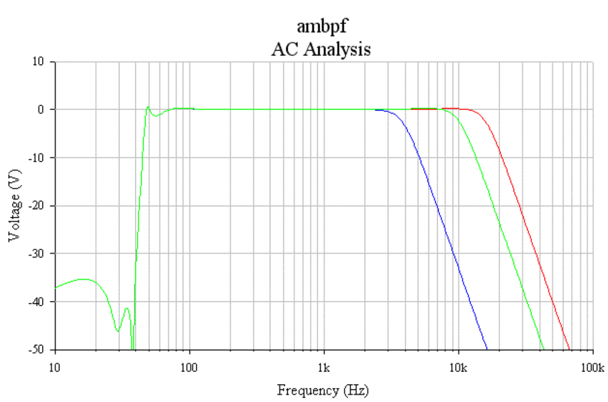


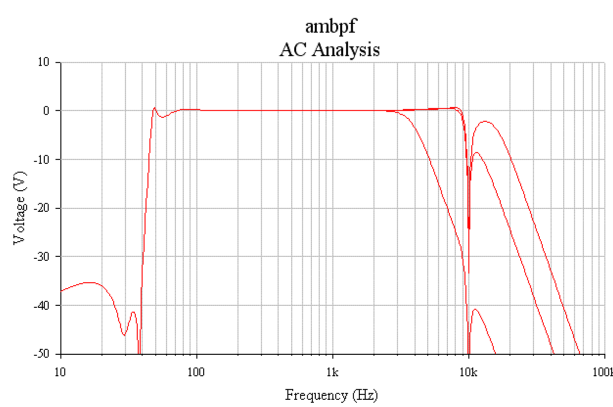
The rumble filter above has been designed to reject audio below 50 Hz. It is a 5 pole elliptical filter with no more than a couple of dB ripple. Since the ripple occurs in the low bass end of the response, it should be inaudible.

Response

This audio processor has several switches to enable and disable functions as needed. The response below has the rumble filter enabled, with the 13 kHz, 8 kHz, and 3 kHz low pass filter progressively each separately enabled. The "flat" position is not shown. The notch filter is disabled in the first response curve, and enabled in the second. 9 kHz and 5 kHz notch response is not shown, but will produce similar results.

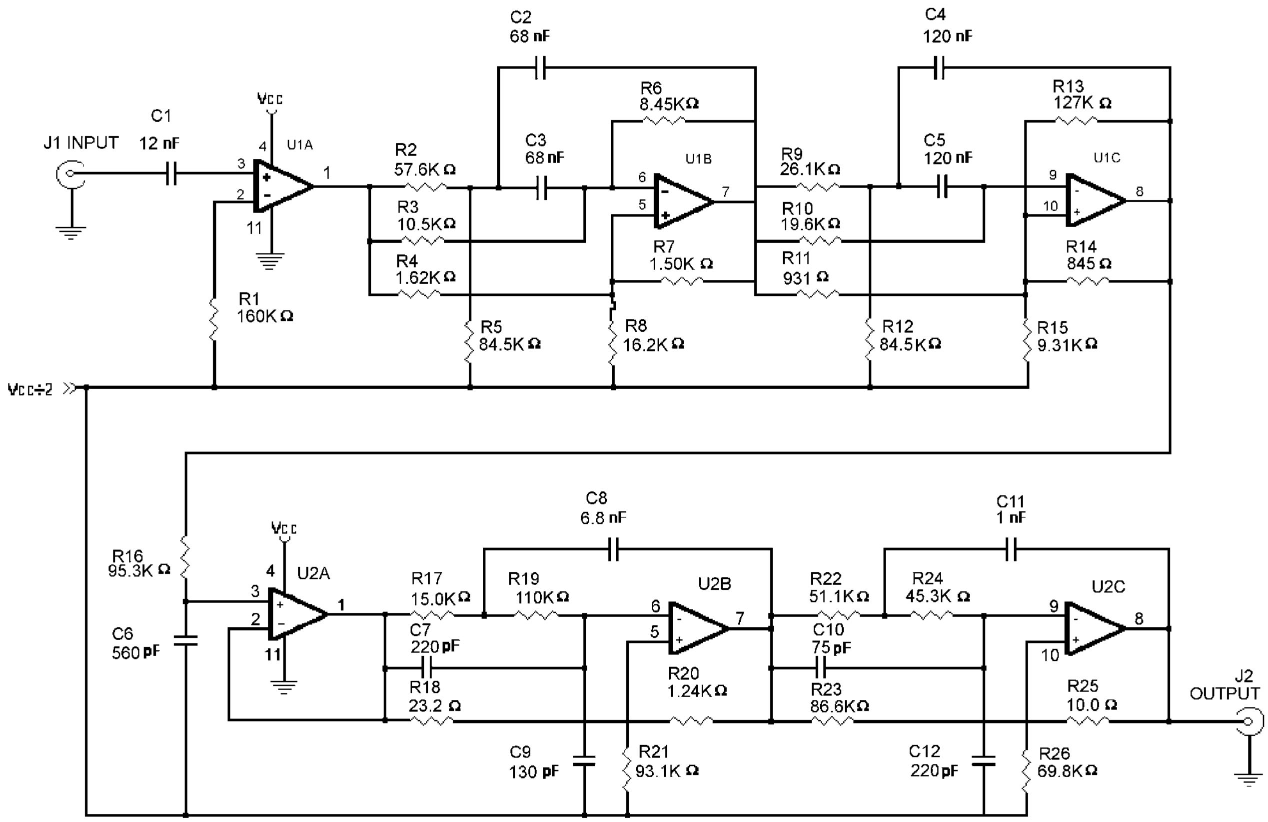
The action of the notch makes the 13 and 8 kHz settings of the low pass filter almost identical, except for a small band of high frequencies above 10 kHz. The 10 kHz notch is not needed for the 3 kHz low pass filter. Even without it, 10 kHz is attenuated more than 30 dB.



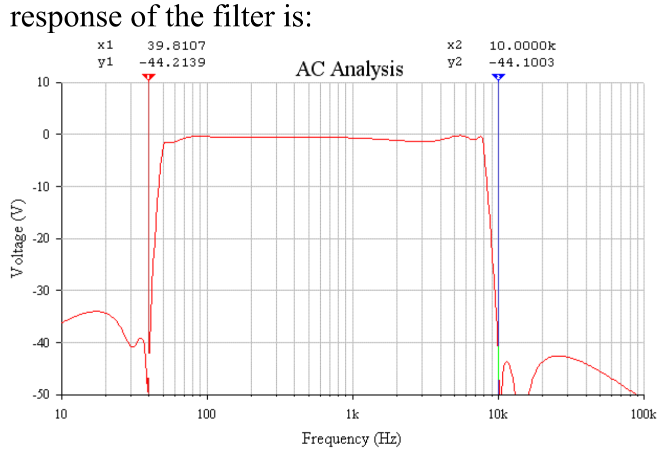


A Simpler Alternative

There is an alternative to the audio processor above. All of the undesirable audio components can be eliminated in a single filter, if high frequency audio (above 10 kHz) can be sacrificed. The audio processor shown below is a 5 pole 50 Hz high pass elliptical filter with 0.5 dB ripple, followed by a 5 pole 8 kHz elliptical low pass filter with 1 dB of ripple. It has over 40 dB of rejection below 40 Hz and above 10 kHz.



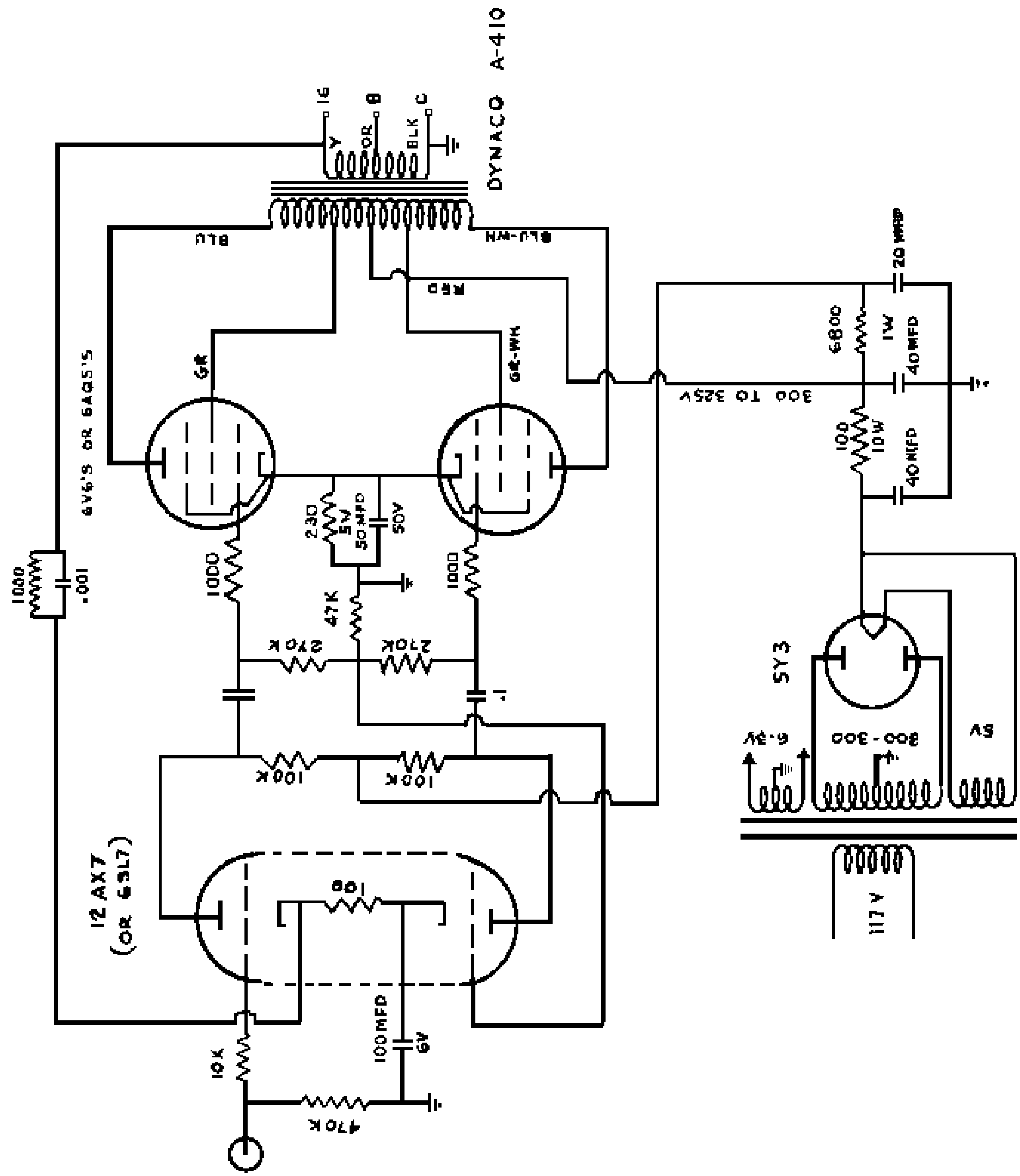
Two of these filters should be constructed for AM stereo. If monaural reception is all that is desired, only one filter is required. Component values are critical. There is no way to convert to 5% resistor values, 1% are absolutely required. If 5% are used, filter characteristics will degrade badly. The response of the filter is:



The high pass filter will remove the pilot difference frequencies as well as the stereo pilot, while allowing bass frequencies of 50 Hz and above to pass unaltered. Because FM radio also rolls off below 50 Hz, the listener should not notice any difference between AM and FM in bass response. High frequency noise and the 10 kHz pilot tones are removed by the low pass filter. This filter also limits the high frequency response of the AM audio to 8 kHz, so the listener may notice a difference compared to FM.

Because high frequency noise only occurs on distant stations in the daytime - the listener should be able to listen to local AM stations with the filter section switched out the circuit.
The circuit above is the rest of the circuit, showing the power supply, the creation of Vcc/2 (U1D), and the bypass switch. If monaural reception is all that is desired, only one set of input / output connectors are needed, and SW2 can be a SPDT.

Dynaco Low Power (6V6/6AQ5) Amp



LOW POWER HIGH QUALITY CIRCUIT WHICH CAN BE USED TO IMPROVE THE QUALITY OF COMMERCIAL AMPLIFIERS

Improving Your Receiver With Frame-Grid R.F. Pentode

BY JOEL BALOGH W3CFA *QST* February 1966

THE many requests for information on the use of the EF 183/6EH7 variable-mu frame-grid pentode mentioned by the author as an r.f. amplifier in an earlier *QST* article[1] have forced the following conclusion: there still exists that breed of radio amateur who would rather rebuild his present equipment than trade it in on a new model to obtain the best possible performance. This article is dedicated to that breed.

A knowledgeable reader of this article may correctly point out that atmospheric and manmade noise levels, will normally exceed the low noise levels in the author's receivers. But that is how it should bet A quick method of checking receiver performance is to' disconnect the antenna and note what happens to the audio-output noise level. It should drop when the r.f. gain is wide open and the receiver is set at its upper frequency limit. It should also be possible to peak up the antenna noise with the antenna trimmer control. If these two checks are marginal, then pull out the r.f. amplifier tube to cause a drop in noise. If the noise stilt doesn't drop you know that the mixer or converter stage is contributing more noise than the r.f. ,amplifier or the atmospheric and man-made noise, levels coming in from the antenna. Improving the receiver is now up to you.

Many articles, on receiver improvements have appeared in the amateur magazines and sections in the ARRL *Handbook* on this subject, so no attempt will be made to give credit to all reference material used in this article.

Some credit is due the old HQ-l20X receiver which has served as a guinea pig during the past decade of rebuilding and experimentation. This particular receiver (a Navy version of the HQ-l20X went through the Second World War and was salvaged from a junk heap around 1951. The first attempts to improve it, were started when W1NXY[2] discussed some changes in the postwar HQ-120 receiver.

This consisted of changing the mixer stage to obtain lower noise operation. Revisions in the r.f. stage were Already underway when W5UZ[3] discussed this and other areas of the same receiver. Not all of the work through the years on the guinea pig resulted in improvements -- the cathode-coupled triodes and cascaded triodes were tried out as r.f. amplifiers were rejected for reasons discussed later on.

A 6BZ6 pentode was used for several years as a r.f.. stage, but it was unstable above 20 Mc. The present 6EH7 has been used as an amplifier in the guinea pig since early 1962, and it will continue in that role until a better tube is invented. This article not only discusses how to use the 6EH7 to obtain better performance, but it also discusses the principles used by the author to evaluate new tubes as they are introduced on the market.

The following list presents most of the factors which should be kept in mind when planning receiver front-end improvements:

1) R.f. stages which were originally designed to operate with remote-cutoff pentodes should be rebuilt with better remote-cutoff pentodes and not sharp-cutoff pentodes or any triodes.

2) The transconductance of the new pentode should be higher than that of the old pentode.

3) The cathode current of the new tube should not exceed 20 milli-amperes,

4) The grid-to-plate capacitance of the new pentode should be equal to, or less than, that of the old pentode.

5) The sum of the pentode's shotnoise and partition noise should be as small as possible.

6) The dynamic plate resistance of the new pentode should not be less than .5 megohm.

7) The cost and availability of the new tube must be such that the average receiver can readily be modified.

8) The heater voltage, must match that of the old tube.

The factors given above are not all independent, so a compromise must usually be reached when considering an assortment of tubes. Various methods of weighing the importance of these factors are discussed in the following sections of this article.

Pentode versus Triode

A number of articles have appeared on the subject of receiver improvement using triodes to replace pentodes, but it is this author's experience and opinion that the resulting loss in r.f. gain and selectivity do not justify the reduction in tube noise. The loss in gain and r.f. selectivity is a result of the low dynamic plate resistance of the triode which swamps the *Q* of the resonate r.f. plate circuit. These plate circuits were originally designed to yield the desired *Q* when operated with the high plate resistance of pentodes. The dynamic plate resistance is parallel with the plate circuit when considering the equivalent circuit of the tube, plus the plate circuit. For this reason a plate resistance which is below .5 megohm is not desirable.

High. Transconductance

If there had to be just one criterion for eva1uating receiver pentodes, it would be for higher transconductance. This primary characteristic determines the obtainable gain as well as the shot noise and partition noise. But in any practical application, higher gain can be utilized only if the stage is stable, and this is primarily a function of the tube's grid-to-plate capacitance. Thus, higher transconductance can be handled only if the C_{gp} is about equal to that of the old tube. Also, in practical applications, higher gain presupposes the end result of detecting weaker r.f. signals, but the minimum detectable signal is determined by the combined atmosphere, man-made and tube noise present at the front end of the receiver. We cannot do anything in the receiver to change the atmospheric and man-made noise levels. The tube noise of a pentode is usually considered to be the sum of the shot noise find the partition noise. The shot noise is reduced by higher values of transconductance. The partition noise is reduced by a, combination of higher transconductance and a smaller ratio of screen current. to cathode current. Some sharp-cutoff pentodes are on the market (the 7788 is one example) which yield low shot and partition noise by having both high transconductance and a low ratio of screen-to-cathode currents.

But such tubes, even if a remote-cutoff version were available, would not be usable for improving a communications receiver because the total cathode current (45 ma. for the 7788) might cause heating in the Litz wire of the coils in the receiver. For this reason, an upper limit on the total cathode current was set at 20 ma. The price of the 7788 would also be a disadvantage for this particular tube.

Evaluating New Tubes

The data on the new tubes which are introduced on the market, seldom include specific mention of shot and partition noise, so recourse must be taken to compute this information from the data which are furnished. Shot and partition noise is usually spoken of in terms of the equivalent resistors which would give rise to the observed noise voltages. This analogy arises from the observable noise voltage which is present across the terminals of any resistor due to the random motion of the electrons which are present in the resistance material. This has the classical name of Johnson Noise. The equations which are in use for computing the values of these equivalent resistors are usually approximations instead of precise equations. This makes the computation easier. But it also result in a variety of approximation equations. The author here prefers to use two different approximations, one from the *MIT* reference [4] and the other from the *Radiotron* reference[5]. The results of both equations are then used to establish a ball-park figure for the equivalent noise resistance of the tube in question. The equations appear below:

$$\text{(MIT)}\ R_{eq} = R_{SHOT} + R_{PARTITION} = \frac{2.5}{gm} + \frac{20 I_{SG}}{g_m\ I_K}\ \text{ohms}$$

$$\text{(Radiotron)}\ R_{eq} = \frac{I_p}{I_K}\quad \frac{2.5}{gm} + \qquad \text{ohms}$$

where gm = transconductance in mhos
I_K = cathode current *in* a11peres
I_p = plate current in amperes
I_{sg} = screen grid current in amperes

A selection of both sharp and remote-cutoff pentodes is listed in Table 1 to show the results of the two equations. Keep in mind that a high gain (high- Transconductance) pentode is desired which has the lowest possible equivalent noise resistance. The other columns in Table 1 are discussed elsewhere. The information on the sharp-cutoff tubes is provided as reference material for use when selecting a low-noise mixer stage. The reader can observe that the 6EH7 has the lowest computed equivalent noise resistance of the remote-cutoff pentodes listed in Table 1.

Ratio of *Transconductance -to-*C_{cg}

As mentioned previously, a high-transconductance pentode will have merit as a r.f. amplifier only if the grid-to-plate capacitance is low enough to give stable operation. The 6EH7 does not have the lowest. C_{cgs}, as shown in Table 1, but it does have the highest g_m. The obvious way to get a relative comparison between r.f. pentodes, is to look at their ratios of g_m/C_{gp}. This is tabulated in one of the columns of Table 1. Again, the 6EH7 shows up as the best tube when the criterion is for the highest *gm -to-* C_{gp} ratio.

Ratio of gm/C_{gp}-to-R_{eq}

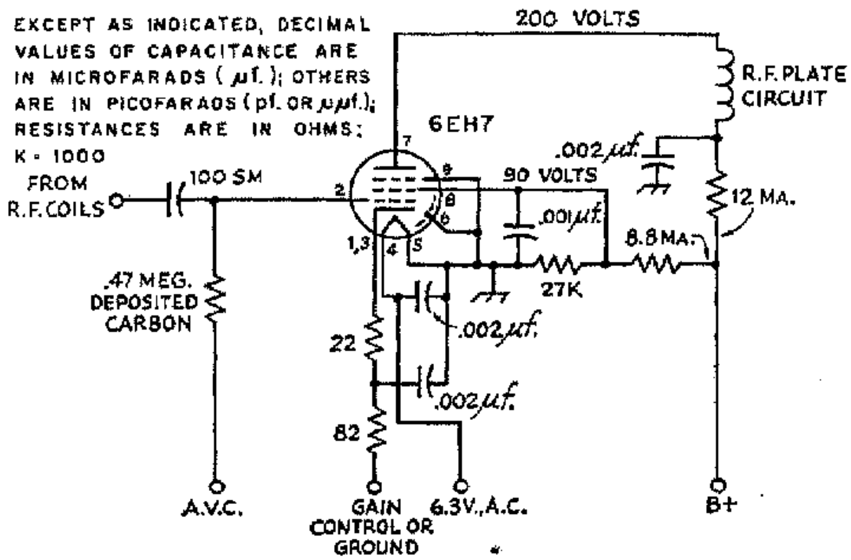
For the case where a new tube may not have the lowest R_{eq}. but does show the highest gm/C_{gp} ratio (or vice versa), a relative comparison based on the ratio of gm/C_{gp}-to-R_{eq}. would be helpful. This latter ratio is tabulated in the last column of Table 1. The lowest value of the two approximated R_{eq}'s was used for computing this ratio. Again, the 6EH7 is the best available tube for use as a gain-controlled r.f. amplifier when the criterion is for the highest gm/C_{gp}-to-R_{eq} ratio.

Selecting a *Mixer Tube*
With a few exceptions, most of the comments on selecting and. tube apply to the task of selecting a mixer tube. Since the plate and grid circuits are not tuned to the same frequency, feedback by way of C_{gp} should not produce instability. Some mixers are desired which have a broad-band response (such as in converters), so a lower value of dynamic plate resistance can be tolerated. This lower value cannot be tolerated, though, if the mixer is in a communications type receiver where high *Q* i.f. transformers follow the mixer stage. The 6EJ7 sharp-cutoff pentode (Table 1) is a good candidate for mixer service. The author uses four of them in this role, in four different receiving systems. The 6EJ7s have also been put to use as Lf. amplifier's in an f.m. tuner and a pre-i.f. noise-silencer.

A R.F. Amplifier Circuit
The 6EH7 has been used as a r.f. amplifier by the author in the HQ-120X receiver as well as an Eddystone 888A ham-band receiver and an RME DB-20 pre-selector.

The circuit shown in Fig. 1 represents an r.f. amplifier which can be adapted for use in any receiver covering all, or parts, of the 0.55-Mc. to 30-Mc. range.

The unspecified plate and screen-dropping resistors must be selected on the basis of the available supply voltages and the required plate and screen circuit currents.



For the 6EH7 in the author's earlier article, the series screen-grid resistor should be 39K ohms when the B+ is 250 volts. There should *not* be a resistor from screen grid to ground as shown in the article. (QST June '66)

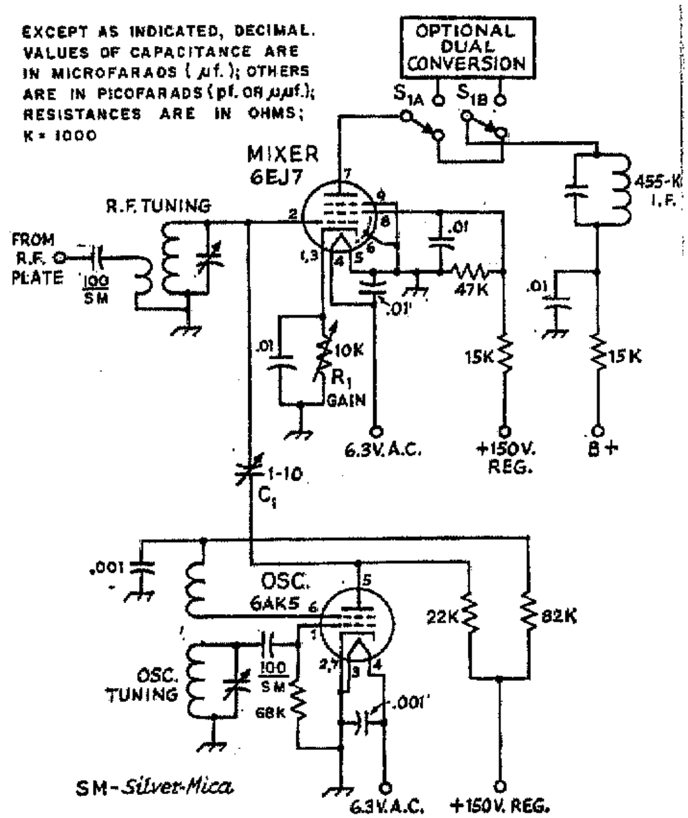
Please note the specified values of cathode and screen bypass capacitors. These values were selected to form series-resonant circuits where the inductance is in the form of the capacitor leads. Do not use larger values of bypass capacitors unless the amplifier is for use only on lower frequencies. A 0.01mfd bypass can be used at 7 Mc. and lower, while a 0.1 mfd bypass can be used at 2-Mc. and lower. The unbypassed 22-ohm cathode resistor is used to compensate for variations in input capacitance and resistance which otherwise would occur when cathode or grid voltages are changed. Pin 6 is used to ground the tube's internal shield. An external shield is also recommended. The heat-dissipating style such as IERC's TR.-6-6020B will assure longer tube life. Something not shown in Fig. 1, but which is always used by the author, is a shield partition which straddles the tube socket. A piece of 1/32-inch sheet brass, which is about 3 inches square, is installed to pass between Pins 1 and 9 and between Pins 5 and 6. Pins 5, 6, and 9, the center post of the socket, and the grounded ends of the bypass capacitors, are then soldered to this brass plate.

