ARRL 2009 HOMEBREW CHALLENGE 50-WATT AMPLIFIER ENTRY:

"The 89-cent Special"

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This amplifier was designed and built by the author as an entry in the ARRL's 2009 Homebrew Challenge contest. The requirements of the amplifier are that it be a 40 meter band 50 watt output amplifier intended for use with a QRP transmitter as driver (less than 5W), constructed at a total material cost of not over $125.00, and meet applicable FCC requirements for amateur radio products, tested in accordance with published ARRL Laboratory test procedures. Other requirements also apply. Hopefully this amplifier will be found to meet or exceed all requirements.

This builder's priorities for this amplifier are as follows, in order of importance:

1. True "homebrew" (no pre-made PC boards, kits, etc).
2. Lowest possible cost of materials. (HB labor is recreational, and therefore FREE, so I always trade off materials for labor). Must be a "cost-effective" design.
3. More than adequate performance.
4. Practicality in construction and in use, with no frills.
5. Usable on other bands.
6. T/R switching fast enough for QSK at usual speeds.
7. "Cosmetics" always seem to be last on my list. Beauty is in the eye of the builder, and for this builder, ugly-style is beautiful!

PHYSICAL CONSTRUCTION (FANCY TERM IS MECHANICAL DESIGN):

Bare copper-clad material, and "ugly construction" methods minimize cost and provide excellent RF grounding of circuitry. This amplifier is a nearly ideal candidate for ugly construction. Only a single side of the copper clad material is used for mounting components, thus either double-sided or single-sided copper clad material (less expensive) may be used. The copper-clad material is drilled for three 4-40 mounting screws threaded into matching drilled and tapped holes in the heat sink. Also, the heat sink is drilled and tapped for two 4-40 threaded holes for the mosfet mounting screws. Last, the copper-clad is drilled with four 4-40 screw clearance holes in the corners, used for four 4-40 screws and threaded spacers to act as supports above the table top when in use. Two rectangular cutouts about 1/8 inch larger all around than the TO-220 device packages, are centered on the base material's length and width, and cut out using a drill and a hand nibbler tool and/or a file. An inexpensive but very effective personal computer Pentium processor heat sink/fan is used. This heat sink/fan maximizes cooling effectiveness required by the relatively poor thermal conductivity of the TO-220AB package of the chosen active devices, the International Rectifier IRFZ24N power mosfet. The key to low cost is the use of these very inexpensive devices, which are intended for switch-mode power supplies and other switching applications, not RF service. As hams have historically done, I am using whatever is readily and *inexpensively* available that will meet my needs. A reasonably stocked junkbox can keep the out-of-pocket cost for this amplifier under $10.

CIRCUIT DETAILS (FANCY TERM IS ELECTRICAL DESIGN):

There is really no original design work here. But for any builder it is important to understand what the components do, and how they were chosen. I see all too many homebrew articles which just present a schematic and skimpy circuit description and do not explain what each part does, how its value is determined, and perhaps how critical its value may be and why.

The circuit used in this amplifier is a conventional and often-used push-pull broadband transformer-coupled architecture. No RF negative feedback is used, for simplicity and to provide high gain while still meeting the minimum IMD requirement of -28 dBc (at least 22dB below 12.5W tones). No complex mechanical assemblies are required and most components are mounted by their leads as usual in Ugly Style construction. I used inexpensive solder-down RCA type connectors for input, output, and P.T.T./Keying, and direct wire connections for a 13.8VDC supply.

