

8/5/09

The John R Leary Series

Reproduction of CQ article January 1979 And Other Sources

Old receivers never die. They collect dust. W9WHM tells how he cleaned up, modified and turned the venerable Hammarlund "Super Pro" into a modern, competitive piece of amateur gear.

Souping Up the Super Pro

**BY JOHN R. LEARY, W9WHM (SK)
438 Hamilton, Fortville IN 46040.**

So you've just bought your new super-duper, ultra. extra-modern, completely solid-state transistorized, up-to-date, best-in-the-world receiver. And you think it's so good that it will last you at least ten years (if WARC 79 allows it!),

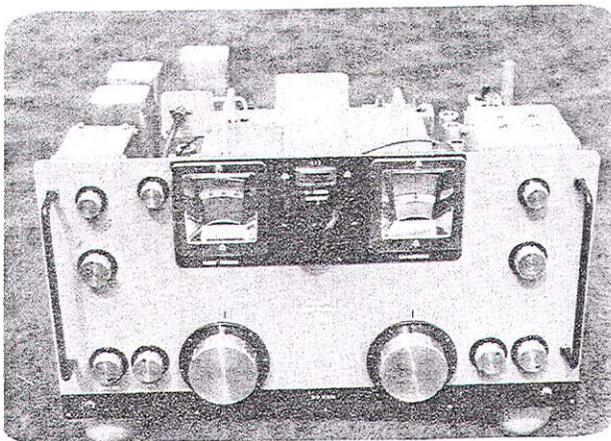
Well, guys like me (and I'll bet thousands of other CQ readers) were born and bred during an era that you young fellers can never really appreciate. Now I don't mean to talk down to you. The fact is, since you were never there, you didn't live it. You see, I was weaned on vacuum tubes. And I guess that the only time that the younger generation ever sees a tube is at a hamfest.

Anyway, about twenty years ago an electronics company called Hammarlund came out with a general coverage receiver that knocked amateurs and SWL's for a loop. It was called the "Super Pro" and it was a magician of a receiver for its time. It became very popular and, as a result, became the subject of many articles in the amateur journals. CQ probably published more information on the "Super Pro" than all the others combined. Names like Geisler, Reed, Lee and others are the fellows who helped to turn the original "Super Pro" into a receiver that will make some of the newcomers say "uncle!"

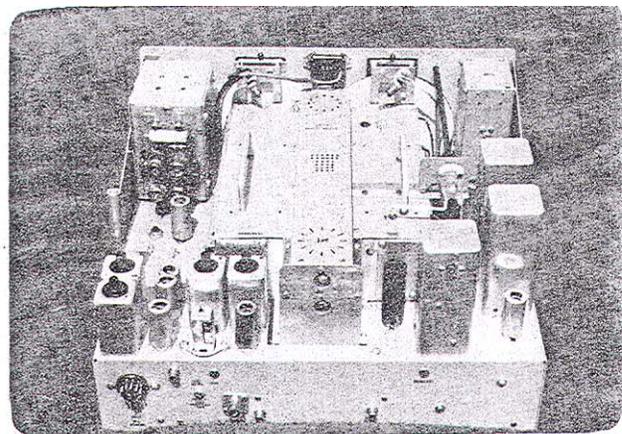
What I did was incorporate the best features of each article I found. It wasn't an easy job. Certainly not a weekend project. It took me twenty years to complete. Painstaking, deliberate, careful and judicious work. But it was a labor of love.

I added an antenna trimmer circuit, 6AB6 r.f. stages, fast and slow a.g.c. circuits, a product detector and an outboard Q5-er (85 kHz i.f. from a BC-453), among other things. All of the original components were removed, cleaned and remounted on a new 17 x 5 x 15 inch chassis.

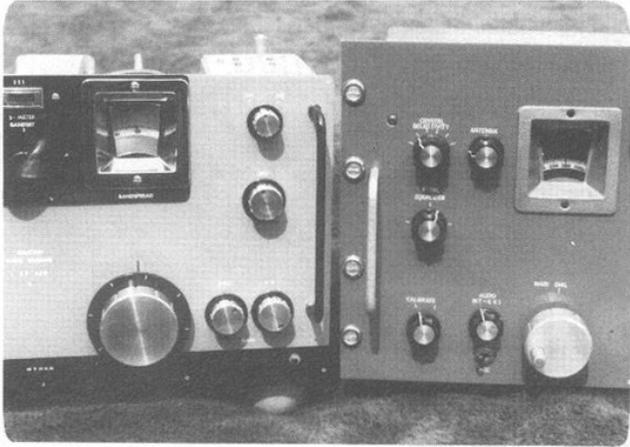
The photos tell a more graphic story than I can write, so I refer you to them.



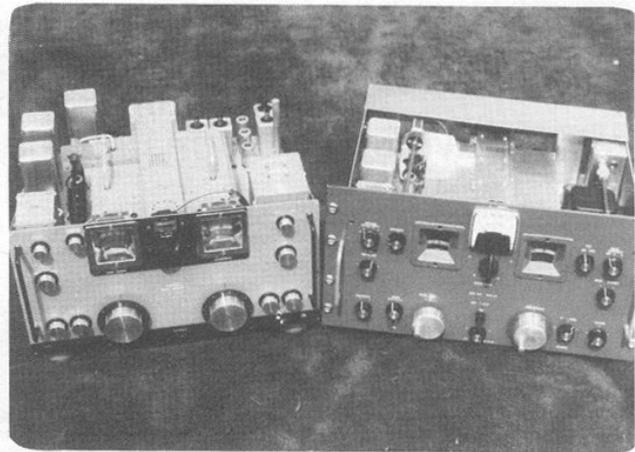
Front view of a restored Super Pro SP-400.



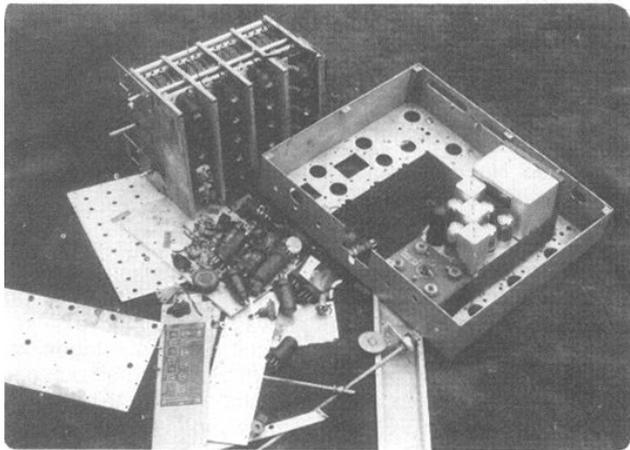
Top rear view of a "Souped-up Super Pro."



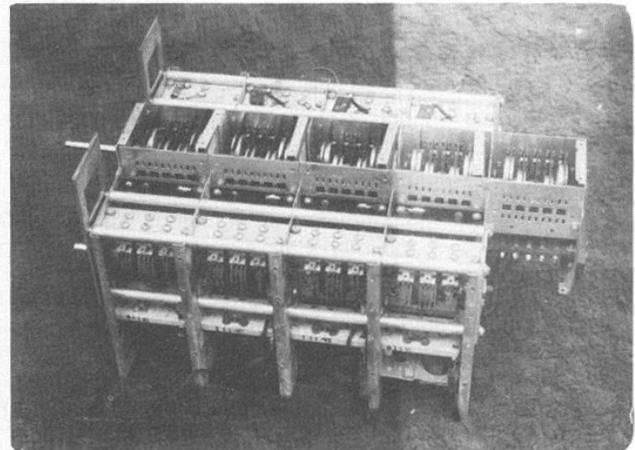
Detailed look at the front panels of two restored units.



Left: A restored SP-400X (540 kHz to 30 MHz).
Right: Restored SP-400SX (1240 kHz to 40 MHz).



Parts used to build the "Souped-up Super Pro."



The famous Super Pro tuning unit.

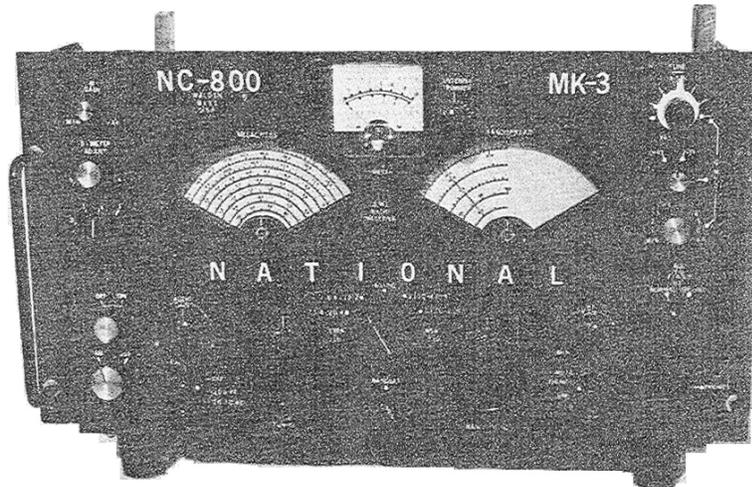
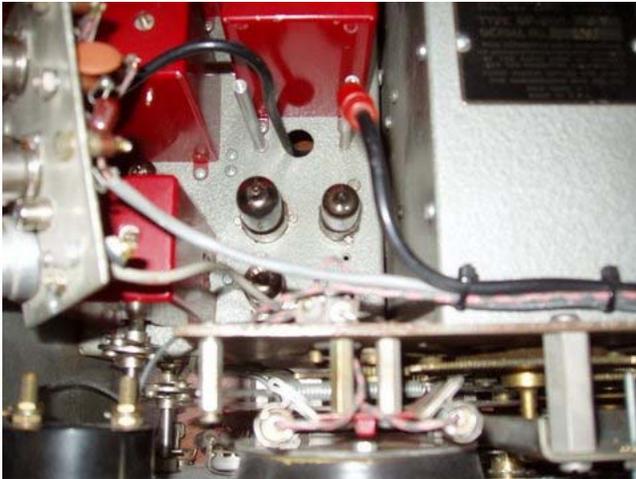


Fig. 8- W9WHM's winter restoration project would truly make a dream receiver for this month's classic rigs.

Fig. 8 shows yet another product of John Leary. W9WHM's handiwork. John built this 21-tube NG-800 from two junked NG-400s in a quite diligent and radical restoration. The front panel was replaced with one cut from scratch, and the dials were moved from the left to the center. It has double tuned IFs. 0 multiplier. 100 kc calibrator, and reflects a full winter's effort in restoration. Beautiful!

One of John Leary's many different SP 600 Modifications

(Courtesy of Les Locklear-present owner)



This is the 2nd conversion section using a 6EJ7 for 2nd mixer, improves reception on high bands



Product detector that sets on top of spacer studs over the 2nd conversion section on chassis



Band in use plate painted by Cecil and the engravings were filled in using a black lacquer stick (was brass and couldn't read it)



Front panel



RF Deck, using a 6BZ6 1st rf amp tube



Product detector mounted in place



OA2 Voltage Regulator tube



Leary's audio section, 6C4, 6BA6, 12AX7 and 6AQ5 like Collins 75A-4 (don't ask)



Solid state rectifier section, diodes and resistors under red cans B+ is perfect.



R.H. side of front panel. Note AM/LSB/USB switch at upper right.



L. H. Side of front panel



Top of chassis

Specs for the 2nd Mixer Tube

6EJ7 Sharp-Cutoff Pentode

Class A Amplifier

Plate Voltage	190 V
Grid No. 2 Voltage	190 V
Grid No. 1 Voltage	-2.35 V
Plate Resistance (approx)	350K Ω
Transconductance	15K μ ^u
Plate Current	10 mA
Grid No. 2 Current	4.1 mA



Front panel

John Leary's Modification Notes

These were copies from John's handwritten notes for his #28 receiver. No attempt has been made to sort or categorize. Some of the comments are bit cryptic and one might have to look at receiver pictures for clarification.

1. Changed in #28. Changed IF screen voltages to 105 volts
2. Changed all sockets to new ceramics.
3. Added speaker terminal on right rear. Added new antenna connector on upper right rear for better access. Also lettered all connections.
3. 200 Kc calibrator. Adjust to WWV. 100 Kc can be used is desired. Near rear. Can be removed for access underneath units for L.F.
4. 1000Kc calibrator near rear panel. No adjustment. Just for locating H F places. Only about 1/2 kc off.
5. 6EJ7 mixer for much better H.F. reception. Cathode adjustment on vertical shaft. Wire nearby was connection for Q multiplier that original owners wanted. Didn't take. Leave alone or clip off if bottom panel is removed.
6. 12AX7 driver to 6AQ5 output. Audio circuit from 75A4. Best ever found. ??? to 6.3 Vac center tapped (transformer) for hum reduction.
7. 6C4 best audio rectifier.
8. Separate bias transformer and filter to relieve current drain on main transformer. Located under main power supply.
9. On rear panel S-meter adjustment. All marked on rear.
10. Separate avc circuit. Diagram is included on a separate page.
12. Complete bias control on rear terminal strip. Adjust -1 volt main receiver "gain" to suit operator. Avc bias to avc rectifier to prevent avc from coming on (for) weak signals. Approximately -3 volts. 3rd. I.F. gain to adjust overall receiver gain to detector.

If very strong signal distorts audio, reduce gain. After adjusting tighten screw on top to lock in place. Figure shown in another location.

13. Crystal phasing on front panel inoperative. It take an oscilloscope which I don't have. I've never found any use for it. Check original manual for alignment. They (crystal phasing controls) have never been very effective.

14. Antenna terminal on upper right side, to move receiver out to try another antenna. (Meaning of this statement unclear.)

15. Wires near antenna terminal on rear of tuning unit, are small capacitor hooking (the) 200-100Kc calibrator to (the) antenna.

16. (The large framework around (the) receiver on top plus tall rear panel on receiver (shown in Number 32) are for one reason only: oscillator stability. Most all complaints of (the) SP 600 when in the 3 Kc or 1.5 Kc position plus c.w. and ssb have been solved by this expensive (hi) addition. (This) was my experiment and it really helped.

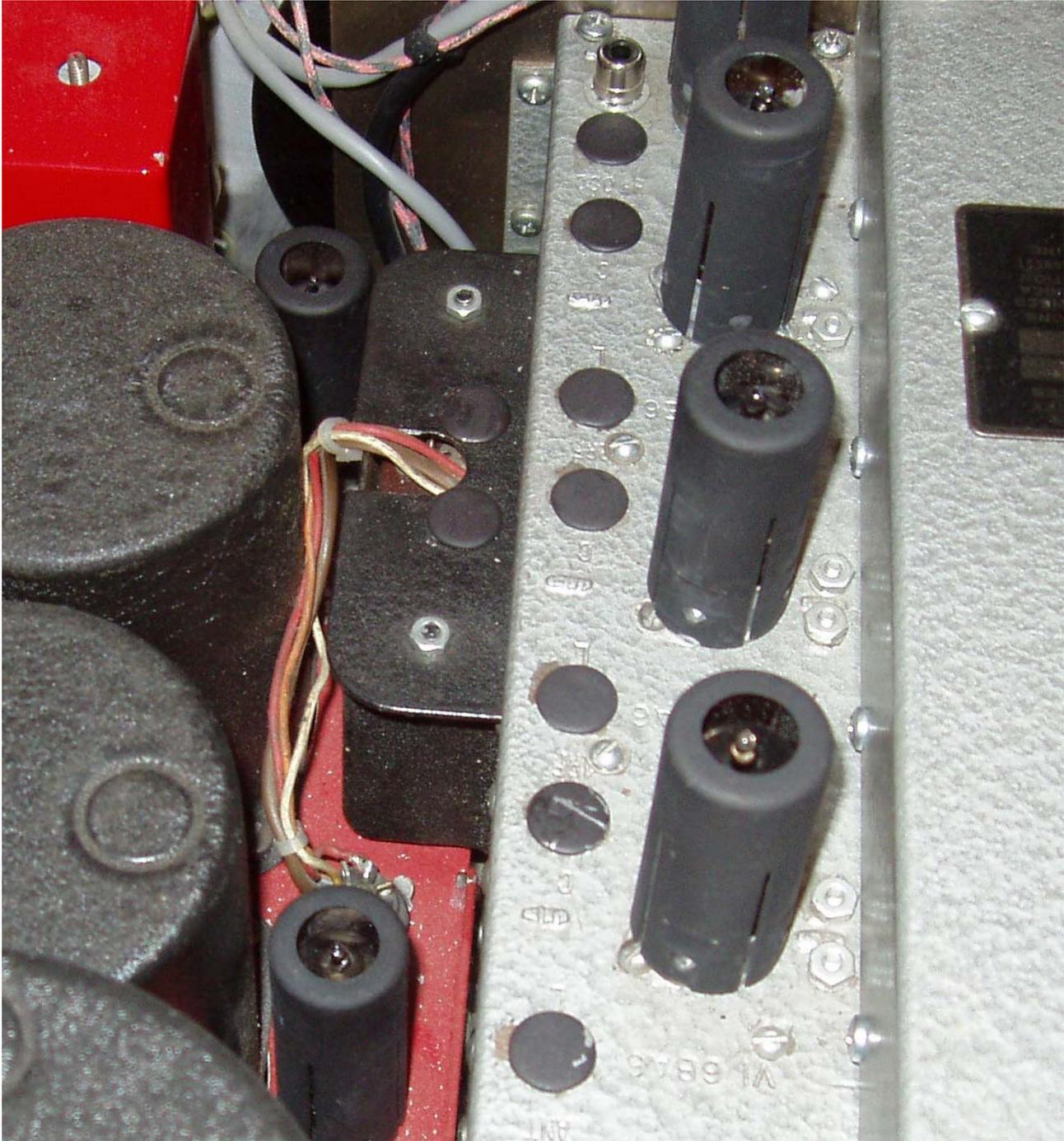
17. May have to align receiver slightly after having been moved and jarred around. But it sure is a wonder.

18. Wiring to top R.F. unit is not through (the) I.F. unit on the side which all wires cause trouble. Voltage (unclear) goes to (the) top unit by a 1/4 ?? hole in (the) rear of (the) tuning unit. To?? Under chassis all there were going through I.F. unit on the side has caused problems. So simple to run these out of (the) rear unit.

19. Audio speaker phone switch on front is because the audio, when switched to phones run audio to large resistor to ground audio (transformer to protect.) I think that this what was written but the writing is small and abbreviated.



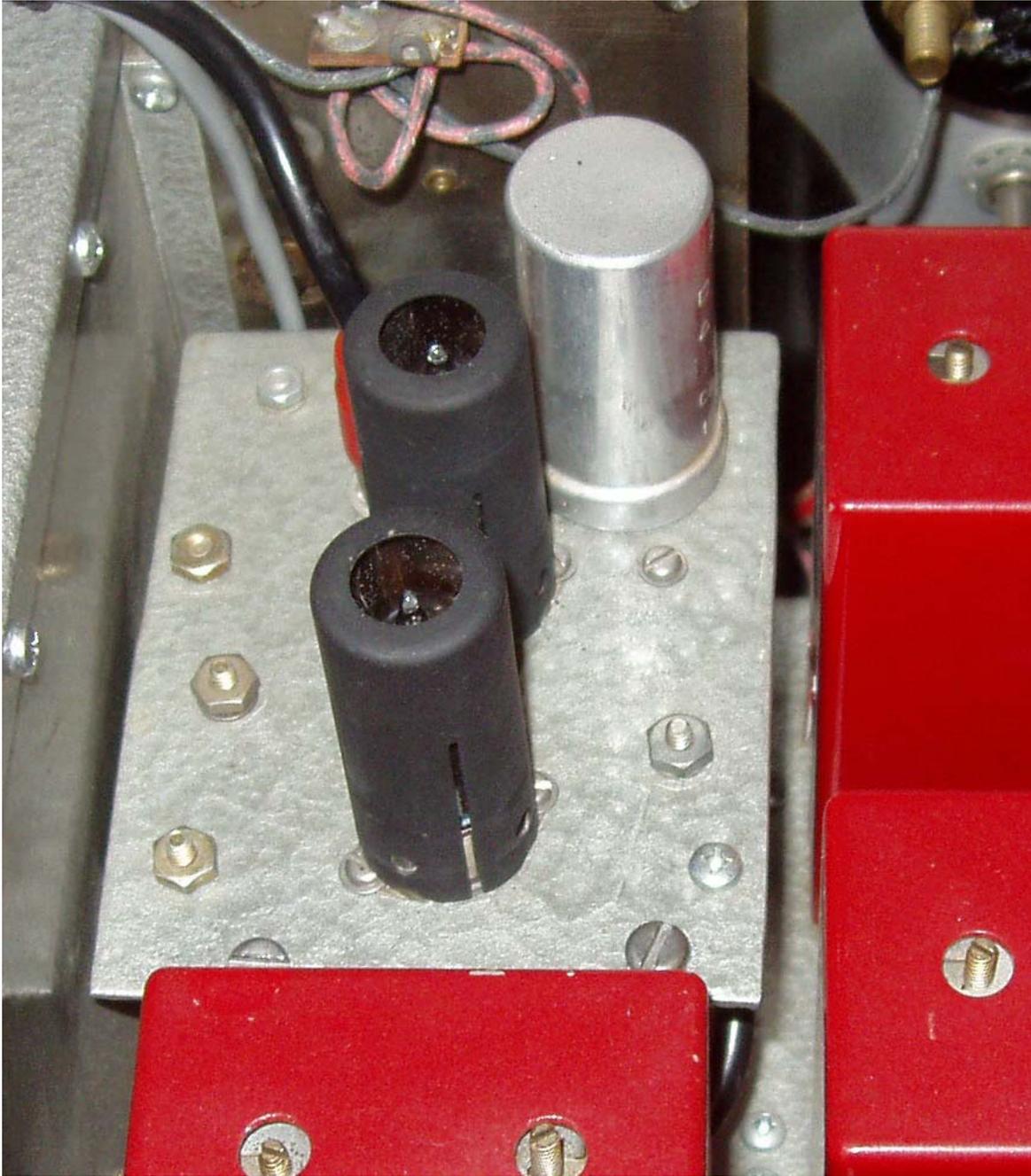
Note the additional improvements added: new antenna connection, extra fuses, AGC level control, and the -1 volt receiver gain control.



A simpler method of bringing out the leads from the 1st RF-Mixer pod. See John's notes for details.



Rear oblique showing from left to right the added filter capacitor, two solid state rectifier units, audio matching transformer, new audio output transformer and AGC transformer. No information was noted for the origin or wiring circuitry of the additional AGC transformer.



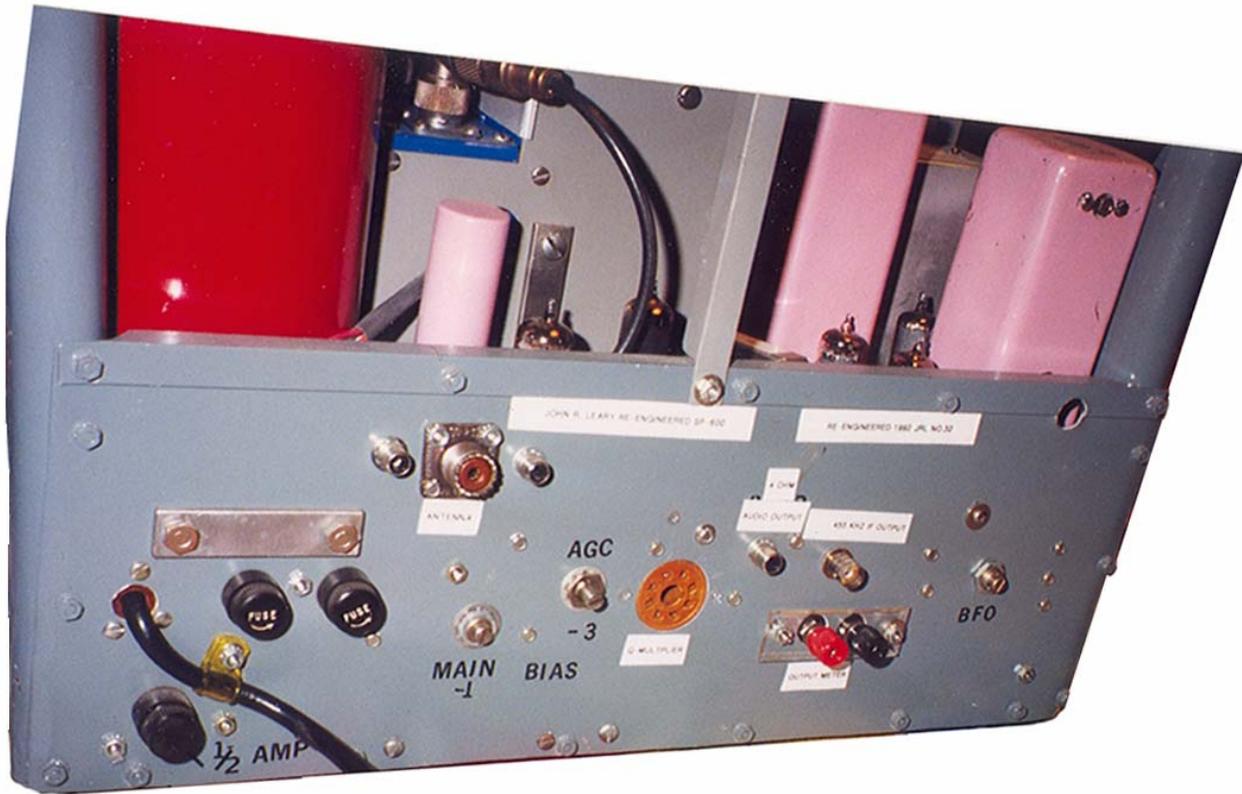
Close up of Product Detector using a 6BA6 and 6BE6, copied from the Hammarlund HQ-180A product detector.