Of course, the r.f. circuits will have to be realigned after the new tube is installed.

A Mixer Circuit Using the 6EJ7 Pentode

The mixer circuit which was used in the HQ120X appears in Fig. 2. A separate local oscillator using a 6AK5 was installed on a sub-chassis underneath the main chassis. The 9-pin socket for the 6EJ7 was mounted on an adaptor plate which replaced the old 8-pin socket. The variable cathode resistor and the trimmer capacitor between the oscillator plate and the mixer grid are adjusted so that the mixer has high gain while remaining stable over the entire frequency range. Too much oscillator injection or too little mixer bias will produce "birdies." A triode local oscillator should *not* be used if pulling of the oscillator frequency is to be avoided on the higher frequency bands.

This is the mixer circuit John Leary used in his SP 600's for the 2nd mixer with the exception of making C_1 a fixed value of 10pf. (He didn't use the 6AK5 circuit .)



Note: On some receivers John used a variable pot that could be adjusted from the from panel. It is unclear whether an adjustable pot was located by the mixer tube and adjusted for optimal performance as stated in this article.

Dual Conversion for the HQ-120X Receiver

The image rejection of this receiver was not very good when operating above 10 Mc. because of its relatively low i.f. of 455 ke. A subchassis Was installed under the main chassis of the HQ-120X which provided dual-conversion capabilities above 10 Me. A three pole, double throw wafer switch was mounted behind the front panel with its shaft coming out just to the left of the sensitivity control and below the send-receive switch.

The schematic of this dual-conversion unit is shown in Fig. 3. It uses a 6U8A triode-pentode, with the triode serving as a crystal-controlled oscillator and the pentode as the mixer. The frequency of this crystal can be between 1855 kc. and 2055 kc. The 1965-kc. unit was obtained from one of the surplus crystal companies. The cathode resistor of the pentode might have to be adjusted a little to obtain stable mixer operation. Of course, the first local-oscillator circuits had to be realigned when the dual-conversion feature was incorporated in the receiver. The fixed-padder capacitors which were in series with the oscillator ooi1s, for the two bands above 10 Mc., were replaced with Arco 307, 350-1180 pf., variable padder capacitors. A signal generator which provided marker signals every 1 Mc., as well as a variable-frequency signal, was useful for the realignment work.

Changing the HQ-120X over to dual conversion entailed a lot of work, but the elimination of the image problem was well worth it. The elimination of the images was especially rewarding when using this receiver as a tunable i.f. in the 14-Mc. to 18-Me. range, which is the output frequency of the 2-meter converter used by the author.

The noise present at the image frequency tended to degrade the performance of the converter by approximately 3 db. when the 455-kc. i.f. was used at 14 Mc.

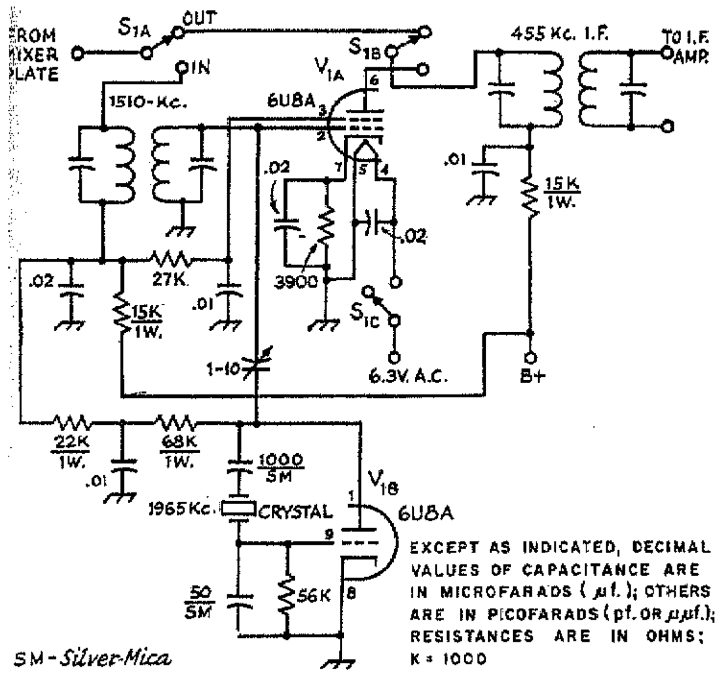


fig. 3-Circuit diagram of the dual-conversion unit for the HQ-120X receiver.

Actual Results

Some reasonable questions come up every time the subject of receiver improvements is discussed. How can I check my present receiver's performance? What improvements can be obtained with these frame-grid tubes? The first question was answered in the introduction to this article.

The noise figure which can be expected with the 6EHF at 30 Me. is approximately 5 db. The improved HQ-120X was able to copy a c.w. 0.03-microvolt signal from a 50-ohm calibrated signal generator at 30 Me. *before* the dual conversion was installed, This 0.03 microvolt level was where the c.w. signal dropped into the noise.

The narrow-band crystal filter was in use during the c.w. test. An a.m. signal, modulated 40 percent with 400 c.p.s., dropped into the noise at .06 microvolts without a crystal filter. The S-meter circuit in the receiver was rebuilt and calibrated for use at 15 Me. with the 2-meter converter.

Table II shows the a.m. test-signal levels present at the antenna jack of the HQ120X from a 50-ohm generator operating at 15 Mc., which produced the various S-meter readings. These measurements were obtained *before* the dual conversion was installed.

The Eddystone 888A ham-band receiver, modified by the author, was checked with the calibrated signal generator on c.w. at 29.5 Me. The signal dropped into the noise at 0.01 microvolts. The Eddystone has dual conversion, a l-kc. i.f. bandwidth and an 80-e.p.s. audio filter for use on c.w.

Table II - S-Meter Readings at 15 Mc.

S-Units	Input μ,v. 50-Ohm Source.
9	50
8	25
7	10
6	3.5
5	1.2
4	0.5
3	0.2
2	0.1
1.5	0.05
1	noise

Table I - Tabulation of Pentode Tube Data.

Tube Type	Plate Voltage	Screen Voltage	Cathode Current ma.	Plate Current ma.	Screen Current ma.	g_m µmhos	C_{gp} pf.	R_p Meg-Ohms	Equivalent Noise Resistance MIT ohms	Radio-tron ohms	Ratio g_m/ C_{gp} 10^{12}	g_m/ C_{gp} 10^9 R_{eq}
6SG7	250	100	11.0	9.0	2.0	1,850	.004	1	3320	10,680	.463	.139
6SK7	250	100	11.8	9.2	2.5	2,000	.003	0.8	3450	11,100	.666	.139
6BJ6	250	100	12.5	9.2	3.4	4,400	.0035	1.3	2160	4,260	1.03	.477
6SG7	250	150	12.6	9.2	3.4	4,000	.003	1	1980	3,100	1.33	.674
6BA6	250	100	15.2	11.0	4.2	4,400	.0035	1	1820	3,520	1.25	.685
6DC6	200	150	12.0	9.0	3.0	5,500	.02	0.5	1370	1,830	.275	.202
6BZ6	125	125	17.6	14.0	3.6	8,000	.015	0.26	822	1,140	.533	.648
6JH6	125	125	17.6	14.0	3.6	8,000	.015	0.26	822	1,140	.533	.648
6HR6	200	115	17.5	13.2	4.3	8,500	.006	0.5	875	1,400	1.42	1.62
6BH6	250	150	10.3	7.4	2.9	4,600	.0035	1.4	1770	2,330	1.31	.739
6SH7	250	150	14.9	10.8	4.1	4,600	.003	0.9	1630	2,850	1.63	1.00
6AK5	180	120	10.1	7.7	2.4	5,100	.02	0.5	1420	1,880	.255	.180
6AU6	250	150	14.9	10.6	4.3	6,200	.0035	1	1590	2,660	1.48	.930
6BC5	250	150	9.8	7.5	2.1	6,700	.02	0.8	1210	1,360	.285	.236
6CB6A	125	125	18.7	13.0	3.7	8,000	.015	0.28	867	1,150	.533	.616
6DE6	125	125	19.7	15.5	4.2	8,000	.015	0.25	847	1,280	.320	.378
6AC7	300	150	12.5	10.0	2.5	9,000	.015	1.0	722	720	.600	.838
6HA6	300	150	12.5	10.0	2.5	9,000	.03	0.5	722	716	.300	.419
6HS6	150	75	11.6	8.8	2.8	9,500	.006	0.5	769	668	1.58	2.37
6FS5	275	135	9.2	9.0	0.17	10,000	.03	0.24	287	278	.333	1.20
6EW6	125	125	14.2	11.0	3.2	14,000	.04	0.2	503	392	.350	.893
6E.J7	200	200	14.1	10.0	4.1	15,600	.0055	0.51	582	352	2.84	8.07
6688	180	150	14.4	11.5	2.9	15,900	.018	0.09	410	308	.884	2.87
7788	185	165	40.0	35.0	5.0	50,000	.035	---	100	79	1.43	18.1

Tubes listed above in the top section are remote or semi-remote cutoff. The tubes listed in the bottom section are sharp cut-off types

The 21st Century R-390A/URR Reference Y2K-R3

Cheaper and Simpler Upgrades for the R-390A HF Receiver

From Electric Radio Magazine August 2004

By Chuck Felton, KD0ZS Felton Electronic Design PO Box 187 Wheatland, WY 82201

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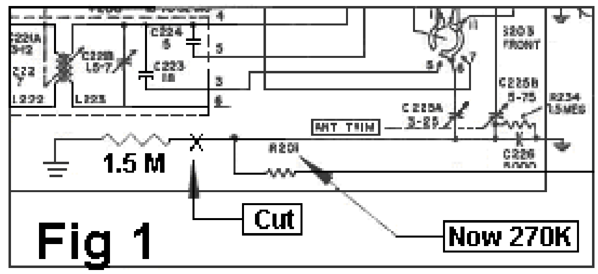
In a few hours it is possible to make some simple and inexpensive upgrades to the R-390A receiver which greatly enhance its basic performance. These changes involve 18 components which are added or replaced, all of which are easy to find. This article will discuss my upgrades in the. text and captions to the various schematics on the pages that follow.

A Better Front End Tube.

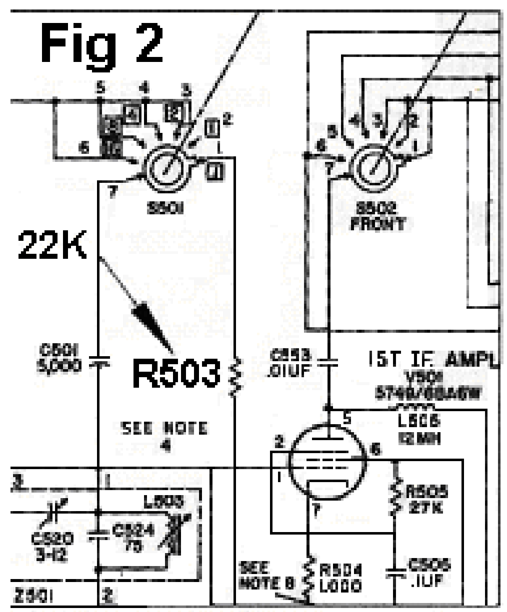
Most of tube manuals list the6DC6 as a remote-cutoff RF pentode. Actually, this is not the case, and the 6DC6 is really a sharp-cutoff type that cuts off or limits at a little over-5V on the signal grid. Many receiver designers used the 6DC6 as the RF amplifier, apparently without ever testing it!

By changing it to a 6BZ6, you will gain a stage that has a much lower noise floor. There are no circuit changes to make. Just remove the 6DC6 and plug in a 6BZ6. The pin-outs are the same and the electrode voltages are correct for both tubes..

Some of the later R-390A's were changed to include a voltage divider on the RF amplifier as a partial solution, If you're receiver has this mod, remove the 1.5M resistor to ground to allow full AGC action as shown in **Figure 1**.



In **Figure 2**, replace R503 with a 22 k, 1/2 Watt resistor. Whatever the value was, replace it with 22 k. The original 100 cycle filter position is too narrow and too lossy. Then new value will give you about 350 cycles of bandwidth at -6dB.



Looking at **Figure 3**, more easy changes are shown. If the shunt across the carrier meter adjustment, R537, is replaced with 6.8 ohms, 1/2 Watt, the adjustment range of the meter will be more reasonable.

If the BFO plate load resistor, R550, is replaced with a 3.6 mH shielded choke in a 1-Watt case size, a slight detuning of the 4th IF transformer {T503} will produce 40 volts of BFO injection. Then adjust C525 for minimum BFO signal at J116. This change makes SSB demodulation much easier. Due to some pulling, it will also produce exalted-carrier AM detection, which is known to counter the effects of selective fading.

Wouldn't it be nice to have a limiter that acts like a real limiter instead of a "distortion injector"? In **Figure 3**, If you add a 10k resistor in series with the wire going to the #7 top limiter switch contact, The limiter will soft limit at turn-on, and will progress to hard limiting when the control is fully clockwise.

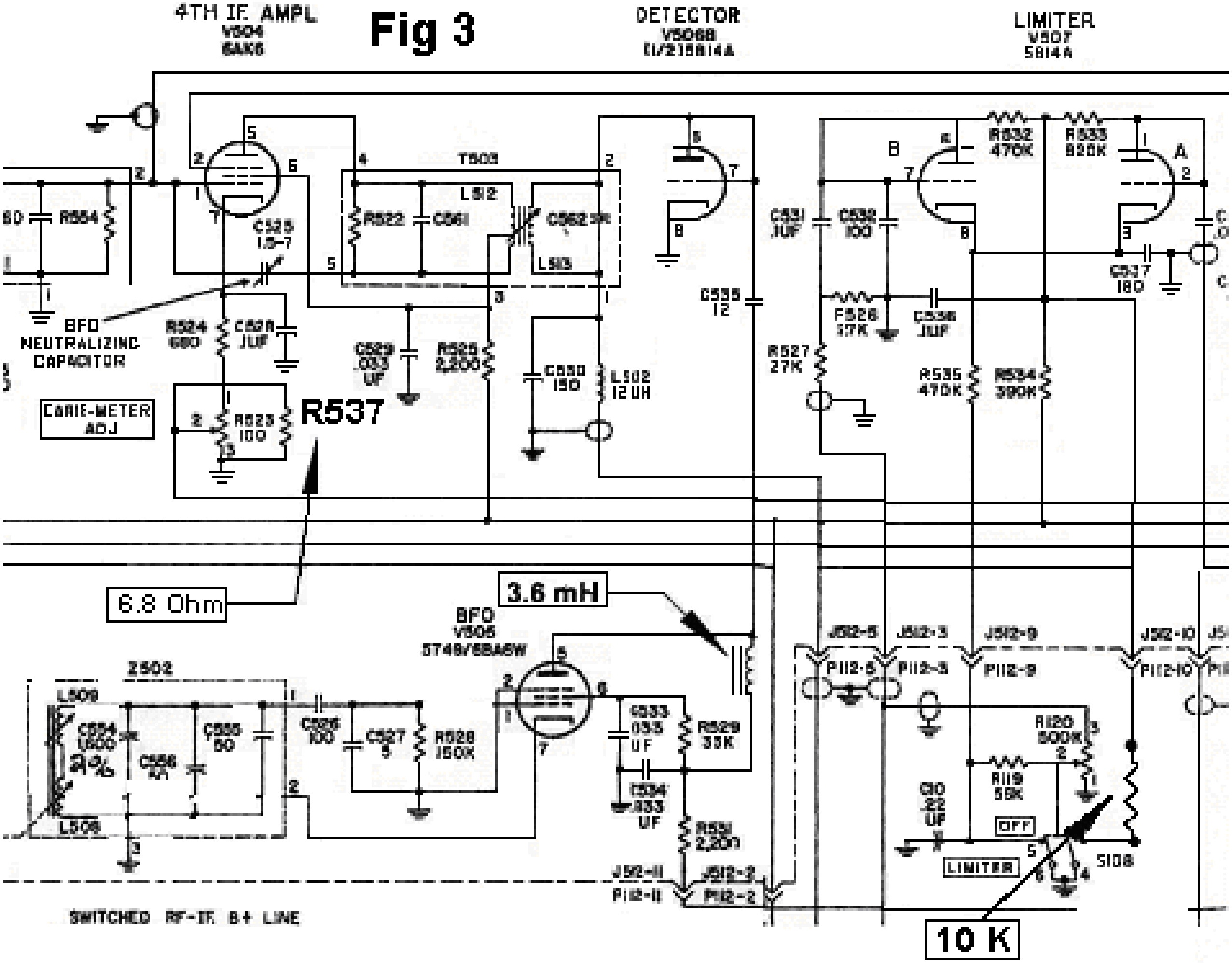


Figure 3: Schematic locations for the changes to the carrier meter adjustment, BFO, and noise limiter.

Figure 4 has some simple AGC changes. Changing the pull-up resistor, R544, to 10 Megohm, metal film, 1/2 Watt will produce some AGC "hang" action, plus an obvious increase in AGC gain.

The maximum detector signal is now about 8 volts. The 100k resistor added in series with C551 increases the AGC time constant in all positions of the AGC switch.

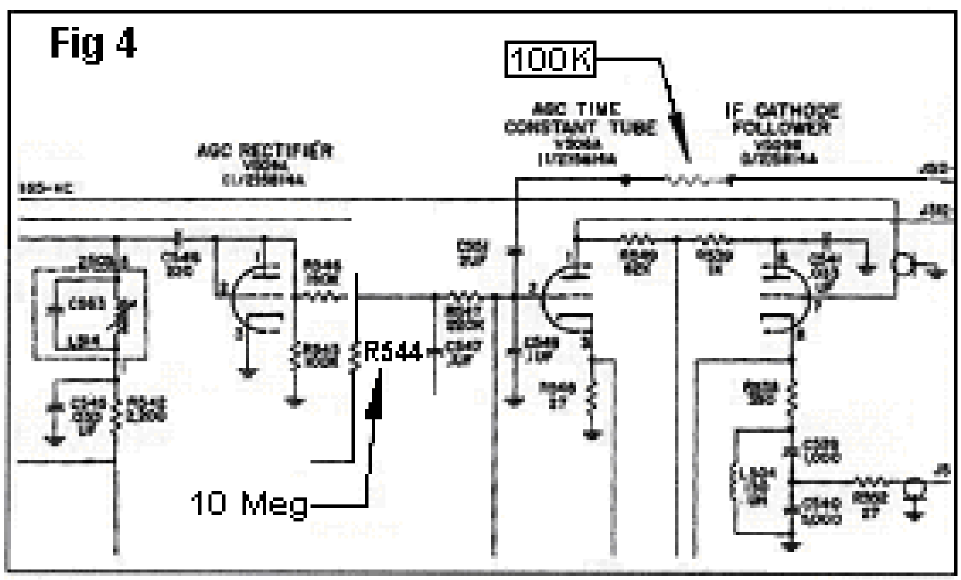
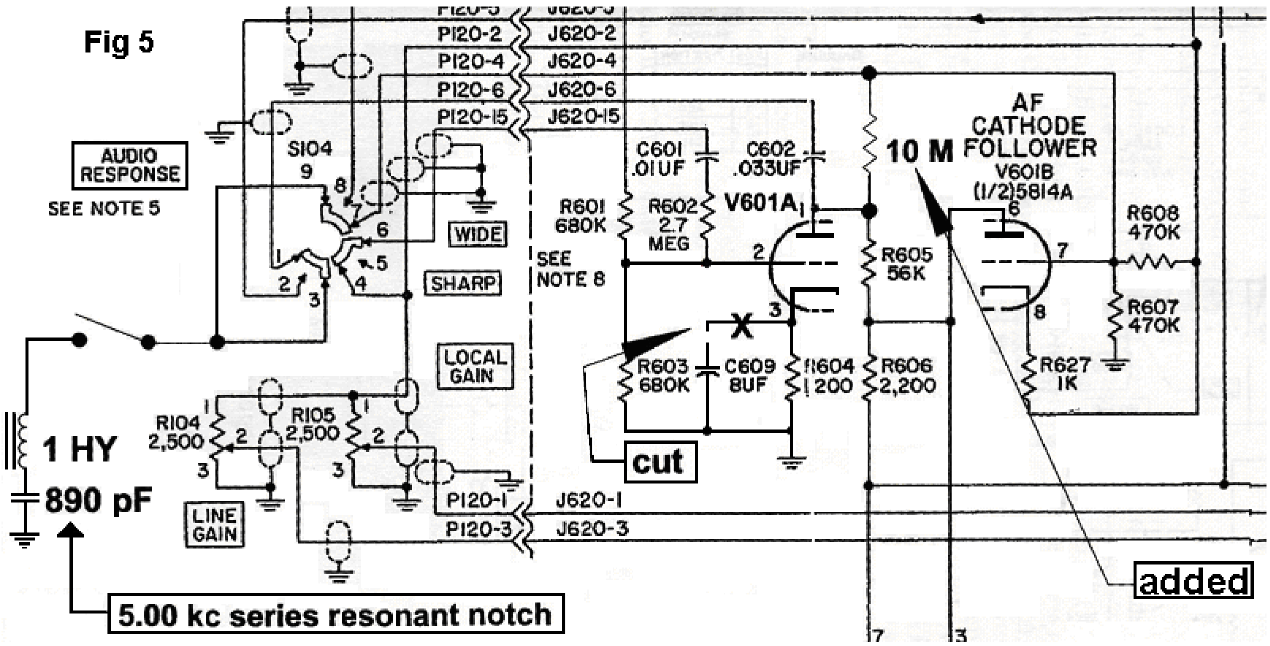


Figure 4: The AGC Changes

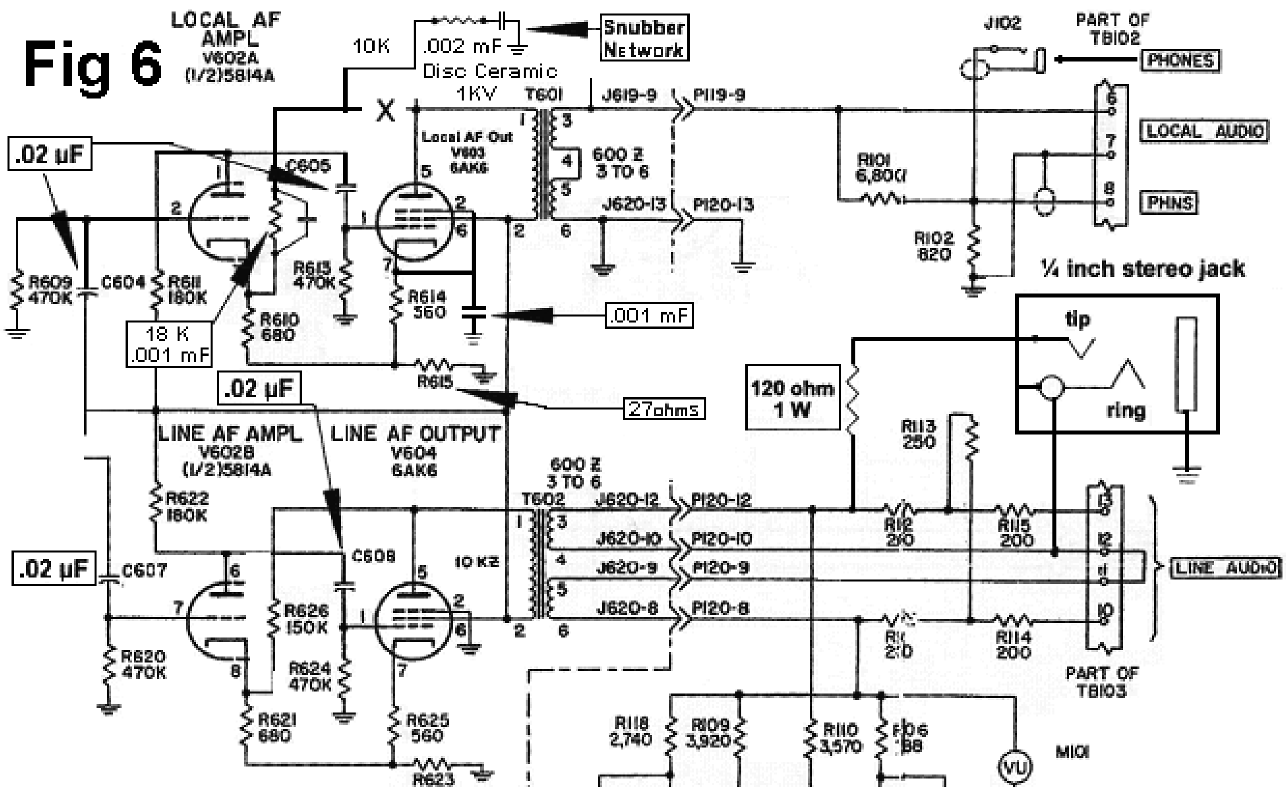
Fig 5



In **Figure 5**, find V601B, the 5814 cathode follower stage. If you add some forward bias to avoid overloading, less distortion from the detector will take place. Simply add a 10 megohm, 1/2 Watt resistor from pin 1, V601A, to pin 7.