Operation is class AB for linear applications (SSB, PSK, etc); or the amplifier is easily re-biased for class B for CW use if desired. A bias control potentiometer for each mosfet allows user variation of operating point to enable optimization of operation for a particular mode or for optimizing performance. Limited temperature compensation of bias point is provided by gluing pairs of 1N4148 (costing as low as 2 cents each from at least one supplier) diodes to the plastic cases of the mosfet devices and utilizing their known forward voltage vs temperature characteristic (about -2mV/deg C) to reduce forward bias as mosfet junction temperature increases. The diodes run at a constant current. This is accomplished by making simple current sources using a PNP transistor at Q1 and Q2 with associated biasing resistors. Why use a current source? Because, if driven from a voltage (ie, low resistance) source, the temperature variation of diode forward voltage will be reduced by the divider formed between the diodes' changing forward voltage and the other resistors in the circuit. With a current source, the current through the diodes is maintained over a wide range of voltages and variation in diode forward voltage, thus preserving the *variation* in voltage due to the diodes' changing temperature. (See LTSpice simulation circuit for the current sources later in this article). The absolute value of the bias voltage depends on the magnitude of the current from the current source, and the setting of the bias pot. How does the current source work? Zener diode D1 maintains a fixed voltage of about 4.8V across the base-emitter of Q1 and Q2, plus emitter resistor of Q1 and Q2 (the zener has a tolerance of +/- 5%, so its voltage can vary from 4.75 to 5.25 volts depending on the accuracy of any given diode). The current flowing in the emitter and collector circuits of Q1 and Q2 is determined by this voltage, minus 0.7 volts drop emitter to base, divided by the emitter resistance. Thus, in the prototype circuit the current is (4.8 – 0.7)/1.2K or 3.3 mA. This value is not critical, and will vary from unit to unit depending on zener tolerance and resistors R1 and R2 tolerances. Potentiometers R4 and R5 adjust gate bias on the mosfets over a range from about 2.9 to 4.5V, which for most IRFZ24N devices will cover operation from Class B (no quiescent drain current), up to class AB (up to a few hundred mA of drain current for each mosfet). The author found it unnecessary to make biasing individually adjustable for each mosfet, after trying several pairs of IRF24ZN mosfets in the circuit, but individual control is included here for accommodating those cases where the gate threshold voltage matching is not so close between the two mosfets. .

The gates of the two mosfets are heavily loaded at RF frequencies by the pairs of 20 ohm ½ watt resistors, connected from gate to RF ground via C2 and C5. These values were chosen somewhat arbitrarily (yes, they were available in the junkbox), and reduce the gain in addition to partially swamping out the high input capacitance of the mosfets. The lower the values of these resistors,the higher the required input drive power (and the lower the power gain) of the amplifier. They also act to help stabilize the input impedance, so that the driver sees approximately a 50 ohm load. Transformer T1 steps the 20 ohms of push-pull gate impedance up to about 50 ohms, thus requiring at 50/20 or 2.5:1 impedance step-up from gates to input. Recall that for any transformer the impedance ratio is equal to the turns ratio squared, ie, Z1/Z2 = (N1/N2)^2. This requires a turns ratio in this case of the square root of 2.5, which is 1.58. The inductance of the secondary or primary must be at least 4 times the working impedance of the circuit. For the secondary, this impedance is 20 ohms. So we need an inductive reactance of at least 4 times 20 or 80 ohms, at the lowest frequency we plan to use, say 1.8 MHz. 80 ohms of inductive reactance at 1.8 MHz works out to an inductance of 7.1 uH (easily determined from a reactance chart; also see calculation method below for T2's primary). Using an Amidon FT-37-43 (readily available at low cost, but also was in my junk box), we find the AL value to be 420 mH/1000 turns. Since turns = 1000 * square root of (inductance in mH/AL), we get 4 turns. For a little margin I chose 5 turns, making the primary 1.58 x 5 or 8 turns, rounded to the nearest turn.