Version Number 31

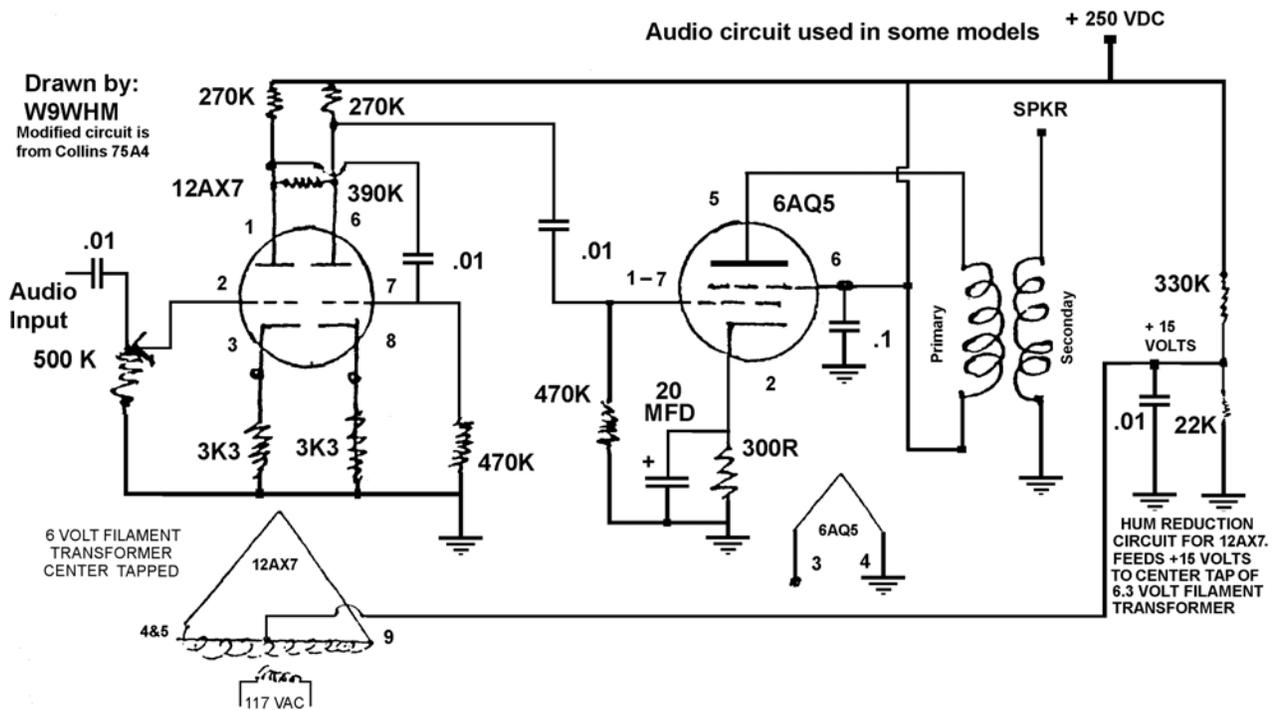


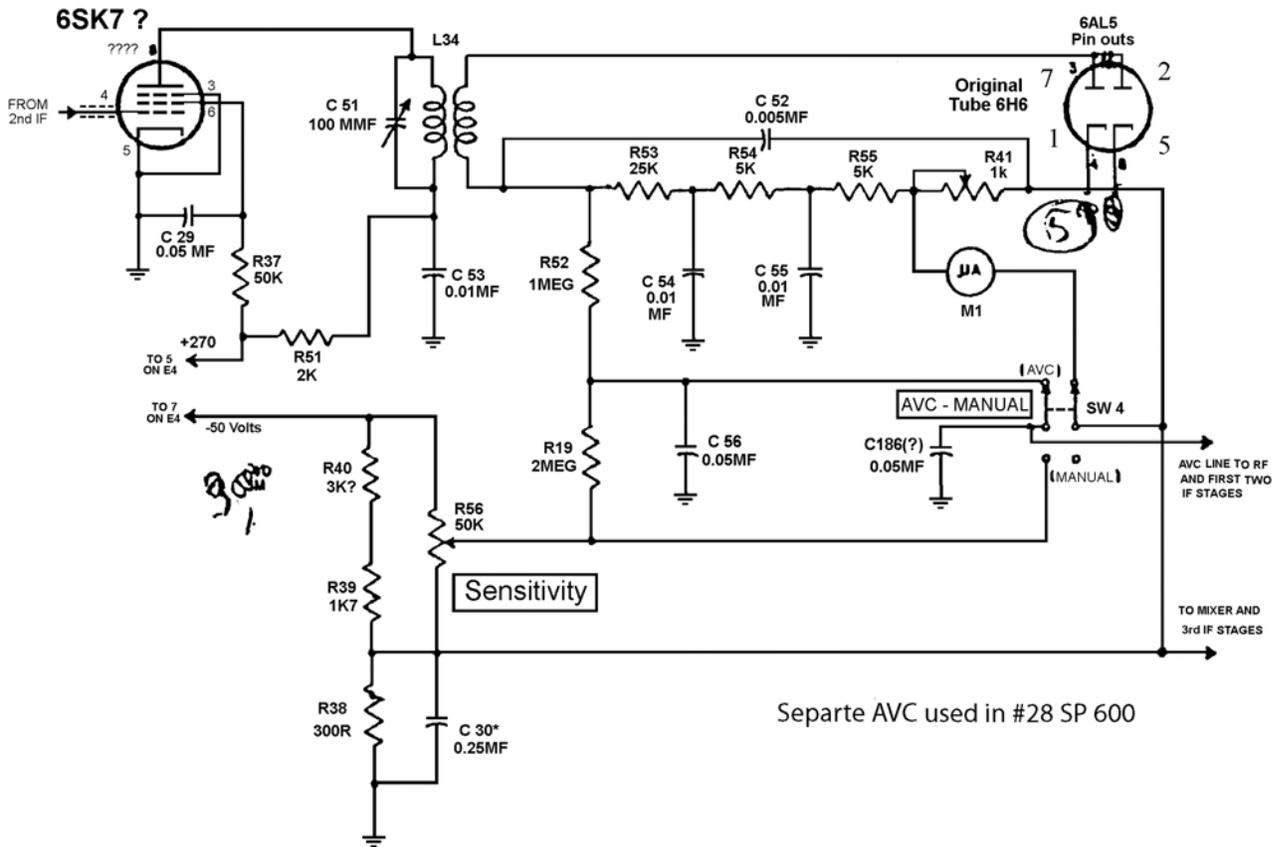
Front view of Version Number 32



Rear view of Version Number 32

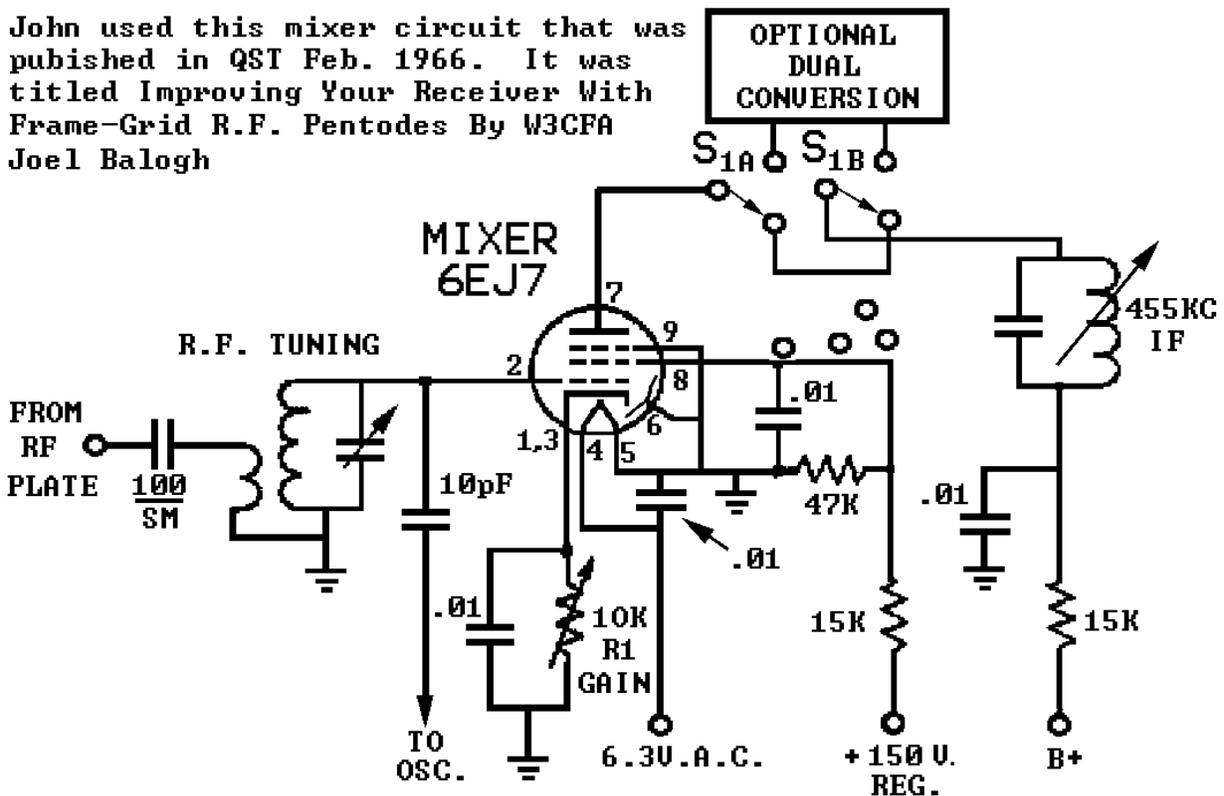
Circuits Used in the SP 600 Conversions





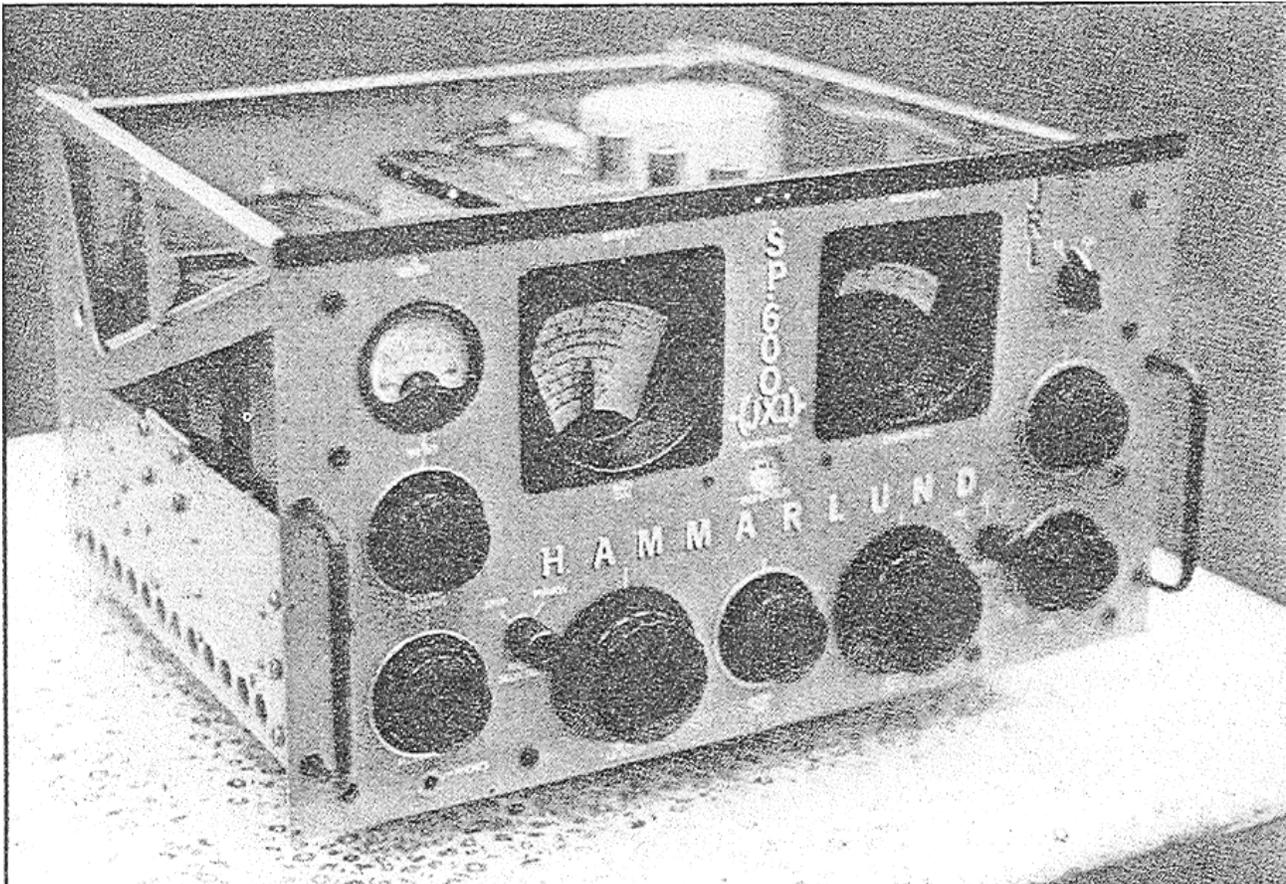
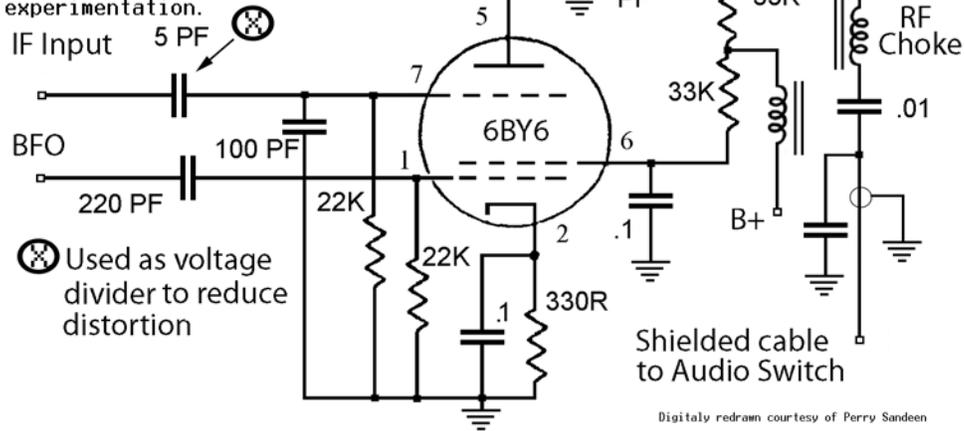
Source of this circuit is unknown. Hand annotation by R40 was not explained either.

John used this mixer circuit that was published in QST Feb. 1966. It was titled Improving Your Receiver With Frame-Grid R.F. Pentodes By W3CFA Joel Balogh



6BY6 Product Detector

The schematic was reproduced from Johns notes. Missing values will have to be determined by experimentation.



A restored – to - new condition SP 600 as pictured in CQ November 1992. John said the additional bracing provided additional frequency stability.

Tube Substitution

BOATANCHORS Digest 3532

by "RJ Mattson" <rjmattson@hvi.net> Date: Tue, 26 Aug 2003 17:12:14 CDT

To: Old Tube Radios <boatanchors@theporch.com> Date: Sun, 24 Aug 2003 06:41:30 -0700
Subject: Re: question about Hollow State Newsletter article

Hi Joe;

.....My understanding is that the author of that article later wrote that it degraded cross-modulation and advised against doing it.

I have two 51J's that I haven't messed with yet and I've wondered about the front end, its design for whip antenna use and its sharp cutoff pentode RF amp. Seems there's lots to do to make the front end right for a decent dipole or long wire antenna including changing to a remote cutoff tube. I'd like to know more about what people think about this subject.

Arden Allen KB6NAX

From: Heinz und Hannelore Breuer <hbreuer@debitel.net> Date: Sun, 24 Aug 2003 16:27:54 +0200
Subject: Re: question about Hollow State Newsletter article

Hi Joe,

I can fax HSN #26 to you if you have not received it yet. Please give me your fax number and I get it on the way.

73 Heinz DH2FA, KM5VT

From: Liles and Naomi Garcia <landn@easystreet.com> Date: Sun, 24 Aug 2003 14:12:52 -0700
Subject: Receiver Front End Information

Good afternoon Arden and Boatanchors Group,

I will only scratch the technical surface on this communications receiver front end discussion, but here is a start. The 51J series of receivers have the 6AK5 sharp cutoff pentode for their first RF amplifier. The Hammarlund SP600's have a 6BA6 remote cutoff pentode for their first RF amplifier. Hallicrafters used the 6DC6 sharp cutoff pentode for the front end of several of their receivers. (I think that 6DC6's are hard to get, maybe someone can check this fact for us.)

In the past years there has been discussion about replacing the 6BA6 in the front end of the SP600 with the 6BZ6 semi-remote cutoff pentode. This replacement improved the SP600's performance in some areas, and was just a plug-and-play replacement. Perhaps the 51J series could benefit from this replacement also. Here is where my information falls short--I do not remember where this replacement discussion is located in written form that would have all of the details and technical information. Perhaps some of you Hammarlund experts could help out here.

Does this 6BZ6 replacement bring back any memories?? Everybody have a great rest of the weekend.

Liles Garcia

To: Old Tube Radios <boatanchors@theporch.com> Date: Sun, 24 Aug 2003 21:55:54 GMT
Subject: Re: Receiver Front End Information

Hi,

Allow me to offer a correction to Liles Garcia's posting. He stated that the 6DC6 was a sharp cutoff tube. It is a semi-remote cutoff tube, and was used as the rf amplifier in the Collins 51S-1 receiver.

Perhaps Liles was thinking of the 6CB6, which is a sharp cutoff tube.

As long as we are talking about front end modifications, has anyone ever used a "Nuvista plug?" It was a plug-in assembly to replace the common 7 pin rf amplifier tubes with a pair of nuvistors in a cascode amplifier circuit.

The Hammarlund engineers tried a Nuvista plug in a SP-600 and liked it very much. See Tech Bulletin 107: <http://www.hammarlund.info/info/sp-600/>

I remember seeing an ad for the "Nuvista plug" in CQ magazine in the early 60's, but I never heard of anyone using it, until I saw the Hammarlund Historian's web site.

73, Ed k4pf@juno.com

From: "Arden Allen" <gumbear@pacbell.net> Date: Sun, 24 Aug 2003 15:00:01 -0700
Subject: Re: Receiver Front End Information

Hi Liles;

I will only scratch the technical surface on this communications receiver front end discussion, but here is a start.

Glad you mentioned those two tubes. The 6DC6 is also a semi-remote cutoff job (used in Collins S line receivers) if I remember correctly. I ran a 6BZ6 in my SP600 for a while and it improved S+N/N sensitivity by 6 DB which, if you think about it, isn't much of an improvement. Worse, the S meter scale gets cramped instead of having more of a logarithmic response over the full scale. The so-called remote cutoff types act more like sharp cutoff types in such an AGC system. One should compare plate curves to see the difference.

Arden Allen KB6NAX

From: WA5CAB@cs.com Date: Sun, 24 Aug 2003 18:41:23 EDT
Subject: FS: 51J-4 Manual Repro (was Receiver Front End Information

Group,

This thread reminds me that several years ago, someone twisted my arm and I made several repro's of the Collins 51J-4 manual. And then of course it was like the Bendix TA-12 plug inserts that I had made. The twister turned out to be the only one for many years who needed one. :-). Anyway, I still have several of the manuals, complete cover to cover, GBC bound, \$20.00 plus postage. They are a copy of an original Collins IB 520 5014 00.

73 Robert Downs - Houston

From: Al Klase <skywaves@bw.webex.net> Date: Sun, 24 Aug 2003 20:44:14 -0400
Subject: Re: Receiver Front End Information

Anchorites,

I've owned two of these receivers, a J-2 that's been passed along and a J-4 that's still here. I'm convinced Collins chose the 6AK5 for the RF amp to achieve the best front-end noise figure possible at the that time. (1948-49ish) There is definitely a problem with front-end overload when connected to a large antenna and receiving strong signals, such as 49-meter broadcasts.

Now, the low noise figure is only important on the higher frequencies where things are quieter. For our purposes it might only matter on 15 and 10 meters.

A proper remote-cutoff tube, like a 6BA6 is definitely a plus on 40 and 80M. Remote-cutoff tubes, allow the gain of a stage to be turned down, by increasing the grid bias, without causing the major amounts of distortion that occur when you try this with a sharp-cutoff device. The 6BZ6 (semi-remote cutoff) is a good compromise, and seems to work OK as a plug-and-play replacement.

If one really wanted to hot rod a receiver of this sort, he'd check out the 6EH7. This is a much later "frame-grid" design, and probably one of the best tubes ever. Unfortunately it's a 9-pin device. Of course, then, you'd start casting a jaundiced eye in the direction of the two 6BE6 mixers, and the whole thing would get out of control. AVOID CREEPING ELEGANCE! Plug in a 6BA6 or 6BZ6 and relax.

There is another problem lurking here. The original receivers, up to the 51J-2, had primary windings on the RF coils to accommodate standard low-Z feedlines. However, starting with the R-388 (51J-3), they dispensed with the primaries and cap coupled to the top of the tuned circuit to get better performance with a short (15 ft.?) whip.

I think there was a article, perhaps in Electric Radio, that told how to reinstall the primary windings.

My two cents, Al Klase - N3FRQ

From: Morris Odell <morriso@vifm.org> Date: Mon, 25 Aug 2003 13:02:24 +1000
Subject: RE: Receiver Front End Information

Hi all,

If one really wanted to hot rod a receiver of this sort, he'd check out the 6EH7. This is a much later "frame-grid" design, and probably one of the best tubes ever. Unfortunately it's a 9-pin device.

I think the 6GM6 is a close 7 pin equivalent to the 6EH7. I have one in the front end of my R390A and it works pretty nicely there - unfortunately the operating conditions are a little different to the 6DC6 so it's not plug & play. You need to pull the RF deck and change a few resistors.

my 0.02 worth73 de Morris VK3DOC

From: "Dan Martin" <dmartin@visuallink.com> Date: Mon, 25 Aug 2003 18:07:19 -0400
Subject: Re: question about Hollow State Newsletter article

My quick comment is that I seem to recall an article in Electric Radio a couple of years back that spoke to Arden's very question - some simple (as I recall) front-end 51J mods to better serve the typical dipole. I shall check around and refresh my memory and join the fray.

Dan WB4GRA

From: "RJ Mattson" <rjmattson@hvi.net> Date: Mon, 25 Aug 2003 22:14:29 -0400
Subject: Re: question about Hollow State Newsletter article

Sounds like a great newsletter. Is there a compendium of past issues available? Is there a web site e-mail address to subscribe? bob...w2ami

I live in a rural mountainous area at 8,900 ft. We have buried power lines that pick up lightning strikes nicely. Also the line voltage is generally 124 volts. We frequently have fraction of a second dropouts. In a nutshell: lousy power for electronic equipment.

I bought a SP 600 on eBay and when it arrived I noticed that the original filter chokes were removed and substituted with much smaller units with no identification, Querying the Ham who sold it to me about the values he replied "that the inductance was sufficient". Also both 6AL5's as well as the 5R4 were replaced with 1N4007 diodes, I moved the line input tap to the 130 volt setting and the radio worked so-so with a 30 ft. antenna.

At this time I came across a solid state electronically regulated power supply in Audio Express magazine using series zener diodes for its reference source. The limitation of this circuit was that it was not adjustable and one had the problem of zener drift. Previously on the R390 Reflector list Dr. Jerry Johnson had made reference to the Texas Instruments TL 431 programmable shunt regulator. For practical purposes it is an amplified "buried zener" on steroids that can sink 200 Ma, the same as an 0A2.

I joined the two circuits using the TI data sheet obtained on the web. The programming formula is shown on the schematic, My main addition is the considerable size increases of the filter capacitors due my particular power line situation. You made not need that much but excellent Nicon electrolytics are cheap.

Although designed for a SP 600 the design is scaleable to any other tube receiver. I chose conservative 25 volt "blocks" for my design although the upper limit is 36 volts. As the SP 600 uses a minus 1 volt for one of its bias voltages and the minimum programmable voltage is 2.5 volts, a voltage divider is required of off the minus 10 volt bias line. Any B+ value can be selected by using fixed blocks and one variable.