If you remove C609 from the cathode of the first audio amplifier, V601A, the stage will have much improved linearity. It is usually a defective part anyway, and won't be missed.

Fig 6



To get rid of the 5 kc heterodyne caused by the channel spacing in the SWL bands, add a series-resonant notch filter. It is made from an inductor, a capacitor, and a switch. If a 1 Henry inductor and an 890 pF capacitor is added at the audio response switch top connection (#9) to ground, you will do just that. The 890 pF capacitor may be adjustable or selected. This change also reduces ringing of the 8 kc mechanical fitter.

Figure 6: For improved low frequency response, replace audio coupling caps C604, C60S, C607, C608, with .02 µF, 400 volt mylars. A snubber network is added to the V603 plate consisting of 10k, 1/2 W and .002 µF 1 kV ceramic disc. To include the original output transformer in the negative feedback loop, disconnect the wire at pin 5 going to R612 and remove it from the wiring harness. Replace R612 (in V602A cathode ckt.) with 18k 1/2 Watt. Connect the disconnected wire to T601 pin 3. Replace R615 with 27 ohm, 1/2 Watt. Parallel he new R612 with .001 µF disk ceramic. Add .001 µF ceramic from V603 pin 7 to ground. The 1 Watt of audio power now available requires a correct matching transformer and an efficient speaker for undistorted audio.

The R-390A/URR and its Squelch facility

By: Prof. Johannes Fischer   Revision 1.1

It seems necessary to add another detail to the R-390A saga: this receiver's inherent SQUELCH ability.

At the beginning of the *Cost Reduction Program,* the engineers at Collins were obliged to reduce the costs of the Collins Radio Receiver R-390/URR and to employ the latest improvements into the subsequent receiver R-390A/URR. A squelch facility was considered as superfluous. After many discussions pro and contra, a compromise between the Collins engineers and the Army was reached. In all the production runs of the new R-390A the factories had to prepare the chassis and its wirings for a later Field Installation of the squelch facility.

But the Program's *Final Progress Report* offers no wiring details and remains kind of vague: *"At the start of the contract it was considered desirable to drop the squelch requirement... Later on, the Signal Corps decided to make this an optional unit retaining the cable wiring and function switch and leaving out the parts in the AF unit."* (p.3, p.12). The Cost Reduction Department at Collins suggests: ".. *.An optional kit of parts can be made available for field installation of the squelch facility.* "(p.14.) As to construction details, the *Report* confirms that *"The audio unit was lengthened somewhat and rearranged to accommodate the optional squelch function." (Phase "E" Construction of "A" Models,* p.26.)

Evidently, the R-390A's AF unit has a bundle of wires underneath its chassis, which end up to no apparent use: white-black (6.3V filament voltage), white-green (from the FUNCTION switch) and white-red-green from the cathode follower V601B (1/2 5814A). One Teflon stand-off is connected to the DIODE LOAD terminal TB103 at the receiver's rear panel.

From a spare AF sub-chassis I removed the lid next to J619 and J620, which covers two factory made cut-outs, one for the insertion of V606 (tube 6C4, as printed on the chassis), and one for the squelch relay K602.

Even the FUNCTION switch holds some surprise. After removing the clamp behind its hex nut, the switch can be turned into a 6th position.

Normally, the clamp arrests the FUNCTION switch at "CAL".

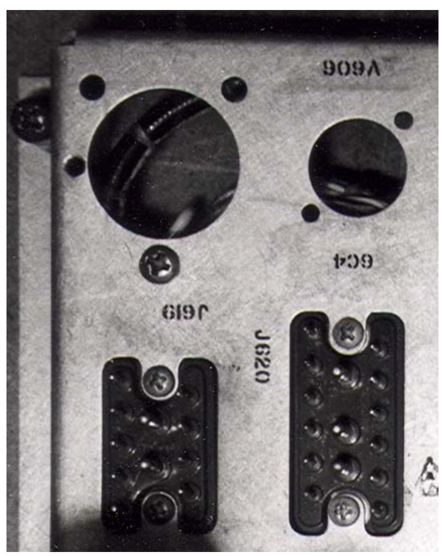


Figure 1. R-390A, Af sub-chassis, top view. Factory cut-outs for tube 6C4 (V606) and relay K602.

Figure 2. *R-390A. Af sub-chassis. top view, front section. Squelch tube V606 (6C4) and squelch relay K602 installed.*

Now, in the new position next to "CAL", the switch conducts 195 volts DC to the AF sub-chassis via the above mentioned white-green wire.

With a "kit of parts", including tube 6C4 with socket, relay K602, a decal "SQUELCH" for the R-390A's FUNCTION switch and the wiring instructions at hand, the conversion process from start to finished job of converting the R-390A into a receiver with the squelch ability is about one hour.

The Technical Manual for the R-390A (TM 11-856A) does not mention any squelch facility, but the schematic diagram of *Figure 106,* p.188, shows two interrupted lines from J 619-8 and J 620-11. I assume that the Collins engineers in their very first drafts, before the discussions started, had designed the squelch tube 6C4, the relay, the resistors and the capacitor C610 into the schematic diagram. When they had to erase the whole squelch section from the diagram, the B+ line from the FUNCTION switch (J619-8) and the DIODE LOAD line (J620-11) from terminal TB 103 were cut off and the place under the AF CATHODE FOLLOWER V601B and the 800 CPS BAND PASS FILTER became an empty area. (see *Figure 106,* p.188 of TM 11-856A). Curiously enough, the Operator's Manual for the R-390A lists the squelch tube 6C4 on p.6: *"Figure 5. Running spares. 6C4, V202, V203, V204 OR V606."* (TM 11-5820-358-10.)

To complete my R-390A with the squelch facility, I decided to use a spare EAC AF sub-chassis. Installing the tube 6C4 is easy. For the appropriate relay I looked in my parts box. Of course, the original squelch relay from a R-390/URR would have been the best choice. But I found three other relays with the right dimensions to fit into the cut-out. Does one of them has the correct ratings?

Comparing the 6C4 (V606) with the 12AU7 tube in the R-390's squelch circuit revealed similar ratings. The R-390's Technical Manual quotes a coil resistance of 10K for the squelch relay (TM 11-856, Paragraph 102. p.104, *Dc Resistances of Transformers and Coils).* One of my relays had a coil resistance of 6.5K, which was acceptable, if inserting a 3K resistor between the plate of V606 and the coil. Then I compared the B+ voltages of the R-390 and the R-390A. When turning the R-390A's FUNCTION switch to the 6th (SQUELCH) position, the unused white-green wire under the AF deck conducts 195 volts DC. The squelch circuit in the R-390 gets 180 volts, the plate voltage of the R-390's squelch tube (1/2 12AU7) is 155 volts. A 10K 2W resistor should make up for the voltage differences. For stability, I soldered the dropping resistor to an unused pin of the relay socket.

Discarding the arresting clamp on the FUNCTION switch and using an appropriate washer, I fastened the hex nut and the control knob. Again I turned the switch into the SQUELCH position.

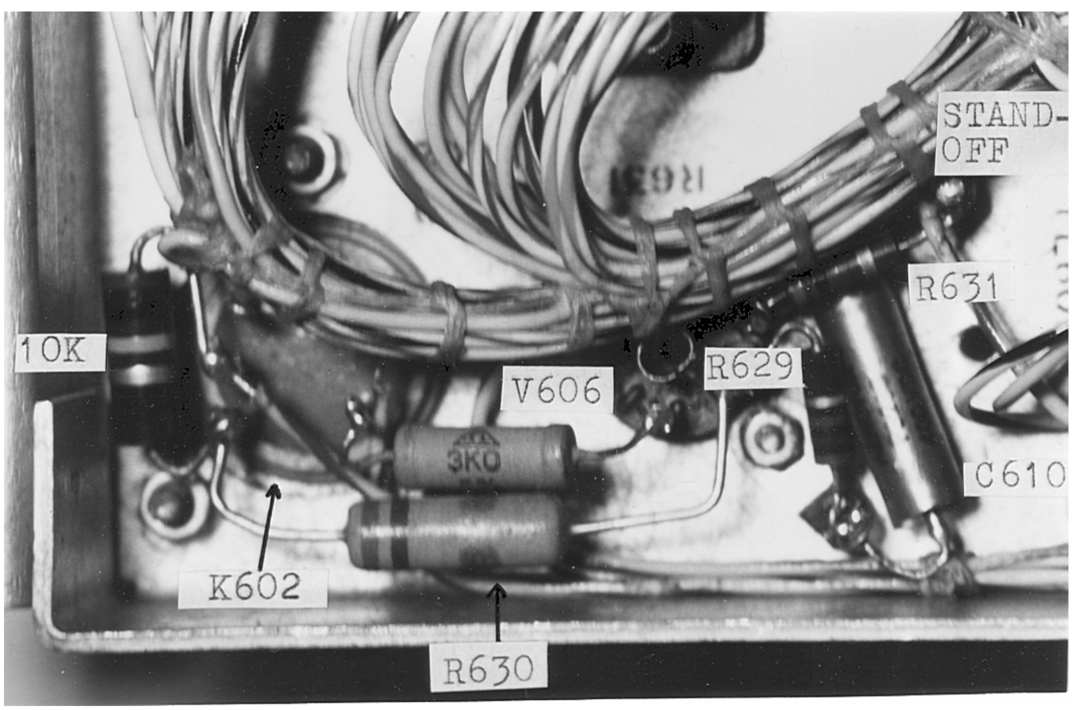


Figure 3. R-390A, Af sub-chassis, bottom view, front section.
Squelch wiring details

With the RF GAIN control at 10 and a relatively strong and steady carrier, the CARRIER LEVEL meter read 60 DB. Turning the RF GAIN control back to 9, 8, 7, 6, etc., the DIODE LOAD became more positive, the squelch tube began to conduct and the relay draw enough current from the plate of the 6C4 and B+, to close its contact, which grounded in return the cathode of AF cathode follower V601B, silencing the AF output.

Now the bad news. I could not reverse the procedure. The substitute relay was factory adjusted for straight action: at a given voltage the contact closes and at OV it opens, which is not the working condition in the squelch circuit.

To investigate into this matter, I had to open the relay can with a fine saw.

Between the coil ends (+ 160V above ground), I measured a delta V of 13V to 17V corresponding with 60 DB to 0 DB on the CARRIER LEVEL meter. On 17V the coil core ("pole face") was magnetized and attracted the armature ("lever blade"). On 3V the core was still magnetized and held the blade, grounding the cathode of V601B.

Fortunately, there were screws to adjust (see *Figure 4*). The lever blade was to be prevented from *touching* the pole face. A distance of l/2 mm or less, in any energized condition, is desired.

Very carefully I adjusted the screws and secured their lock nuts.

Now the relay responded freely to the RF GAIN control and the different carrier levels. I completed the R-390A's schematic diagram on page 188 of *Figure 106* with the squelch wiring details.

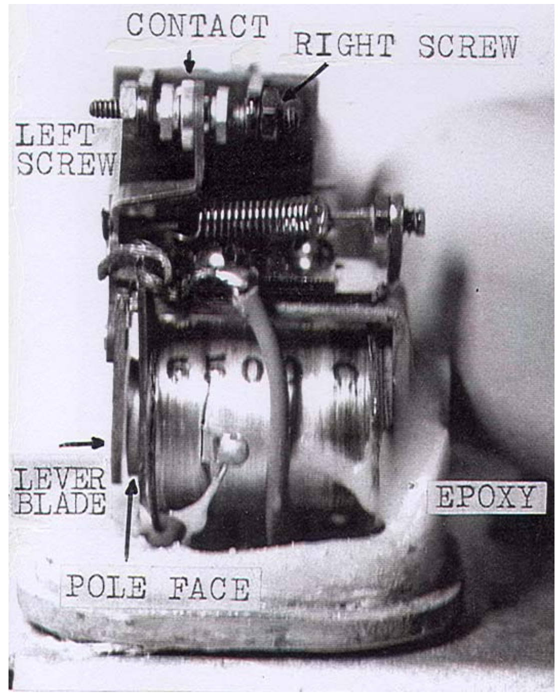


Figure 4. R-390A, squelch relay K602 (not original), internal view

Fair Radio Sales might sell the original squelch relay from the R-390, no mechanical adjustments would be necessary. Since the original relay has the nominal 10K coil resistance as shown in *Figure* 5, the coil is to be connected directly to the plate of V606. The squelch circuit is shown with no or small negative voltage fed to control grid ofV606.

Warning. The tube OA2WA, contains radioactive materials, which are harmful when contacted or inhaled. For that reason, I prefer the OA2WA's predecessor OA2. See Change 4 of the Operator's Manual.

References:

- Radio Receiver R-390/URR, Technical Manual TM 11-856. (January 1955)
- Radio Receiver R-391/URR, Instruction Book, NAVSHIPS 93766. (23 October 1953)
- Cost Reduction Program for Radio Receivers, R-390/391(*)/URR,* Final Progress Report.
-Radio Receiver R-390A/URR, Technical Manual TM 11-856A. (January 1956)
- Radio Receiver R-390AfURR, Field and Depot Maintenance Manual TM 11-5820-358-35. (8 December 1961)
- Radio Receiver R-390AfURR, Operator's Manual, TM 11-5820-358-10. (16 January 1961) Change No. 4. (7 April 1975)

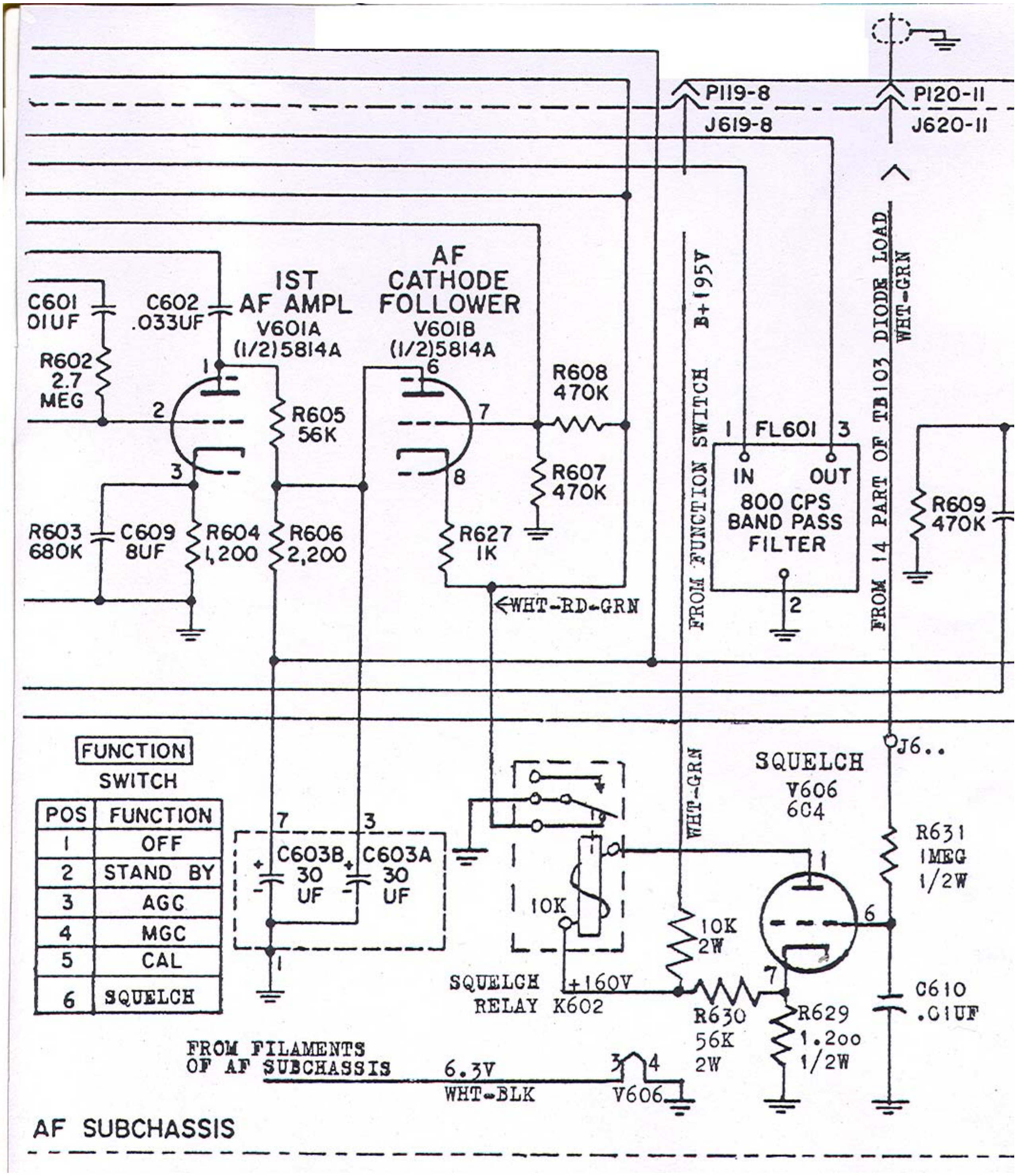
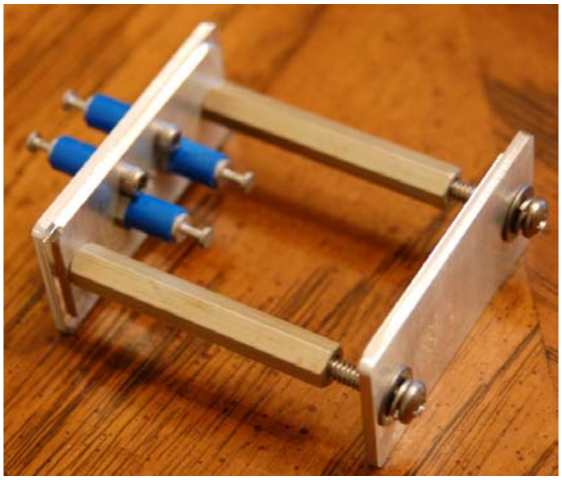


Figure 5. R-390A, squelch circuit inserted into Figure 106, p.188 of Technical Manual TM 11-856A.

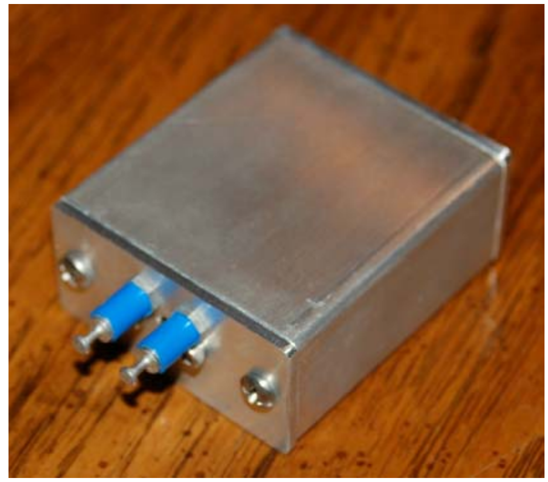
Making A New C551

By XXXXXXX

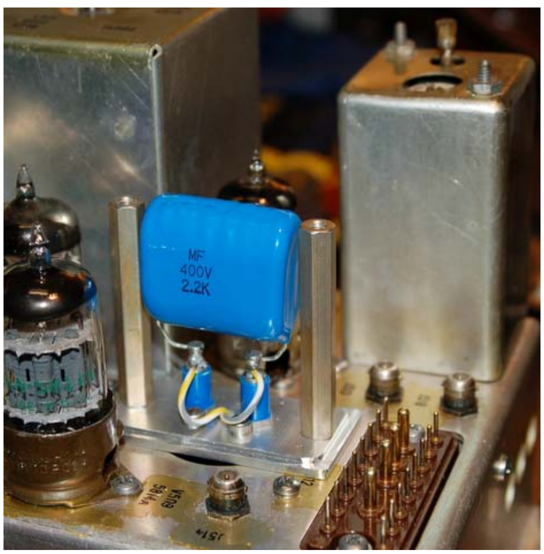
A very elegant replacement.



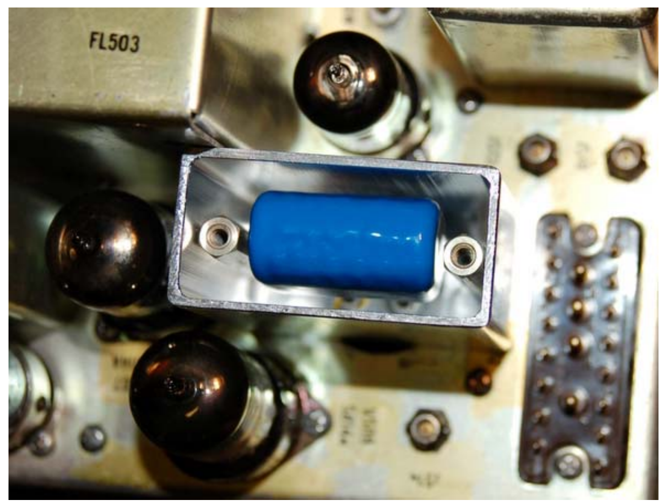
The basic frame made from scrap aluminum



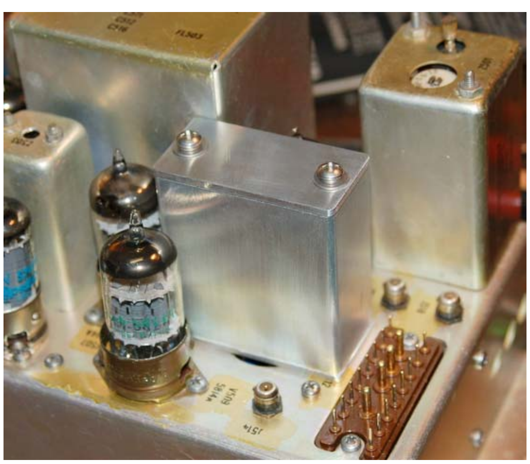
Completed Assembly



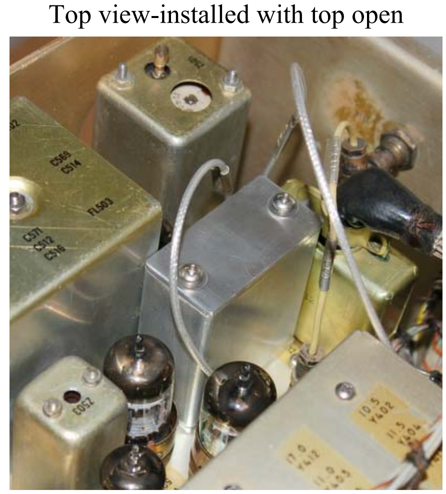
A installed without covers



Top view-installed with top open



All buttoned up



IF Module installed in receiver

DIY R-725

Conversion of the R-390 IF Deck for Use in R-390A For Improved Audio, Or Make Your Own R-725

Thomas F. Marcotte, N5OFF

As many of the readers of Electric Radio know, the US Army created a modified R-390A version for direction finding known as the R-725/URR. I wrote about this modified radio in this publication a few years ago.