The output transformer T2 is determined similarly, but we need to know what drain-to-drain load resistance we need. This is easily determined from a consideration of the supply voltage and output power that the mosfets can sustain. For a push-pull Class AB or B amplifier, the drain to drain impedance can be calculated from the formula Rdd = 2*(Vsupply)^2/Pout (see Note 1, also see sidebar for derivation). In this case we calculate Rdd to be [2*(13.8)^2]/50 or 7.6 ohms. The voltage swing on the mosfet drains will be about twice the supply voltage, or about 28V. (The rated 55 V maximum drain-source voltage of the IRFZ24N's thus provides adequate margin, *as long as the load is connected to the amplifier*. Some mismatch is tolerable of course, but it is good practice to assure that the amplifier is looking into a relatively low impedance at all times). Since transformer T2 has a single-turn primary, the available ratios are integer ratios only. The closest integer ratio is 1:3 which provides a 1:9 impedance step-up, (or 9:1 impedance step-down the other way, from output to input). With a 50 ohm load resistance, this works out to 50/9 or 5.6 ohms at the input, which is close enough to 7.6 ohms for our purpose. Why is a single turn chosen for T4's primary? We need a primary inductive reactance greater than 4 times the drain-to drain impedance so as not to significantly load the circuit with the transformer's primary inductance, at the lowest frequency of operation. This means no less than 5.6 ohms x 4 or 22 ohms of inductive reactance, at 1.8 MHz. Since XL =2*PI*F*L (or use a reactance chart), we need a primary inductance of at least 22/(2*PI*1.8MHz) or 1.9uH. Using the Amidon BN-43-3312 ferrite two-hole balun core (also available for about $2 but also available salvaged from another project in my junk box), and looking up the AL value in the Amidon data sheet, we find an AL value of 5400 mH/1000 turns. Since turns = 1000 * square root (Lmh/AL), plugging in the inductance of 1.9 uH, which is .0019 mH, gives us 0.6 turns. The closest we can get is a single turn, so that should do, even for 160M.

The method of building T4 is nothing new, using the braid from a short piece of RG-58 coax, and looping the braid through to make the single turn primary (a single turn in a binocular core passes through both holes). The braid does not have to be insulated, because the ferrite core is a non-conductor of electricity for all practical purposes. The braid is kept open and a hole is enlarged through it at the center of the output side to enable threading three turns of insulated wire through the inside of the braid, around the input holes (staying always inside the braid), and with the final turn exiting the braid through the hole. This takes longer to describe than becomes immediately apparent when seeing a sketch, which is shown below. Another way is to use the common method of two brass tubes that fit the core holes to make the primary, shorting the ends at the output side, and connecting the mosfet drains to the two tubes at the input side. Then the secondary is threaded through the brass tubes. Either way is equivalent, but the braid method requires no brass tubing to be cut and a short to be fabricated from PC stock with holes drilled to short the output end. As you can see, there is a simple rationale for everything, as it should be: Keep It Simple.

For those wondering about why there is no RF feedback, as is often used, the use of small value source resistors and shunt RF feedback was investigated. Both can make the amplifier more linear, reducing IMD further, but drive power increases and peak output power becomes insufficient to meet the 50W CW requirement, at least while limited to a 13.8 VDC supply voltage. There is just not enough "head room" to avoid peak rounding (amplifier compression) with a 13.8VDC supply. The amplifier simply saturates at a lower power level when using feedback, particularly source resistive feedback. (A higher supply voltage may enable use of feedback while still obtaining sufficient peak power output). The peak RF output limitation with a 13.8V supply is probably due to the ON-state resistance of the mosfets. At peak RF drain currents of say 10 amperes or more, the voltage drop across the Rdson resistance (rated at .07 ohms at 25 deg C) of the mosfet can become a significant portion of the 13.8V supply voltage, subtracting from it and resulting in insufficient remaining drain voltage swing to provide the desired 50W of peak output. Mosfets designed for RF power amplifier service tend to have higher ON resistances (and much lower capacitances) and they almost always require a higher supply voltage for proper operation. This is why 14V powered RF power amplifiers at near the 100W level tend to use bipolar transistors instead of mosfets.