Normally the SP 600 is a capacitive input filter but the previous owner made it a choke input due to the SS rectifier. More on this later.

Circuit explanation

The heart of the supply is the TL 431 regulators. Set up in series blocks they feed the MOSFET source follower through a 10 Meg and 10 Mfd time constant to give a delayed B+. The filter capacitor values are not critical and based on what I had on hand. Any voltage rating 1 Kv over the B+ diode output should be fine. Bleeder resistors were added to all capacitors for safety. "Flameproof" metal oxide or WW types are recommended, .01 1Kv disc capacitors on all the rectifier diodes are for RFI suppression.

The input circuit compared to the original wiring was completely re-configured. Besides 3 wire polarization it is now routed so that there is no power to any of the transformer components until the power switch is turned on. A Corcom line filter replaces what is usually C152 & C168 bypass caps used on the line input. A MOV is used to suppress input spikes while the CL-90 does double duty by in-rush current limiting and preventing the large value of capacitance from blowing the fuses. The CL-90 Needs to be mounted in a location where its heat won't cause a problem. It has to be warm to work properly. Each lead should be crimped, not soldered to the line leads.

I have not built this circuit at this time but the math comes from the TI data sheet.

Hammarlund SP-600 Dial & Gear Train Alignment

FROM THE HAMMARLUND HISTORIAN

This section is provided thru the graciousness of Andy Moorer, and is used here with his permission.

OK - here's the deal - throw out all the instructions. It is simpler than it seems.

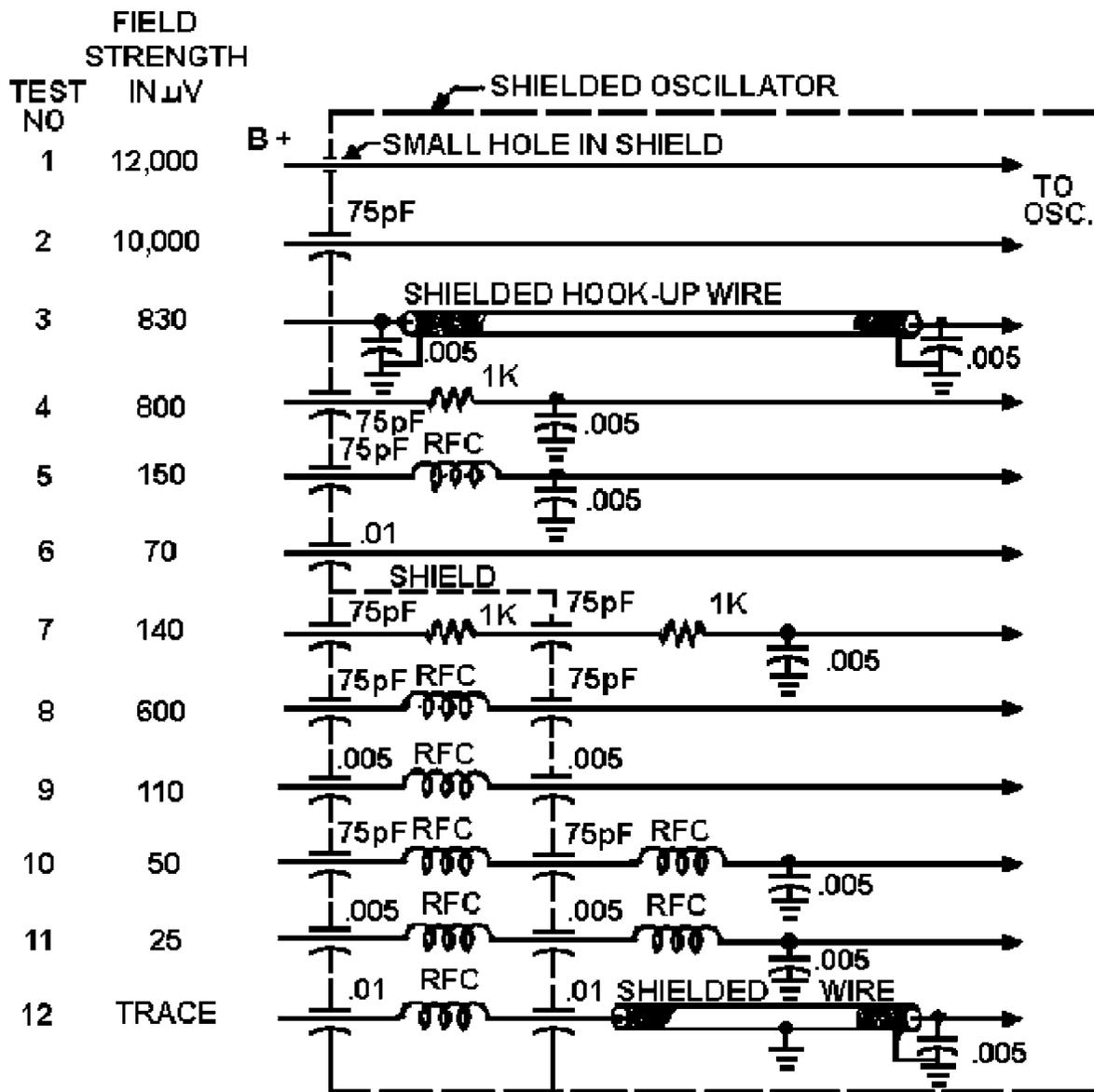
Manually rotate the shafts until you are at the lowest possible frequency. That will be when the vanes of the main tuning capacitor are fully enmeshed - that is, you can't get the moveable vanes any further into the static vanes. The little rotating brass piece that has the stop peg has to be up against the stop that is on the right-most brass gear. You have to fiddle with the gears until these things happen at the same time - capacitor fully closed and the rotating peg against the stop. If you are really adventurous, you can check it by winding all the way to the other extreme - the high-frequency stop should be when the capacitor is fully open.

Note that it won't want to stay there. Here is where it takes 3 hands. Rotate it back to the low-frequency position, against the stop, capacitor fully closed. At this point, the fiducial (the little black pointer up top that bolts onto the faceplate on two hex spacers) and the frequency disc should be positioned so that it points right to a little mark on the disc to the left of the lowest frequency. Note that BOTH the disc and the fiducial (pointer) have some freedom and can be moved around some amount. Put the disc on with the screws loose - put the fiducial on with the screws loose. Holding the peg up against the stop, move each around to align them.

Note that you might not be able to align them - that is, there may not be enough motion on the tuning disc. If this happens, you have to re-fiddle the gears. It means that the hub that the tuning disc bolts to needs to be rotated. You have to disengage it from the other gears and move it around a bit. Then go back to the beginning and get the stop, the capacitor, and the tuning disc and the fiducial all in the right place at the same time. After the third try or so, have a beer and come back tomorrow. It will go together the first time tomorrow. The vernier (the right-hand disc) lines up with its fiducial at zero at this same alignment position. The vernier dial doesn't have a special alignment mark - you just position it to zero when the tuning dial is at its alignment mark. Since that is the one with the stop on it, you generally don't have any problem with the hub being in the wrong position. Note that you have a fair amount of flexibility as to when the capacitor is fully "closed." That is, it doesn't have to be really, really closed. You are going to follow this whole thing with an alignment, which will adapt the rest of the receiver to whatever you end up with as the low-frequency position. It is more important to get the dial alignment mark to line up with the fiducial at the place the stop engages. That is, you don't want the operator to be able to tune lower than the alignment mark. If it does, you might not be able to get the high frequencies to align properly.

Hope this helps. The instructions are often more confusing than helpful. If you get the disc, the stop, and the capacitor to all end up in the right place, then the radio can be aligned and everything will be wonderful. All other considerations are secondary.

James A. (Andy) Moorer
www.jamminpower.com



PARTS USED

- | | |
|-----------------------------------|-----------------------------|
| 1K CARBON RESISTER | .005 DISK CERAMIC |
| RFC - OHMITE Z-50 (7 μ Hy) | .01 SPRAGUE HI-PASS |
| 75 μ F CERAMIC FEED - THROUGH | .005 CERAMIC FEED - THROUGH |

LEAD LEAKAGE WITH VARIOUS LEAD FILTERING SYSTEMS
(COURTESY W1DBM)

Original drawing courtesy Roy Morgan Digital remanufacturing courtesy Perry Sandeen

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.

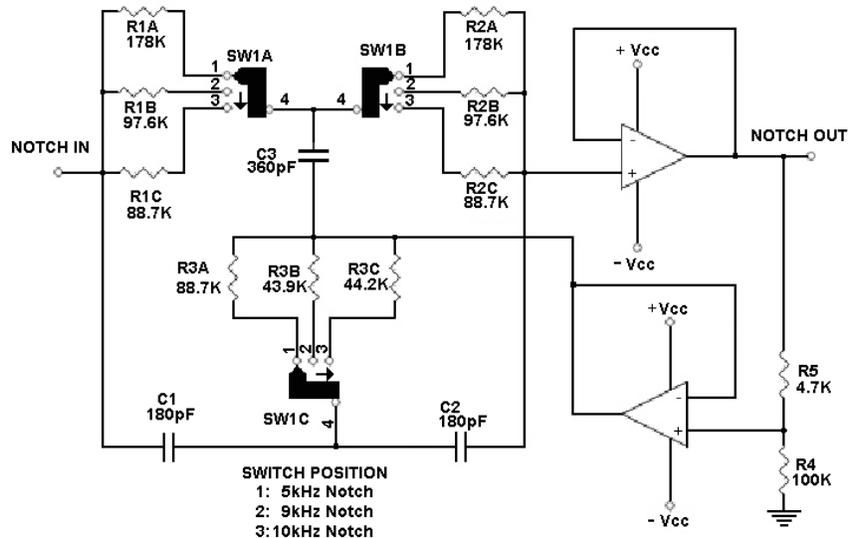
The 10 kHz spacing of AM stations causes another problem. The channel spacing is close enough that the carrier frequencies from adjacent channels are audible. These carriers produce audible tones in the audio - 10 kHz in North America, and 9 kHz in Europe and much of the rest of the world. Because AM 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).

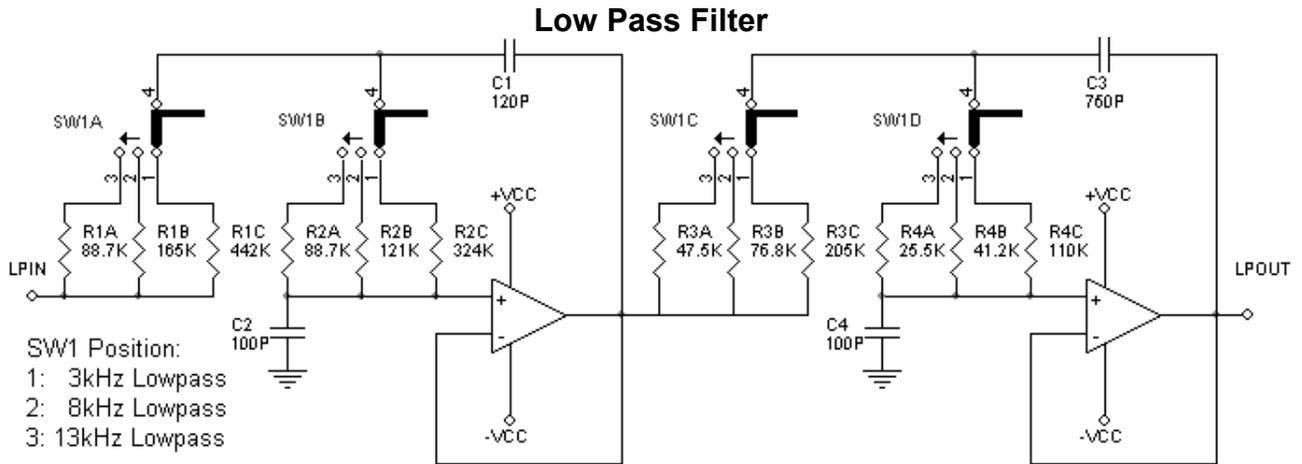
Whistle Filter

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.

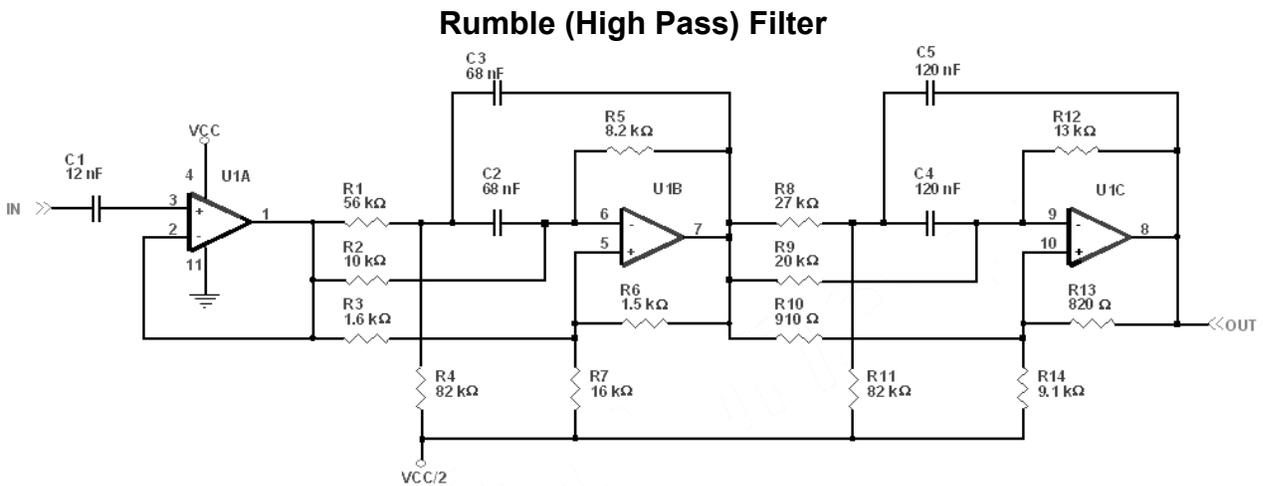


The schematic above is the notch filter section of the AM signal processor. It is extremely difficult to form a high Q notch filter at high audio frequencies. The Twin T notch configuration was chosen because it is unity gain, and relatively easy to tune. All passive components should be 1% except for 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 $1/2\pi R1C1$. $R2 = R1$, $C2 = C1$. 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.



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.



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.

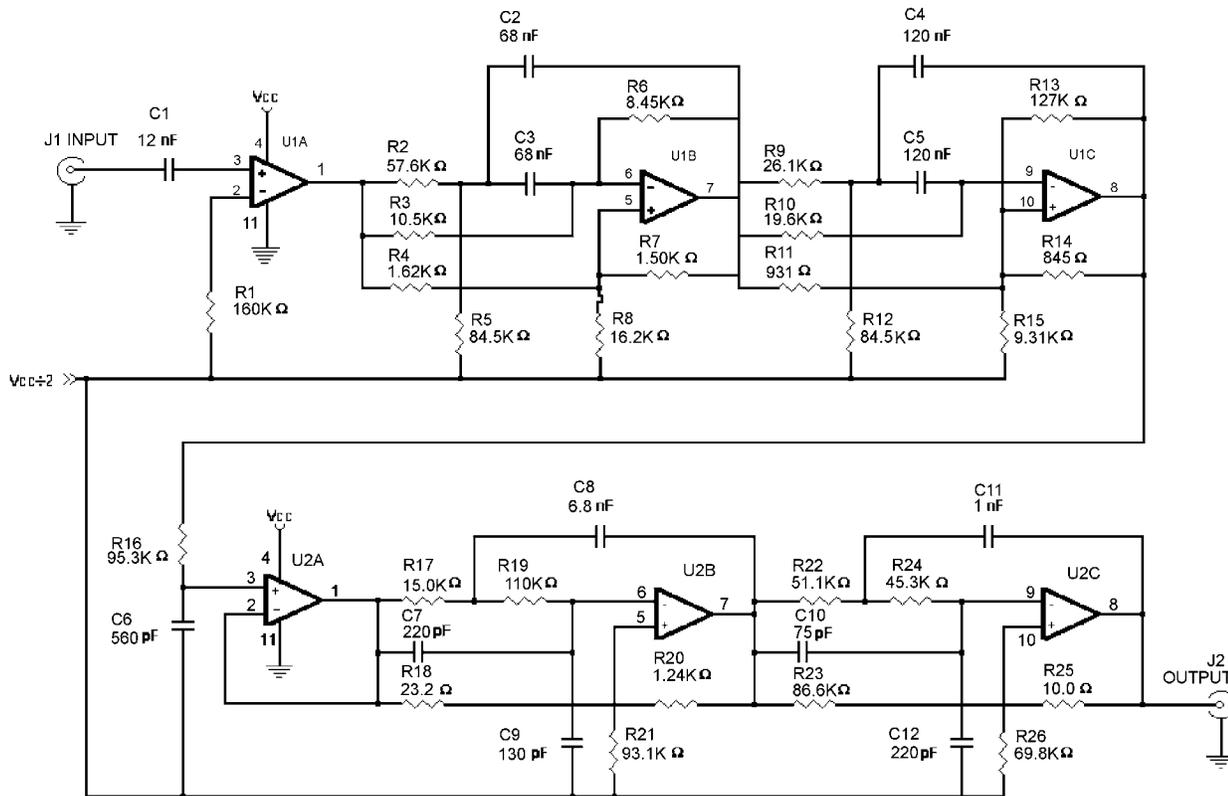
Response

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.

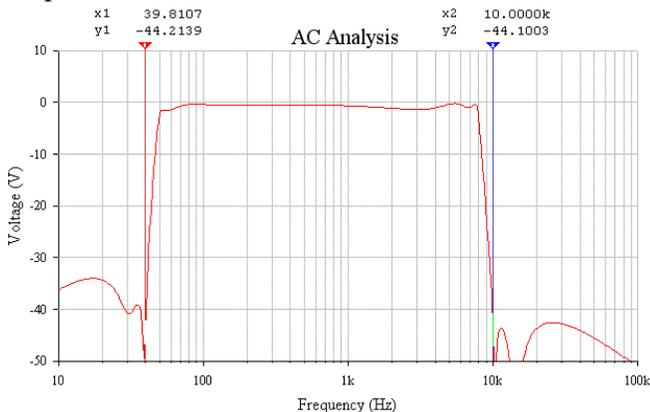
The action of the notch makes the 13 and 8 kHz settings of the low pass filter almost identical, except for a small band of high frequencies above 10 kHz. The 10 kHz notch is not needed for the 3 kHz low pass filter. Even without it, 10 kHz is attenuated more than 30 dB.

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.

MOSFET 6V6 or 6AQ5
Adapted from Electronics World

Unable to find a replacement 6V6 audio power valve for his radio, Dave Allen set about making a MOSFET substitute which, he says, works better than the valve it replaced. Here's how he did it...

After giving me many years of good service, my faithful Murphy wireless recently fell silent. This was due to a failed 6V6 audio output valve.

The valve's demise was caused by the coupling capacitor feeding its input grid on pin 5 going short circuit. This resulted in a positive voltage being applied to the valve's grid, which in turn caused excessive current between the anode and cathode.

Apart from the dead valve and capacitor, the rest of the radio was in good order. Replacing the 0.1uF/400V coupling capacitor was easy but finding a replacement valve was not. In view of this. I devised a direct plug-in replacement for the valve based on an IRFI830G high-voltage MOSFET (**now only widely available, but still cheap, as a IRFI830GPBF [RHOS]**)- of which I had plenty.

In addition to the FET, all that was needed were a few peripheral components and the valve base from the dud valve.

Can this technique be applied to other valves?

It should be possible to build similar plug-in modules to valves other than the 6V6, like the 6PI or KT61. This assumes that you can find information on the pin connections for the valve you want to replace, and that you have a suitable spare base to hand.

Such plug-in modules can directly replace a variety of octal-based output valves that are wired in a parallel heater chain without further modification to the original equipment.

Details for the 6V6
Connections for the 6V6 valve are shown in Table 1. Figure 1 shows a typical single ended Class-A output stage using a 6V6 tetrode or similar valve, incorporating cathode bias.

In Fig. 2 is the output stage - minus valve - and the relevant connections shown for use with the plug-in module.

Figure 3 shows the circuit for the plugin replacement module.

Table 1. Pin connections for the 6V6 tetrode

Pin	Description
1	No connection. This pin may be connected to the suppressor grid, 93, on certain pentode valves and can be ignored.
2	Heater
3	Anode
4	Screen grid (g2)
5	Input grid (g1)
6	No connection
7	Heater
8	Cathode

The circuit

The circuit is based on an IRFI830G MOSFET (now **IRFI830GPBF**) wired as a triode. In this application, no screen grid connections to the valve base holder are involved.

For the MOSFET to operate as an amplifier it has to be biased into conduction. This is achieved by applying a small positive bias to its gate. Bias voltage for the FET is derived from a potential divider fed from the HT rail.

Now insert the plug-in module into its socket and connect a meter switched to its 20V DC range across R1. Switch on your receiver and let it warm up for about ten minutes.

Slowly rotate VR1, until the MOSFET springs to life. This will happen quite suddenly. Finally adjust VR1 for a drop of 2V across R1. This corresponds to a current of 20mA flowing through the input stage which works well with my particular receiver.

Power dissipation considerations

As the MOSFET in this case is biased in class-A, it is constantly dissipating power - or heat if you like - so a good heat sink will be required. The sink needs to be at least 9.9°C/W. In my particular case. The voltage across the MOSFET was measured at 277V at a current of 20mA so the power being dissipated was 277×0.02 , which is 5.54W.

In use

My original plug-in module has been in use for about two years now with no problems. As a bonus, the audio is much improved - especially at the high frequency end..

Parts List

R1 100R .5W

R2 1Meg .5W

VR1 = 120K .5W & 100K,

Multi-turn 10 or 20 turn

C1,C2 .01uF 63V

IRF1830GPBF ([RHOS]),

This is a isolate tab unit

18V/400mW Zener

Heat sink

SP 600 General Upgrade and Mod List

This should be considered a starting point, NOT the end of options

Power Supply

- 1 Replace power cord with IEC 3 wire filter. Install 3 MOV's.
- 2 Remove 110 volt accessory outlet if installed.
- 3 Remove .01 bypass capacitors on AC line, if installed.
- 4 Add CL-10 input rush limiter
- 5 Remove rectifier tube socket and replace with plate. Rewire silicon rectifiers to perf board by power xformer. Change to choke input and replace OEM filter capacitors
- 6 Remove V20 6AL5 bias rectifier wiring, dropping resistors and filter capacitors to perf board by power xformer and add silicon diodes. Socket may be used for product detector.
- 7 If so equipped, bring R72 & R73 divider network to perf board. Mount R85 (2K5 10W) on the top side of the chassis for better cooling of underside components. After a B+ regulator is installed adjust R 85 value to achieve 5 Ma. current flow for maximum life of OA2 tube.

RF Amp

Use 6BZ6 for first RF amp tube.

Consider frame grid tube modification. *Note: This requires great mechanical as well as electronic skills. Only for the advanced.*

Mixer

Remove first mixer AGC control, if present. Set to 2nd mixer gain.

Consider John Leary's frame grid tube modification.

IF

Add manual IF gain pot if not present

Stagger tune IF. The first transformer is set at 455.7, the 2nd at 456.2, the 3rd at 456.7kc.

Detector/AGC

- 8 Replace V14 and V15 6AL5's with 1N4148 diodes on terminal strips. Sockets may be used for product detector or modified audio output circuit.

Consider the Hammarlund site for an AGC/Product detector.

Product Detector

Consider the Hammarlund site for an AGC/Product detector.

Audio

Do Chuck Ripple's audio mods as a minimum.

Consider replacing the OEM audio output xformer and install a 6AQ5 push-pull audio circuit and xformer. 6V6 socket can be replaced with metal plate that can hold an additional tube. Consider using SS 6AQ5's.

Tube Shield Dissipation Measurements

From: Tisha Hayes <tisha.hayes@gmail.com> Date: Fri, 17 Apr 2009 12:09:01 -0500
Subject: Re: [R-390] Black Tube Shields, IERC

A few years ago I went through the process of measuring envelope temperatures on a wide range of tubes in a SP-600 with different tube shields. It did turn out that the black IERC tube shields did make a difference in envelope temperature.

Generally speaking I was seeing 20-60 degrees F difference across the range of shields. I had tested with the following;

Factory default, silver cam-lock tube shields with the loading spring to keep the tube down in the socket

No shield at all, just using ambient air and radiative cooling from the tubes.

IERC tube shields

IERC tube shields with a dab of thermally conductive paste applied to the finger-stock grippies inside of the IERC shield (to improve thermal conduction between the envelope and the shield).

I had strictly controlled air-flow and room temperature and would let the radio stabilize for an entire day before making measurements. This was also in a room where I was not moving around in so there was a bare minimum of air movement. Room temperature was at 70 F. I measured temperature with an optical pyrometer and had put a dab of flat black paint on the top of the tube (so I could get consistent temperature readings without gluing thermocouples everywhere).

By far the worst was the silver tube shields. If anything these kept the heat on the envelope with some temperatures well above 250 F

When I used the IERC's I could get the temperatures in the 150 to 160 range. If I used the thermal paste it would knock the temperature down another 5-10 F on average.