The main difference between a R-725 and the R-390A is that the R-725 utilizes a custom built IF deck that is very similar in construction to the R-390 IF deck. It has tuned circuit selectivity instead of mechanical filters. The mechanical filters of the R-390A created distortion when that radio was employed for radio direction finding use. In an internal Collins engineering report published in 1952, Lou Couillard wrote of the improved R-390A, which at that time was called the R-390(XC-3), "Although the superior shape factor of the mechanical filter IF is desirable in most applications, consideration should also be given to an alternate tuned circuit design for use in special applications. The excellent shape factor of the mechanical filters precludes the possibility of linear phase shift across the passband. Where a linear phase characteristic is desired such as in direction finding equipment, a tuned circuit IF is necessary."

Employment of the tuned circuit IF is exactly what was done in the R-725. In addition to the new IF deck, the R-725 was given a new filament transformer and a minor modification for the PTO to provide improved stability.

Motorola was awarded a contract in 1956 (476-PH-56-91) to prototype the R-725. I know of a couple of these Motorola sets existing today.

Packaging of modified sets for quality DF use were handled by Arvin Industries and Servo Corporation of America. Approximately three hundred R-390A's were modified to the R-725 configuration. These sets are relatively tough to find today. New IF decks were manufactured by the modification companies (actually salvaging some of the components from the now junker R-390A IF decks) and installed in existing R-390A's. The new IF decks were named "SERIES 500 IF STRIP ASSY." They looked almost just like R-390 IF decks (see the comparison photos), except that the IF connectors were relocated to match the cables and connectors in the R-390A chassis. The circuits were designed to plug and play in the R-390A instead of the R-390. The decks are not interchangeable (until now that is, after performing the modification described herein).

Notwithstanding DF capabilities, a side benefit of the SERIES 500 deck is that it provides a smoother sound than does the stock R-390A IF deck. Mechanical filters are said to "ring" and after a while can be fatiguing to the listener. The purpose of this article shall be to describe how one may "roll his own" SERIES 500 IF deck from a surplus R-390 IF deck. Please note that I don't advocate trashing of a good R-390 to do this mod. The IF deck I started with came from a Motorola junker. I would urge you to likewise find a junker R-390 as a source of an IF deck for this project . Make sure that the deck is in working condition prior to beginning the modification.

The R-390 IF deck was designed to operate with one 25V filament supply (unfortunately the 25V supply available in the R-390A is insufficient to power all of the filaments in the R-390 IF deck). The R-390A deck was designed to operate with a combination of 6.3V and 25V filament supplies.

The task involved in this conversion is to rewire the R-390 filaments to comply with the voltages available in the R-390A and provided at the main IF deck connector, plug and play, without the addition of any new power transformers. Each of the twelve tubes in the R-390 IF deck must be addressed for full compliance with the voltages available from the R-390A. Refer to the schematic in Fig 1 as to the final filament wiring configuration. We will also drop the B+ a bit.

Plug P112 of the R-390A shall be plugged-in to the R-390 IF deck at jack J517. There is much commonality here, except for the connections mentioned herein.

In general, you will be converting 25V series connections into 6.3V (herein referred to as 6V) parallel connections for most tubes, and moving the connections of the BFO/PTO/ballast tube series to a different connection point in jack J517. The 6V filament supply shall be provided to the R-390 deck by pin 20 of P112 from the R-390A.

General instructions: Refer to the schematic for the original R-390 as the "before" schematic, and Figure 1 as the "after" schematic for filaments. Use the best soldering technique you can in this limited access space. Don't insulate or bind any wires until instructed to do so. You will be utilizing some of the new 6V supply connections more than once. Make sure you can recognize your new wires. I used black wire for the 25V supply, red wire for the 6V supplies, and green wire for new grounds.

The first task is to install a B+ dropping resistor to better match the 180V B+ that the R-390 IF deck is expecting. To do this, locate inductor L503 under the IF deck. This will be found snapped into a holder right above pin 2 of J517. Disconnect one end of the coil, and install in series with it (the equivalent of) a 470 ohm 2 watt resistor. This will tame the B+. The first tube circuit we'll work on is the ballast tube circuit.

V508 (5749) and RT512 (3TF7) These must be supplied by the R-390A 25V filament supply. To do this, sever the connecting wire at pin 8 of jack J517 (underneath the deck) to free this slot up (hint: save access to the connector end of the wire as you will use it to wire supply to V509). Then, sever the connection at pin 2 of RT512 and wire this pin to pin 8 of J517 of the R-390 deck with a long piece of new wire. The filament return connection remains unmodified. This modification will make the BFO/PTO/ballast tube series connections identical to the R-390A 25V filament supply connections. As mentioned above, this 25V supply is insufficient to supply the remainder of the tube filaments, thus the need to employ the 6V supply for this task.

The following 6V tubes shall have filaments wired from the 6V R-390A supply. The filament pins of these tubes are pins 3 and 4. Don't sever any connections unless instructed to do so. The modification will use as much existing R-390 IF deck wiring as possible (and thus may seem a bit screwy to you until finished).

1 V504 (6BJ6) 6V will come from its existing connection at pin 4. Ground will come in the next step.

2 V503 (6BJ6) Wire pin 3 of V503 to pin 4 of V504 for 6V supply. Ground V503, pin 4.

3 V502 (6BJ6) 6V supply will come from an existing connection at V503, pin 3. Ground V502, pin 3.

4 V501 (6BJ6) Sever ground connection at V501, pin 3 and wire pin 3 to V502, pin 4.

5 V505 (6AK6) 6V supply will come from existing connection at pin 4. Ground will come in the next step.

6 V506 (6AK6) Ground pin 4 of V506. Wire pin 3 of V506 to pin 4 of V505 for 6V supply.

7 V509 (6BJ6) Locate the free wire which was cut from underneath J517, pin 8, and connect it to the 6V filament supply at J517, pin 20.

The following tubes are 12AU7's wired in various series schemes in the R-390. They must be rewired according to their 6V option for use in the R-390A. Note two of the connections require dropping resistors on the 6V source of V507 and V510 to obtain the desired 5.3V filament voltage.

1 V511 Sever ground connection at pin 5.

2 Connect pins 4 and 5 together for 6V supply, ground pin 9.

3 V507 Sever connections at pins 4, 5 and 9, including the two resistors (one 120 ohm and one 22 ohm).

4 Wire 6V supply from your previous work at V505, through the deck opening for variable capacitor C525, to pins 4 and 5 through a 3.9 ohm, 1 watt resistor. Ground pin 9.

5 V 510 Sever connections at pins 5 and 9. Ground will come from existing pin 4 connection.

6 Remove 120 ohm resistor between pins 4 and 9. Wire pin 5 to pin 4.

7 Wire 6V supply from J517, pin 20 to V510 pin 9 through a 3.9 ohm, 1 watt resistor.

There are no changes to any of the other connections in the R-390 deck.

Prior to installing the modified R-390 deck in your R-390A, you must check your work.

Perform continuity checks from J517, pin 8, with RT512, pin 2.

Perform filament supply and ground continuity checks as follows. Note: There will be more than one ground connection at various tube sockets, but the filaments should have continuity exactly as shown.

Filament supply is checked from J517, pin 20.

1 V501, filament, pin 3; ground, pin 4.

2 V502, filament, pin 4; ground, pin 3.

3 V503, filament, pin 3; ground, pin 4.

4 V504, filament, pin 4; ground, pin 3.

5 V505, filament, pin 4; ground, pin 3.

6 V506, filament, pin 3; ground, pin 4.

7 V507, filament, pins 4 and 5; ground, pin 9.

8 V509, filament, pin 4; ground, pin 3.

9 V510, filament, pin 9; ground, pins 4 and 5.

10 V511, filament, pins 4 and 5; ground, pin 9.

Now insulate any bare connections, and use mini-tie wraps to secure the new wires to sturdy nearby points.

To enable final installation of the deck in your R-390A, you'll need to make two adapter cables. These cables shall consist of jumpers (RG-59 is OK, approximately eight inches in length) with BNC's on each end.

You'll also need two adapters of the type found on the back of the frame of the R-390A at the IF OUT jack (AMPHENOL 47200). This will provide crossover from MB connection (R-390A standard) to BNC (R-390 standard). Connect P-218 of the R-390A to J-526 of the R-390 IF deck with one of the cables.

Connect P-213 of the R-390A to J-525 of the R-390 IF deck with the other cable. It is a good idea to label these cables. See the photo for reference.

Install the deck in your R-390A. You will notice that the screw holes are the same as for the R-390A IF deck, however the screws of the R-390 deck are of larger diameter. I did not change these screw as they are captive into the deck. The BFO, BANDWIDTH, and power connector of the deck will hold it in place, however I would not install it in a Jeep this way. Changing these screws is optional.

When you turn on the power, make sure your dial lamps light up normally. If they don't, you have a filament supply problem so turn off the set immediately and troubleshoot.
For great sound, instead of using the built in audio deck, I prefer to tap the audio from the diode load jumper at the back of the set. Through a 0.1 uF or larger capacitor, feed this signal into your line audio amp of choice, and enjoy the tuned circuit audio of the new R-725, errrr, R-390A with tuned circuit IF. You'll get the smooth sound of the R-390 and R-725, but have the parts availability and support common to the R-390A for the balance of the set. I've used my modified IF deck in two different R-390A frames, and it worked equally well in both. In my opinion, it makes the long term listening experience much more enjoyable.

If you have any questions about the mod, please feel free to write. 73 DE N5OFF QRT

Thomas Marcotte is a registered professional engineer working in Lafayette, LA.

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Special thanks to Wally Chambers and George Rancourt.

ACG Modifications for the R-390A

By Ron Deeter, K6FSB K6fsb@arrl.net

Originally published in ER magazine Oct. 2005

I have always marveled at the performance of the R390-series receivers, in particular the R-390A. They have proven to be reliable and excellent performers, no wonder the military ordered so many and production lasted for more than three decades. There is, however, one major flaw-SSB reception. Despite the number of published modifications, none in my opinion made SSB reception acceptable. This is primarily due to the lack of addressing proper AGC action. The original Collins concept was to use an external SSB demodulator, the CV-591. This does a wonderful job however, is not what I consider a satisfactory solution for a radio: that is capable of outperforming the modern equipment of today

The R-390A that I had acquired many years back had been modified with a nonworking product detector (tube type 7360). It was a nonreversible modification with plenty of ventilation (nice way to say" unauthorized holes"). There were many other issues as well, most were not difficult to solve. This particular receiver was a complete, very low serial numbered Collins unit. Luckily, it was not a depot dog, just a butcher job. Finally, after getting everything fixed, the radio played great except on SSB. I later acquired another R-390A, and it has become my favorite as an R-725 because it has an R-390 IF sub-chassis.

The goal of this article is to address the AGC problem while complying with three requirements: no holes, the mod must be portable (it stays in the IF deck), and should have the ability to be restored. It is presented as a procedure combining the modifications by Ray Osterwald, N0DMS, from ER #150, and myself by changing the circuitry and time constants of the AGC. I have left out several pages of calculations; they can be generated by use of the formulas in the RCA Radiotron Designers Handbook. 4th Edition. The result will place the AGC in the category of fast attack and slow decay, providing a smooth AGC action that will work for SSB, AM, and CW. The AGC times should be as follows; fast attack times of 12 ms, 73 ms and 2 sec with decay times of I, 3.5, and 6 seconds respectively for settings of fast, medium, and slow. The carrier level meter will also reflect this by being more active and register far differently than before, becoming useful and functional.

Read everything first, take your time, think, pick off the varnish if any, and keep the heat down to a minimum. A copy of the original schematics and parts locator will be helpful. With the IF deck upside down, "down" is near the tube sockets and "up" is closer to you. A de-soldering tool is not necessary. There is no reason why it can't be done with braid. There are times that a de-soldering tool will not fit, let alone all the tools. It is going to get a bit crowded at times (especially on the '390), so use insulation on bare wires as necessary, and double-check your work

Overview of the R-390A Changes

Begin the mod by making sure the IF deck is working as originally designed. Replace all leaky caps, this means the Black and Brown Beauties all gotta go! Use high-quality metallized-polyester caps with an epoxy coating such as Sprague Orange Drops or similar. Please refer to Figure 1, and the simplified original R-390A schematic.

More than likely, C551 is leaky. If by chance it is good, you may wish to leave it alone, but the delay times will be a bit longer on "medium" and" slow." Five of the ones I've checked were bad (.5 to 7 mA leakage), so plan on rebuilding it and putting in a l.5-uf, 400-volt, metallized poly capacitor in the can. Take your time; drill a small #68 hole in the bottom to let the vapor out and then gently heat it to unsolder the bottom (Please think *"SAFETY FIRST"* and don't breathe the vapor.).

Do this upside down to keep the PCB oil in the can. There is not much oil, about a teaspoon, just enough to make a big mess! I used a small torch to melt the solder and clean up the can. Remove and dispose of the innards properly, solder in the new capacitor, and then either tack solder the bottom back together or use screws (0-80 will work).

- C551 is hopefully already changed to 1.5uF, 400V.
- C548 is replaced with .47pF, 400V and 33k, 1/2W in series.
- R547 and R544 are removed.
- C547 is replaced with .022uF, 400V or .033pF, 400V (each IF deck is slightly different, use .01 to .047 uF).
- R546 is replaced with a IN4148 diode from V509 pin 1 to V506 pin 2.
- Add a 2.0-Meg resistor from V506, pin 2, to ground. .
- Add a 30-ohm 1/2 watt resistor in series with carrier level meter.

Following Ray's AGC mods from ER #150:

- V504 pin 2 to ground.
- V508, connect pin 2 to pin 7.
- Remove the ground on V509 pin 3 and replace with IN4733 in parallel with .01 uF to ground.
- Add 15 Megohm, V509, pin 3 to B+.

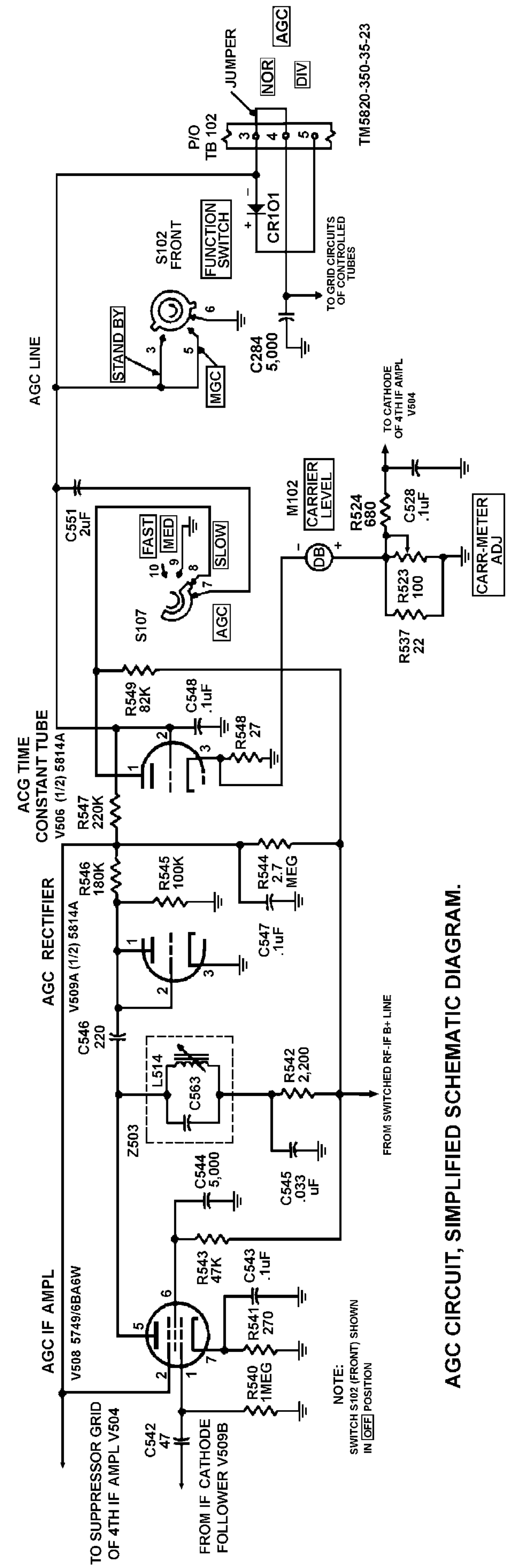
Regarding the zener, the resting AGC voltage will depend on your receiver.

You may need to shift the resting voltage one way or the other. The same goes for C547's replacement, as it is somewhat subjective.

The R-390A Working Details

Remove the BFO bellows and C535 (12 pf) and replace it with a 47 pf. Of course, insulate the leads as necessary. Reassemble the bellows.

Find the terminal next to V506; completely remove everything: R547 (220k), R544 (2.7Meg), the wire going to pin 2 of V504 (connect this to the ground lug closest to V504) and the white wire with green and brown stripes (it will no longer be used), and insulate the end.



AGC CIRCUIT, SIMPLIFIED SCHEMATIC DIAGRAM.

Remove the wire attached to pin 3 of V506, white with orange and blue stripes. This wire runs back to J512, pin 12, and verify its routing. Attach it to the now-empty terminal. Install a 30ohm, 1/2-w resistor from pin 3, V506 to the terminal (this may be changed for different meter deflection).

Find the terminal next to V509 and remove everything. Insulate the wire that led back to the terminal next to V506. See the previous paragraph, the other end is also insulated.

Now look at V508, pin 2. Disconnect everything and remove C547 completely, but save the wire because it will be reused. Connect pin 2 to pin 7 on V508 using a short, insulated jumper.

Remove C548. Reusing the wire, connect it from the C551 terminal that had C548 attached to the terminal next to V509. Connect a 2-Meg resistor from this same point on C551 to the most convenient ground lug.

Remove the ground on V509, pin 3. The replacement for R546 (180k) is the diode, with the cathode (banded end) going to V509, pins 1 and 2. Attach the anode (no band) to the terminal.

Attach the .022-uF or .033-uF cap to the terminal and lay it along side C541, attaching the other end to the ground lug on the side above V509. Actually, you will want to install C541 last just in case you wish to try a different value. The space is needed for the parts layout.

Next, attach one end of the 33k (which is in series with the .47uf cap) to the terminal, and then the capacitor lead goes to the ground lug.

Connect one end of the .01-uF cap, 1N4733 diode (band end), and the 15 Meg resistor (use insulation on the un-grounded leads) combination to pin 3 of V509. All three can lie along side C541. The other end of the .01-uF capacitor and zener diode connect to the ground lug.

The other end of the 15 Meg resistor connects to the tie point on TB501 closest to the ground lug; it has a wire colored white with red striping attached to it.

Be sure to double-check everything against the schematic, as well as all solder work.

Touch up the alignment on C503 and neutralize the BFO with variable capacitor C525.

Voila, you are done. Enjoy the nice AGC action for all modes.

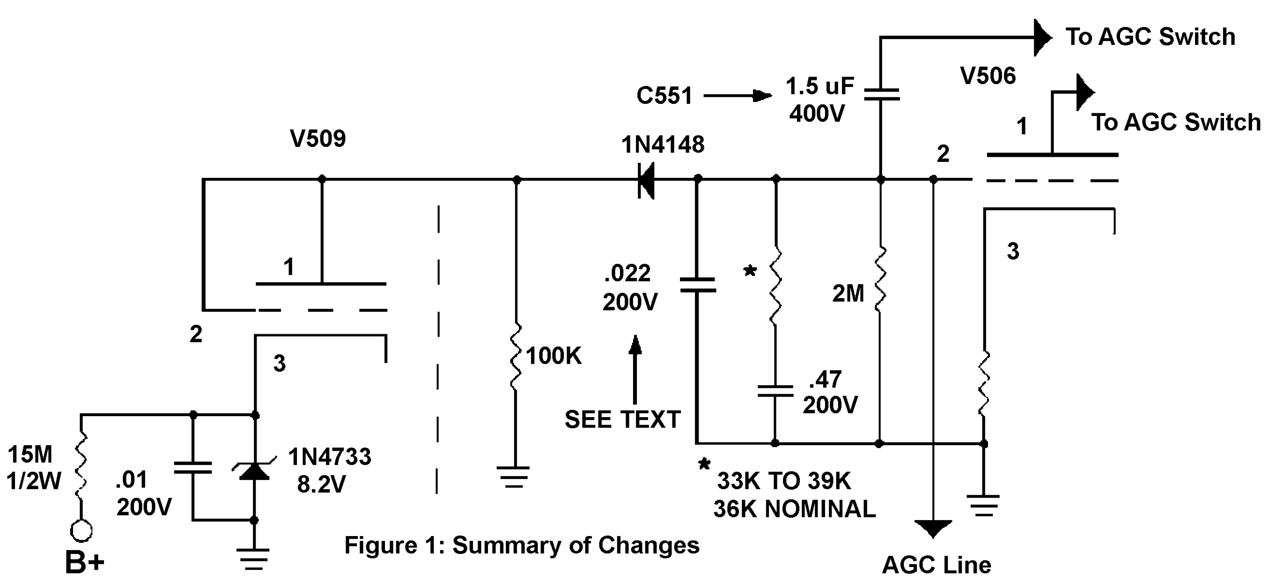


Figure 1: Summary of Changes

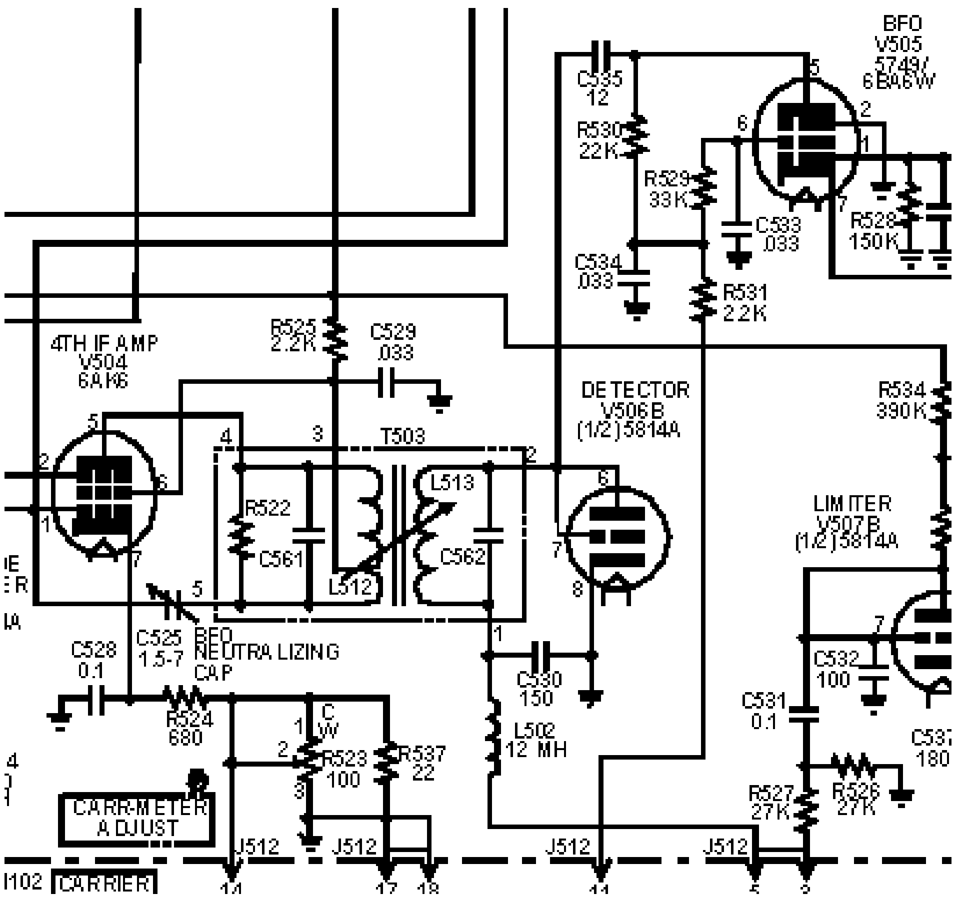
Parts List

5 ohm, 1/2w
27 ohm, 1/2w
33k-ohm,1/2w
2 Meg, 1/2w
15 Meg, 1/2w
150-uh choke (Mouser Electronics, Fastron, P /N 434-22-151)

.001 uF, 400V
.0047 uF. 400V
.01 uF, 400V .
.022 uF, 400V
.1 uF, 400V .
47-pf mica cap, 500V.

1 or 2 each, .47uF at 400V, or 1 or none if 1.5uF @400V is installed. See article.