Push-pull operation tends to cancel even harmonics if well balanced, but odd harmonics are still significant. (This is a price we pay for untuned, broad-band amplifier circuits like this). The usual 40-

meter “half wave” low-pass filter is included here for further harmonic suppression; other bands will require different low-pass filter component values. These are easily determined and available components (the reactance of each L and C is chosen to be equal in to the load impedance of 50 ohms), see table below for list of components for filters on other bands. If you have a selective antenna system and/or antenna tuner, if you calculate how much harmonic suppression you have with the total radiating system, you may find that on some bands you will have sufficient harmonic filtering due to an appropriate antenna tuner/antenna combination’s selectivity, to eliminate or greatly simplify the low pass filtering of harmonics. This is an often-overlooked capability. I have a two-element yagi on 40 meters for example, which is a very high Q (narrow band) radiator, offering significant attenuation outside of the band and poor radiating (and receiving) efficiency at harmonic frequencies. Running an antenna simulation using Eznec or other programs, can aid in determining the selectivity of your antenna system. We are only interested in total *transmitted* harmonic suppression, so you can take the whole transmitter/tuner/antenna system selectivity into account.

Simple relay T/R switching is provided. On “make” (key-down), the small Omron relays chosen close after a less than 7 millisecond (mS) delay after key-down (“make”), and, with a 22uF capacitor across the coils to slow the current decay in the coils, open again after 20 mS of delay following key-up (“break”). This enables QSK at up to about 30 WPM with no hot switching in the amplifier if the exciter does not have RF build-up before about 7 mS from key-down, and if the RF output does not last more than 20 mS after key-up. This can usually be arranged with the exciter (especially if the exciter is also homebrew!). Use of a lower-rated coil voltage relay, operating at 13.8VDC, can speed up the “make” time in some relays. This is no problem if key-down time is limited. (Do not be afraid to “mis-use” a component in a known and intentional way; we are not building an amplifier for mass production here; we just need reliable operation for this one to work. Again, think of *simple* ways to solve a problem and/or use an existing junk-box part to advantage). The higher voltage (and hence, coil current) on “make” can speed up the armature movement to the closed condition. The T/R relay switching was also modeled using LTSpice, after measuring the relay resistance and inductance. It turns out the inductance is insignificant and the relay opening (“break”) time stretch is determined by the resistance and the coil drop-out current, in conjunction with the shunt capacitor value. (The capacitor value is easily determined experimentally with an oscilloscope without any coil measurements or simulating of the circuit, but I have become attached to using LTSpice and find any excuse to model something :o)). With the values shown, the maximum QSK CW speed is limited to about 30 WPM. QRQ ops will need shorter switching times. (This can also be done very inexpensively with a homebrew PIN diode T/R switch, eliminating all relays, which the author built for his QRO QSK rig, using inexpensive 5-cent 1N4007 rectifier diodes, but that is outside the scope of this article).

Operation with key-down for over 20 minutes at a time has been demonstrated at the 50W continuous output power level on CW or 2-tone output at 12.5W per tone on 40 meters, at stabilized heat sink temperatures (see thermal test results graph below). The inexpensive CPU cooler does an excellent job of dissipating the heat from the two TO-220 devices, but they must be mounted with silicon grease and/or good quality insulators. Mica is ideal. The inherently rather high (about 1.8 deg C per watt) thermal resistance from junction to case of the TO-220AB package also limits the maximum safe power dissipation even with a very good heat sink

BAND	INDUCTANCE	CONSTRUCTION	C9, C12	C10+C11
160M	4.2uH	18T ON T106-2	1500pF	3000 pF
80M	2.5uH	14T on T106-2	820pF	1500pF
30M	1.2uH	15T on T-50-2	330pF	560pF
20-17M	0.58uH	12T on T-50-6	180pF	330pF
15-10M	0.33Uh	9T on T-50-6	100pF	180pF