Interestingly the chassis temperature went up when using the IERC shields due to the mechanical connection at the base of the IERC shield and the radiative cooling off of the shields. If you have a concern about capacitor/ resistor aging this may offset your worries about tube temperatures.

In every instance, the use of a small computer-fan to move air across the chassis really helped out in lowering the temperature of the tubes and the chassis. This does not need to be a gigantic fan.

If someone was really interested I could dredge up my notes and put them into a human- readable format. At the time I was doing my little experiment to justify the cost of the IERC shields. (I did end up finding IERC shields for every tube operated device I have)

Ms. Tisha Hayes

From: Tisha Hayes <tisha.hayes@gmail.com> Date: Fri, 17 Apr 2009 14:20:14 -0500
Subject: Re: [R-390] Tube Shields

I think what confuses many is the design of the IERC tube shield (and others of the same family) where the shield has a mechanical attachment (albeit by spring pressure) to the glass of the tube. Transferring heat by conduction cools the tube much more efficiently than trying to remove the heat by radiation. Whenever there is a tube shield that does not contact the glass, the only heat transfer mechanism is radiative with maybe a small amount of convective heat transfer by the random air currents inside of the shield.

Depending upon radiative cooling alone would be similar to disconnecting all of the cooling fans on your car's motor and sitting in traffic for a few hours. While some heat will be transferred by radiation (as the motor gets smokin-hot the hood heats up) the motor will seize up quickly (unless you believe Castrol television commercials). In a car, the cooling comes from moving a much cooler fluid through the hot motor, then giving that heat energy a place to go in the radiator (where it becomes forced convection transfer to the air) where the cooling fans or driving can dilute the heat into the atmosphere.

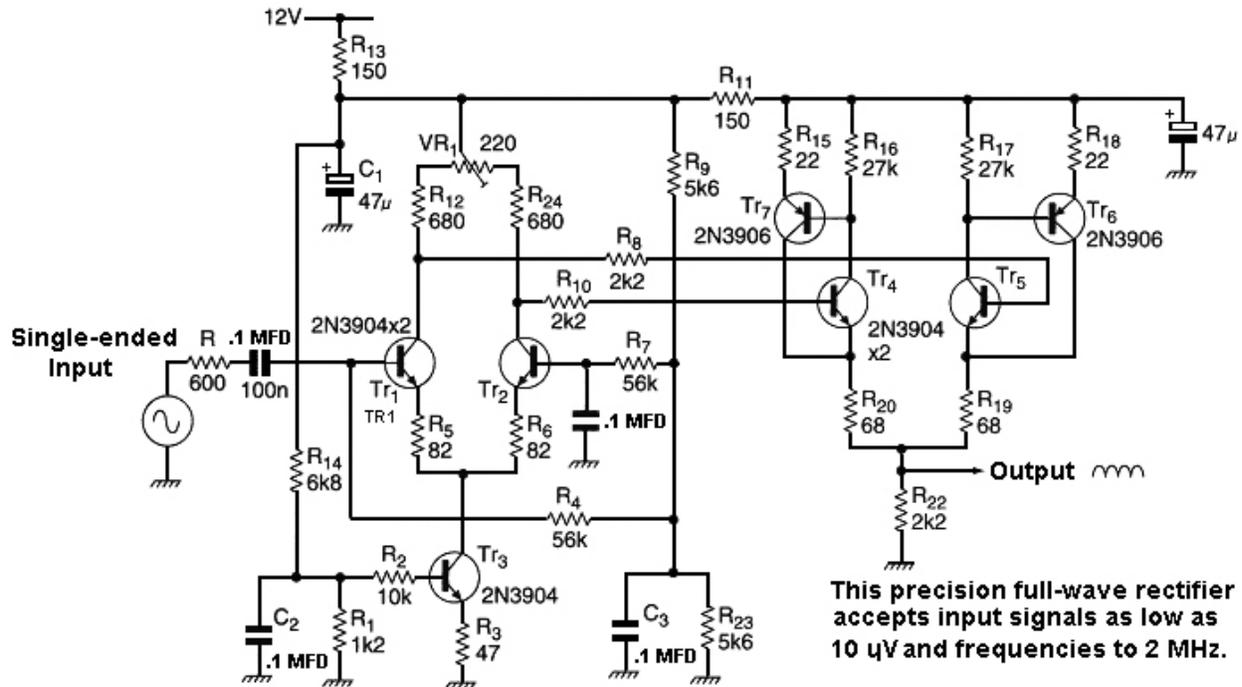
Air is a pretty good insulator of heat, look at how close your finger needs to get to a soldering iron tip before you get burned. You may feel the heat at less than an inch (radiative) but there is a dramatic temperature difference from being 1/4" away and touching the iron (conductive). Tubes suffer from the same problem where there is this tiny air-gap between the glass and the metal.

The IERC tube shields actually touch the glass in several places and the spring fingers are under slight compression. The heat transfer is from the contact. My crazy experiment was to slightly increase the thermal conduction of the spring contact area by using a small dab of thermal grease (what they use on CPU's, commonly sold at Radio Shack) to increase the contact surface area. Ideally the glass envelope and the IERC shield would be at the same temperature (yea, glass does not conduct heat evenly, that makes my brain hurt). BTW, the thermal grease idea makes a mess if you are constantly pulling tubes and has a tendency to remove tube lettering.

Where the IERC fingers come down and imperfectly "dock" with the bayonet-socket (from the silver shields) gives another conductive escape path for the heat to the chassis. To remove heat further would be to passively remove it from the radio by natural convection (heat rises) or forced convection (fans).

Ms. Tisha Hayes

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 be 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 VR_1 to balance the rectification action with the aid of an oscilloscope. The whole circuit runs from a single 12V supply, 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' !

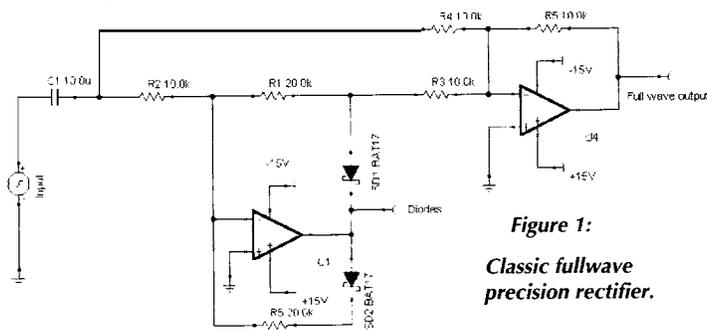


Figure 1:
Classic fullwave precision rectifier.

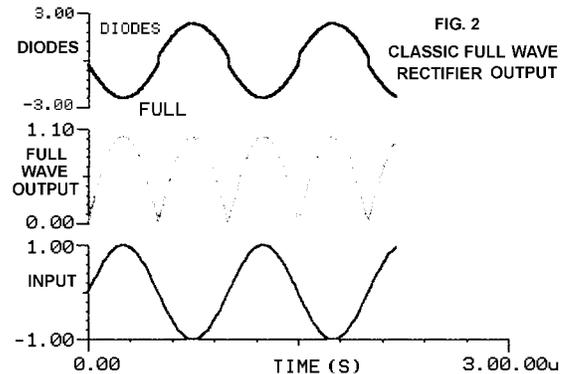


FIG. 2
CLASSIC FULL WAVE RECTIFIER OUTPUT

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 R19 and R20. At the quiescent bias condition, the resistors equally share the output current.

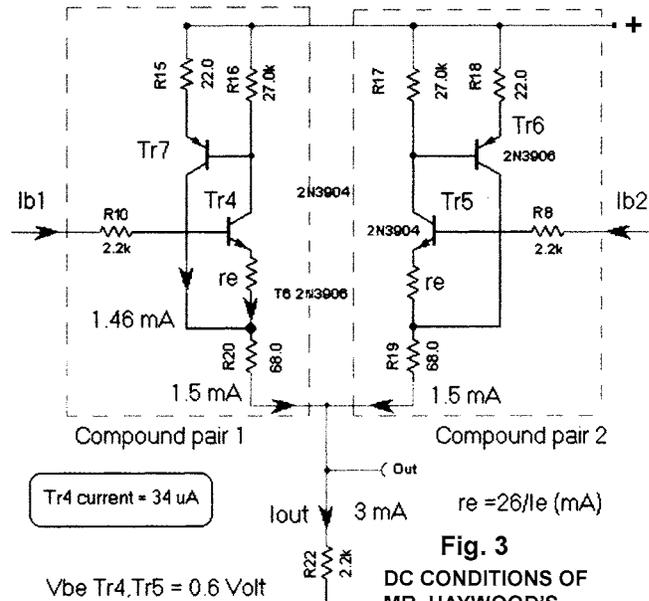
However, at either signal extremity all the output current will pass through one of the emitter resistors while the other compound pair is virtually turned off. This variation in current shifts the turn on threshold at each input, due almost entirely to the volt drop across R19 and R20.

Consider the **Figure 3** Model of the output stage DC requirements. Assuming a typical beta of 100, V_{be} for silicon of 0.6 Volt, and using the rule of thumb for transistor intrinsic emitter resistance $r_e = (26 \text{ } \mu\text{V} / I_e)$ (based on Schottky's diode law at room temperature where the emitter current I_e 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/le (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;



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).

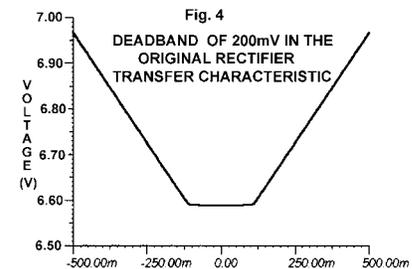
Delta Vin = I(R10/Ai + R20). Iout/2 As the compound pair current gain Ai, is very large, we can ignore the volt drop across R10, simplifying the change in Vin to: Delta Vin = R20.(Iout/2)

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 = ±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.

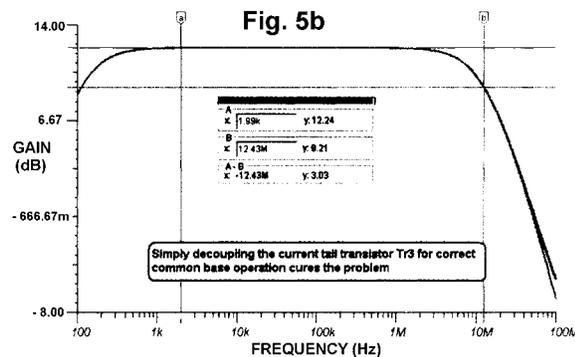
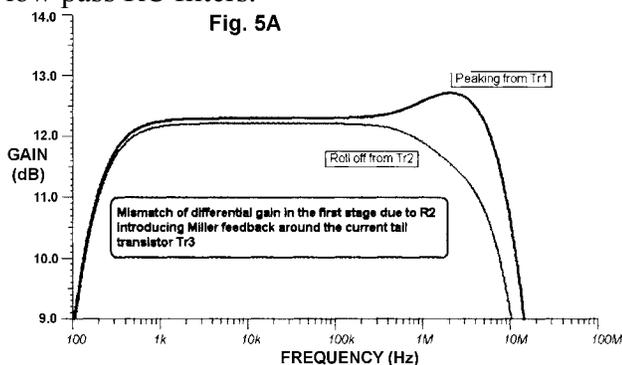
The third improvement would be to ac couple the long tail pair preamplifier to the output stage.

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.

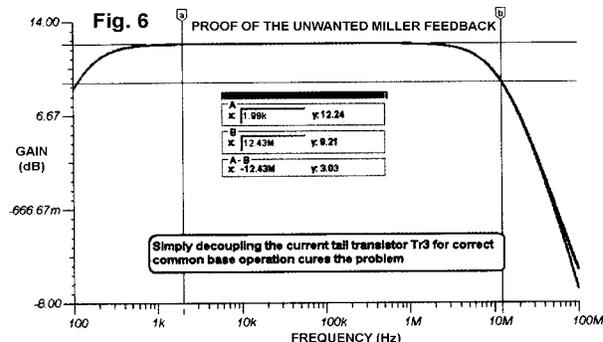


There is also a second Miller feedback effect from the base collector capacitance of Tr1. Miller feedback is minimal at Tr2 due to its de-coupled base and common base operation (with the long tail pair driven single ended). This second Miller effect can easily be demonstrated by driving the input from 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.

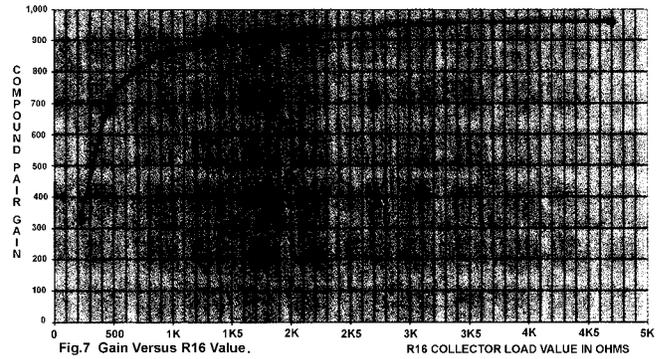
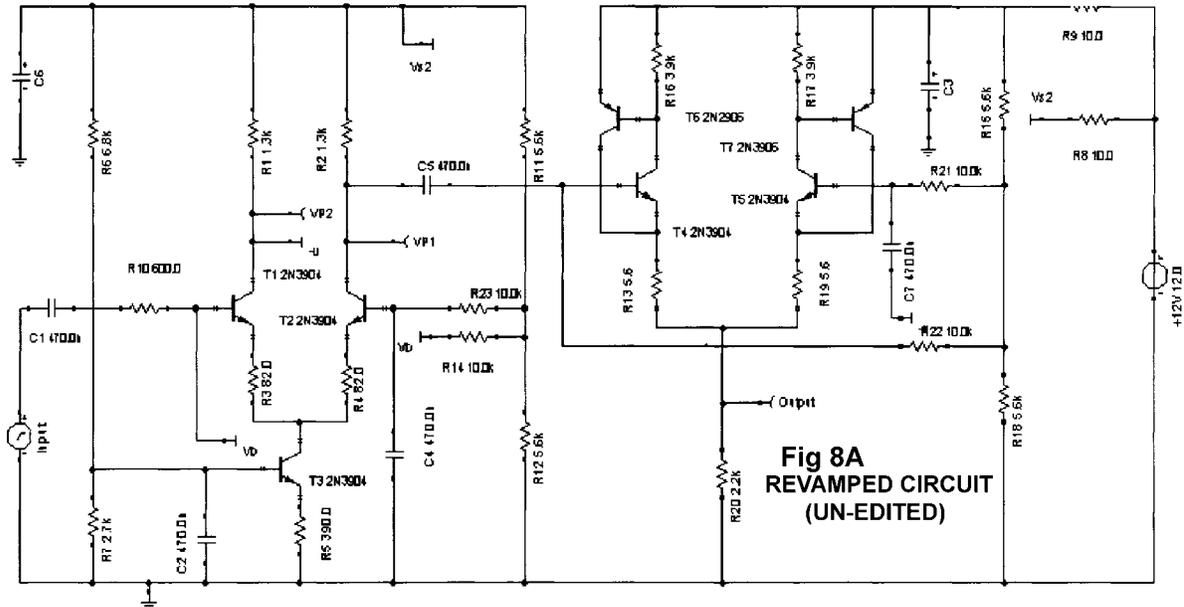


Fig.7 Gain Versus R16 Value.



Further development.

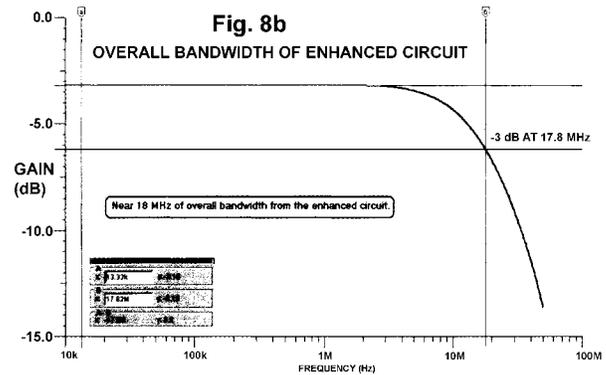
With bandwidth limited by the Miller effect in the first stage and wishing to find a simpler output stage, my thoughts turned to designing a rectifier output stage, which would use the Cascode principle. The 'Cascode' was an amplifier topology invented back in 'valve days' to overcome Miller feedback.

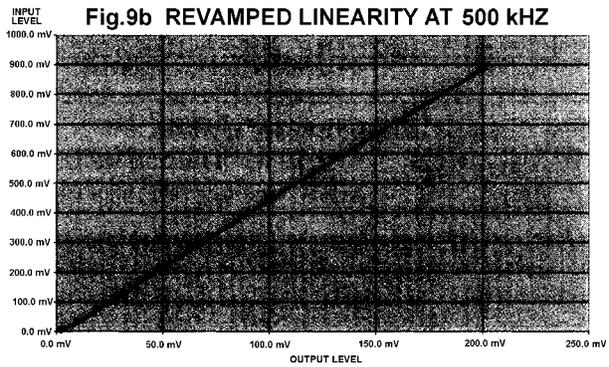
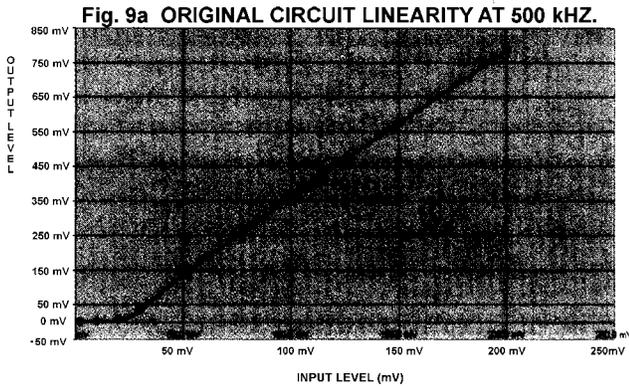
Voltage generation of the signal at the amplifier's first stage output was avoided and hence.

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.





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.

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.

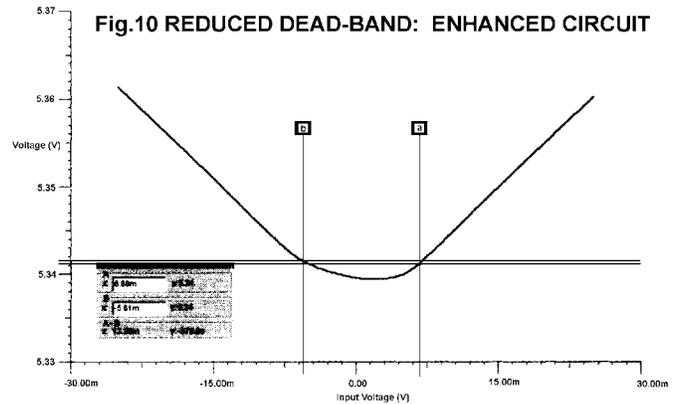
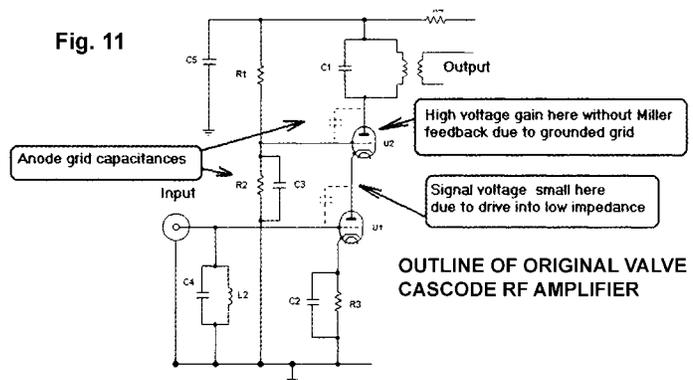


Figure 10, comparison of frequency response with the original circuit.

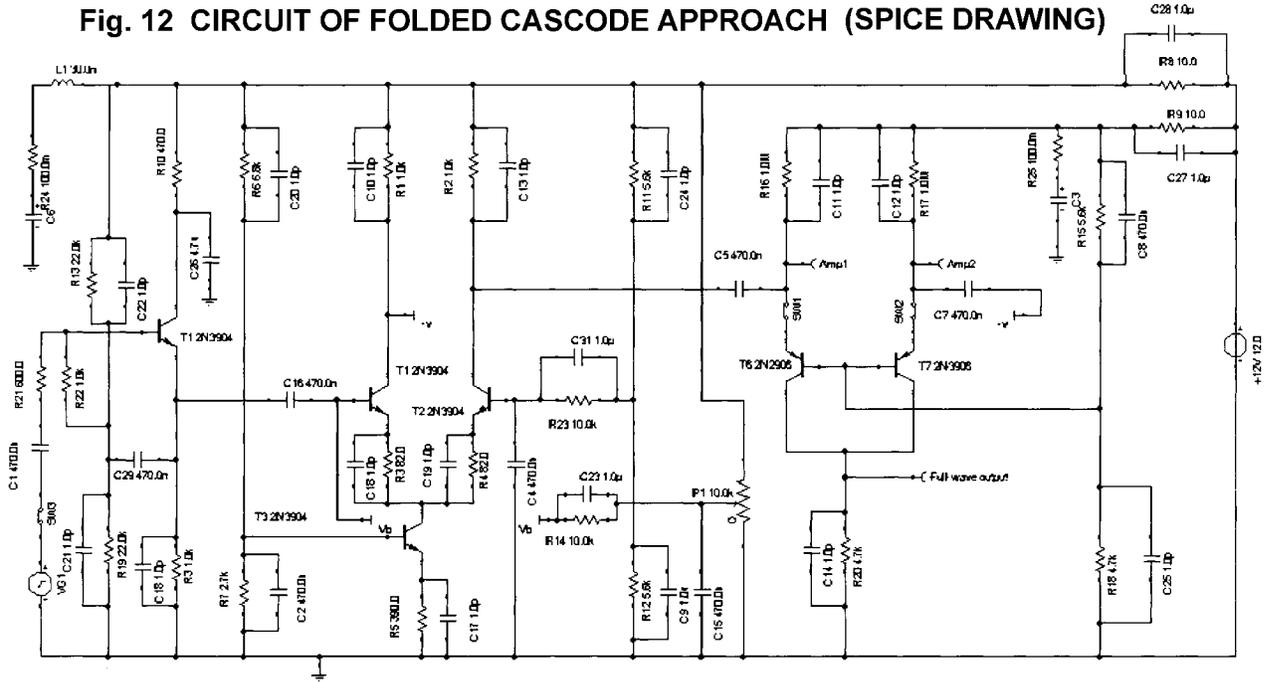


As soon as each amplified signal from the first stage swings positive the relevant PNP transistor conducts and the signal current flows on into the output resistive load. Common base operation ensures the PNP transistors will have very closely matched gains. Because $h_{fb} = h_{fe} / (h_{fe} + 1)$ and with h_{fe} 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 r_e 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.

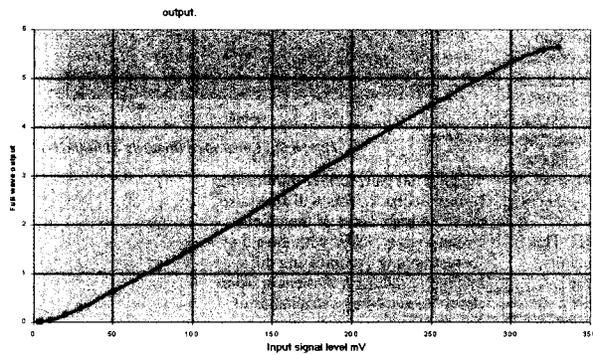
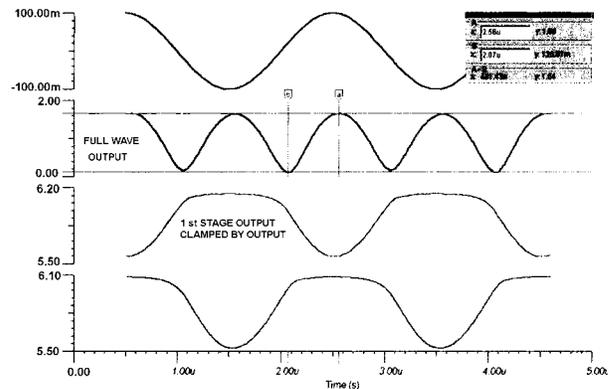


Figure 13: Linearity at 500kHz of the enhanced circuit using the PNP rectifier output.

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 A_v of 100 and ideal infinite input resistance with zero output impedance. It has unwanted feedback capacitance between the input and output terminals of C_f . 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 $V_{in} - (-A_v \cdot V_{in})$ or $V_{in}(1 + A_v) = 101 V_{in}$.