1N4148 diode and a 1N4733 zener diode



Partial schematic of the R-390A BFO circuit. BFO coupling cap C535 is in the top of the drawing.

R-390A ILO AM Synchronous Detector Mod

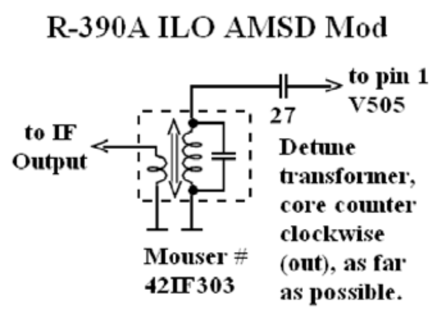
Dallas Lankford, 6/1/06

Here is a simple but effective AM synchronous detector for the R-390A which reduces fading and flutter distortion about as good as the best of them. It converts the R-390A BFO to an injection locked oscillator (ILO).

The injection source is the R-390A IF output. The injection source is stepped up to a higher voltage with a sub miniature IF transformer, 5K to 20K, black core, 3rd IF type, unloaded Q of 70, loaded Q of 35, fixed tuning capacitor of 180 pF, slot adjust core, Mouser catalog no. 42IF303. The stepped up source is injected via a 27 pF capacitor into pin 1 of V505, the 5749 BFO tube. For widest lock range the IF transformer should be adjusted by detuning the core counter clockwise (out) as far as possible. This ILO also works equally well with my product detector mod which uses a 5750 tube. A 500 volt silver mica is suggested for the 27 pF capacitor. My ILO AMSD was motivated by Graham Maynard's ILO AMSD which was published in Issue # 17 of The Hollow State Newsletter, Fall 1987 (more information may be found in subsequent issues). Back issues can be downloaded at http://www.hollowstatenews.com/ . My ILO has a wider lock range than Graham's ILO, about 200 Hz vs. 75 Hz. The R-390A ILO BFO can be offset to either side of 455 kHz to provide single sideband synchronous AM detection if desired. This, of course, requires that the R-390A be off tuned so that one or the other sideband falls within the IF filter in use. When the signal is off tuned as far as possible consistent with good lock and good audio, you will observe that by turning the BFO off and on (and hence the ILO AMSD off and on) that there is little or no fading distortion in either case. This is because maximum off tuning with an ordinary AM detector gives about the same amount of fading distortion reduction as center tuning with an AMSD. In other words, you can hear for yourself that selectable sideband AMSD is of little or no value. Nevertheless, some may wish to use selectable sideband AMSD anyway just in case it helps a little. With strongly fading SW signals these ILO type AM synchronous detectors ***will*** occasionally lose lock briefly during deep fades, manifested by a brief growl, ***if*** the AM carrier is not tuned precisely so that it is very close to the BFO frequency in the IF passband. Neither the main nor the BFO R-390A tuning has enough resolution to do such precise tuning reliably. My BFO vernier fine tuning mod with about 200 Hz per turn resolution is recommended to provide the fine tuning necessary to adjust the main tuning and BFO so that the ILO does not lose lock. With the BFO vernier fine tuning mod installed in your R-390A, adjust the BFO tuning, or the main tuning, or both so that lock is maintained throughout the BFO fine tuning range, and then set the BFO fine tuning pot to mid range. The R-390A BFO vernier fine tuning mod can be downloaded from The Dallas Files at http://www.kongsfjord.no/ .

One way to implement this R-390A ILO is shown here. The braid of miniature Teflon coax was soldered directly to the IF transformer shield, and the center conductor was soldered to the appropriate link coupling input.

Do not use excessive heat; it can melt internal parts of the IF transformer.

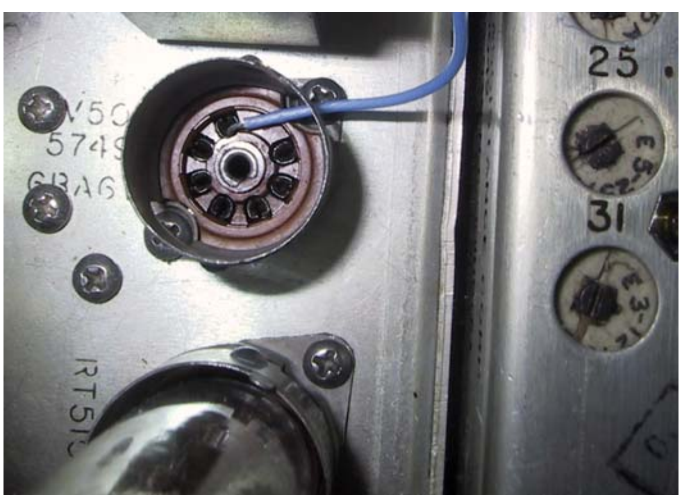


The two appropriate grounded pins of the IF transformer were connected to one of the IF transformer shield lugs with #24 solid tinned copper wire and soldered. A 4.7 Meg ohm resistor was attached to the IF transformer shield as shown. One lead was soldered to the IF transformer shield.

Three (3) turns of #24 solid tinned copper wire were wrapped around the resistor and one lug of the IF transformer shield and soldered. This provided an insulated standoff for attaching one lead of the 27 pF capacitor and a short length of #22 stranded insulated (blue color here) silver plated Teflon wire which was connected to pin 1 of the V505 tube socket as will be shown below. The other lead of the 27 pF capacitor was soldered to the appropriate IF transformer pin. The finished IF transformer assembly should be insulated with heat shrink tubing or some other good quality insulating material.

The assembly should be tested before insulation is applied. If heat shrink tubing is used, be careful not to apply too much heat. There are several options for connecting the miniature coax to the IF output. If you have a (used) miniature BNC connector like those used to connect to the IF output connector on the R-390A IF module, then you can attach it to the end of the miniature coax and connect to the IF output connector on the R-390A IF module (after detaching the miniature BNC connector with the miniature coax which goes to the IF output connector on the inside of the rear panel). If not, then you may use a regular size BNC connector and connect the end of the miniature coax to the BNC IF output connector on the outside of the rear panel. The BNC connector is small enough to pass through the hole in the rear panel for aligning the oscillator shaft.

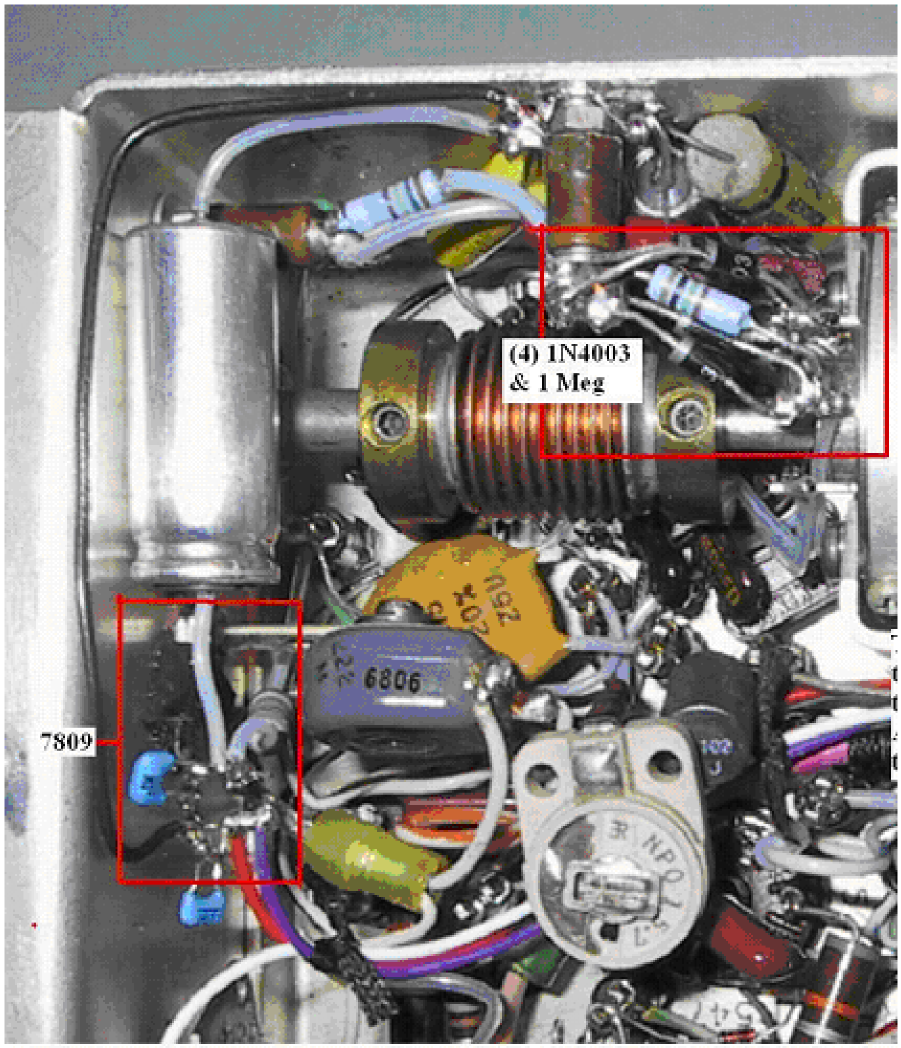
Below is a photo of the connection of the insulated stranded wire from the 4.7 meg ohm insulated standoff to pin 1 of the V505 tube socket. First, the V505 tube was removed. Next, about 5/16 inch (8 mm) of insulation was removed, a right angle in the strands was formed, the end of the wire with the right angle strands was inserted through a small hole in the base of the tube shield support, and the end of the strands inserted into the 1 position of the tube socket. Finally, if the strands are not twisted too tightly, the V505 tube can be re-inserted. The ILO can also be installed underneath the chassis of the IF module because the only alignment required is the BFO frequency. After the R-390A has warmed up for an hour, select the 2 kHz BW and tune a steady AM signal so that it is in the center of the passband. Turn on the BFO and adjust the BFO tuning shaft so that zero beat occurs when the tuning knob is at 12 o'clock (0). The BFO is now re-aligned and you are ready to play with your ILO. When tuning selectable sideband AMSD you should set the BFO frequency (+ or -) to a little less than half the filter bandwidth. For example, when using the 4 kHz BW filter, single AMSD sideband is gotten by setting the BFO frequency to about + or – 1.75 as shown on the front panel, and then tuning AM signals without changing the BFO frequency, except slightly as necessary to adjust the ILO lock range. Of course, as I said above, with offset tuning like this, you will not hear much fading distortion, if any, in either case, with the BFO on (AMSD) and the BFO off (ordinary AM diode detection). The case where AMSD will give substantial reduction in fading distortion is when the signal is tuned near the center of the passband and the BFO is set near 0.



R-390A BFO Fine Tuning Dallas Lankford 12/24/03

It is very difficult to tune SSB well with an unmodified R-390A. The same can be said for tuning AM via ECSS. Neither the main tuning KCS not the BFO tuning (BFO PITCH) are slow enough.

Here is a BFO fine tuning mod which uses voltage tuned diodes. It works very well. Be sure to get a Clarostat type J 50K ohm pot from Newark. Or a NOS AB pot would be equally good. I mounted the pot in place of the dial lock mechanism.

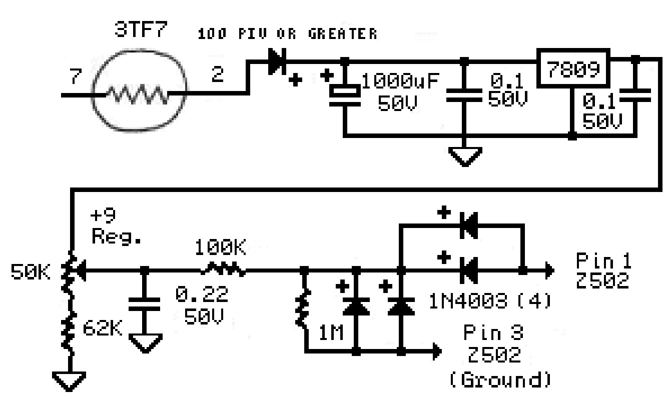


The pot tab hole on the back side of the front panel. The pot tab did not mate with the dial lock tab hole on the back side of the front panel, but the pot tab can be bent appropriately.

The picture and schematic show most of the details. The 7809 is heat sinked by mounting it against the of the IF deck using the mounting bracket for L502. When mounting the 50K pot make sure the wiring does not contact nearby moving parts. All grounds should be solder joints and should be returned to pin 3 of the BFO PTO for best frequency stability.

The BFO fine tuning frequency range is about 200 Hz.

The there wires to the 50K ohm pot go through the holes in the IF chassis which pass the wires to the GAIN ADJ and the CARRIER METER ADJ. Some recycled blue plugs were used so that these three wires could be quickly disconnected.



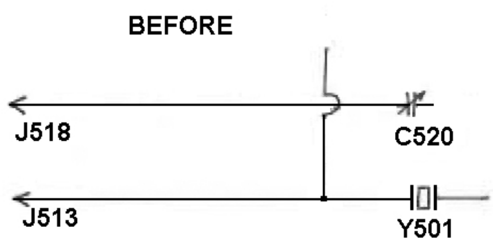
Note: In one IF deck mod, BFO frequency instability was observed. The problem was traced to one or more (the exact number is uncertain) capacitors in the 5749 tube circuit which was modified to a 5750 tube product detector. For this reason, all capacitors in the BFO circuit should be replaced with new high quality capacitors.

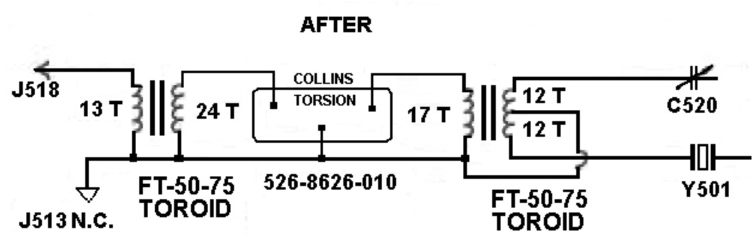
R-390A Roofing Filters

Dallas Lankford 9/1/2004

There are two reasons to put a roofing filter in an R-390A: (1) to improve the close spaced 3rd order intercepts, and (2) to provide a more appropriate wide AM bandwidth. The close spaced 3rd order intercepts for 2, 4, 8, and 16 kHz BW's without a roofing filter is about -20 dBm, which can definitely cause 3rd order intermod in high RF environments. With a 6 kHz BW roofing filter placed at the input of the IF deck, the close spaced 3rd order intercepts are improved to better than 0 dBm. The 8 and 16 kHz BW's are too wide for general AM listening in most cases, so a 6 kHz roofing filter kills two birds with one stone.

You could put a roofing filter in the RF deck, immediately after the 3rd mixer, and the performance would be the same. But removing and reinstalling an RF deck is an order of magnitude more difficult than removing and reinstalling an IF deck. Since the signal path from the RF deck to the IF deck is via two minicoax cables and two quick release miniBNC connectors, you could implement the roofing filter external to the IF deck, provided you had appropriate connectors. But it seems simpler to put it in the IF deck. The general idea behind the mod is quite simple. Below are "Before" and "After" schematics for the mod that I do.





This mod is designed for 2000 ohm source and load filters, so you could also use a 6 Khz ceramic filter, like the LF-H4S which typically has a 6.5 kHz BW. But I would recommend a so-called low cost Collins torsion filter (they are definitely not low performance). The part number above is for the old style low cost torsion filter (with a metal case). Contact Collins filter division for current prices and availability of low cost torsion filters. The last time I checked prices, they were about $100 each.

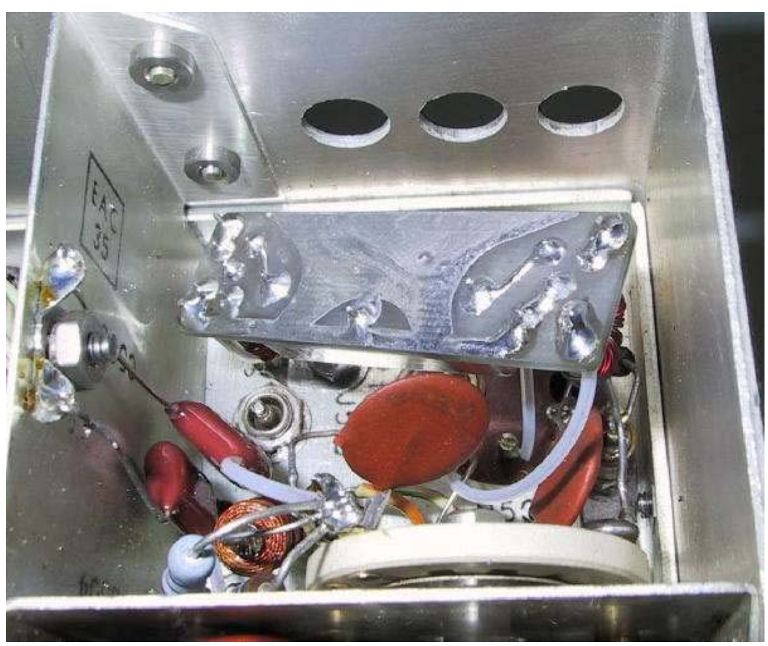
If you look at an R-390A schematic, you will see that the 1st 455 kHz IF transformer (immediately after the 3rd mixer) has a high impedance LC tuned primary, and untuned push-pull outputs which are connected to the IF deck through minicoax using miniBNC connectors. I measured the output impedance of each output as about 600 ohms. To match 600 ohms to 2000 ohms requires a turns ration of about 1.826 (the square root of 2000/600), which is closely approximately by 13 to 24 turns (24/13 = 1.846). In the original R-390A circuit, the push-pull output of the 1st 455 kHz IF transformer worked into a high impedance load. Presumably the number of turns in the output windings of the 1st 455 kHz IF transformer were not chosen to match impedances (since the impedances were not matched), but to establish the desired signal level at the input to the IF deck. The 17 to 12 + 12 turn center tapped output transformer is an artifact from a previous roofing filter design and perhaps could be improved upon (a few more turns on the output? ... a few less?), but it seems satisfactory, so we have left well enough alone. Trying to use a high impedance standard size Collins mechanical filter is probably not a good idea because it would require more complicated impedance matching and it would be difficult to fit everything in the available space. The following photos illustrate the space constraints.

The PC Board Length Is About 2" Or 5.5 CM

In the photo to he right, the two wires from J513 to Y501 and from J518 to C520 have been removed, two long #22 solid tinned copper wires added to the Y501 and C520 lugs, and a ground lug added between J513 and J518. Two shorter wires with Teflon insulation will be added to J518 and to the ground lug; these two wires will be soldered to one end of the PC board. Teflon insulation of an appropriate length will be added to the two longer wires which will be soldered to the other end of the PC board. A 100 ohm half watt resistor has been added in parallel with R504 560 ohms to recover the filter mod insertion loss.

IF Deck Rear Compartment Prepared For Filter Board

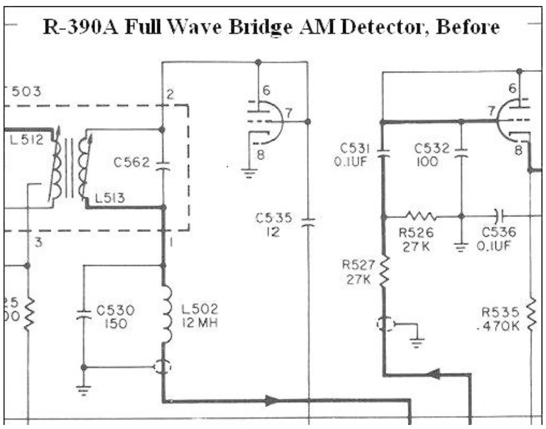
Observe that the PC board is slanted from front to back and from top to bottom. It is probably not feasible to use a larger filter for this mod

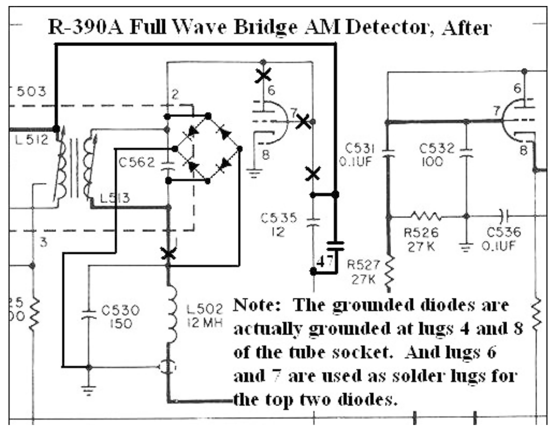


Completed Roofing Filter Mod

R390A Full Wave Bridge Detector

Recently I had the good fortune of using my friend Bjarne's Teleftmken E1501 communications receiver. Bjarne had sent it to me for repair, which I did, and after I repaired it I had the pleasure of DXing with it for a few days. On thing I noticed right away was that the E 1501 AM detector sounded much better than an ordinary detector, almost as good as an AM synchronous detector. Later, when I returned it to Norway, Bjarne said the same thing. The E1501 manual merely said that the AM detector was an envelope detector. Inspection of the schematic revealed that it was not a simple diode detector. In fact, the E 1501 did not resemble any detector I had ever seen before. I looked through some electrical engineering books and discovered that this kind of detector is called a push-pull AJ\1 detector; cf. Electronic Designers' Handbook, by R Landee, D. Davis, and A. Albrecht, McGraw-Hill, 1957. Almost immediately I started wondering if a push- pull AM detector could be developed for R-390A's. A few days later, after several failures, I succeeded in developing a push-pull AM detector for one of my R-390A's.





But my design had a substantial defect, decreased audio output. Ideally the secondary of T503 should be rewind as a bifilar center tapped transformer. Rather than attempt such a complex change, I decided to split the interna168K resistor (not shown on R-390A schematics) into two 33K halves. In effect this provides the necessary push-pull input to two 1N4148 silicon diodes. Later I used 1N34A geranium diodes which worked equally well. Rather than change the detector tube socket wiring, I cut pins 6 and 7 off the 5814A tube. C530 was rewired from the lug of T503 and a short jumper was added from that lead of C530 to the lug of L502. After these changes the push-pull AM detector worked very well, except that audio output was down about 6 dB. I was concerned that the push-pull AM detector might not work for SSB and CW, but it seemed to be at least as good as a single diode detector in that regard.

I didn't like the audio loss of my push-pull AM detector, but couldn't immediately think of any way to prevent it other than rewinding the IF transformer T503, which I wasn't about to do. A day later as I was driving in my car the solution popped into my head... replace the 33 K ohm resistors with diodes. I implemented this new mod and it worked fine for AM signals, with no audio loss. All of a sudden, in a delayed reaction, I recognized my new AM detector circuit for what it is... a full wave bridge rectifier, but with an IF transformer for the source AC voltage, and operating at RF frequencies instead of ELF frequencies which power supplies usually operate at. When I tried my new detector on SSB it had terrible distortion. I mentioned this to my good friend Wally K50P during a phone conversation shortly after implementing the full wave bridge AM detector. He replied that the BFO injection was too low. Later I rerouted the BFO injection to the primary of the IF transformer T503 and SSB was fine. **Years ago, following K50P's advice, I had already increased the BFO injection for my unmodified R-390A diode detectors in all of my**

R-390A's by paralleling a 47 pF 500 volt silver mica across C535 12 pF. This mod was retained when I moved the BFO injection to the primary of T503. It was necessary to add a splice to the ends of C535 and the 47 pF cap in order to reach the appropriate lug of T503. The polarity of the diodes in the R-390A full wave bridge AM detector schematic above should be observed. Other wise, when the R-390A noise limiter is turned on there will be little or no audio. I know because I made this mistake in some of the early versions of my R-390A full wave detector. When I started these mods I did not know that the original R-390A diode detector is part of the noise limiter circuit and that any R-390A detector mod must be fictionally equivalent to the original with respect to the noise limiter.

I do not know of any other reference to a push-pull AM detector other than the brief mention in the book by Landee, et al. Push-pull AM detectors may be related to full wave AM detectors which have been discussed by K2CU; see www.amwindow.orgl/tecWhtm/alowdisdet.htm. At one point I thought they were equivalent, but now I don't know. There are a few internet references to full wave AM detectors, but none of those resemble the push-pull or full wave bridge AM detectors described here. I do not know of any reference to a full wave bridge AM detector.

So what does a push-pull, full wave, or full wave bridge AM detector do for an R-390A? It sometimes reduces part of the distortion you hear on fading AM signals. The improvement is most noticeable for some, but not all, weak signals fading in and out of the ambient noise floor and for some, but not all, quickly fading signals. However, a push-pull or full wave AM detector does not eliminate distortion caused by total or almost total fading of the carrier, but neither does an AM synchronous detector despite claims to the contrary.