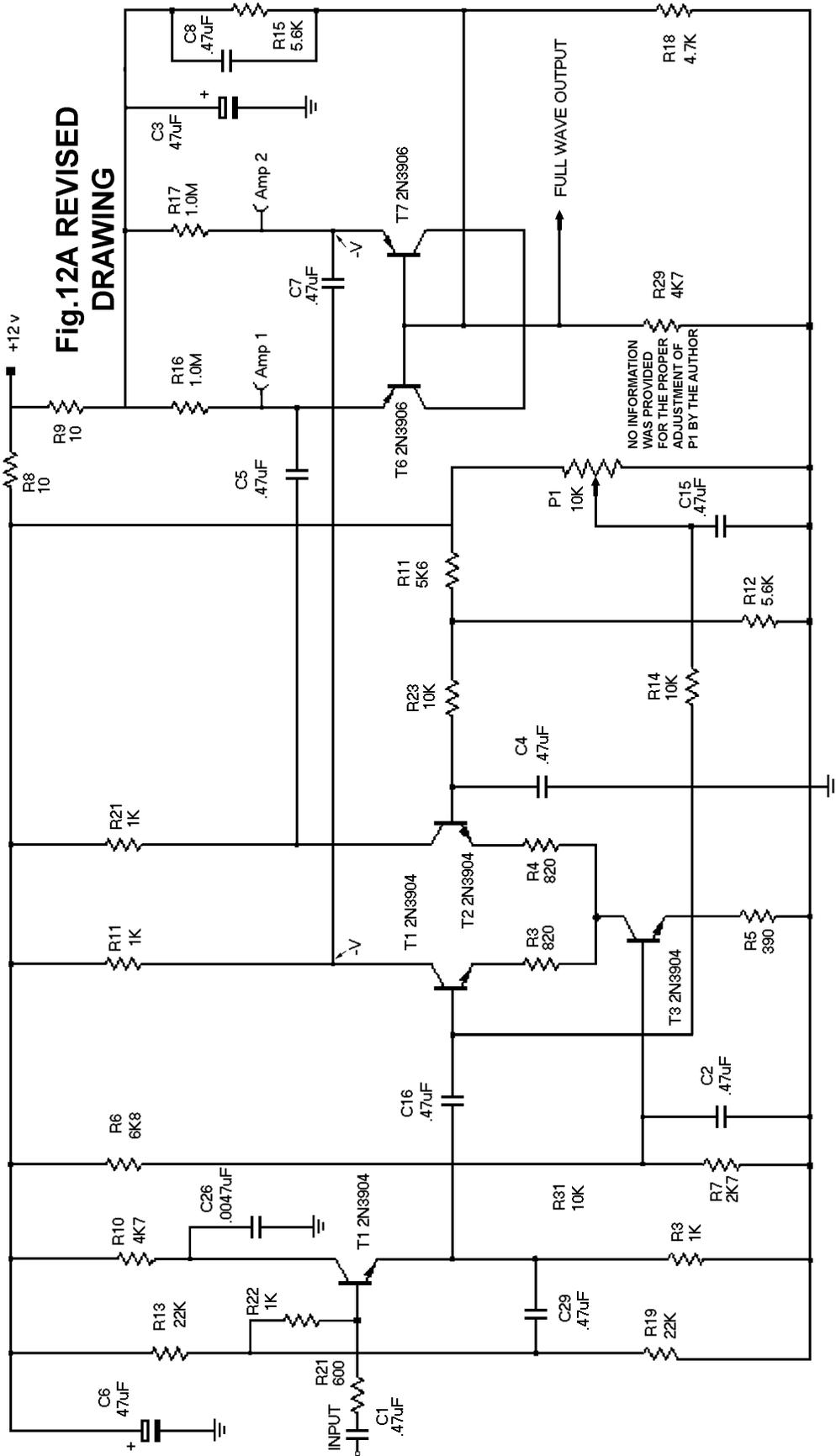
If the capacitor volt drop has increased 101 times due to the amplifier then the feedback capacitive current will be 101 times larger for the same input voltage. The feedback capacitance C_f , now looks 101 times bigger at the amplifier input! The problem occurs when we drive the amplifier input from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

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 r_{bb} 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.

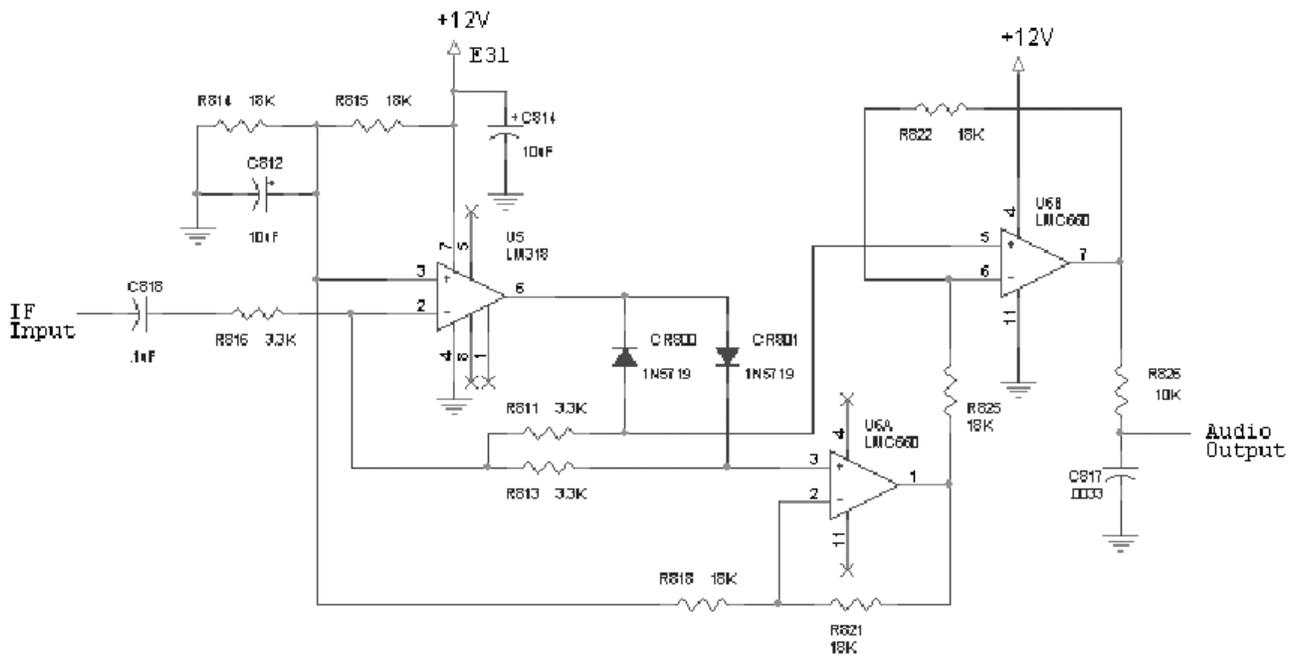
6. Last but not least, I made a great effort not to make errors in my drawing BUT CHECK IT YOURSELF!



An Improved, Precision, Full Wave AM Detector

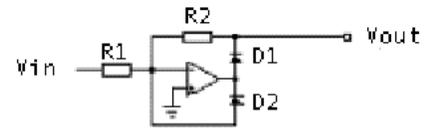
By Rob Schenck, K2CU K2CU@arrl.net

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; $V_{out} = V_{in}$.

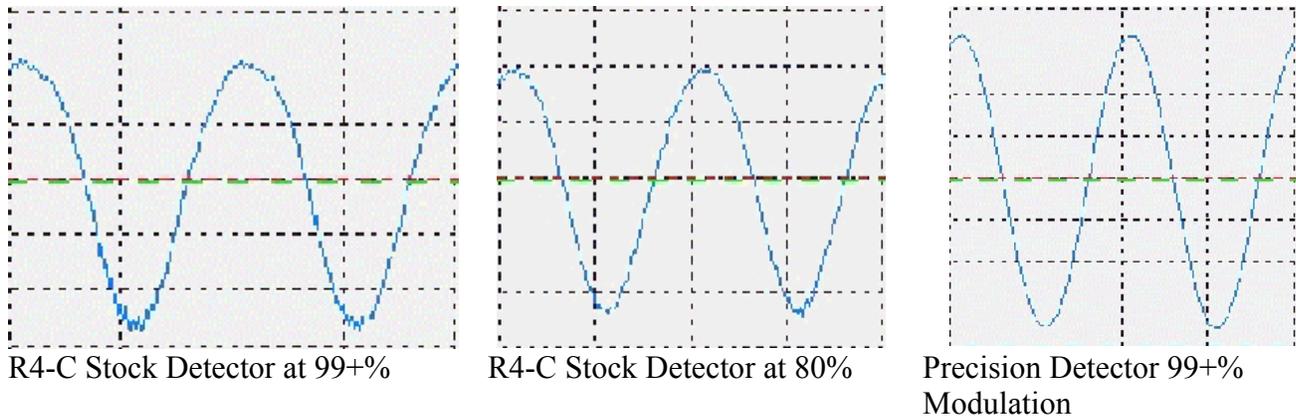
Similarly, the positive going input cycle produces a feedback current through D2. Any non linear voltages developed across the diodes is not relevant as the output is derived from the feedback resistor only.

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



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.

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

The simplistic AM modulated signal is:

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

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.

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

$$\text{Where; } \cos(W_c) t = a \qquad \frac{m}{2} \cos(W_c - W_m) t = b \qquad \frac{m}{2} \cos(W_c + W_m) t = c$$

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:

$$2 (\cos(W_c)t) (\frac{m}{2}\cos(W_c-W_m)t) = \frac{m}{2} [\cos(2W_c - W_m)t + \cos(W_m)t]$$

$$2 (\cos(W_c)t) (\frac{m}{2} \cos(W_c+W_m)t) = \frac{m}{2} [\cos(2W_c + W_m)t + \cos(W_m)t]$$

Which results in double sideband AM at twice the carrier (2Wc) 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:

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

The $\frac{m}{4} \cos(2W_m)t$ 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, W_c , 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 W_c the function is:

$$f(X) = \frac{1}{2} + \frac{2}{\pi} \cos W_c t + \frac{2}{3\pi} \cos 3W_c t + \frac{2}{5\pi} \cos 5W_c t +$$

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, $\cos W_c t$. 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:

$$V_{out}(t) = \{A \cos(W_c) t + \frac{1}{2} AB \cos(W_c - W_m) t + \frac{1}{2} AB \cos(W_c + W_m) t\} * \frac{2}{\pi} \cos W_c t$$

The demodulated sideband terms are:

$$V_{out}(ty) = AB/2\pi * \cos(W_m) t + AB/2\pi * \cos(W_m) t, \text{ or}$$

$$V_{out}(ty) = AB/\pi * \cos(W_m) t$$

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.

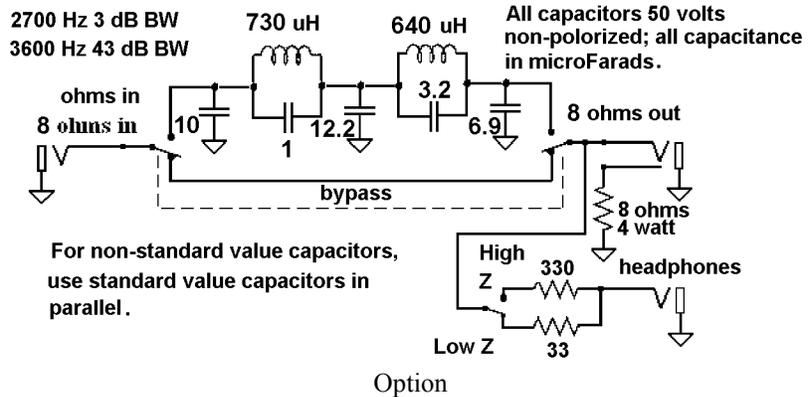
Elliptic Low Pass Audio Filters

Dallas Lankford 7/4/05 Rev 7/10/09

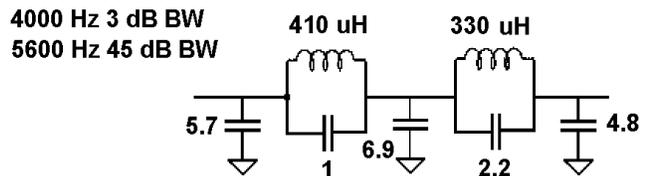
There are a number of different accessories for improving recovered audio from strongly fading MW and SW signals, including audio filters and AM synchronous detectors, for which designers, sellers, and users have often given glowing praise. But I have heard only modest improvement due to these kinds 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. 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.

Elliptic Low Pass Audio Filter

DL 7/4/05 Rev 7/10/09



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 3rd option as I did for two of the prototypes.



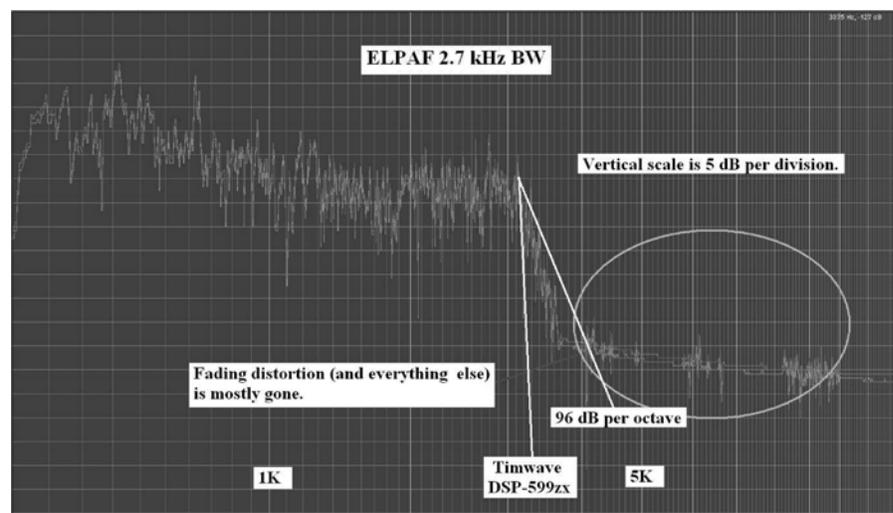
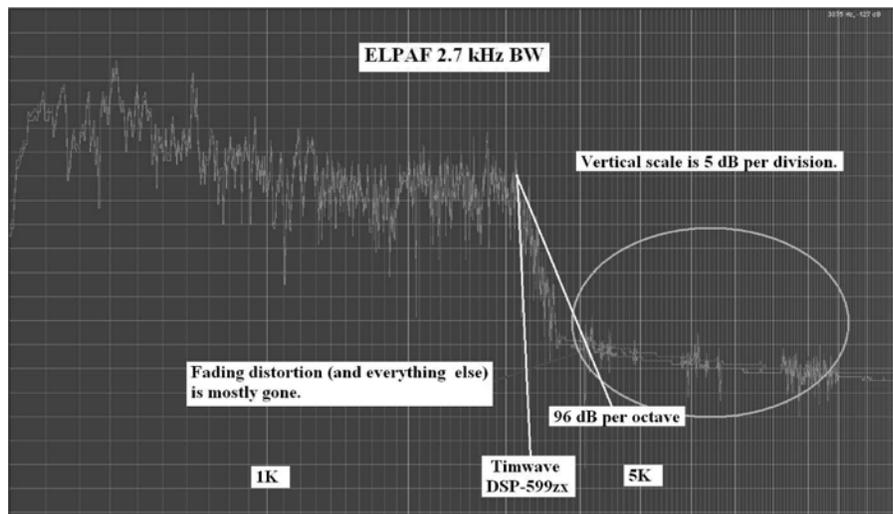
After considering possible kinds of audio filters, it seemed to me that elliptic filters, if they could be scaled to practical audio frequencies, might provide the sharp cutoff I wanted. Also, elliptic filter tables (in the ARRL Radio Amateur's Handbooks) allowed me to easily determine component values 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 ($\mu = 125$) 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.

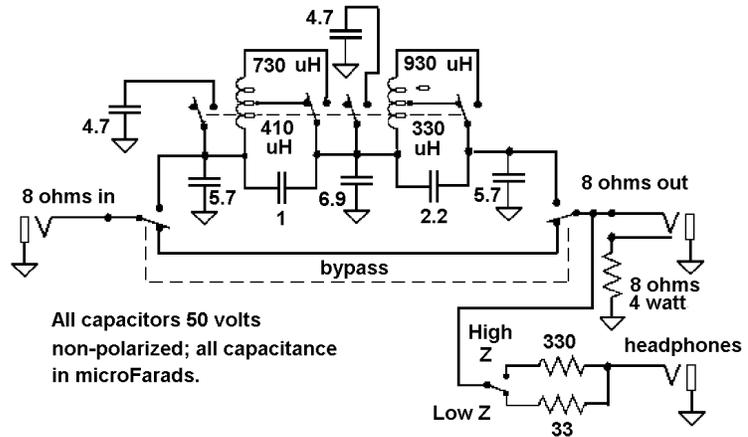
But I could find none in my catalogs. A high quality 6 pole double throw rotary switch is available for about \$30, Electroswitch part # D4G0603N (Mouser 690D4G0603N), so the switched filter can be implemented that way. By taking some minor liberties with the filters designs, a 4 pole double throw toggle 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.

A Simlified Switched 2.7 And 4.0kHz BW ELPAF
DL 11/6/05 Rev 7/10/09



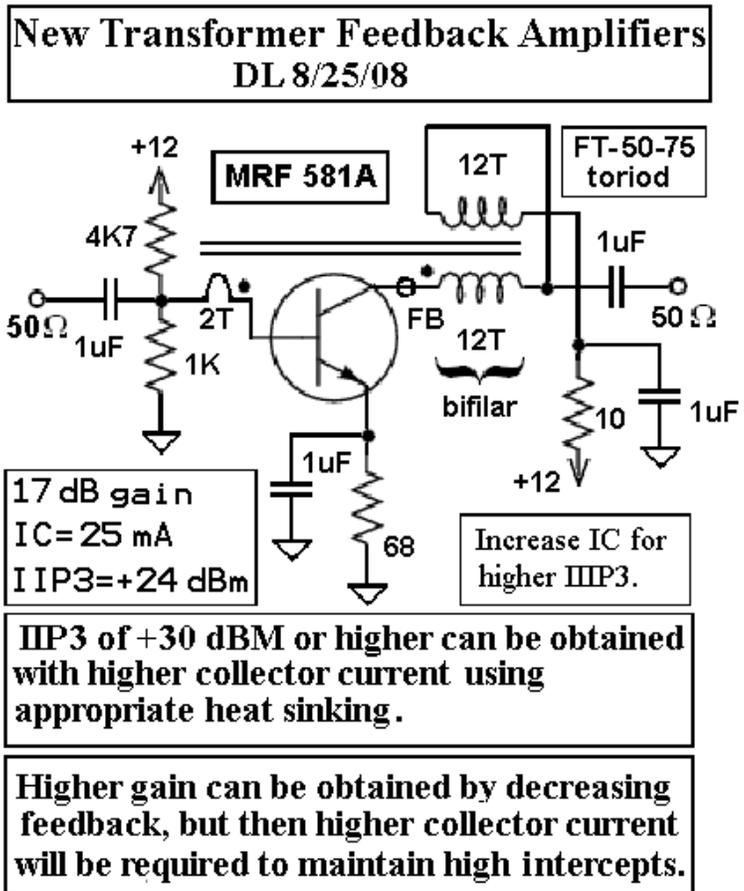
New MRF581A Transformer Feedback Amplifiers

Dallas Lankford, 8/26/08 Rev 7/10/09

The new MRF581A transformer feedback amplifiers described here were developed for use with insensitive low impedance antennas when used with relatively insensitive receivers, and for DXing situations where a high intercept preamplifier with more than normal gain is needed.

The single MRF581A amplifier at right was used with an LC tuned MW whip antenna with low impedance output together with a Grundig G5 portable radio to determine man made noise levels at remote sites.

For that application a high Q inductor consisting of 54 turns of #22 enameled copper wire on an Amidon FT-82-61 toroid with a four turn link was used.

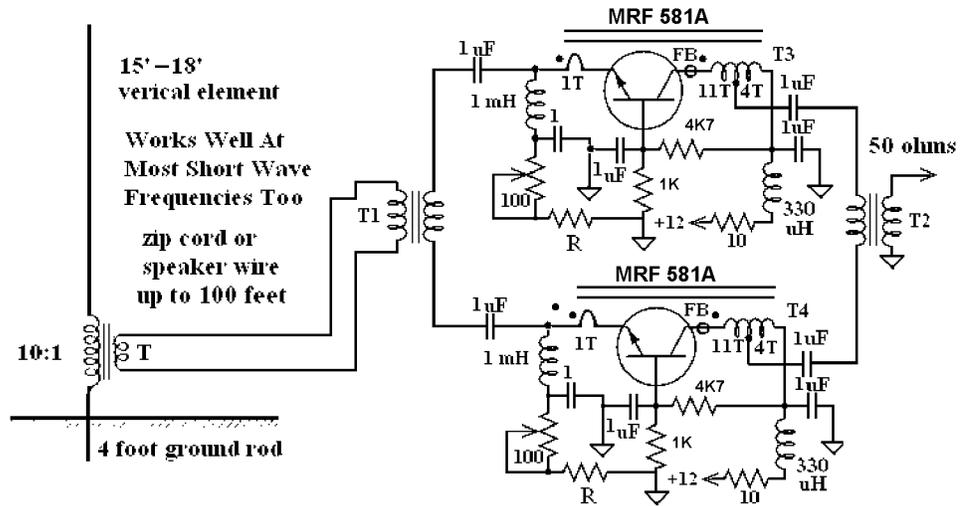


For broadband antennas a push-pull implementation using two of these amplifiers is recommended in order to obtain higher 2nd order intercepts. Substituting BJT's is not recommended. Sensitivity to DC coupled man made noise was observed when using an AC-DC power supply and was subsequently eliminated by a common mode choke consisting of 30 bifilar turns of #22 enameled copper wire on an Amidon FT-114-75 or -J at the DC input to the amplifier.

Short Amplified High Performance MW, LW, And SW Vertical Antennas
 Dallas Lankford, 10/10/06, rev. 5/9/08 rev. 7/10/09

The amplifier is a push-pull Norton amplifier. It is described in several articles in The Dallas Files at www.kongsfjord.no

Antenna T may be an Amidon FT-114-75 or FT-114-J with 80 close-spaced turns of #26 enameled copper wire primary and 8 turns wire of #26 enameled copper secondary at the grounded end of the primary.



It is important to insulate the toroid T with, say, Scotch Glass Cloth Electrical Tape to prevent possible- shorting of turns (high permeability material is a semi-conductor).

If constructed with appropriate hardware, such as a 12' telescoping MJF J whip, a DX Engineering 5' nominal threaded aluminum mast section and coupler, and a Fire-Stick model K4 Stud mount, then it is easy to put up and easy to take down. The- Fire Stick or' equivalent whip element mount should be mounted on a heavy duty plastic electrical junction box, with the transformer T inside. If a metal box is used, water surface tension can cause the whip element to short during and after rain due to acid content of the rain or other ion contamination. A Delrin Antenna Insulator made by GeoTool may be used with a metal box if desired; see: www.geotool.com.

If you don't wish to build the amplifier yourself, you can buy one or two from Kiwa Electronics

The Best Small Antennas For MW, LW, And SW

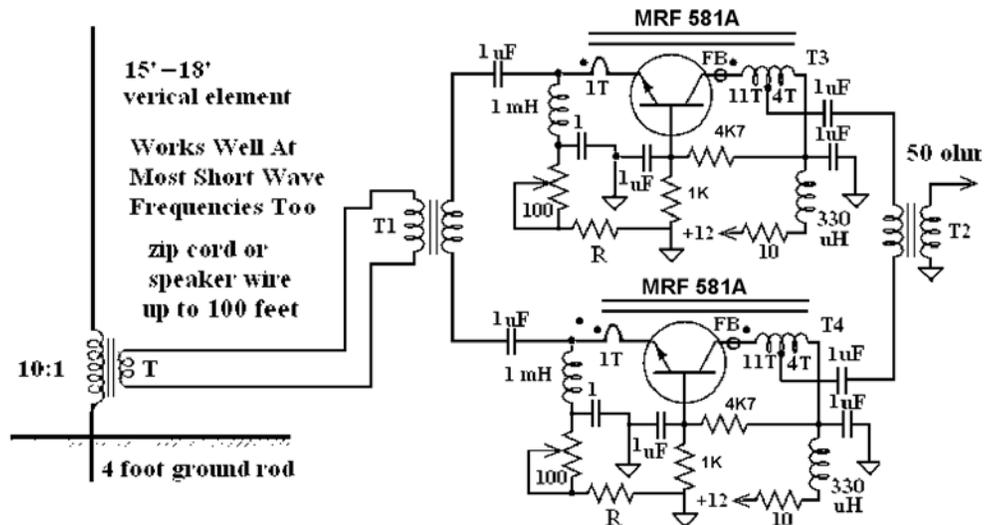
Dallas Lankford, 5/5/08, rev. 3/30/09 rev. 7/10/09

The amplified vertical antenna below was developed during experiments to see how short I could make a noise reducing vertical antenna while maintaining good sensitivity using only a 10.8 dB gain push-pull Norton amplifier (discussed in detail in other articles in The Dallas Files) to bring signal levels back up to a good level. The amplified 15 foot noise reducing vertical antenna is not an active antenna. The push-pull Norton amplifier, which is located at the receiver, is a low impedance device (as opposed to the high impedance FET's used in most active antennas) and consequently does not have the common mode noise problems which active whips and active dipoles sometimes have.

This amplified short noise reducing vertical was tested with twin lead up to 100 feet in length. A pair of these separated by about 60 feet makes a good MW phased array. If you are not a builder, you can buy an equivalent Norton amp from Kiwa Electronics for about \$110 plus shipping (as of May 2008). The gain of the 15 foot noise reducing vertical is about -15 dB, and its 2nd and 3rd order intercepts are typically greater than +120 dBm and +60 dBm respectively in the MW band. When used with a push pull Norton amplifier the cascaded input 2nd and 3rd order intercepts are greater than +95 dBm and +50 dBm respectively in the MW band. It is, in my opinion, the best small omni directional LW-MW-SW receiving antenna, period. A previous version used relay switching for improved performance at higher SW frequencies. Increasing the antenna transformer turns made relay switching unnecessary. I use two of them as my current phased receiving array. The antenna is now excellent for LW, MW, and SW. All of my longer and higher passive inverted L's and verticals and all of my active antennas have been permanently retired.

The amplifier is a push-pull Norton amplifier. It is described in several articles in The Dallas Files at www.kongsfjord.no

Antenna T may be an Amidon FT-114- 75 or FT-114-J with 80 close- spaced turns of #26 enameled copper wire primary and 8 turns wire of #26 enameled copper secondary at the grounded end of the primary.



It is important to insulate the toroid T with, say, Scotch Glass Cloth Electrical Tape to prevent possible- shorting of turns (high permeability material is a semi-conductor).

If constructed with appropriate hardware, such as a 12' telescoping MJF J whip, a DX Engineering 5' nominal threaded aluminum mast section and coupler, and a Fire-Stick model K4 Stud mount, then it is easy to put up and easy to take down. The- Fire Stick or' equivalent whip element mount should be mounted on a heavy duty plastic electrical junction box, with the transformer T inside. If a metal box is used, water surface tension can cause the whip element to short during and after rain due to acid content of the rain or other ion contamination. A Delrin Antenna Insulator made by GeoTool may be used with a metal box if desired; see: www.geotool.com.

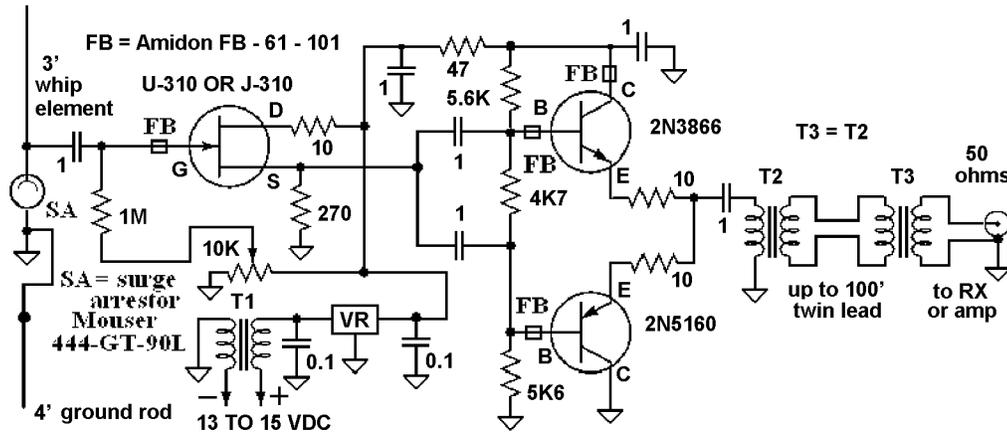
If you don't wish to build the- amplifier yourself, you can buy one or two from Kiwa Electronics.

At least one person has claimed that noise reducing antennas are noisy. But when I quizzed him about his implementation, it turned out that he had not implemented the antenna correctly. If you do not follow the instructions, then you may end up with a noise increasing antenna like he did

Although I have retired my active whips, there still seems to be a place for a high performance LW – MW – SW active whip antenna, such as for phased arrays at temporary locations, or as a compact substitute for the 15 foot noise reducing antenna above.

Revised Simplified complimentary Push-Pull Output Active Amplifier
For Active Whip Antennas With Simple IIP2 Adjustment

Dallas Lankford, 12/7./07 rev. 7/10/09



T1 = 30 bifilar turns #22 enameled copper wire on insulated Amidon FT-114-J.

T2 = 8 bifilar turns #24 enameled copper wire on insulated Amidon FT-50- 75.

VR = 12 volt 1 amp low dropout voltage regulator, Mouser 511-L4940V12.

*****+*****

10 Ohm resistors Should be 1% all other resistors 5% or better, adjust 10K: ohm pot for 0.200 voltage across 10 ohm resistor using 2% accuracy or better DVM.

While this method of adjusting IIP2 is not ideal, it does has the advantage of being simple. A 12 volt voltage regulator is used to establish a constant DC supply voltage because IIP2 of the active whip is sensitive to changes in its DC supply voltage. A number of FET's were installed one at a time in a given active whip circuit and several parameters for each of the FET's was measured. No components other than the FET's of the whip were changed. A 10 ohm 100 resistor was inserted in series with the FET drain to simplify measuring the total FET current. The FET current was found to be the most accurate measure of IIP2. For U-310's maximum IIP2 was $19.5\text{mA} \pm 0.1\text{mA}$. (0.195 ± 0.001 DC). For Motorola J-310's maximum IIP2 was $20.0 \text{ mA} \pm 0.1 \text{ mA}$. For Fairchild J-310's maximum IIP2 was $20.7\text{mA} \pm 0.1 \text{ mA}$. Measurements were made with a DVM having 2% accuracy. The voltage regulator output was measured as 12.04 VDC*. The 511-L4940V12 regulator is rated as $12.00 \pm 0.25\text{VDC}$ '. We picked 0.200 V'DC as the adjustment voltage because it is all approximately the mean of the voltages we measured.

Yon can use the voltage for the particular FET yon use. However our sample was not large enough to determine if that is a better strategy. (*but 11.86 is another example)

Low Noise Active Antennas AC/DC Power Supplies

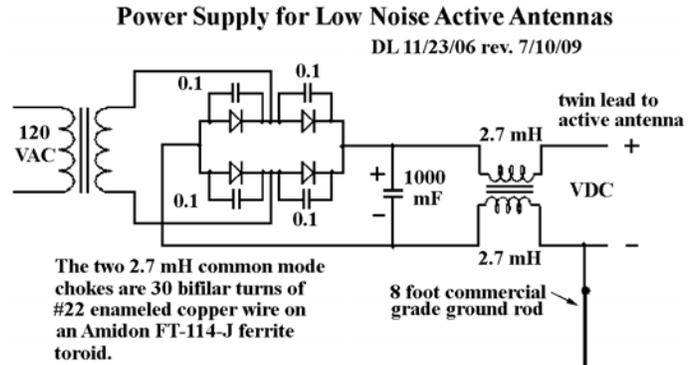
Dallas Lankford, 11/26/06, rev. 2/17/07

By using battery power it can be shown that AC/DC power supplies for active antennas often introduce noise into the signal path. With the power supplies I have used the noise usually begins in the NDB band and increases in intensity as frequency decreases. For some AC/DC power supplies the noise begins in the MW band and may even begin in the SW bands. This noise and the difficulty of eliminating it are the main reasons I have not used active antennas in the past as my main antennas.

Common mode chokes often eliminate the noise in active whips due to AC/DC power supplies and other reasons for MW and higher frequencies; see "Some of my favorite small antennas for MW and LW" in [The Dallas Files](#). For active antennas I have used common mode chokes both at the antenna, as shown in the article above, and at the power supply, as shown in the schematic here. However, common mode chokes alone usually are not sufficient to eliminate noise due to AC/DC power supplies for the lower NDB band and VLF band. For several AC/DC power supplies I tried much larger common mode chokes, as large as 1 Henry, but obtained no additional reduction in noise over the 2.7 mH specified in the schematic above

Recently I bought a Radio Shack # 273-1690 18/24 volt AC-AC adapter, thinking I was buying an AC/DC adapter. Fortunately, there was adequate space inside to convert it to a AC/DC supply, which I did. I used the 18 VAC output ($1.4 \times 18 = 25.2$ VDC nominal, 25.7 VDC measured, which is close enough to 24 VDC), a 100 volt 1.4 amp full wave bridge rectifier Radio Shack # 276-1152, and a 1000 mF 50 volt electrolytic capacitor. Talk about noisy. The noise started at the low end of the MW band and increased as frequency decreased. Bypassing the negative DC lead to the AC neutral lead moved the beginning of the noise down into the middle of the NDB band, about the same as for other AC/DC power supplies I have used with active antennas. But again, using a larger common mode choke did not reduce the noise further.

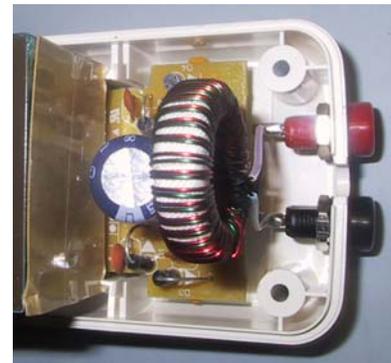
I happened to mention this noise problem to Terry Fugate, who advised me to parallel each diode of the bridge with a 0.1 capacitor. And he sent along a reference from which I quote from as follows. *I have also run into that [noise] problem [...] on LW. The mechanism that causes all of that noise is the rectifier diodes. If you use diodes that have a long minority carrier lifetime, no problem, but with the faster diodes that are being used nowadays, it sounds as if florescent lights are running, once you get below 600KHz. The cure is quite simple; I have posted it on this NG [rec.radio.shortwave] before. First of all, you need to connect .01uF capacitors across each of the rectifier diodes, so that they do not rectify RF.*



Next, you need to bypass each leg of the secondary of the power transformer with a 1uF non-polarized capacitor; a mylar or polyester would do fine. This will give you a power supply that is quiet, all the way down to 10KHz or so. I have done this mod on the power supplies for the Yaesu FRG-100, the ICOM R75, the AOR3030, the AOR7030, the Palstar R30, the Collins KWM-380, and even that old Radio Shack long range AM receiver that was sold in the 70s. Doug DeMaw gets the credit for bypassing those power supply diodes; I came up with the idea for bypassing the power transformer secondary. Pete KE9OA.

I modified my AC/DC power supply as Terry recommended, using 0.1 caps across each diode of the bridge as shown on the schematic above, and low and behold most of the noise was gone. Next I unsoldered the secondary bypass to AC neutral in order to evaluate that part of the mod. Curiously, bypassing the secondary to the AC neutral seemed to increase noise slightly, so I removed that part of the mod. Some additional noise reduction was obtained by grounding the negative DC lead on the antenna side of the power supply common mode choke to an 8 foot commercial grade ground rod.

I intended to use the 9 VAC part of a dual 9/12 VAC radio shack transformer for a 12 VDC transformer ($1.4 \times 9 = 12.6$ VDC nominal), but my local Radio shack did not have any. So I used a 12 volt AC/DC # 273-1776. It used individual rectifiers for the full wave bridge and a PC board which made it relatively easy to modify using small 1 mF 50 volt ceramic capacitors as shown in the photo here. The capacitors and their leads do not touch the PC board traces, but are about 1/32 inch above the trace surfaces which leaves adequate clearance between the bottom of the PC board and the inside bottom of the case. It turned out that the diodes in this power supply were not noisy like the diodes in the bridge I used for the 24 VDC supply because bypassing them with the 1 mF capacitors did not reduce noise at the lower end of the NDB (there was very little excess noise to begin with). Also, the dual 2.7 mH common mode chokes did not reduce noise at the low end of the NDB band either (again, there was very little excess noise to begin with). However, grounding the negative DC lead with an 8 foot commercial grade ground rod did reduce noise at the low end of the NDB band. I considered omitting the common mode chokes in the 12 VDC supply, but decided to include them in case common mode noise makes its way into my house from the outside. On both power supplies I used insulated banana jacks, Radio Shack # 274-725B. The flimsy two wire (white) DC power cords of both power supplies were removed and discarded. The PC board solder pads where the DC power cords were removed were used for the mods. Banana plugs were used to connect the twin lead power feed of the active antennas to the power supplies banana jacks. Two photos of the 24 volt AC/DC power supply are included above.



While testing the low noise 12 VDC power supply about 2345 on 11/25/06 using my relatively new active whip (see below) I came across 183 kHz Felsberg, Germany in French relaying Europe 1 from Paris. The 183 kHz 2 relay of Europe 1 is uncommon here and almost never produces any understandable audio. So it was a first for me to hear it producing clear audio.

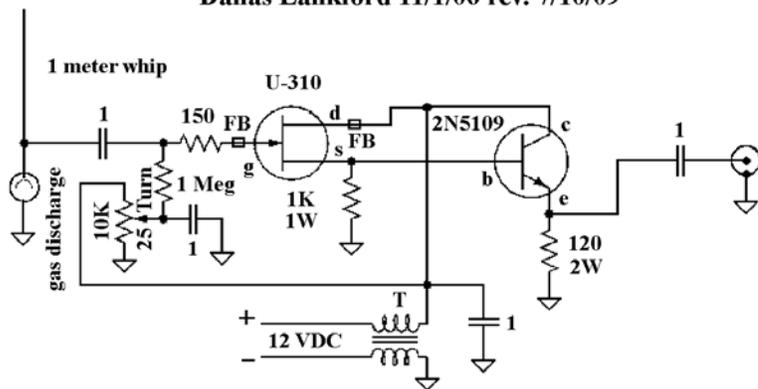
I switched over to my 45 foot noise reducing vertical and got another surprise; there was hardly any clear audio from the big antenna... business as usual. Just for fun I gave my ALA 100 clone a shot at

183 Europe 1; nope, no audio at all. I switched back and forth among the three antennas multiple times and got virtually identical results every time. After 183 closed down just after 0000, I tuned down to 162 Allouis, France. They are also uncommon here, rarely producing understandable audio. But on the whip they produced excellent audio, just beautiful, while the 45 foot vertical was very poor, and the ALA 100 clone was only slightly less poor. The receiver used was a modified IC-746 Pro using a 3.6 kHz BW and ECSS, USB or LSB depending on which had less interference, with the automatic notch engaged.

Even though the active whip was only a few meters away from the 45 foot noise reducing vertical and ALA 100 clone, the active whip placement turned out to be the reason it produced clear audio while the larger antennas did not. However, without the modified AC/DC power supply, even with the lucky placement I doubt the active whip would have produced any audio at all from 183 Germany and 162 France. Listeners on the ECNA, and especially New England, probably hear these signals most every evening, and often like locals. So they may not care whether their active whip antenna power supply is low noise. But I do because hearing these LF signals clearly in North Louisiana is a rare occurrence, and would not occur here on an active whip without a low noise AC/DC power supply.

Here is a relatively new 12 volt whip / dipole which I used to evaluate the 12 VDC AC/DC power supply mods described above. Some might consider the 3rd order intercept somewhat low, at about 34 dBm, but the 2nd order intercept is quite good, about 69 dBm. Also, this one uses a common and inexpensive U-310 FET, and no transformers for the whip. For a dipole, build a second whip, oppose the whips, and join the two outputs with a 1:1 balun. This version is

More New Active Whip And Dipole Antennas
Dallas Lankford 11/1/06 rev. 7/10/09



shown with coax output, but I always use a 1:1 balun and twin lead. If you use coax for the signal lead-in, you may need to use a common mode choke on the coax. The 25 turn pot is adjusted for minimum 2nd order intermod. This occurs at about 60 mA current drain for the 2N5109, so a heat sink is advisable. The U-310 draws about 18 mA, and while it runs quite warm to the touch, a heat sink is not really necessary for it. I used separate 1 mH common mode chokes for T instead of a 2.7 mH bifilar choke. The 24 VDC AC/DC power supply mods were evaluated using an ultra high intercept whip (+50 / +100 dBm II3 / II2) which I have agreed not to disclose.

Continuing the antenna comparisons the next day on MW groundwave signals, the active whip man made signal to noise ratio was often not quite as good as the 45 foot noise reducing vertical. For one signal, the 45 foot vertical produced clear audio, while the whip produced mostly noise. The ALA 100 clone signal to man made noise ratio was often not quite as good as the active whip, but on a few signals was better, due to the ALA 100 clone nulling local man made noise. On a few other signals the ALA 100 clone produced no audio at all compared to clear audio produced by the whip, due to the ALA 100 clone null being pointed at the desired signal, which illustrates one disadvantage of a large air core loop. The bottom of the whip for these comparisons was 2 meters above the ground. Placing the whip at a higher location might improve its signal to man made noise ratio. We will see.

As a result of these experiences I have revised my opinion of active whips. When used with a low noise AC/DC power supply, they are quite respectable, and with some placements and at some frequencies are superior to full size antennas.

Receiver noise figure sensitivity and dynamic range - what the numbers mean

From Ham Radio Magazine October 1975 Author: James R. Fisk, W1DXY (SK),
Ham Radio ceased publication in 1990 Repaginated and edited by Perry Sandeen Sept. 2008

A complete discussion of receiver sensitivity, intermodulation distortion, and gain compression, and what they mean in terms of performance

When it came to receivers, the earliest amateur operators were concerned primarily with sensitivity and experimented almost endlessly with different types of crystals, trying to find the one that was the most sensitive. Then came DeForest's Audion, and Armstrong's regenerative detector, and amateurs who could afford the tubes found they had all the sensitivity they could use. However, as the hobby grew, and more and more amateurs started populating the band below 200 kHz, the simple regenerative detector simply wasn't up to the task. Selectivity, with simple tuned input circuits, was practically nonexistent, and the regenerative detector hopelessly overloaded in the presence of strong signals.

In the early 1920s amateurs worked to improve their tuners, but even the so-called "Low-Loss" tuners were only marginally acceptable. Although several superheterodyne designs were described in the amateur magazines, it wasn't until low-cost, commercial i-f transformers became available in the late twenties that the superhet saw widespread amateur use. Selectivity against interfering signals was still a problem, however, and James Lamb revolutionized receiver design in 1932¹ with his "single-signal" CW circuit which used an i-f stage with extremely high selectivity - provided by regeneration or a simple crystal filter.

The single-signal, single-conversion superhet of the late 1930s suffered from poor RF image response at the higher frequencies, but it wasn't too severe on 14 MHz and few amateur receivers of the day, in fact, tuned much above 18 or 20 MHz (15 meters was not yet assigned to amateur use and most 10-meter operators used specialized receivers or converters). When the 10- and 15-meter bands opened up after the war, however, the poor RF image response of the single - conversion superhet with a 455-kHz i-f had to be faced - it was solved by going to a double-conversion layout with a first conversion to 2 or 3 MHz to minimize RF image response, and a second conversion to 455 kHz or lower for adjacent channel selectivity.

Although amateur radiotelephone operation in the 1930s was relatively limited, the huge growth of a-m activity after the war demanded improved adjacent-channel phone selectivity. While the crystal filter provided excellent selectivity for CW operation, it was of little or no use on a-m or ssb and some phone operators started using a Q5er an outboard 80-kHz i-f strip - for improved phone selectivity. This led to the triple-conversion superhets which were the rage of the 1950s.

As pointed out by Goodman², however, the multiple-conversion design had many shortcomings, including high-selectivity i-f which made it practically impossible the large number of stages between the antenna and the to attenuate strong, adjacent signals.

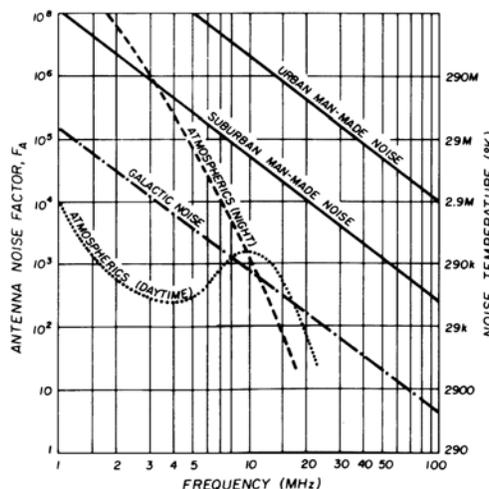


fig. 1. Receiver sensitivity is limited by the external available noise power which varies with frequency. For quiet, rural locations galactic noise is the limiting factor down to about 18 MHz, and atmospheric noise dominates below 18 MHz.

And, with at least three oscillators running at the same time, it was difficult to avoid the many spurious signals which were generated within the system. He advocated a return to the single-conversion superhet using highly-selective, high-frequency, crystal lattice filters which were just then becoming commercially available.

The 1960s saw a return to single conversion designs, the use of high frequency, crystal-lattice filters and the widespread use of semiconductors. With modern devices receiver sensitivity was no longer limited by the RF amplifier (or mixer) stage, but by the external galactic and man-made noise. Cross modulation and overload, on the other hand, were becoming a serious problem as more and more amateurs started using high-power linears and large, directive antennas.

Modern communications receivers, therefore, in addition to meeting stringent frequency accuracy, stability, sensitivity, and selectivity requirements, must provide freedom from cross modulation, intermodulation distortion and blocking. Some modern, solid-state solutions to these design goals were discussed recently by Rohde³.

The specifications for a typical modern communications receiver might list sensitivity of 0.5 μ V for 10 dB signal plus-noise-to-noise (S+N/N) ratio, intermodulation distortion of -65 dB, "wide" dynamic range, and "virtual elimination" of overload from adjacent signals.

However, is 0.5 μ V sensitivity for 10 dB S+N/N adequate for operation on the high-frequency bands? For satellite communications on 10 meters? What is - 65 dB intermodulation distortion in terms of signal strength? "Wide dynamic range" and "virtual elimination of overload" are obviously advertising superlatives without definition but what, exactly, can you expect from a high quality, modern receiver design? Perhaps, if these performance data were defined, and amateurs understood what they meant, manufacturers would be encouraged to use no-nonsense numerical data. Only then can amateurs compare the dynamic range and cross-modulation performance of one receiver against that of another.

sensitivity

The minimum usable signal or sensitivity of a receiver is determined by the noise in the receiver output. This can be noise generated within the receiver, thermal noise generated by losses in the transmission line, or atmospheric, manmade or galactic noise picked up by the antenna. As shown in fig. 1, external noise sources are likely to be the limiting factor up to 100 MHz or so.⁴ In urban areas man-made noise predominates and measurements indicate the average level of man-made noise in suburban areas is about 16 dB lower. In a quiet, rural location which has been chosen with care the man-made noise may be near the galactic noise level, but few amateurs are so fortunate.

Atmospheric noise usually predominates in quiet locations at frequencies below about 20 MHz and is produced by lightning discharges so the level depends upon a number of variables including frequency, weather, time of day, season and geographical location. This type of noise is particularly severe during rainy seasons near the equator and generally decreases at the higher latitudes. More complete data on high-frequency atmospheric noise is given in reference 5.

Galactic or cosmic noise is defined as RF noise caused by disturbances which originate outside the earth or its atmosphere. The primary causes of this noise, which extends from 15 MHz well into the microwave region, are the sun and a large number of noise sources distributed chiefly along the Milky Way. Solar noise can vary as much as 40 dB from "quiet" sun levels (low sunspot activity) to periods of "disturbed" sun (high sunspot activity).

Galactic noise from the center of the Milky Way is about 10 dB below the noise from a "disturbed" sun, whereas noise levels from other parts of the galaxy can be as much as 20 dB lower. This is important in satellite communications and will be discussed later.

thermal noise

The free electrons in any conductor are in continuous motion - motion that is completely random and is the result of thermal agitation. The effect of this electron motion is to cause minute voltages which vary in a random manner to be developed across the terminals of the conductor.

$$e^2 = 4kTBR \tag{1}$$

where e^2 = mean square noise voltage
 k = Boltzmann's constant = 1.38×10^{-23} joules/o K
 T = absolute temperature, ° K
 B = bandwidth, Hz
 R = resistance, ohms

Since this phenomenon was first demonstrated by J. B. Johnson in 1928⁶, thermal noise is sometimes known as Johnson noise. At the same time, H. Nyquist showed, on the basis of the statistical theory of thermodynamics, that the mean square noise voltage generated in any resistance can be expressed as ⁷

Note that the noise voltage is dependent upon the bandwidth across which it is measured. This implies that noise is evenly distributed across all frequencies which, for all practical purposes, it is. *

*** At extremely high frequencies statistical mechanics is no longer valid, and eq. 1 must be revised on the basis of quantum theory. This equation is valid, however, to at least 6000 GHz.⁸**

Although noise bandwidth is not precisely the same as the 3-dB bandwidth of a receiver, in modern receivers with high skirt selectivity the 3-d B bandwidth can be used in eq. 1 with little error.