For many kinds of fading AM signals an appropriately slow release time AGC (2.0 - 3.0 seconds) provides much more reduction of distortion and noise than a push-pull or full wave bridge AM detector or an AM synchronous detector. Another technique for reducing fading distortion of AM signals which many are not aware of is to tune mostly one sideband or the other. This eliminates distortion caused by incoherent sidebands. But the best technique I have found for reducing AM distortion due to fading is to use an appropriate elliptic low pass audio filter. This method will be discussed in detail in a future article.

Some have claimed that a phase locked loop AMSD (like the SE-III, AMSD-l, and AMSD -2) is better than (has less distortion than) a synchro phase AMSD (like the Racals... RA6790/GM, RA6793(A), and RA6830). But I have used both types side by side for over 6 months now and have observed little or no difference in improved audio between the two types. I am also into my second day of using a modified RF notch filter as a fading simulator. With it I can generate realistic sounding fades with realistic sounding distortion. Whether or not my simulated fades are accurate approximations to the real thing I do not know. But they surely sound like the real thing. I have been using my fading simulator with many of my receivers... R-390A with external AMSD -2, unmodified R-390A, R-390A with internal full wave bridge AM detector, modified (MW and LW attenuator removed, Preamp 1 enabled) IC-746 Pro, unmodified R8B, unmodified RA6790/GM, and modified (grounded tuning encoder shaft for microprocessor RFI elimination) WJ-871IA. It has been most instructive. None of these receivers eliminates much fading distortion in the case of an AM carrier which is mostly or completely eliminated during the fade. If I slow down the fade (slow motion fading!), I believe I can hear that phase locked loop AMSD's hang onto the AM carrier longer than synchrophase AMSD's. Having said that, let me add that the distortion produced by a fading carrier is about the same in either case when a fade takes place at the usual rates. The bottom line, as I have said on many other occasions, is that one will not hear anything using an AMSD that one will not hear almost as well without an AMSD. The same is true for push-pull and full wave bridge AM detectors.

In my opinion, a full wave bridge AM detector, without other techniques such as slow AGC, tuning mostly one side band or the other, and elliptic low pass audio filtering, is a marginally worthwhile modification for an R-390A or other receiver, similar to adding an AM synchronous detector. Nevertheless, it is a mod which will remain in my R-390A and which I will do again because in difficult DX situations every little bit of improvement in recovered audio helps.

60 dB Wide Dynamic Range, Low Frequency AGC Circuit Using a Single VGA

From Analog Devices AN-934 by James Staley Used with permission Edited by Perry Sandeen

Introduction

Low frequency automatic gain control (AGC) circuits are used in audio and power equipment for applications such as sensitive microphone preamplifiers (preamps) and regulators. An AGC circuit, a closed-loop feedback system, is shown in Figure 1. The loop consists of a controllable gain element, a detector, a stable reference and a comparison circuit.

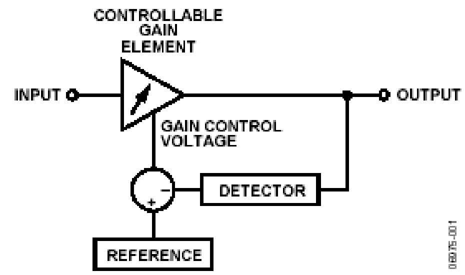


Figure 1. AGC Circuit Using a Variable Gain Amplifier

This application note describes a low frequency AGC circuit using a wide dynamic range AD8336 variable gain amplifier (VGA) as the gain control element, an AD736 rms-to-dc converter as the detector, a low cost rail-to-railAD8551 op amp, and an ADP3339 LDO as the reference. Because of its wide controllable gain range and circuit flexibility, the AD8336 is featured in this application note.

Controllable Gain Element

A VGA is a special type of amplifier, which controls its gain by electronic means instead of by a set of fixed resistors, as is the case with the classic op amp circuit. VGAs are the familiar and preferred solution for automatic gain control circuits in a variety of communications applications.

VGAs operate at frequencies from hundreds of kilohertz up to hundreds of megahertz. An ideal VGA performs as a linear amplifier, without introducing distortion or otherwise corrupting the desired signal.

When a VGA is used, the gain element is an amplifier combined with electronic volume control. In this example, the controllable gain element is further reduced to an electronic potentiometer and a fixed gain amplifier and it adjusts the loop gain by attenuating the input signal, without contributing significant distortion.

The other fundamental elements oft he loop are the detector, a stable reference, and a summing circuit that senses the state of the loop, compares it to the stable reference, and adjusts the output accordingly.

A functional block diagram of the AD8336 is shown in Figure 2.

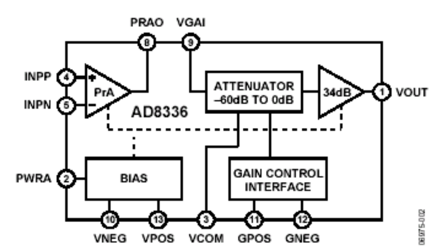


Figure 2. AD8336 Functional Block Diagram

Circuit Design

An audio AGC requires the following features.

- Wide dynamic range, which is the ability to amplify very low level signals and very large signals.
- Amplification with low distortion over the entire operating range.
- A means of adjusting the minimum and maximum gain limits.

The AD8336 described in this application note uses the Analog Devices, Inc. exclusive X-AMP architecture, consisting of a ladder network with multiple taps spaced in equi-resistive increments, and accessed by an array of differential amplifiers. See Figure 3.
A passive resistor ladder network performs the gain control function, introducing no distortion. The gain element is a fixed gain op amp.

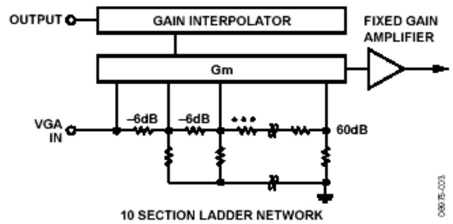


Figure 3. AD8336 Ladder Network

Because the gain of the op amp remains unchanged, the application benefits from constant bandwidth, distortion, and overload performance optimized over a wide range of operating conditions.

The AD8336 features a wide gain range (60 dB) an e,-tended supply voltage, capable of operating with power supplies up to $\pm$ 15V. It features an uncommitted preamplifier and permits inverting, non-inverting, or differential input configurations.

The preamplifier and VGA sections are completely independent, and the VGA can be used as a standalone element if no preamp is needed The gain control inputs are fully differential Figure 4 shows the gain characteristic for the VGA, for two values of preamplifier gain.

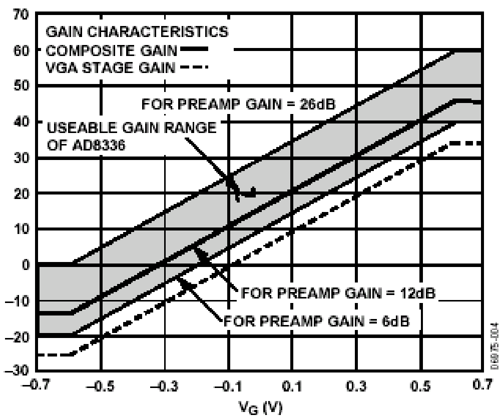


Figure 4. Gain Characteristics of the AD8336 for Various Operating Conditions

AGC Circuit Design Example Signal Voltage levels

The range of signal voltages to be controlled, the supply voltages and input and output voltage levels, all highly interactive factors, influence the topology of an AGC circuit. In this example, the goal is to fully exploit the full 60 dB gain control range of the AD8336. First, assume power supply voltages of 5 $\pm$V.

With a known power supply voltage, the stabilized output voltage is established. Because saturation in either the pre-amplifier or the 34 dB fixed gain stage limits the available output swing to about 7V p-p, a nominal maximum swing of 5V p-p is easily attainable. With a preamp output voltage swing of 5V p-p and the X-AMP attenuator set for -16 dB (0.05x), the output voltage is 250 mV p-p. If the preamp gain is set at -1x (unity inverting gain (equivalent to a noise gain of 2x), the maximum input voltage is 5 V p-p. Finally, with a gain range of 60 dB, the minimum input voltage is 5 mV p-p. The AGC circuit operates with an input voltage range of 60 dB (5 mV p-p to 5 V p-p), with a fixed output voltage of 250 mV p-p.

Control Voltage Levels

The differential gain control input of the AD8336 performs any level shifting required for the available control voltages considerably simplifying the gain control drive circuitry. In this example, the GNEG input (Pin 11) is biased at 0.75 V and the gain range voltage at GPOS is l.5V.

Detector

The detector is an AD736 rms-to-dc converter and provides an accurate dc control voltage directly proportional to the rms value of the output signal The output of the AD736 drives the inverting input of an op amp connected for very high de gain for accurate loop control.

Comparison Circuit

The AD8551 is a single-supply rail-to-rail op amp with a very low, offset voltage. The voltage applied to the non-inverting input is the reference voltage, and establishes the rms value of the output. T he voltage to be compared is the detector voltage from the rms-to-dc converter. When the comparison input falls below the reference the comparison output voltages increases to restore the output to its nominal level.

AGC Circuit Operation

Table 1 lists the data for the AGC control for six frequencies from 1 mV to 2 V rms input. Refer to Figure 6 for a plot showing the flat output level vs. input for the typical audio frequency range of 20 Hz to20 kHz. The output level is flat over the 2 mV rms to 2 V rms range.

Table 1.

EIN (V rms)	EOUT (mV p-p) 20 Hz	100 Hz	1 kHz	5 kHz	10 kHz	20 kHz
0.001	125	130	136	135	140	140
0.002	245	255	253	253	260	265
0.003	251	250	251	253	257	258
0.005	250	250	250	251	256	258
0.01	250	250	250	251	255	255
0.1	250	250	250	251	254	254
1	250	250	250	251	254	254
1.5	250	250	250	251	254	254
1.8	250	249	250	250	254	254
2	250	256	261	266	266	266

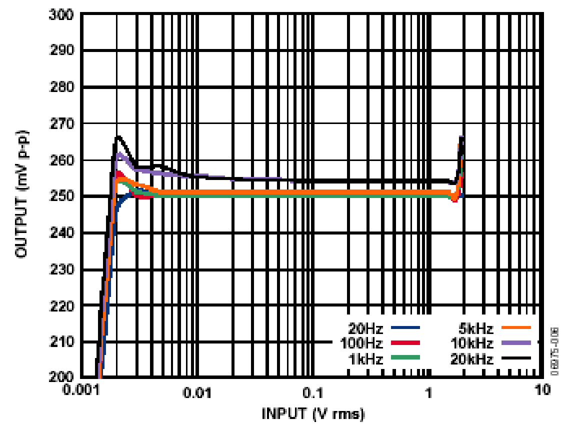


Figure 6. AGC Performance at Various Frequencies

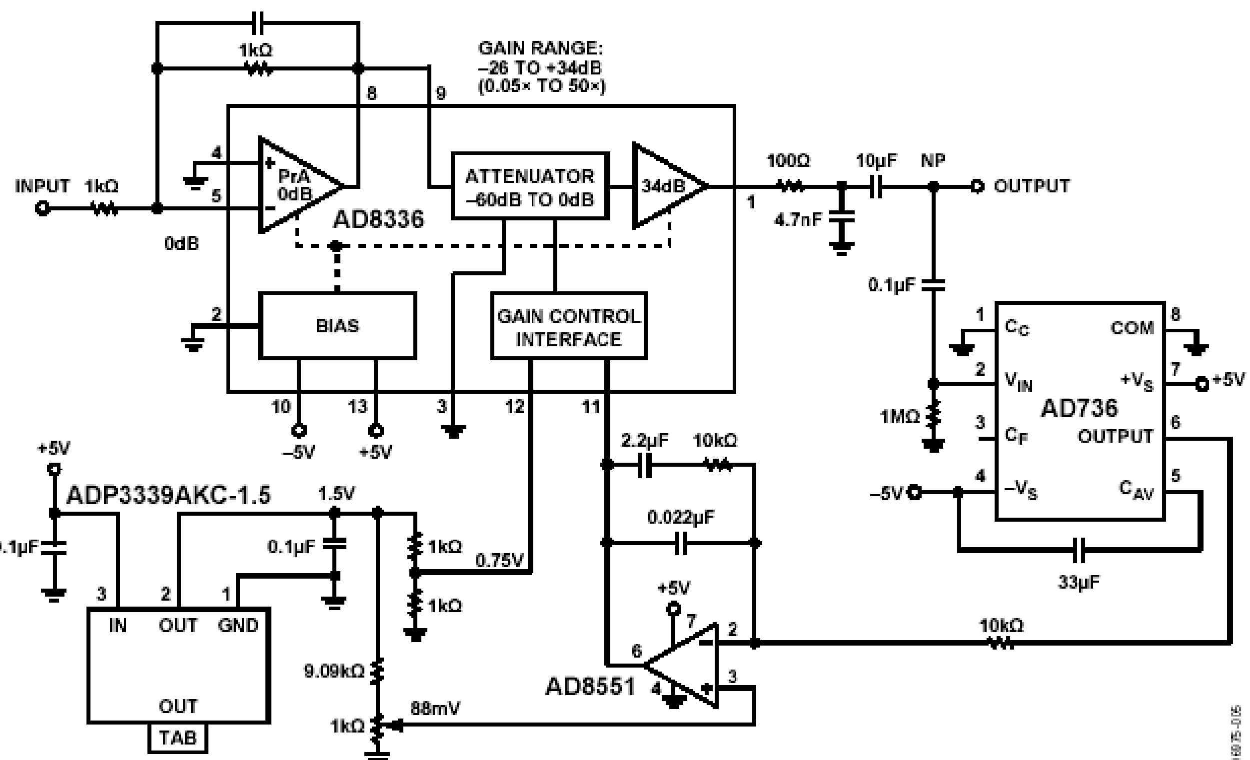
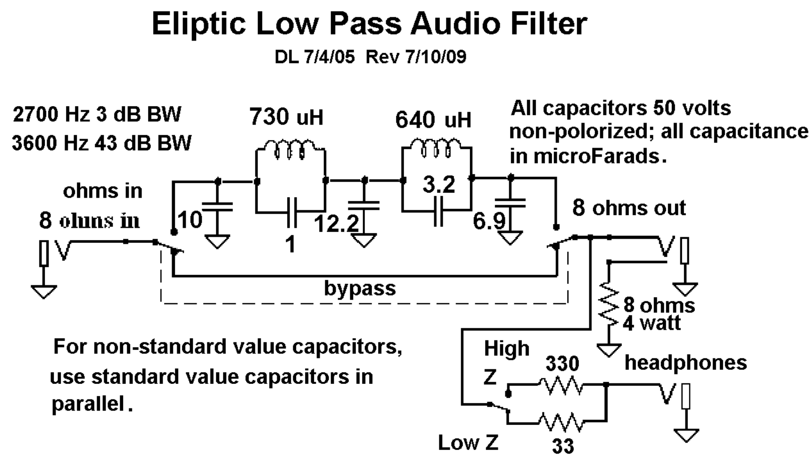


Figure 5. AGC Circuit Using the AD8336

Elliptic Low Pass Audio Filters

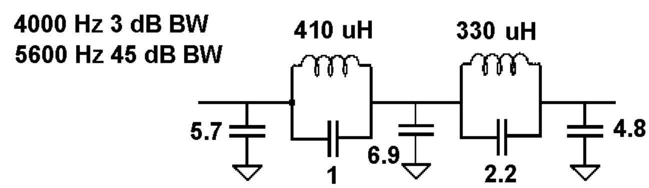
Dallas Lankford 7/4/05 Rev 7/10/09



Option

There are a number of different accessories for improving recovered audio from strongly fading MW and SW signals, including audio filters and AM synchronous detectors, for which designers, sellers, and users have often given glowing praise. But I have heard only modest improvement due to these kinds of devices when the receiver used with the devices had a suitably slow AGC release time. The distortion which one hears from strongly fading MW and SW signals manifests itself as high frequency sound akin to noise. So an appropriate low pass audio filter should substantially improve the audio quality of strongly fading MW and SW signals. However, the audio filters I tried did not.

A wider BW filter does not improve recovered audio as much as the 2700 Hz BW filter above. Nevertheless, below is a 4000 Hz BW filter for those who desire one. It can be used in place of the bypass, or a 3 rd option as I did for two of the prototypes.

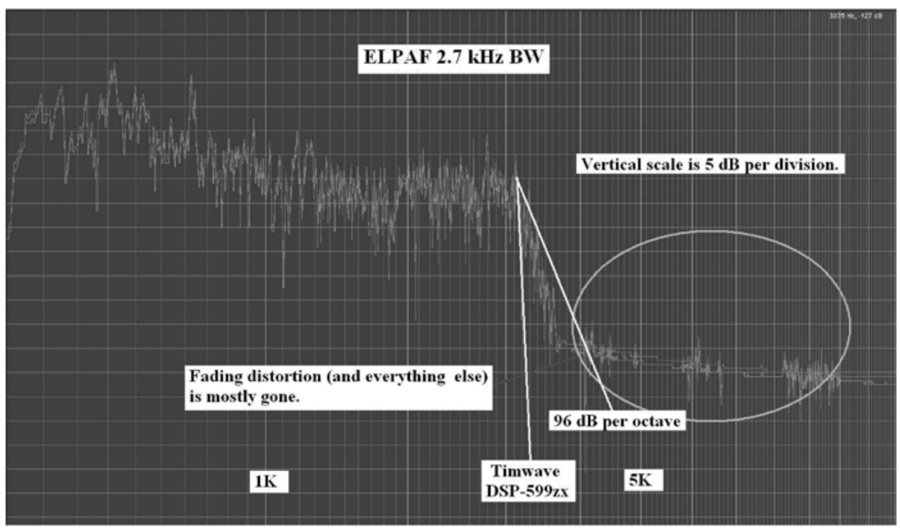


I suspected that either the cutoff frequencies of the filters were not appropriate, or the shape factors (roll off) of the filters were not appropriate, or both, or other factors were responsible for the lack of audio improvement. Also, many audio filters roll off the low frequencies, giving the recovered audio a tinny sound which degrades audio quality in my opinion. And there are no digital artifacts with the elliptic low pass audio filter as there were with a Timewave DSP-599zx filter which I used some time ago. Rather than spend thousands of dollars (Stanford Research 640) for audio filters with variable cutoff frequencies and steep attenuation roll off, which still might not be satisfactory and which might require additional circuits, I decided to design and build such audio filters myself; see the schematics above.

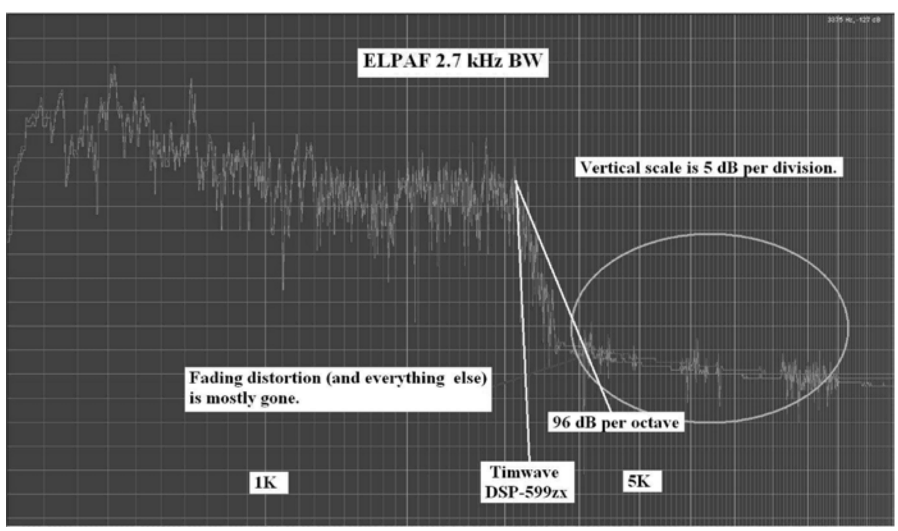
After considering possible kinds of audio filters, it seemed to me that elliptic filters, if they could be scaled to practical audio frequencies, might provide the sharp cutoff I wanted. Also, elliptic filter tables (in the ARRL Radio Amateur's Handbooks) allowed me to easily determine component values for whatever cutoff frequencies I wanted, and to scale the values for 8 ohms. As it turned out, the component values for 2.7 kHz and 4.0 kHz cutoff frequency elliptic low pass filters were practical. The voltage for 2 watts into 8 ohms is 4 volts. I used 50 volt non-polarize capacitors for greater power handling capability. The current for 2 watts into 8 ohms is 500 mA. No high Q and high current off-the-shelf inductors were found. So the inductors were wound on Amidon FT-82-61 toroids (u = 125) using #24 enameled copper wire. There is no formula for calculating the number of turns for multi-layer toroids. So the numbers of turns were found by trial and error using an Almost All Digital Electronics inductance meter (the current version is L/C Meter IIB; www.aade.com).

I used between 2 and 3 close wound (on the inside circumference) layers of #24 enameled copper wire, about 95 turns for 740 uH and about 89 turns for 630 uH. I used between 5 and 6 feet of #24 wire for each inductor, which provided excess wire (better too much than not enough). For 410 uH and 330 uH the turns were 72 and 65 respectively. The 8 ohm resistor at the output (which is switched by the 1/4 inch plug) provides the necessary 8 ohms filter termination when a speaker is not used. The switched resistors in the headphones line are for high and low impedance headphones. The filter input must be connected to the 8 ohms speaker output of the receiver; otherwise the filter performance will be degraded.

An audio spectrum analyzer was used to examine the shape factors of the elliptic low pass audio filters and to study the nature of fading distortion. The best audio spectrum analyzer we found for these purposes was WavePad; see www.nch.com.au/wavepad . Two "snapshots" of the WavePad spectrum display are given below. The first spectrum snapshot, without filtering, shows where most of the fading distortion occurs, in the frequency range above about 3500 Hz. The second spectrum snapshot shows the audio typically obtained with the 2700 Hz bandwidth elliptic low pass audio filter. The shape of the 2700 Hz bandwidth filter is better than 96 dB per octave (as indicated on the snapshot), but not as good as the Timewave DSP-599zx (also indicated on the snapshot). Also, the stopband of the 2700 Hz filter is not nearly as good as the DSP-599zx. Nevertheless, the 2700 Hz elliptic low pass audio filter stopband has been more than adequate for substantially reducing, if not totally eliminating, fading distortion.



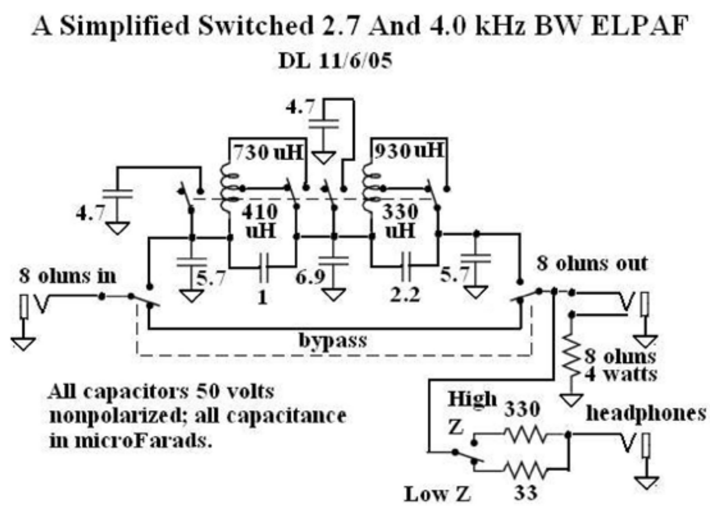
Winding 4 toroids with between 65 and 96 turns of #24 enameled copper wire is quite tedious. So I thought about how I could implement the 2.7 and 4.0 kHz BW filters using only two toroids with taps. As it turned out it can easily be done in an obvious manner (using two tapped toroids, and switching capacitors) with the with a 6 pole double throw toggle switch.



But I could find none in my catalogs. A high quality 6 pole double throw rotary switch is available for about $30, Electroswitch part # D4G0603N (Mouser 690D4G0603N), so the switched filter can be implemented that way By taking some minor liberties with the filters designs, a 4 pole double throw toggle switch can be used.

I used Allied Electronics 6763280 large lever C&K toggle switches, about $16, rated at 100,000 make and break cycles

The bypass shown in the schematic to the right is optional. Because a 5th switch position is not available with the toggle switch, the 2.2 mF cap cannot be switched to 3.2 mF; instead the 640 uH inductor is increased to 930 uH to give the same resonant frequency for that parallel LC. And because there is no 6th toggle switch position, the output cap cannot be switched; instead the intermediate value of 5.7 mF is used in place of switching 4.7 mF and 6.9 mF.



Adjustable Frequency Notch Filters For 1 MHz, 100 kHz, 10 kHz And 5 kHz

From Analog Applications Journal High-Speed Notch Filters **By Bruce Carter** (Email: r-carter5@ti.com) Low-Power Wireless Applications
Reproduced with permission by Perry Sandeen 8/09

Introduction

Active notch filters have been used in the past for applications like elimination of 50- and 60-Hz hum components. They have proven to be somewhat problematic from the standpoints of center frequency (f_0) tuning, stability, and repeatability. The advent of high-speed amplifiers opens the possibility of higher-speed notch filters—but are they actually producible? This article will show what is presently possible and what design trade-offs a designer will face with real-world components. As a review, the reader should remember some characteristics of the notch filter:

• The depth of the notch obtainable in simulations like that shown in Figure 1 is *not* the depth that can be achieved with real-world components. The best that the designer can hope for is 40 to 50 dB.

• Instead of focusing on notch depth, the designer should focus on center frequency and Q. The Q for a given notch filter is the –3-dB point, *not* the notch depth or a point 3 dB above the notch depth, as shown in Figure 2.

Remember that the designer's objective is not a notch filter but the rejection of a specific interfering frequency. Any filter that does not reject that interfering frequency because it misses the frequency or has too little rejection at that frequency is not much use.

Figure 1 Simulated notch depth

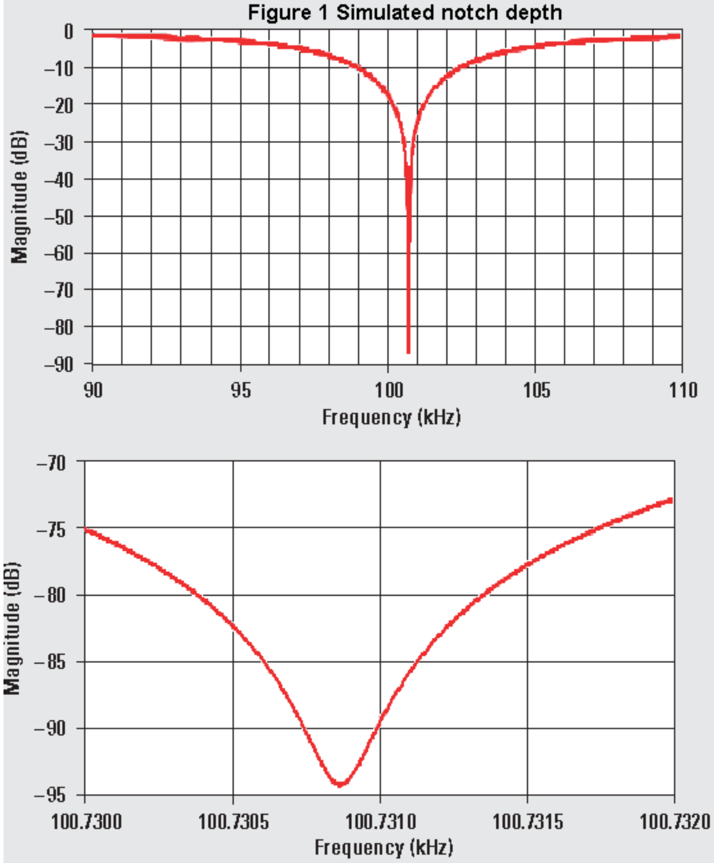
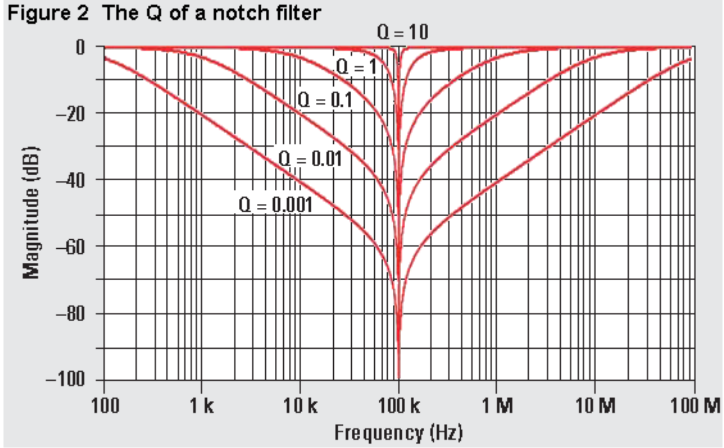


Figure 2 The Q of a notch filter



The best way to avoid missing the interfering frequency is to select the best values of R and C from the start. The RC Calculator under "Filter Design Utilities" in Reference 1 should be used to find the correct values of R0 and C0 for the circuits in the following discussion.

Topology

A number of notch-filter topologies were explored. Some design goals are a topology that:
• produces a notch (as opposed to band rejection);
• uses a single op amp;
• can be easily tuned with independent adjustments for center frequency and Q;
• can operate from a single-supply voltage; and • can be adapted to fully differential op amps.

• It requires six high-precision components for tuning, and two of those are ratios of the others. If the designer wants to get away from ratios, eight precision components are required. R0/2 = two R0 in parallel, and 2 × C0 = two C0 in parallel.

Unfortunately, it was not possible to achieve all of these, although some desirable circuits can be constructed that can meet some of these goals.

The twin-T topology of Figure 3 deserves an honorable mention here, because a notch filter can be implemented with a single op amp. It is not as flexible as one would hope, because the center frequency is not easily adjustable. Trimming the center frequency involves simultaneous adjustment of the three R_0 resistors. This is a concern because triple potentiometers are large, expensive, and may not track very well—especially the section that has to be one-half the value of the other two. Mismatches in the R_0 resistors will very quickly erode notch depth to less than 10 dB.

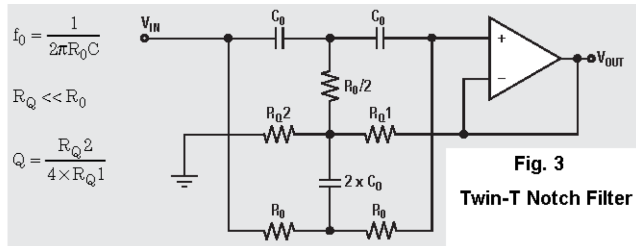


Fig. 3
Twin-T Notch Filter

The circuit has some other disadvantages as well:

• The twin-T topology is not easily adaptable to single supply operation and cannot be used with a fully differential amplifier.

• The spread of resistor values becomes large due to the requirement of RQ << R0. The spread of the resistor values has a bearing on the depth of the notch and on center frequency.

Nevertheless, for applications where only a single op amp can be used, the twin-T topology is quite usable if the designer matches components or buys very high-precision components.

Fliege notch filter

The Fliege notch topology is shown in **Figure 4**. The advantages of this circuit over the twin-T are as follows:

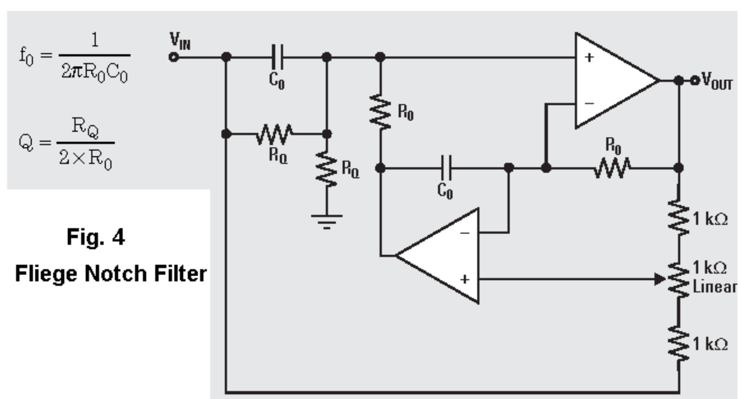


Fig. 4
Fliege Notch Filter

• Only four precision components—two R_S and two C_S— are required for tuning the center frequency. One nice feature of this circuit is that slight mismatches of components are okay—the center frequency will be affected, but not the notch depth.

Table 1. Component values for the Fliege notch filter

Q	1 MHz			100 kHz			10 kHz			5 kHz		
	R_0 (kΩ)	C_0 (pF)	R_Q (kΩ)	R_0 (kΩ)	C_0 (nF)	R_Q (kΩ)	R_0 (kΩ)	C_0 (nF)	R_Q (kΩ)	R_0 (kΩ)	C_0 (pF)	R_Q (kΩ)
100	1.58	100	316	1.58	1	316	1.58	10	316	118	270	2 M
10	1.58	100	31.6	1.58	1	31.6	15.8	1	316	118	270	
1	1.58	100	3.16	1.58	1	3.16	15.8	1	31.6	118	270	

0.01uF / MFD 10nF 10000pF (MMFD) 0.001uF / MFD 1nF 1000pF (MMFD)

• The Q of the filter can be adjusted independently from the center frequency by using two non-critical resistors of the same value.

• The center frequency of the filter can be adjusted over a narrow range without seriously eroding the depth of the notch.

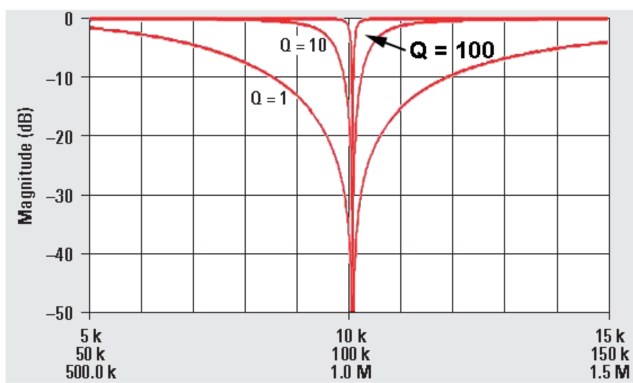
Unfortunately, this circuit uses two op amps instead of one, and it cannot be implemented with a fully differential amplifier.

Simulations

Simulations were first performed with ideal op amp models. Real op amp models were later used, which produced results similar to those observed in the lab. Table 1 shows the component values that were used for the schematic in **Figure 4**. There was no point in performing simulations at or above 10 MHz because lab tests were actually done first, and 1 MHz was the top frequency at which a notch filter worked.

A word about capacitors: Although the capacitance is just a value for simulations, actual capacitors are constructed of different dielectric materials. For 10 kHz, resistor value spread constrained the capacitor to a value of 10 nF. While this worked perfectly well in simulation, it forced a change from an NPO dielectric to an X7R dielectric in the lab— with the result that the notch filter completely lost its characteristic. Measurements of the 10-nF capacitors used were close in value, so the loss of notch response was most likely due to poor dielectric.

Fig. 5 Simulation Results Before Tuning



The circuit had to revert to the values for a Q of 10, and a 3-M R_Q was used. For real-world circuits, it is best to stay with NPO capacitors. The component values in Table 1 were used both in simulations and in lab testing. Initially, the simulations were done without the 1-k potentiometer (the two 1-k fixed resistors were connected directly together and to the non-inverting input of the bottom op amp). Simulation results are shown in **Figure 5**.

There are actually nine sets of results in **Figure 5**, but the curves for each Q value overlie those at the other frequencies. The center frequency in each case is slightly above a design goal of 10 kHz, 100 kHz, or 1 MHz. This is as close as a designer can get with a standard E96 resistor and E12 capacitor.

Consider the case of 100 kHz:

$$f_0 = \frac{1}{2\pi R_0 C_0} = \frac{1}{2\pi \times 1.58\ k\Omega \times 1\ nF} = 100.731\ kHz$$

A closer combination exists if E24 sequence capacitors are available:

$$f_0 = \frac{1}{2\pi R_0 C_0} = \frac{1}{2\pi \times 4.42\ k\Omega \times 360\ pF} = 100.022\ kHz$$

The inclusion of E24 sequence capacitors can lead to more accurate center frequencies in many cases, but procuring the E24 sequence values is considered an expensive (and unwarranted) expenditure in many labs. While it may be easy to specify E24 capacitor values in theory, in practice many of them are seldom used and have long lead times associated with them. There are easier alternatives to selecting E24 capacitor values. Close examination of **Figure 5** shows that the notch misses the center frequency by only a small amount. At lower Q values, there is still substantial rejection of the desired frequency. If the rejection is not sufficient, then it becomes necessary to tune the notch filter.

Again considering the case of 100 kHz, we see that the response near 100 kHz is spread out in **Figure 6**. The family of curves to the left and right of the center frequency (100.731 kHz) represents filter response when the 1-k potentiometer is inserted and adjusted in 1% increments.

When the potentiometer is exactly in the middle, the notch filter rejects frequencies at the exact center frequency. The depth of the simulated notch is actually on the order of 95 dB, but that is not going to happen in the real world. A 1% adjustment of the potentiometer puts a notch that is greater than 40 dB right on the desired frequency. Again, this is best-case with ideal components, but lab results are close at low frequencies (10 and 100 kHz).

Fig. 6. Tuning For Center Frequency

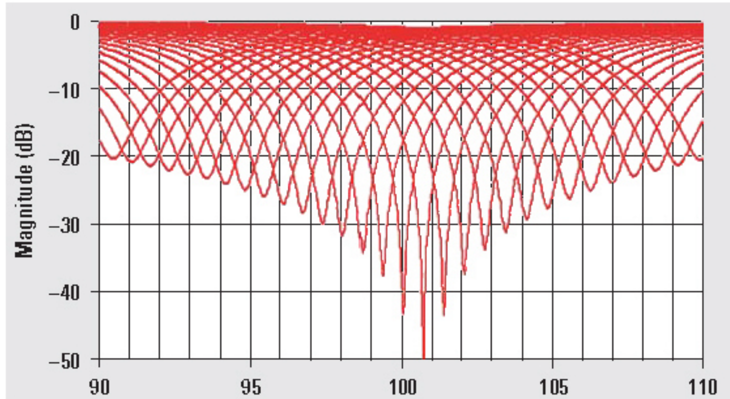
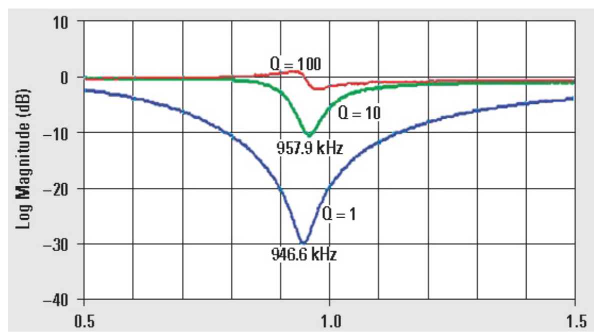


Figure 6 shows that it is important to get close to the correct frequency with R_0 and C_0 from the start. While the potentiometer can correct for frequency over a broad range, the depth of the notch degrades. Over a small range (±1%), it is possible to get a 100:1 rejection of the undesirable frequency; but over a larger range (±10%), only a 10:1 rejection is possible.

Lab results

A THS4032 evaluation board was used to construct the circuit in **Figure 4**. Its general-purpose layout required only three jumpers and one trace cut to complete the circuit. The component values in Table 1 were used, starting with those that would produce 1 MHz. The intention was to look for bandwidth/slew-rate restrictions at 1 MHz and test at lower or higher frequencies as necessary.

Fig. 7. Lab Results At 1 MHz



Results at 1 MHz

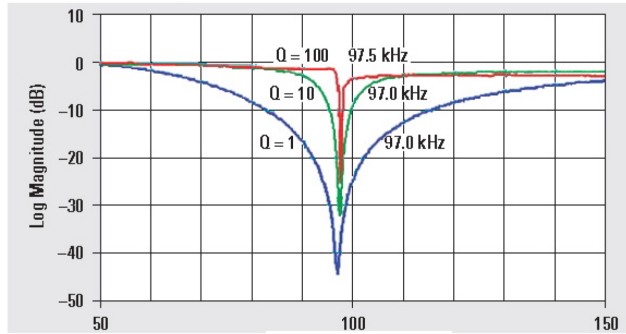
Figure 7 shows that there are some very definite bandwidth and/or slew-rate effects at 1 MHz. The response curve at a Q of 100 shows barely a ripple where the notch should be. At a Q of 10, there is only a 10-dB notch, and a 30-dB notch at a Q of 1. Apparently notch filters cannot achieve as high a frequency as one would hope, but the THS4032 is only a 100-MHz device. It is reasonable to expect better performance from parts with a greater unity-gain bandwidth. Unity-gain stability is important, because the Fliege topology has fixed unity gain. If the designer wishes to estimate what bandwidth is required for a notch at a given frequency, a good place to start is the gain/bandwidth product given in the datasheet, which should be 100 times the center frequency of the notch. Additional bandwidth will be required for higher Q values. There is a slight frequency shift of the notch center as Q is changed. This is similar to the frequency shift seen for bandpass filters. The frequency shift is less for notch filters centered at 100 kHz and 10 kHz, as shown in **Figure 8** and later in **Figure 10**.

Results at 100 kHz

Component values from **Table 1** were then used to create 100-kHz notch filters with different Qs. The results are shown in Figure 8. It is immediately obvious that viable notch filters can be constructed with a center frequency of 100 kHz, although the notch depth appears to be less at higher values of Q.

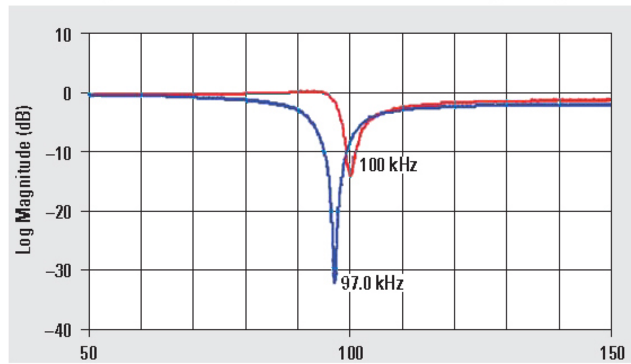
Remember, though, that the design goal here is a 100-kHz—not a 97-kHz—notch. The component values selected were the same as for the simulation, so the notch center frequency should theoretically be at 100.731 kHz; but the difference is explained by the parts used in the lab. The mean value of the 1000-pF capacitor stock was 1030 pF, and of the 1.58-k resistor stock was 1.583 k.

Fig. 8. Lab Results At 100 kHz



When the center frequency is calculated with these values, it comes out to 97.14 kHz. The actual components, however, could not be measured (the board was too fragile). As long as the capacitors are matched, it would be possible to go up a couple of standard E96 resistor values to get closer to 100 kHz. Of course, this is probably not an option in high-volume manufacturing, where 10% capacitors could come from any batch and potentially from different manufacturers.

Fig. 9. Tuning For Exact Center Frequency



The range of center frequencies will be determined by the tolerances of R0 and C0, which is not good news if a high Q notch is required.

There are three ways of handling this:
• Purchase higher-precision resistors and capacitors;
• lower the Q requirement and live with less rejection of the unwanted frequency; or
• tune the circuit (which is explored next).

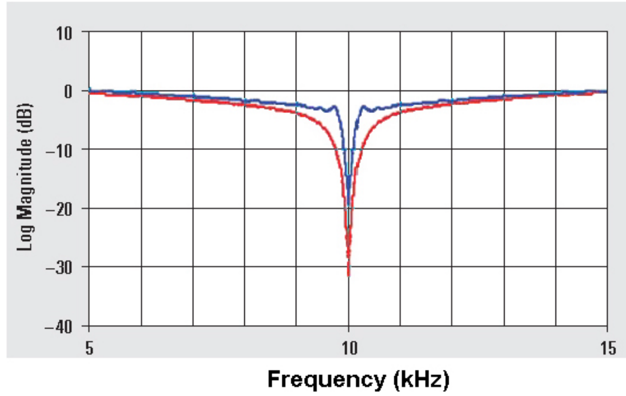
At this point, the circuit was modified to have a Q of 10, and a 1-k potentiometer was added for tuning the center frequency (as shown in Figure 4). In real-world design, the potentiometer value selected should slightly more than cover the range of center frequencies possible with worst-case R0 and C0 tolerances. That was not done here, as this was an exercise in determining possibilities, and 1 k was the lowest potentiometer value available in the lab. When the circuit was tuned for a center frequency of 100 kHz as shown in Figure 9, the notch depth degraded from 32 dB to 14 dB. Remember that this notch depth could be greatly improved by making the initial f0 closer to ideal. The potentiometer is meant to tune over only a small range of center frequencies. Still, a 5:1 rejection of an unwanted frequency is respectable and may be sufficient for some applications. More critical applications will obviously need higher-precision components. Op amp bandwidth limitations, which will also degrade the tuned notch depth, may also be keeping the notch depth from being as low as possible. With this in mind, the circuit was retuned for a center frequency of 10 kHz.

Results at 10 kHz

Figure 10 shows that the notch depth for a Q of 10 has increased to 32 dB, which is about what one would expect from a center frequency 4% off from the simulation (Figure 6). The op amp was indeed limiting the notch depth at a center frequency of 100 kHz! A 32-dB notch is a rejection of 40:1, which is quite good.

So even with components that produced an initial 4% error, it was possible to produce a 32-dB notch at the desired center frequency. The bad news is that to escape op amp bandwidth limitations, the highest notch frequency possible with a 100-MHz op amp is somewhere between 10 and 100 kHz. In the case of notch filters, "high-speed" is therefore defined as being somewhere in the tens or hundreds of kilohertz.

Fig. 10. Lab Results At 10 kHz*



A good application for 10-kHz notch filters is AM (medium-wave) receivers, where the carrier from adjacent stations produces a loud 10-kHz whine in the audio, particularly at night. This can really grate on one's nerves when listening is prolonged.

*Some artistic liberties were taken with this plot. The laboratory instrument displays values only down to 10 kHz, so the left-hand portion of the plot is a mirror image of the right-hand portion. The laboratory instrument also has some roll-off at frequencies below 100 kHz, which was artistically eliminated from this plot.

Fig. 11. AM Reception With And Without Notch Filter

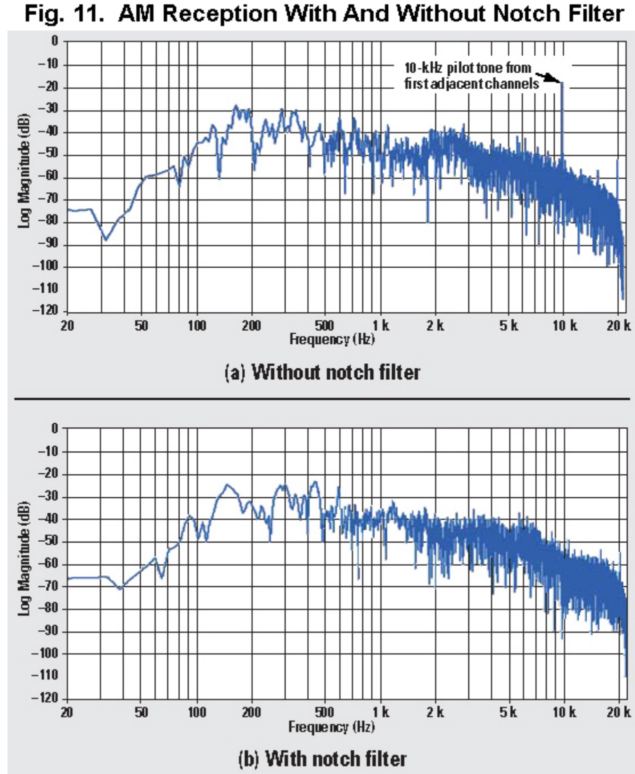


Figure 11 shows the received audio spectrum of a station before and after the 10-kHz notch was applied. Note that the 10-kHz whine is the loudest portion of the received audio (Figure 11a), although the human ear is less sensitive to it. This audio spectrum was taken at night on a local station that had two strong stations on either side. FCC regulations allow for some variation of the station carriers. Therefore, slight errors in carrier frequency of the two adjacent stations will make the 10-kHz tones heterodyne, increasing the unpleasant listening sensation. When the notch filter is applied (Figure 11b), the 10-kHz tone is reduced to the same level as that of the surrounding modulation. Also visible on the audio spectrum are 20-kHz carriers from stations two channels away and a 16-kHz tone from a transatlantic station. These are not a problem, because they are attenuated substantially by the receiver IF. A frequency of 20 kHz is inaudible to the vast majority of people in any event.

Figure 12 shows the same spectrum on a waterfall diagram. In this case, the sample window is widened, and the 10-kHz carrier interference is shown as a string of peaks that vary in amplitude. When the notch is applied, the 10-kHz peaks are eliminated, and there is only a slight ripple in the received audio where 10 kHz has been notched out. For European readers who want to have a more pleasing medium-wave listening experience, the component values are C_0 = 330 pF, R_0 = 53.6 k, and R_Q = 1 M. Shortwave listeners will benefit from a two-stage notch filter, one stage being the 10-kHz previously described, and the other stage being a 5-kHz notch filter with component values of C_0 = 270 pF, R_0 = 118 k, and R_Q = 2 M.