The equivalent circuit of any impedance as a source of noise voltage is shown in fig. 2A. Note that the thermal noise voltage is dependent only on the resistive component and is independent of any reactance in the circuit. As might be expected, maximum noise power is transferred from a thermal noise source when the load impedance presents a conjugate match to the source impedance. This is represented in fig. 2B where the load resistance, R_L , is equal to the source resistance. Since $R = R_L$, the noise voltage developed across the load is $e/2$, and from Ohm's law:

$$P = \frac{E^2}{R} = \frac{(e/2)^2}{R} = \frac{e^2}{4R} \text{ watts} \tag{2}$$

Substituting the value of e^2 from eq. 1 into eq. 2, the power

which can theoretically be transferred under such conditions is called the available noise power and is given by

$$P_n = kTB \tag{3}$$

The factor of $4R$ has cancelled out so the available noise power does not depend upon the value of the resistance. This is significant because it means that the available noise power of any resistor (or any noise source), if measured over the same bandwidth, can be represented by a resistor at temperature T . Thus, every noise source has an equivalent noise temperature.

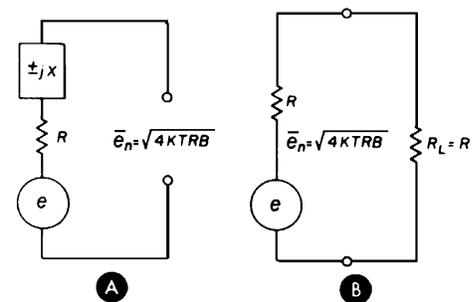


Fig. 2. Mean noise voltage depends on temperature, resistance and bandwidth, and is completely independent of reactance as shown in (A). Maximum noise power is transferred to the load when the load resistance is matched to the source resistance (B).

The actual noise power dissipated in the load resistance may be affected by loss in the connecting leads, noise power generated in the load resistor itself, or a less than perfect match to the original resistance. This property is sometimes used in low-noise uhf amplifiers by creating a deliberate (but carefully determined) mismatch between the input termination and the detection device so that something less than the available termination noise power is coupled into the detector.

signal-to-noise ratio and receiver noise figure.

The relation of signal amplitude to noise is commonly referred to as the signal-to-noise (S/N) ratio. Unfortunately, this ratio has not been well standardized and is often used interchangeably to mean the ratio of rms signal voltage to rms noise voltage, the ratio of peak signal voltage to peak noise voltage and, in pulse systems, the ratio of peak signal power to average noise power. Therefore, when discussing SIN ratio, it's important to determine exactly which ratio is being referred to.

Although the minimum discernible signal (MDS) that can be heard above the receiver noise level is sometimes used as an indication of receiver sensitivity, it is extremely subjective because it differs many dB from measurement to measurement, and from one operator to another (some experienced weak-signal operators can detect signals as much as 20 dB below the noise level while other operators may have difficulty discerning signals which are equal to the noise level).⁹

Receiver sensitivity has also been defined in terms of a signal-to-noise ratio of unity (signal equals noise)* or equivalent noise floor, but this is difficult to measure unless you have a calibrated signal generator and a spectrum analyzer.

***This is sometimes erroneously referred to as tangential sensitivity. Tangential sensitivity, however, corresponds to a signal-to-noise ratio of 6.25 and is about 8 dB higher.¹⁰**

Noise figure or noise factor, on the other hand, is less susceptible to measurement errors than sensitivity and, since its introduction in 1944 by Friis,¹¹ it has become the accepted figure of merit for receiver sensitivity. Noise figure, NF, is simply noise factor, F, expressed in dB.

$$NF = 10 \log F \text{ (dB)} \tag{4}$$

The concept of noise factor allows the sensitivity of any amplifier to be compared to an ideal (lossless and noiseless) amplifier which has the same bandwidth and input termination. As far as noise is concerned, that part of a receiver between the antenna and the output of the i-f amplifier can be regarded as an amplifier. The fact that the mixer stage shifts the frequency of the noise does not change the situation – it merely causes the noise to lie in a different place in the spectrum from the input noise. The only exception is when the receiver has poor RF image rejection. In this case the noise figure of the receiver is 3 dB worse than it would be if the same receiver had good RF image rejection because the image noise appears at the output along with noise associated with the desired received frequency. This effectively doubles the noise at the output of the i-f amplifier.*

The noise factor, F, of a receiver is defined as:

$$F = \frac{S/N \text{ (ideal receiver)}}{S/N \text{ (practical receiver)}} = \frac{S_i/N_i}{S_o/N_o} \quad \begin{array}{l} S_i = \text{available signal input power} \\ N_i = \text{available noise input power} \\ S_o = \text{available signal output power} \\ N_o = \text{available noise output power} \end{array} \tag{5}$$

Using this definition, it can be seen that an ideal receiver adds no noise to a signal so its output signal-to-noise ratio is the same as that at the input and the noise factor, $F = 1$.

***The noise figure is always defined at the input of the final detector(i-f output) because the noise output of a detector (but not of a mixer) is affected by the presence of a signal. An fm signal, for example, will suppress weak noise but will be suppressed itself by strong noise.**

Since the available noise input power, N_i is defined as kT_oB in eq. 3, and the power gain of the system $G = S_o/S_i$, eq. 5 can be rewritten as

$$F = \frac{N_o}{GkT_oB} \tag{6}$$

Where T_o is 290⁰ kelvin (IEEE definition). With the receiver noise factor defined in terms of noise output power, N_o , power gain, G , and noise input power, kT_oB , noise factor can be easily correlated to receiver sensitivity. Consider the case where the output signal to-noise ratio, S_o/N_o , is unity.

$$S_o = N_o$$

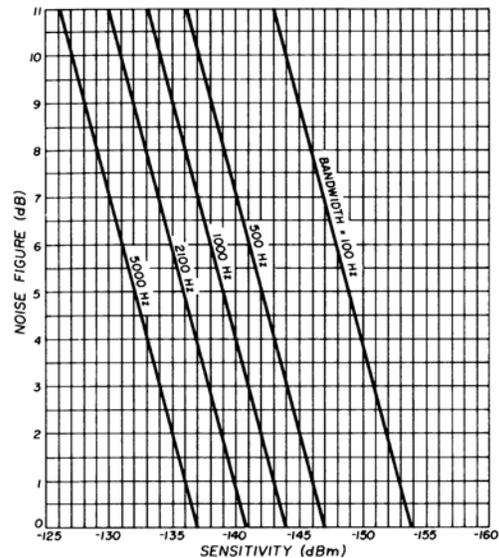


Fig. 3. Receiver sensitivity (-dBm) vs receiver noise figure and bandwidth for unity signal to-noise ratio (signal equals noise). Add 10 dBm for a SIN ratio of 10 dB. Table 1 lists microvolt sensitivity for 50- and 75-ohm systems in terms of dBm sensitivity.

For this specialized case, eq. 6 can be rewritten

$$F = \frac{S_i}{kT_oB} \tag{7}$$

When the temperature is 290⁰ K and the bandwidth is in kHz, $kT_oB = 4 \times 10^{-15}$ mW per kHz. Rewriting eq. 7 in terms of dBm (dB referenced to 1 mW)

$$S_i = 10 \log kT_o + 10 \log B + NF = -144 + 10 \log B_{kHz} + NF \text{ (dBm)} \tag{8}$$

This function is plotted graphically in fig. 3 for bandwidths commonly found in amateur communications receivers. For example, assume a high-frequency receiver has an 8 dB noise figure at 14.2 MHz and a bandwidth of 2.1 kHz. From eq. 8 or fig. 3, the noise floor of the receiver at 14.2 MHz is at about -133 dBm. An input signal of -123 dBm (10 dB greater) would be required for a 10 dB SIN ratio.* To convert dBm to microvolts, recall that:

$$E = \text{the square Root of } RP \tag{9}$$

where E is in volts, R is resistance in ohms and P is power in watts. Since -123 dBm is 5.01×10^{-16} watts, -123 dBm is equivalent to 0.16 μ V across a 50-ohm input termination. However, for a matched signal source, as shown in fig. 4, where the source resistance, R_s , is equal to the load resistance, R_L the source voltage must be twice the voltage across R_L because of the voltage dividing effect of the two series resistors in the network. For a matched 50-ohm source, therefore, an input signal of -123 dBm requires a source voltage of 0.32 μ V. A chart of dBm vs microvolts for matched 50- and 75-ohm systems is presented in table 1.

*This is the signal-to-noise ratio in ssb and CW reception. The SIN ratio of a-m and nbfm signals is somewhat less because a-m (and nbfm) detection use only the envelope as a useful output and the SIN ratio must be reduced by a factor which is related to percentage of modulation (or modulation index).

Because of this two-to-one voltage dividing effect, you must be very careful when comparing the sensitivity of one receiver against that of another. An input of +119 dBm, for example, implies an input directly at the receiver terminals and is 0.25 μV rms across 50 ohms. Sensitivity, on the other hand, implies the use of a matched signal generator so sensitivity of 0.25 μV corresponds to an input of -125 dBm.

This is a 6 dB difference. Since most amateur receiver manufacturers tend to use sensitivity specifications, there is no advantage, but the difference must be considered when you calculate the receiver noise figure and dynamic range. In this article inputs will be stated in dBm as this eliminates the 6 dB conversion factor - **table 1** can be used to convert to *sensitivity*.*

cascaded stages

A relatively simple equation for the noise factor of a receiving system, in terms of the individual stage gains and noise factors is

$$F_T = F_1 + \frac{F_2 - 1}{G_1} + \frac{F_3 - 1}{G_1 G_2} \tag{10}$$

- F_T = overall system noise factor
- F_1 = noise factor of the first stage
- F_2 = noise factor of the second stage
- F_3 = noise factor of the third stage
- G_1 = power gain of the first stage
- G_2 = power gain of the second stage
- G_3 = power gain of the third stage

	μV dBm 50 ohms	μV 75 ohms	P.V μV dBm 50 ohms	75 ohms	
-76	70.8	86.7	-111	1.26	1.54
-77	63.2	77.4	-112	1.12	1.38
-78	56.2	69.1	-113	1.00	1.23
-79	50.2	61.5	-114	0.90	1.09
-80	44.8	54.9	-115	0.80	0.97
-81	39.8	48.7	-116	0.71	0.87
-82	35.6	43.6	-117	0.63	0.77
-83	31.6	38.7	-118	0.56	0.69
-84	28.2	34.5	-119	0.50	0.62
-85	25.2	30.9	-120	0.45	0.55
-86	22.4	27.4	-121	0.40	0.49
-87	20.0	24.5	-122	0.36	0.44
-88	17.8	21.8	-123	0.32	0.39
-89	15.8	19.4	-124	0.28	0.35
-90	14.2	17.3	-125	0.25	0.31
-91	12.6	15.4	-126	0.22	0.27
-92	11.2	13.8	-127	0.20	0.25
-93	10.0	12.3	-128	0.18	0.22
-94	9.0	10.9	-129	0.16	0.19
-95	8.0	9.7	-130	0.14	0.17
-96	7.1	8.7	-131	0.13	0.15
-97	6.3	7.7	-132	0.11	0.14
-98	5.6	6.9	-133	0.10	0.12
-99	5.0	6.2	-134	0.09	0.11
-100	4.5	5.5	-135	0.08	0.10
-101	4.0	4.9	-136	0.071	0.087
-102	3.6	4.4	-137	0.063	0.071
-103	3.2	3.9	-138	0.056	0.069
-104	2.8	3.5	-139	0.050	0.062
-105	2.5	3.1	-140	0.045	0.055
-106	2.2	2.7	-141	0.040	0.049
-107	2.0	2.5	-142	0.036	0.044
-108	1.8	2.2	-143	0.032	0.039
-109	1.6	1.9	-144	0.028	0.035
-110	1.4	1.7	-145	0.025	0.031

Table 1. Microvolt sensitivity vs dBm for matched 50- and 75-ohm receiving systems.

fig. 4. When a signal source is matched to a load, the voltage across the load is one-half the source voltage because of the voltage dividing effect of the source and load resistors. When making receiver sensitivity measurements in the laboratory, a G-dB attenuator is placed in the line so sensitivity can be read directly from the signal generator's internal calibrated attenuator.

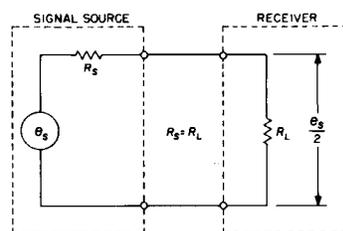
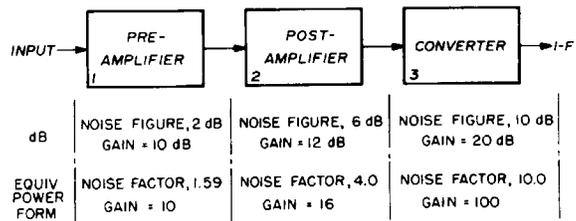


fig. 5. Noise factor of cascaded stages can be calculated by using equation 10 where all quantities are in power ratios. Noise factor at the input of the three cascaded stages shown here is 1.946 (noise figure = 2.9 dB).



Note that all terms are in power ratios. If the gain of the first stage is high, and the noise factor of the second stage is low, the overall system noise factor is determined primarily by the first stage and the third term of eq. 10 may be dropped.

For example, consider the block diagram of the first three stages of a typical vhf receiving system shown in fig. 5. The overall system noise factor is:

$$F_T = 1.59 + \frac{4 - 1}{10} + \frac{10 - 1}{10 \cdot 16}$$

$$= 1.59 + 0.3 + 0.056 = 1.946$$

For this receiver the overall system noise figure is 0.9 dB higher than that of the preamplifier alone. Raising the preamplifier gain to 13 dB or

$$NF_T = 10 \log 1.946 = 2.9 \text{ dB}$$

dropping the noise figure of the second stage to 4 dB ($F = 2.5$) would reduce the system noise figure to approximately 2.45 dB ($F = 1.75$). Depending on the frequency and the application, this may be a worthwhile improvement.

noise temperature

It is often convenient when working with very low noise uhf and microwave receivers to represent the noise figure of the receiver as an equivalent noise temperature. This is because the noise temperature of a receiving system varies over the range from 0^0 to 174^0 K as the noise factor varies from 1.0 to 1.6 (0 to 2 dB noise figure) and noise calculations using equivalent noise temperatures will provide better accuracy.

As mentioned above, the noise figure of an ideal receiver is 1, so the component of receiver noise figure which is due to internally generated noise, N_r , is $F - 1$. Therefore $N_r = (F - 1) GkT_oB$ (11)

The internally generated noise, N_r , can be represented by a noise power, GkT_rB , where T_r is the equivalent noise temperature. Substituting into eq. 11, this equivalent noise temperature can be expressed in terms of the reference noise temperature, $T_o = 2900\text{K}$. $T_r = (F - 1) T_o$ (12)

Eq. 12 can be easily rearranged to express noise factor as a function of the receiver noise temperature, T_r , and the reference noise temperature, T_o

$$F = 1 + \frac{T_r}{T_o} \quad (13)$$

Although noise temperature is seldom used by amateurs, it is a more basic unit than noise factor and is actually easier to deal with, both in understanding concepts and making practical noise calculations. For a more complete discussion of noise temperature, see reference 12.

*The usual procedure for measuring receiver sensitivity is to place a 6 dB attenuator between the signal generator and the receiver. Receiver sensitivity can then be read directly from the signal generator's calibrated output attenuator.

transmission lines

The transmission line’s contributions to receiver noise come from a common source: line losses. The first of these is the more obvious. When a signal travels down a lossy transmission line, the signal is attenuated. This reduces the signal-to-noise ratio and is equivalent to increasing the noise factor of the receiver. This increase can be calculated by introducing a loss factor, L , which is the loss of the cable expressed as a power ratio.

The second effect is due to the noise factor of the transmission line. The fact that the line has losses implies that there is a loss resistance associated with it (which is distinct from characteristic impedance). Since the line is warm it generates noise due to thermal agitation.

The noise factor of the line, F_t , is related to the loss factor, L , and the physical temperature of the line, T_t , by the following equation:

$$F_t = \frac{(\frac{1}{L} - 1) T_t}{290} + 1 \tag{14}$$

The degradation of the receiver noise figure due to transmission line contributions may be calculated by considering the transmission line and the receiver as cascaded stages and using a form of eq. 10

$$F_{tr} = F_t + \frac{F_r - 1}{L} \tag{15}$$

- where F_{tr} = noise factor of the receiver and transmission line
- F_r = noise factor of the receiver
- F_t = noise factor of the transmission line
- L = loss factor of the transmission line

For example, a receiver with a noise factor of 4 (NF = 6 dB) is used with a transmission line which has a loss factor of 0.63 (2 dB loss). The physical temperature of the line is 300⁰ K ($F_t = 1.61$). What is the combined noise figure?

When line losses are low but receiver noise figures are 3 dB or greater, line loss is the predominate contributor to increased noise figure. When receiver noise figure is very low, the thermal effect predominates.

$$F_{tr} = 1.61 + \frac{4 - 1}{0.63} = 6.37$$

$$NF_{tr} = 8.04 \text{ dB}$$

antenna noise

Of all the contributions to system noise, antenna noise is probably the least understood. Assuming the antenna is built of good conducting materials, it contributes virtually no thermal noise of its own to the receiving system. The noise power the antenna does deliver to the receiver depends almost entirely on the temperature and other physical characteristics of the material lying in the antenna's field of view.

Table 2

frequency	noise factor at antenna	noise figure	external available noise power	receiver input signal for 10 μ V dB S+N/N	acceptable noise figure
1.8 MHz	15.8	12.0	- 93 dBm	15.3 μ V	45 dB
3.5 MHz	16.2	12.1	-101 dBm	12.6 μ V	37 dB
7.0 MHz	16.7	12.2	-111 dBm	4.0 μ V	27 dB
14.0 MHz	17.6	12.5	-113dBm	3.1 μ V	24 dB
21.0 MHz	18.3	12.6	-118 dBm	1.8 μ V	20 dB
28.0 MHz	18.9	12.8	-123 dBm	1.0 μ V	15 dB
50.0 MHz	20.9	13.2	-129 dBm	0.5 μ V	9 dB
144.0 MHz	26.9	14.2	-139 dBm	0.2 μ V	2 dB

Table 2 Performance of a receiver with 0.5 μV sensitivity for 10 dB S+N/N with 100 feet (30.5m) of RG 8 A/U transmission line is shown in first two columns. Third column lists external available noise power for quiet receiving locations on each of the amateur bands. Fourth column shows receiver signal (50-ohms) required for 10 dB S+N/N on each of the bands (based on external noise). Last column lists acceptable noise figure for each of the bands (see text). Bandwidth = 2.1 kHz.

A 432-MHz moonbounce receiving antenna looking out into space, for example, may deliver only as much noise power as a resistor at 10°K (noise factor = ".03). If this same antenna is rotated so that the warm earth comes into its field of view, the antenna noise temperature would rise to about 300°K (noise factor = 2).

A ten-meter Oscar receiving antenna which is pointed into "cold" space, on the other hand, will see a noise temperature of about $10,000^\circ\text{K}$ minimum (noise factor = 35). Pointed at the horizon, however, the antenna noise temperature may be ten or fifteen times higher, depending upon the amount of man-made noise.

There is little that can be done to improve the situation of a terrestrial radio circuit, but an antenna that looks at the sky, such as a satellite antenna, deserves careful design. This is because the effect of the earth is still present, and any sidelobe that sees the earth will pick up thermal noise. This is sometimes quite serious and sidelobes are of major concern in many deep-space communications and radio astronomy systems. Careful attention to antenna design with respect to sidelobes can provide antenna temperatures significantly under 50°K , while poor design can result in much higher values.

minimum usable sensitivity

With an understanding of receiver noise factor and its relationship to signal-to-noise ratio, it's now possible to determine the minimum usable sensitivity (MUS) of a receiving system, and how the performance of your own equipment affects your ability to receive weak signals. Let's first consider that modern communications receiver mentioned earlier which had a specified sensitivity of 0.5 μV for 10 dB S+N/N ratio.*

From table 1 a 0.5 μV sensitivity for 10 dB S+N/N is equivalent to a sensitivity of -119 dBm in a 50-ohm system (-129 dBm noise floor). Assuming a bandwidth of 2.1 kHz, the noise figure of the receiver is about 11.8 dB (noise factor = 15.1). Assuming 100 feet (30.5m) of RG-8A/U transmission line at 300°K , the noise factor at the antenna terminals may be calculated from **eq. 15**, and is shown in **table 2** for the six high-frequency amateur bands, (calculated on the basis of 0.5 μV sensitivity for 10 dB S+N/N on all bands, which may be optimistic). Even at 28 MHz, where the line loss has increased the noise factor by 25 per cent, the system noise factor is still well below the available noise power seen by the antenna (see **fig. 1**).

At 50 MHz, however, the system is limited by receiver noise and a lower noise figure would be required for weak signal work (system noise at 50 MHz in this example is about 2.4 dB higher greater than external noise for a quiet location). If the receiver was connected directly to the antenna terminals to eliminate transmission line losses the system noise figure would be essentially that of the receiver alone and a 0.5 μV signal would provide the desired 10 dB S+N/N ratio. Although 100 feet (30.5m) of RG-8A/U coaxial cable has only about 1.35 dB loss at 50 MHz, it degrades the noise figure sufficiently that the system is no longer limited by external noise sources. This points up the importance of using low-loss transmission lines (or mounting a receiving preamp at the antenna).

Assuming a quiet, rural location that is limited primarily by galactic noise down to about 18 MHz, and atmospheric noise below 18 MHz, what is the minimum usable receiver sensitivity for terrestrial communications?

As can be seen from **table 2**, rather poor receiver sensitivity is acceptable on 40, 80 and 160 meters because the external noise at these frequencies is very high. This also explains why the simple receivers of the 1920s were relatively successful. The high external noise levels also make it possible to use rather inefficient receiving antennas on the lower frequencies.¹³

It's important to note that a 0.5 μ V signal is of little practical use on 160, 80 or 40 meters because it would be buried in the noise level.

The required sensitivity on 20, 15 and 10 is not difficult to obtain with modern devices, but receivers which are optimized for the lower frequencies may not offer top performance on 10 meters. It should be pointed out that the "acceptable" noise figure in the last column of table 2, is somewhat arbitrary and is based on setting the receiver noise floor about 3 dB below the external noise floor. This is probably adequate 90 per cent of the time, but since noise varies randomly, a statistical analysis indicates there may be times when a lower noise figure may be desirable. However, it is generally agreed that a 10 dB noise figure is more than adequate up to 22 MHz and an 8-dB noise figure may occasionally prove useful on 10 meters. Why design a high frequency receiver for extraordinary sensitivity when its performance is limited by external noise over which you have no control? A very sensitive receiver is more prone to intermodulation and cross-modulation effects, and these may be more important.

At vhf the external noise levels are much lower and low-noise receivers are required for good weak-signal performance. Since it's relatively easy to build low-noise receivers for 50 MHz with modern semiconductors, there's no excuse for being limited by system noise figure on this band.

A receiver with a 5 dB noise figure at 50 MHz, for example, when used with 100 feet (30.5m) of RG-8A/U transmission line, will provide a system noise factor of 4.38 (NF = 6.4 dB) at the antenna terminals. This is well below the external noise.

The 144-MHz example in **table 2** is hopelessly inadequate and represents at least 12 dB degradation over what can be obtained in practice. A receiver with a 1.5 dB noise figure on this band, when used with the 100 feet (30.5m) of RG-8A/U, will provide a system noise factor of 2.54 (NF = 4.1 dB) at the antenna terminals which is still inadequate. A transmission line with 0.7 dB loss would bring receiver noise figure within acceptable limits, but it might be easier and less expensive to install a low noise preamp at the antenna.

As pointed out earlier, the noise temperatures of antennas that are pointed into space for satellite communications (or EME) are much lower than for terrestrial communications where the antenna is pointed at the horizon. This means that the receiver noise figures must be lower for maximum performance. Some parts of the sky are noisier than others due to the presence of noise sources, but the noise figure of the receiver should ideally be low enough that the system is galactic-noise limited. Following are the receiver noise figures to shoot for when designing receivers or converters for satellite communications on vhf.

These figures are based on a 2.1-kHz bandwidth and assume a lossless transmission line. For more accurate calculations at low noise figures, the use of noise temperatures is recommended.

frequency	galactic noise floor	noise figure
28 MHz	-125 dBm	8 dB
50 MHz	-130 dBm	5 dB
144 MHz	-139 dBm	1 dB
220 MHz	-140 dBm	0.7 dB
432 MHz	-141 dBm	0.2 dB

Although the topic of noise figure measurement is beyond the scope of this article, the simplest method of making the measurement is to compare receiver noise to the noise generated by a temperature-limited vacuum diode.

This technique is easily applied in the home workshop and has been discussed many times in the amateur radio magazines.^{14,15,16} Guentzler also described a noise measuring system which used a pilot lamp as a noise source.¹⁷

intermodulation distortion

Amplitude distortion occurs in an amplifier when the magnitude of the output signal is not exactly proportional to the input signal. Although amplifiers can be designed to be nearly perfectly linear over a portion of their operating range, every amplifier has nonlinearity which can cause distortion products or harmonics of the driving waveform. Intermodulation distortion or IMD is a type of amplitude distortion which occurs when a nonlinear amplifier is driven by more than one discrete frequency. Although the discussion here is limited to IMD in receivers, this is the same distortion which is used to define the linearity of ssb linear power amplifiers.

When an rf signal with varying amplitude is passed through a nonlinear device, many new products are generated. The frequency and amplitude of each component can be calculated mathematically since the nonlinear device can be represented by a power series expanded about the zero-signal operating point.¹⁸ Although many products are generated, the ones of primary concern are the second and third. This can be demonstrated with a two-tone signal with outputs at f_1 and f_2 at 14001 and 14003 kHz.

$$f_1 = 14001 \quad 2f_1 = 2802 \quad 3f_1 = 42003 \quad f_2 = 14002 \quad 2f_2 = 28004 \quad 3f_2 = 42006$$

Although each of the harmonics fall well outside the passband of a receiver which is tuned to pass 14001 and 14002 kHz, the harmonics mix together to produce intermodulation products which do fall within the passband.

The third order products consist of:

The fifth-order products consist of:

$$2f_1 - f_2 = 14000 \text{ kHz} \quad 2f_2 - f_1 = 14003 \text{ kHz} \quad 3f_1 - 2f_2 = 13999 \quad 3f_2 - 2f_1 = 14004 \text{ kHz}$$

The output spectrum is shown in **fig. 6**. Unless the nonlinearity of the amplifier is particularly severe, fifth-order IMD is not usually a problem and can be ignored in receiver applications.

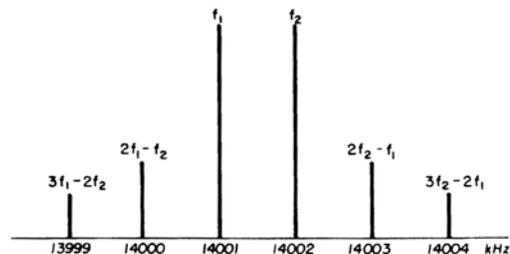


fig. 6. Third- and fifth-order intermodulation products generated by input signals at 14001 and 14002 kHz. In receiver stages fifth-order IMD is usually small enough to be neglected.

Although the IMD distortion products which are generated by two discrete frequencies are used here because they're easy to visualize, exactly the same sort of thing occurs with complex speech waveforms.

In a receiver rf amplifier or mixer stage IMD may be caused by two adjacent CW signals or by a ssb signal. Furthermore, in a mixer where the input must be wideband (such as the double-balanced mixer which is currently finding wide use). Two input signals, f_{R1} and f_{R2} may mix with the local oscillator (f_L) to produce in-band, two-tone, third-order intermodulation products $(2f_{R1} - f_{R2}) \pm f_L$ and $(2f_{R2} - f_{R1}) \pm f_L$. Third-order inter-modulation products also occur at $(2f_{R1} - f_{R2})$ and $(2f_{R2} - f_{R1})$.

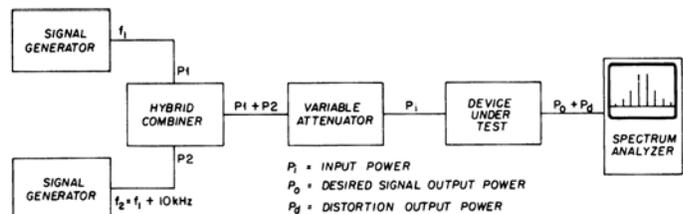


fig. 7. Block diagram of the test setup for evaluating the IMD performance of an amplifier or mixer with a spectrum analyzer.

Two input frequencies at 14210 and 14230 kHz, for example, with a 5.2-MHz local oscillator (9 MHz i-f). will produce two-tone, third-order intermodulation products at 8990 and 9050 kHz (i-f passband) and 14190 and 14250 kHz (rf passband).

Third-order IMD is measured in the laboratory with a spectrum analyzer using the test setup in fig. 7. However, the concept of the third-order intercept is finding increased use to describe the IMD response of mixers, and can also be used to describe the linearity of amplifiers.* The third-order intercept is the theoretical point where the two-tone, third-order response is exactly equal to the two-tone input. Amplifiers and mixers are not operated at this level in practice, but the intercept point offers an internationally recognized figure of merit for comparison of devices, both active and passive.

In addition, the intercept point permits comparison of amplifiers and mixers where the intermodulation specifications are given at different two-tone levels. Once the intercept point is known, you can calculate the two-tone, third-order response at any input level by simply remembering that every 1-dB change in the two-tone input produces a 3-dB change in the third-order output. With this information it is possible to predict the maximum rf input level which is allowable.

*Class A amplifiers. Linear class AB or B amplifiers often exhibit two-tone, third-order intermodulation products which follow an S-shaped curve that both increases and decreases with additional input signal level so they cannot be compared by the intercept point method.

With each 1 dB decrease in the IR input level, for example, the third-order product is decreased an additional 2 dB. As shown in **fig. 8**, a high-level double balanced mixer will suppress third-order products about 65 dB when both signals are at zero dBm (224 mV across 50 ohms) and 85 dB when both input signals are at -10 dBm (71 mV across 50 ohms). The third-order intercept point for these mixers is +27.5 dBm, relative to the output. Relative to the input, the intercept point is at +32.5 dBm. This is 17 dB higher than the intercept point for a low-level double-balanced mixer such as the Minilabs SR1A or Anzac MD108. The 3-dB compression point shown on the graph is a combination of both conversion compression and desensitization.

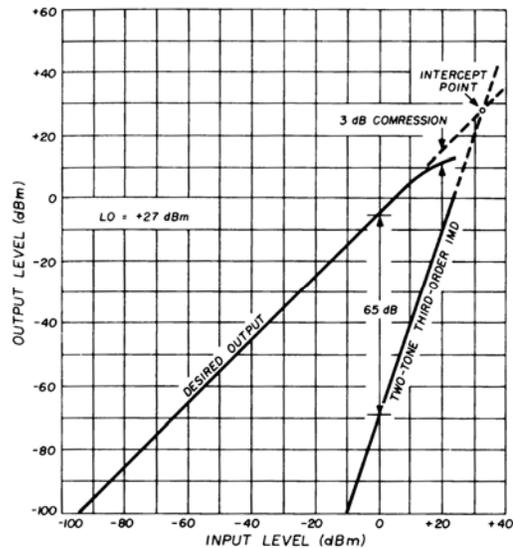


fig 8 Third-order intercept point of high level double balanced mixer (SRA 1H) at 50 MHz is +32.5 dBm relative to the input, 3 dB compression occurs at +20 dBm and third-order IMD is suppressed 65 dB when both input signals are 0 dBm (224 mV across 50 ohms).

intercept point

The third-order intercept point, IP, can be calculated from the relationship

$$IP = 1/2(P_o - P_d) + P_i \tag{16}$$

IP = third-order intercept, dBm

P_o = desired output, dBm

P_d = third-order distortion products, dBm

P_i = input power, dBm

Since the third-order IMD is defined as (P_o - P_d). eq. 16 can be rewritten as

$$IP = 1/2 IMD + P_i \tag{17}$$

For the spectrum display shown in **fig. 9**, for example, the third-order inter-modulation distortion products with two input signals of 4 mV (-35 dBm) are 80 dB down, and the third-order intercept is $IP = 0.5 + 80 - 35 = +5 \text{ dBm}$.

Most amateurs don't have spectrum analyzers, but if good intermodulation distortion information is provided on the receiver data sheet, the intercept point can be easily calculated with eq. 17 (in all too many cases, however, amateur receiver manufacturers ignore IMD completely, and when they do provide IMD data, it is incomplete).

How ever, assume the specifications for an amateur-band receiver list -75 dB IMD for an input of 1 mV (-47 dBm). The third-order intercept is $IP = 0.5 + 75 - 47 = -9.5 \text{ dBm}$

Once the intercept point is known, the IMD performance at any input level can be found by rearranging eq. 17.

$$IMD = 2(IP - P_i) \tag{18}$$

For an intercept point of -9.5 dBm, for example, the IMD at various input levels is shown to the right.

input signal	IMD
100 μV (-67 dBm)	115 dB
500 μV (-53 dBm)	87 dB
1000 μV (-47 dBm)	75 dB
5 mV (-33 dBm)	47 dB
10 mV (-27 dBm)	35 dB

Compare this with the state-of-the-art receiver front end described on page 17 of this which has -74 dB IMD at an input of 100 mV (-7 dBm). The intercept point is at +30 dBm.

input signal	IMD
100 μV (-67 dBm)	194 dB
500 μV (-53 dBm)	166 dB
1000 μV (-47 dBm)	154 dB
5 mV (-33 dBm)	126 dB
10 mV (-27 dBm)	114 dB

The superiority of this receiver is obvious - for all but the strongest signals the IMD products are at or below the receiver noise level. Assuming a 10 dB noise figure and 2.1-kHz bandwidth, an input signal of -23.7 dBm (14.7 mV across 50 ohms) will produce IMD products just equal to the noise level. In the receiver with an intercept point at -9.5 dBm, however, the IMD products are already 3 dB greater than the noise with an input of -49 dBm (800 μV). This will be discussed further under the subject of dynamic range.

double-balanced mixers

Often a mixer data sheet does not specify the third-order intercept point, but a rule-of-thumb estimate can be easily made by examining the 1-dB compression point. As the rf input is increased, the i-f output should follow in a linear manner. However, after a certain point, the i-f output increases at a lower rate until the mixer output becomes fairly constant. The point at which the i-f output deviates from the linear curve by 1 dB is called the 1 dB compression point. At this point the conversion loss is 1 dB greater than it was when the rf input was smaller.

The importance of the 1-dB compression point is its utility in comparing the dynamic range, maximum output and two-tone performance of various double-balanced mixers. As a rule of thumb, the third-order intercept point is approximately 10 to 15 dB higher than the 1-dB compression point¹⁹ (about 15 dB at the low frequencies and 10 dB at higher frequencies). This is shown in **fig. 10**.

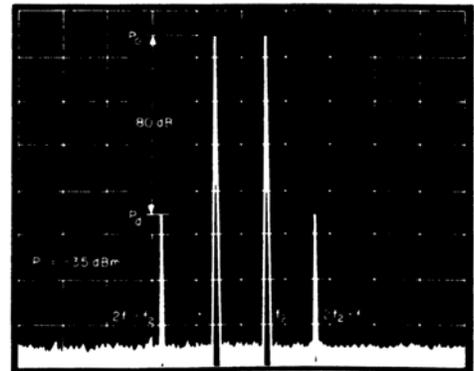


fig 9. Spectrum analyzer display of communications receiver shows IMD is 80 dB down with a two-tone input of -35 dBm (4 mV across 50 ohms). Intercept point is at +5 dBm.

To properly use a double-balanced mixer, it is necessary to relate the two tone input and third-order output levels to avoid generating excessive distortion which would compromise the final design. This is equally valid for amplifiers. Also important, but not as obvious, is the effect higher operating frequencies have on the double-balanced mixer's two-tone, third-order distortion characteristics. Performance is usually better at the lower frequencies and drops off as frequency is increased. For typical high-frequency double-balanced mixers with a maximum frequency specification of 500 MHz, performance starts to fall off somewhere between 50 and 100 MHz.

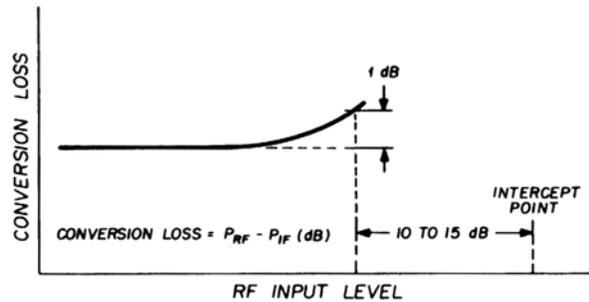


fig. 10. When working with double-balanced mixers, the third-order intercept point may be estimated by using the rule of thumb that the intercept point is 10 to 15 dB above the 1-dB compression point (see text).

It is now possible, using a double balanced mixer in the front end, to build high-frequency communications receivers with a third-order intercept at +30 dBm. Using the rule of thumb that the 1-dB compression point is 15 dB below the intercept point, 1-dB compression occurs at an input of approximately +15 dBm or 1.25 volts across 50 ohms.

By comparison, the 1 dB compression point of many commercial amateur receivers is in the vicinity of -20 dBm (22 mV across 50 ohms) and some solid state receivers with bipolar rf amplifiers go into compression at -40 dBm (2 mV across 50 ohms).

cross modulation

Another type of amplitude distortion which can occur in tuned amplifiers is *cross modulation*. This is related to IMD and is produced when the modulation from an undesired signal is partially transferred to a desired signal in the passband of the receiver. The 3-dB compression point in fig. 8 describes the start of cross-modulation effects.

The cross-modulation effect is independent of the desired signal level and is proportional to the square of the undesired signal amplitude. Because of this relationship, an RF attenuator which lowers the signal level at the input to the receiver may provide a great improvement in cross-modulation performance. A 6 dB attenuator at the receiver input terminals, for example, will reduce cross modulation by 12 dB. If the desired signal is at least 6 dB above the level at which the receiver provides a satisfactory S/N ratio this results in a marked improvement in received signal quality.

Cross modulation is measured in the laboratory by setting one signal generator to deliver a CW output and another generator is set up for 30% amplitude modulation. The output of the a-m generator is increased until 1% modulation appears on the CW signal as measured with a spectrum analyzer. This represents a cross-modulation ratio of about 30 dB (cross-modulation level 30 dB below the reference level).

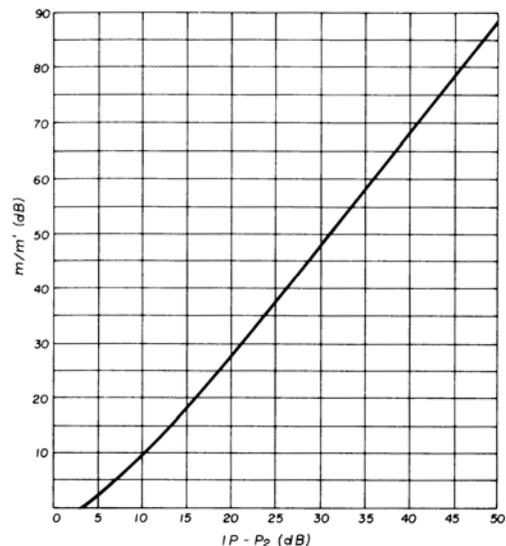


fig 11. Graph of cross modulation vs the ratio of the intercept point to the interfering signal level in dB.

Cross modulation is related to the intercept point by the relationship $m/m^1 = (P_{ip}/4P_c) - 1/2$ (19)

m/m^1 = ratio of cross modulation transferred from a large signal to a smaller one

P_{ip} = intercept point power

P_c = interfering signal power

Cross modulation in dB is simply: $m/m^1 (dB) = 20 \log (m/m^1)$ (20)

In **fig. 11** cross modulation is plotted against the difference between the intercept point and the cross-modulating signal in dBm. Cross modulation of 30 dB, for example, corresponds to 21 dB difference between the intercept point and the signal producing the cross modulation. For a receiver with an intercept point at +30 dBm, a modulated input signal at a level of +9 dBm (630 mV across 50 ohms) will produce 30 dB cross modulation. For a receiver with an intercept point at -9.5 dBm (the more usual case), an interfering signal level of -30.5 dBm (6.7 mV) will produce 30 dB cross modulation.

gain compression

When a receiver is tuned to a weak signal, a strong, adjacent signal may cause an apparent decrease in receiver gain. This is called compression or desensitization and occurs when the input voltage from the undesired signal is large enough to exceed the bias on an RF amplifier or mixer and drives the base (or grid) into conduction. This reduces gain, as shown by the compressed curve of fig. 8, and increases distortion. The rectified base (or grid) current can also be coupled back to the receiver's agc system which results in a further reduction in overall receiver gain.

Compression is measured by setting one signal generator to produce a CW signal and another generator, at a given frequency spacing, is adjusted to depress the desired signal a certain amount, usually 3 dB. Like cross modulation, however, a compression specification has little meaning if the frequency separation between the two signals is not specified.

Since both cross modulation and compression are caused by strong, undesired signals which are adjacent to the receiver passband, they can be controlled to a certain extent by the selectivity at the front end of the receiver. High receiver sensitivity, of course, is the antithesis of good cross-modulation and compression performance - this reinforces the argument for receiver noise figures on the order of 10 dB for the high frequency range.

dynamic range

The front end of a receiver is subjected to a multiplicity of input signals which tend to intermodulate to produce a level of distortion products which is dependent on the magnitude of the incoming signals. Therefore, the upper end of a receiver's dynamic range is defined by the input signal level which produces third-order IM 0 products just equal to the receiver's noise level.

At the lower end dynamic range is limited ultimately by the noise figure of the receiver. Also important, however, is the way the receiver handles weak signals. Some linear RF amplifiers and mixers give good performance in the middle of their operating range but exhibit transfer curves that introduce considerable distortion at low signal levels. With proper design, however, this is not a problem, and the dynamic range of a receiver is usually defined as the spurious-free dynamic range where the maximum input signal is as defined above and the minimum input signal is at the noise floor of the receiver (**eq. 8**).

$$DR = 2/3 (IP - N_o) \quad (21)$$

where DR = spurious-free dynamic range, dB

IP = intercept point, dBm

N_o = receiver noise floor, dBm

The dynamic range of a receiver is important because it allows you to directly compare the strong-signal performance of one receiver against that of another. On today's crowded bands, and the high incidence of kilowatt transmitters and directive antennas, strong signal performance is usually much more important than sensitivity. Although dynamic range can be used as a figure of merit, it's also useful to know the maximum input signal level, $P_{i(max)}$, which will produce third-order IMD products just equal to the noise level. This can be calculated from

$$P_{i(max)} = 1/3 (2IP + N_o) \tag{22}$$

$P_{i(max)}$ = maximum input signal, dBm
 IP = intercept point, dBm
 N_o = receiver noise floor, dBm

For example, assuming a noise figure of 10 dB and 2.1-kHz bandwidth, the spurious-free dynamic range and maximum input signal of a receiver with an intercept point at +30 dBm are

$$DR = 2/3[30 - (-131)] = 107.3 \text{ dB}$$

$$P_{i(max)} = 1/3[2 \times 30] + (-131) \\ = -23.7 \text{ dBm}$$

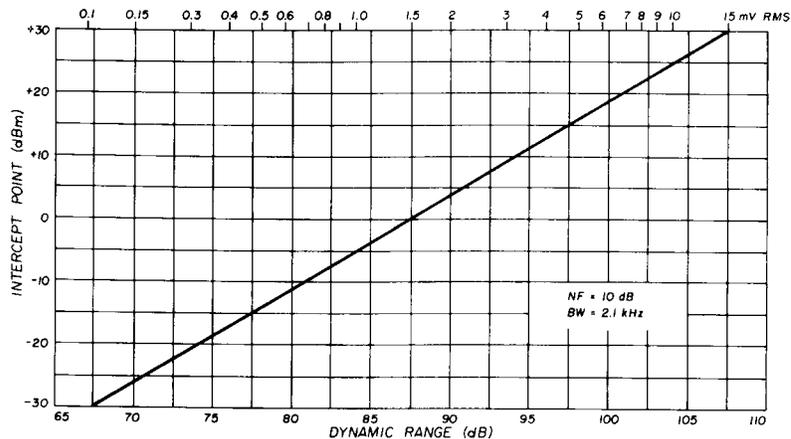


fig. 12. Spurious-free dynamic range vs intercept point (based on 2.1.kHz bandwidth, 10 dB noise figure). Maximum input signal (millivolts rms across 50 ohms) is shown at top.

Thus, the IMD products will be well below the noise level for all input signals below -23.7 dBm (14,000 μ V or about S9+43 dB)! As a comparison, consider a receiver with a third-order intercept at -9.5 dBm (10 dB noise figure, 2.1-kHz bandwidth):

$$DR = 2/3[-9.5 - (-131)] = 81 \text{ dB}$$

$$P_{i(max)} = 1/3[2(-9.5) + (-131)] = -50 \text{ dBm}$$

With this receiver the distortion products are equal to the noise level with an input signal of about 710 μ V or S9+17 dB. This may represent adequate strong-signal performance if you live out in the country, but it's doubtful. If you live in an urban area, you'll have a lot of trouble digging weak signals out of the morass of IMD products which effectively raise the noise floor of the receiver. A graph of intercept point vs dynamic range and maximum input signal is presented in fig. 12 for a receiver with a 10 dB noise figure and 2.1-kHz bandwidth (typical for modern amateur communications receivers).

summary

Although modern amateur receivers are no longer performance limited by noisy vacuum tubes or poor selectivity, the published performance specifications have changed little since the 1940s and are still limited essentially to data on sensitivity and selectivity. Specifications on intermodulation distortion, cross modulation, desensitization and dynamic range, if they're mentioned at all, provide insufficient information for direct buyer comparison. Purchasing a new receiver under these conditions is a bit like buying a new car without knowing gas mileage or how many passengers it will carry.

The performance specifications for the high-frequency receiver shown in table 3 leave no question as to receiver performance and are recommended as a guide for receiver manufacturers to follow in the future. Well informed amateurs should demand nothing less.

table 3. Specifications for a high-performance, high-frequency communications receiver provide a good format for amateur equipment manufacturers to follow. These specifications give complete reference and qualifying data and leave little question as to actual receiver performance.

Frequency range	500 kHz to 30 MHz
Tuning accuracy :	± 500 Hz relative to the frequency of the desired signal
Sensitivity	CW and ssb: $0.5 \mu\text{V}$ for 10 dB SIN ratio in a 2.4-kHz bandwidth (11 dB noise figure)
I-f selectivity	2.1 kHz at -6 dB, 4.2 kHz at -60 dB (2.0 shape factor)
Intermodulation products	Out of band: With two 20 mV signals separated and removed from the desired signal by not less than 25 kHz the third-order IMD products are not less than 90 dB below either of the interfering signals. Intercept point = + 24 dBm. In band: Two In-band signals of 20 mV will produce third order IMD products not greater than -50 dBm
Dynamic range	102 dB
Cross modulation	With a desired signal greater than $100 \mu\text{V}$ In a 2.4-kHz bandwidth, an unwanted signal, 30% modulated, removed not less than 25 kHz, must be greater than 175 mV to produce an output 30 dB below the output produced by the desired signal.
Compression	With a desired signal of $500 \mu\text{V}$, an unwanted signal more than 25 kHz removed must be greater than 300 mV to reduce the output by 3 dB
Spurious response	External signals 25 kHz or more removed from the desired signal must be at least 85 dB above the level of the desired signal to produce an equivalent output Internal spurious signals are not more than 3 dB above the noise level measured in a 2.1-kHz bandwidth
AGC range	An increase in input of 110 dB above $1 \mu\text{V}$ will produce an output change of less than 6 dB

references

1. J. J. Lamb, W1AL, "Short-Wave Receiver Selectivity to Match Present Conditions," QST, August, 1932, page 9.
2. B. Goodman, W1 OX, "What's Wrong with Our Present Receivers?" QST, January, 1957, page 11.
3. U. L. Rohde, "Eight Ways to Better Radio Receiver Design," Electronics, February 20, 1975, page 87.
4. "World Distribution and Characteristics of Atmospheric Radio Noise," CCIR Report 322, International Radio Consultive Committee (CCIR), ITU, Geneva, 1963.
5. C. R. Graf, W5LFM, and M. R. Clinch, K2BYM, "High-Frequency Atmospheric Noise," QST, October, 1971, page 42; February, 1972, page 16.
6. J. B. Johnson, "Thermal Agitation of Electricity in Conductors," Physical Review, July, 1928, page 97.
7. H. Nyquist, "Thermal Agitation of Electric Charge in Conductors," Physical Review, July, 1928, page 110.
8. S. Goldman, Frequency Analysis, Modulation and Noise, McGraw-Hill, New York, 1948, page 394.
9. R. H. Turrin, W2IMU, "Simple Super Selectivity," QST, January, 1967, page 48.
10. G. E. Tralle, "A Guide to Noise Figure," Microwaves, June, 1962, page 46.
11. H. T. Friis, "Noise Figures of Radio Receivers," Proceedings of the IRE, July, 1944, page 419.
12. J. R. Kennedy, K6MIO, "Noise Temperature - The Real Meaning of Noise Figure," ham radio, March, 1969, page 26.
13. R. L. Nelson, K6ZGQ, "Receiving Antennas," ham radio, May, 1970, page 56.
14. L. N. 14Anciaux, WB6NMT, "Accurate Noise-Figure Measurements at VHF ," ham radio, June, 1972, page 36.
15. R. E. Guentzler, W8BBB, "Noise Generators," QST, March, 1972, page 44.
16. J. A. Huie, K2PEY, "A VHF Noise Generator ," QST, February, 1964, page 23.
17. R. E. Guentzler, W8BBB, "The 'Monode' Noise Generator," QST, April, 1967, page 30; "Additional Data on the Monode Noise Generator," QST, Technical Correspondence, August, 1969, page 50.
18. F. E. Terman, Electronic and Radio Engineering, McGraw-Hill New York, 1955, chapter 10.
19. "Get the Most from Mixers," Mini-Circuits Laboratory, 837-843 Utica Avenue, Brooklyn, New York 11203.