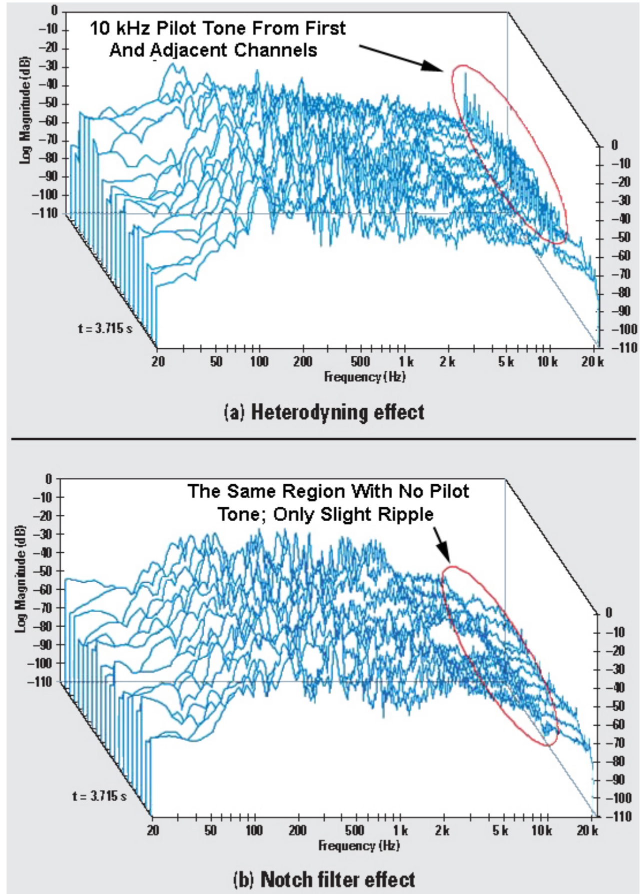
Applicability

Although testing described in this article was performed on the THS4032, the application circuits are usable with all single-ended, unity-gain, voltage feedback op amps. A key specification is unity gain bandwidth, which should be from 100 to 1000 times the center frequency. The Fliege notch filter cannot be constructed from current-feedback amplifiers or from fully differential op amps.

Conclusion

High-speed op amps have been used to produce low-pass and high-pass filters up to the tens of megahertz with fairly good success. Narrow bandpass filters and notch filters are much less understood and much more critical applications. While the tolerance of a capacitor might change the cutoff frequency of a low-pass filter or produce ripple in the passband, that same tolerance can produce dramatic changes in the center frequency and notch depth of a notch filter.

Fig. 12. Heterodyning And Notch Filter Effects



(a) Heterodyning effect

(b) Notch filter effect

With a Fliege notch topology, the number of critical components is reduced to four—two identical Rs and two identical Cs. Fortunately for the designer, there is an inherent matching that occurs when devices are manufactured at the same time, so it is possible to construct notch filters from them even if the tolerance given in the datasheet does not imply matching. There is good, independent control over the center frequency and Q, with the possibility of tuning over a narrow range, which compensates for the initial tolerance errors.

A 1-MHz, Q = 1 notch filter constructed with a 100-MHz op amp showed poor performance at higher values of Q. The same op amp did better at 100 kHz but still showed degradation at higher Q values, particularly when the center frequency was tuned. It was not until the center frequency was decreased to 10 kHz that performance close to simulation results was obtained. Limiting the notch filter to high tens to low hundreds of kilohertz (for faster parts) eliminates many applications. These frequencies, however, represent the state of the art in design for these unusual filters.

With a Fliege notch topology, the number of critical components is reduced to four—two identical Rs and two identical C_S. Fortunately for the designer, there is an inherent matching that occurs when devices are manufactured at the same time, so it is possible to construct notch filters from them even if the tolerance given in the datasheet does not imply matching. There is good, independent control over the center frequency and Q, with the possibility of tuning over a narrow range, which compensates for the initial tolerance errors.

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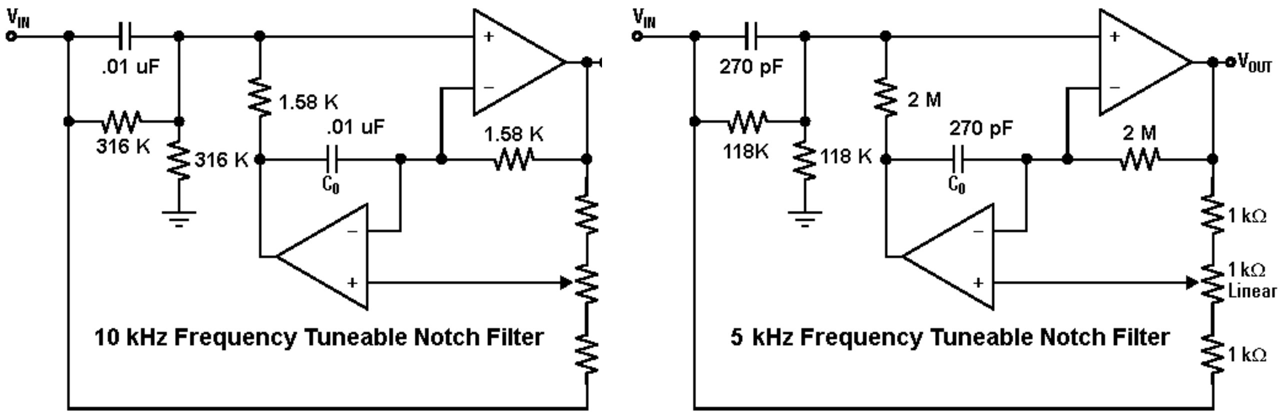
Reference
1. "Amplifiers and Linear Engineering Design Utilities," www.ti.com/amplifier_utilities
Related Web sites   amplifier.ti.com   www.ti.com/sc/device/THS4032



BCB Filter

Dealing With AM Broadcast Band Interference to Your Receiver
TECHNOTE No. 6 Joe Carr's Radio Tech-Notes
Universal Radio Research 6830 Americana Parkway Reynoldsburg, Ohio 43068

If you live anywhere near an AM broadcast band (BCB) station, then you might have serious reception problems, even with a high quality receiver. Although one of the things you get when you pay the premium price for a high quality receiver is superior overload protection, the signal from a local AM BCB station might overwhelm its defenses.

In the USA we use the 530-1700 Khz medium wave AM BCB. Most stations operate with 1,000 to 10,000 watts of RF output power (although a few 250 to 500 watt local fizzlers also exist). A few stations are designated "clear channel" stations, and operate with 50,000 watts, 24-hours a day. These stations (e.g. WSM Nashville, 650 Khz) are on frequencies that are not assigned to other stations for a distance of, I believe, 1300 miles radius. If you live within a few hundred yards of an AM BCB station, then it's possible to see more than one volt of RF appearing at your receiver antenna terminals (a laboratory measured 4 volts in one case!). Given that your receiver likes to see signals in the dozens of microvolts level, then you can understand the problem.

The Problem

So what is the problem? Your receiver, no matter what frequency it receives, is designed to accept only a certain maximum amount of radio frequency energy in the front-end. If more energy is present, then one or more of several overload conditions results. The overload could result from a desired station is too strong. In other cases, there are simply too many signals within the passband for the receiver front-end to accommodate. In still other cases, a strong out-of-band signal is present. Figure I shows several conditions that your receiver might have to survive. Several different receiver problems result from the various types of overload, all of which are species intermodulation and/or crossmodulation.

If you tune across the shortwave bands, especially those below 10 or 12 Mhz, and note an AM BCB signal that seems like it is hundreds of kilohertz wide, then you are witnessing blanketing. If drives the mixer or RF amplifier of the receiver clean out of its mind, producing a huge number of spurious signals, and apparently a very wide bandwidth.

Your receiver can only accommodate a certain amount of RF energy in its front-end circuits. This level is expressed in the dynamic range specification of the receiver, and is hinted by the third-order intercept point (TOIP) and -I dB compression point specifications. The strong out-of-band signal takes up so much of the receiver's dynamic range "head room" that only a small amount of capacity remains for the desired signal. The signal level of the desired signal is thereby reduced to a smaller level. In some cases, the overload is so severe that the desired signal becomes inaudible. If you can filter out or otherwise attenuate the strong out-of-band signal, then the head room is restored, and the receiver has plenty of capacity to accommodate both signals.

One thing that's important today is what happens when signals are received that are much stronger than the input signal that produces the flattening of the response in the output-vs-input curve. One unfortunate factor is the generation of harmonics that were not present in the original signal. The harmonics are integer multiples of the input signal frequency, so will appear at higher points on the frequency dial. The harmonics may fall within the passband of your receiver, and are seen as valid signals even though they were generated in the receiver itself!

The strong intermodulation products are created when two of these signals heterodyne together. The heterodyne ("mixing") action occurs because the receiver front-end is non-linear at this point. The frequencies produced by just two input frequencies (Fl and F2) are described by mF1 $\pm$ nF2, where m and n are integers. As you can see, depending on how many frequencies are present and how strong they are, a huge number of spurious signals can be generated by the receiver front-end.

So, what about IF selectivity? You have an IF filter of 270 Hz to 8 Khz (depending on model and mode), so why doesn't it reject the dirty smelly bad-guy signals? The problem is that the damage occurs in the front-end section of the receiver, before the signals encounter the IF selectivity filters.

The problem that causes all of these problems is an overdriven RF amplifier, mixer or both. The only really effective way to deal with these problems is to reduce the level of the offending signals. In this paper we will investigate the use of front-end attenuators, high-pass filtering and single-frequency wavetraps.

The Attenuator Solution

Some modem receivers are equipped with one or more switchable attenuators in the front-end. Some receivers also include an RF gain control that sometimes operates in the same manner. Some receiver operators use external in-line single-range or switchable attenuators for exactly the same purpose. The idea behind the attenuator is to reduce all of the signals to the front-end enough to drop the overall energy in the circuit to below the maximum level that can be accommodated without either overload or intermodulation occurring at significant levels. The attenuator reduces both desired and undesired signals, but the perceived ratio is altered when the receiver front-end is de-loaded to a point where desensitization occurs, or intermods and harmonics pop up.

The Antenna Solution

The antenna that you select can make some difference in AM BCB problems. Generally, a resonant HF antenna with its end nulls pointed toward the offending station will provide marginally better performance than a random length wire antenna (which are popular amongst SWLs).

Also, it is well known that vertical HF antennas are more susceptible to AM BCB because they respond better to the ground wave electrical field generated by the BCB station (DeMaw).

The Filter Solution

One of the best solutions is to filter out the offending signals before they hit the receiver front-end, while affecting the desired signals minimally. This task is not possible with the attenuator solution, which is an "equal opportunity" situation because it affects all signals equally. A signal that is outside the passband of a frequency selective filter: it is severely attenuated. It does not drop to zero, but the reduction can be quite profound in some designs.

Signals within the receiver's passband are not unaffected by the filter. The loss for in-band signals is, however, considerably less than for out-of-band signals. This loss is called **insertion loss**, and is usually quite small (1 or 2 dB) compared to the loss for out-of-band signals (lots of dB).

Several different types of filters are used in reducing interference. A *high-pass* filter passes all signals above a specified cut-off frequency (Fc). The *low-pass* filter passes all signals below the cut-off frequency. This filter is similar to the type of filter that hams using HF transmitters place between the transmitter and antenna to prevent harmonic radiation from interfering with television operation.

A *bandpass filter* passes signals between a lower cut-off frequency (FL) and an upper cut-off frequency (FH). A *stop-band* filter is just the opposite of a bandpass filter: it stops signals on frequencies between FL and FH, while passing all others. A *notch* filter, also called a wave trap, will stop a particular frequency (Fo), but not a wide band of frequencies as does the *stopband* filter. In all cases, these filters stop the frequencies in the designated band, while passing all others. More or less.

The positioning of the filter in your antenna system is shown in Fig. 1 below. The ideal location is as close as possible to the antenna input connector.

The best practice, if you have the space at your operating position, is to use a double-male coaxial connector to connect filter output connector to the antenna connector on the receiver.

A short piece of coaxial cable can connect the two terminals if this approach is not suitable in your case. Be sure to ground both the ground terminal on the receiver and the ground terminal of the filter (if one is provided). Otherwise, depend on the coaxial connectors' outer shell making the ground connection.

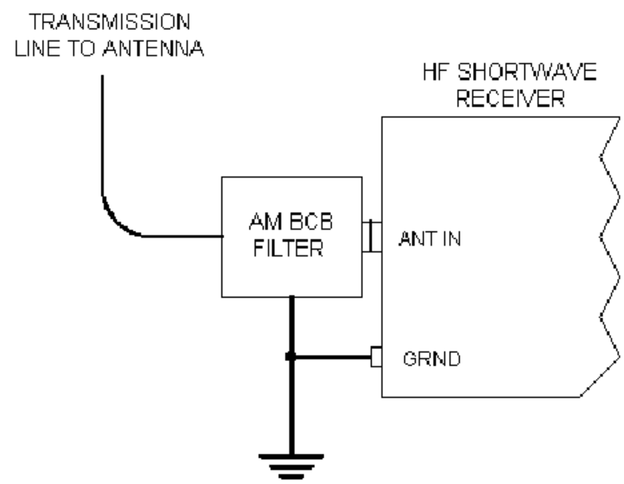


FIG. I

Different Forms of Filters: Wave traps

A wave trap is a tuned circuit that causes a specific frequency to be rejected. Two forms are used: *series tune*d (Fig. 2A) and *parallel tuned* (Fig. 2B).

The series tuned version of the wavetrap is placed across the signal line (as in Fig. 2A), and works because it produces a very low impedance at its resonant frequency and a high impedance at frequencies removed from resonance. As a result, the interfering signal will see a resonant series-tuned wave trap as a short circuit, while other frequencies do not see it at all.

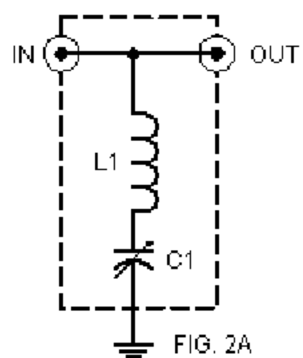


FIG. 2A

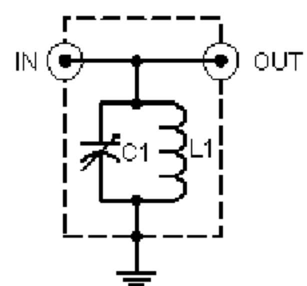


FIG. 2B

The parallel resonant form placed in series with the antenna line (as in Fig. 2B). It provides a high impedance to its resonant frequency, so will block the offending signal before it reaches the receiver. It provides a low impedance to frequencies removed from resonance.

The wave traps are useful in situations where a single station is causing a problem, and you don't want to eliminate nearby stations. For example, if you live close to an MW AM BCB signal and don't want to interrupt reception of other MW AM BCB signals or LW AM BCB signals. Even AM BCB DXers often have wavetraps tuned to the frequency of a nearby station in order to either increase the available dynamic range or eliminate other problems.

The values of components shown in Figs. 2A and 2B are suitable for the MW AM BCB, but can be scaled to the LW BCB if desired. A good starting point is 365 pF for the variable capacitor, and 200 µH or 220 µH for the inductor. Both components should be variable in order to make tuning easier. However, if one component is fixed, the filter will do its job just as well.

If there are two stations causing significant interference, then two wave traps will have to be provided, separated by a short piece of coaxial cable. In that case, use a parallel tuned wave trap for one frequency, and a series tuned wave trap for the other. Otherwise, interaction between the wave traps will cause problems.

High-Pass Filters. One very significant solution is to use a high-pass filter with a cut-off frequency between 1700 and 3000 Khz. It will pass the shortwave frequencies, and severely attenuate AM BCB signals in both MW and LW bands, causing the desired improvement in performance. Figure 3 shows a design used for many decades. It is easily built because the capacitor values are 0.001 µF and 0.002 µF (which some people make by parallelling two 0.001 µF capacitors).

The inductors are both 3.3 µH, so can be made with toroidal cores. If the T-50-2 RED cores are used (AL = 49), then 26 turns of small diameter enameled will suffice. Or if the T-50-15 RED/WHITE cores are used (AL = 135), then 15 turns are used. The circuit of Fig. 3 produces pretty decent results for low effort.

A number of readers contacted me with success stories when this circuit was published once before, a result that gives me pleasure

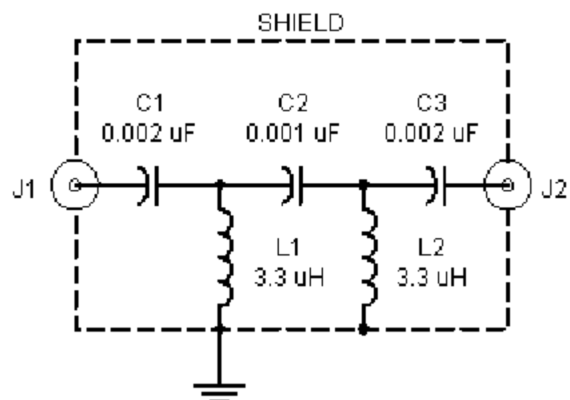


FIG. 3

. But there is a better way....

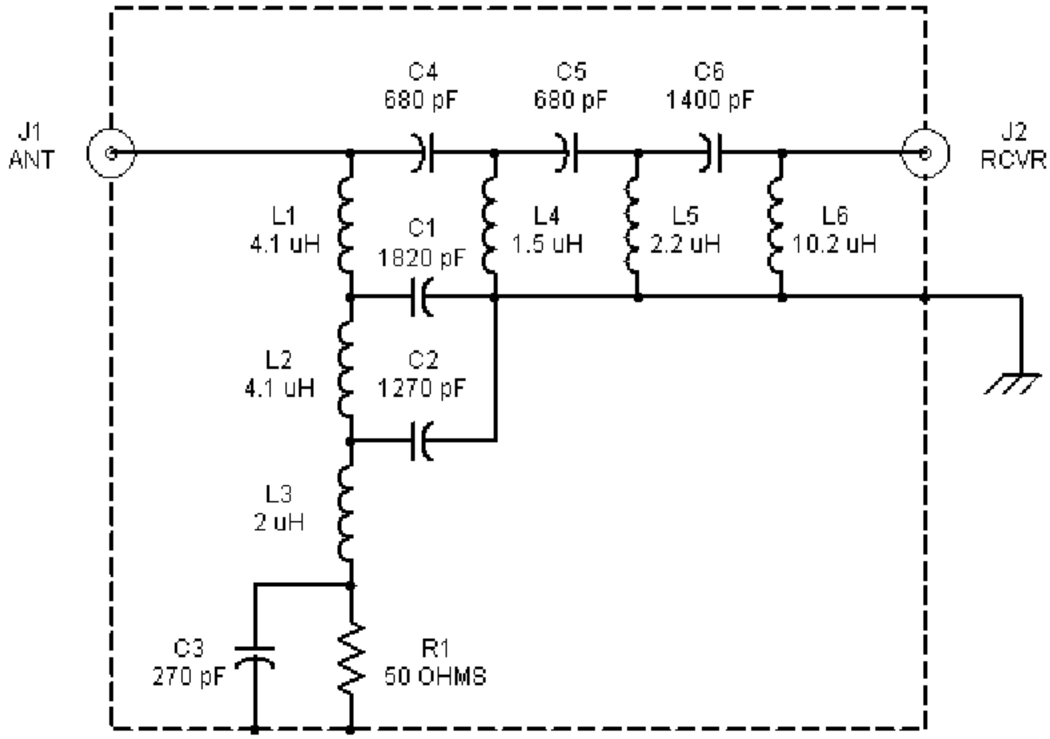
Absorptive Filters. The <u>absorptive filter</u> (Orr 1996 and Weinreich/Carroll 1968) solves a problem with the straight high-pass filter method, and produces generally better results at the cost of more complexity. This filter (Fig. 4) consists of a high-pass filter (C4-C6/L4-L6) between the antenna input (J1) and the receiver output (12). It passes signals above 3 Mhz and rejects those below that cut-off frequency. It also has a low-pass filter (Cl-C3/Ll-L3) that passes signals below 3 Mhz. What is notable about this filter, and from which it takes its name, is the fact that the low-pass filter is terminated in a 50 ohm dummy load. This arrangement works a little better than the straight high-pass filter method because it absorbs energy from the rejected band, and reduces (although does not eliminate) the effects of improper filter termination.

Some of the capacitor values are non-standard, but can be made using standard disk ceramic or mica capacitors using combinations in Table 1:

Table 1

C1: 1,820 pF	Use 1000 pF (0.001 MF or 1 nF) in parallel with 820 pF
C2: 1,270 pF	Use 1000 pF and 270 pF in parallel
C6: 1,400 pF	Use 1000 pF, 180 pF and 220 pF in parallel.

The other capacitors are standard values.



The coils are a bit more difficult to obtain. Although it is possible to use slug tuned coils obtained from commercial sources (e.g. Toko), or homebrewed, this is not the preferred practice. Adjusting this type of filter without a sweep generator might prove daunting due to interactions between the sections. A better approach is to use toroid core homebrew inductors. The toroidal cores reduce interaction between the coil's magnetic fields, so simplifies construction. Possible alternatives are shown below in Table 2:

Table 2

Coil	Value	Core	AL Value	Turns
Ll	4.1 µH	T-50-15 RED/WHITE	135	17
L2	4.1 µH	T-50-15 RED/WHITE	135	17
L3	2 µH	T-50-15 RED/WHITE	135	12
L4	I.5 µH	T-50-2 RED	49	18
		T-50-6 YEL	40	20
L5	2.2 µH	T-50-2 RED	49	21
		T-50-6 YEL	40	24
L6	10.2 µH	T-50-2 RED	49	46
		T-50-6 YEL	40	51

For all coils use wire size to #24 to #30 A WG enamel insulation.

The dummy load used at the output of the low-pass filter (RI in Fig. 4) can be made using a 51 ohm carbon or metal fi1mresistor, or two 100 ohm resistors in parallel. In a pinch a47 ohm resistor could also be used, but is not preferred. In any event, use only non-inductive resistors such as carbon composition or metal film 1/4 to 2-watt resistors.

If you would like to experiment with absorptive filters at other cut-off frequencies than 3 Mhz, then use the reactance values in Table 3 to calculate component values:

Component **X (X_l or X_c)**

Table 3

L1 = 28.8 Ω	L2 = 78.4 Ω	L3 = 38 Ω	L4 = 28.8 Ω	L5 = 42 Ω	L6 = 193 Ω
C1 = 28.8 Ω	C2 = 42 Ω	C3 = 193 Ω	C4 = 78.4 Ω	C5 = 78.4 Ω	C6 = 38 Ω

The exact component values can be found from variations on the standard inductive and capacitive reactance equations: Microhenrys and, picofarads

These component values are bound to be non-standard but can be made either using coil forms (for inductors) or series-parallel combinations of standard value capacitors.

Shielding

Shielding is a non-negotiable requirement of filters used for the QRM reduction task Otherwise, signal will enter the filter at its output and will not be attenuated. Use an aluminum shield box of the sort that has at least 5-6 mm of overlap of the flange between upper and lower portions. I used a tinned steel RF box for this purpose when building the prototype for this filter.

Expected Results

If the correct components are selected, and good layout practice is followed (which means separating input and output ends), then the absorptive filter offers stopband attenuation of -20 dB at one octave above Fc, -40+ dB at two octaves and -60 dB at three octaves. For a 3 Mhz signal, one octave is 6 Mhz, two octaves are 12 Mhz and three octaves are 24 Mhz. My results were slightly less than these figures because some of my components were ill-matched (e.g. slug-tuned commercial inductors were used rather than toroid core coils).

A design suitable for the US television bands is provided by Weinreich/Carroll (1968). The same source also shows a number of other absorptive filter designs for those hams and SWLs who would like to experiment. SWLs who consult that source should remember that the roles of the receiver and dummy load must be reversed for the filter to be useful.

Conclusion

The problem of MW/LW AM BCB interference to HF shortwave receivers can be daunting, indeed, because of the high power level of signals encountered. Even good quality receivers will balk at handling those powerhouse sluggers. The use of either wave traps, high pass or absorptive filters will do the trick for all but the most severe cases. I've even seen it work inside the antenna/transmitter room of an AM station operating with 5,000 watts (the on-duty engineer was a ham, and was goofing off).

References

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