Table 1. Output low-pass filter values for other HF bands. Wire sizes are not given, as they are non-critical. It is best to use the largest wire size (or equivalent-area multifilar smaller gauge wires) for which the required number of turns will fit as a single layer on the toroidal core. The two largest cores, used on 80 and 160 meters, are overkill size-wise; they were available, naturally, in the author's junk box so were used. T-80 cores are a better fit, and turns will more fully fill the core. As with winding any toroidal inductors, the required turns count is calculated according to the published AL values for the cores chosen, and the total turns count is how many turns pass through the central hole. The first capacitor, C9, can be adjusted (ie, "tweaked") in value somewhat, with minimal effect on low pass performance, but it can improve the higher-order IMD products by reflecting a varying reactive load back into the primary of T2, helping to cancel a little of the higher-order IMD energy. This is why C9 is shown as 680pF (instead of the calculated 470pF) on the schematic. However the amplifier should still meet the minimum -28dBc IMD requirement even with the calculated value.

TURN-ON AND TESTING:

There is not much to do before turn-on except verify all connections and parts values; look carefully, in good light. Measure the resistance of the 13.8VDC line to ground to check for shorts, set bias potentiometers to their minimum (counterclockwise, if wired similar to the original, but check this beforehand!); the bias voltage should read less than 3V on the two gates. It is useful to be able to vary the 13.8V supply up from zero slowly, watching the current. That way any problems can be corrected with no damage to the circuit. Short the PTT line to ground to energize the relays. At 13.8VDC, with the PTT grounded, the fan and relays draw about 200mA total, so this will be the minimum current drawn before adjusting the bias pots. With 13.8V supply, adjust the bias potentiometers R4 and R5 one at a time carefully such that 13.8V current increases by about 100 mA for each potentiometer. Total idling supply current will then be around 400 mA.

The amplifier is now ready for use for linear operation. For CW-only, you can set the bias at about 2.95V or just at the point where you start to draw mosfet drain current (Class B). Connect a load reasonably close to 50 ohms, preferably a 50W or greater dummy load through a directional wattmeter to start. Connect a driver and apply drive power carefully, not to exceed 5W and preferably much less (and preferably adjustable; a simple homebrew attenuator is useful for this), observing output on whatever directional wattmeter you normally use. On 40 meters it should not require over 2W of RF power for 50W output if biased for AB1 (100mA per mosfet), or 3W for Class B (0 mA). On CW you can drive the amplifier to about 63W output with 5W of drive in Class B. Higher drive powers will increase the dissipation in the gate loading resistors, so caution should be exercised to not hold drive power above about 3 watts for extended periods of time due to heating of these resistors, but they are quite forgiving.

PERFORMANCE:

This amplifier was tested using a regulated 13.8V DC bench supply capable of up to 10 amperes, an oscilloscope, the usual DVOM, and, for documenting the IMD, a 35 year old HP141T / HP8553B spectrum analyzer. However, adjustment for optimum performance can be done without recourse to a spectrum analyzer with a little practice "reading" the two-tone RF output envelope on the oscilloscope. The amplifier is usable on all HF bands 160M through 10M, but optimum performance (and full output (50W) is reached only on the 80M through 20M bands, with IMD of -30 dBc or better. On 40 meters, IMD approaching -36 dBc was measured. Maximum output on 80 through 20 meters meets or exceeds 50 W with up to 5 W of drive. This makes this amplifier a very useful addition to any QRP-only station when more power (QRO) is required for the QSO and propagation path at hand. In normal use there should be no problems with overheating. A key-down test was made with the CPU cooler's fan turned off, and mosfet temperatures rapidly rose to 100 deg C in less than 3 minute's time. So keep that fan running! Detailed performance figures are illustrated in the graphs below, based on the test data taken on the prototype. Output of 50 watts or better is available from 80 through 20 meters, with reduced output on 160 meters (about 25W), 15 meters (about 40W) and 10 meters (only about 12 watts). The relatively high capacitances and consequent long switching times of these inexpensive mosfets limit the output on the highest HF bands. Even more output could be obtained on several bands by using higher supply voltages (respecting the 55 V maximum breakdown voltage of the device) and optimizing components for another band. There is much opportunity for further experimentation with this design; no doubt the output can be improved on 160 meters, for example.

Total material cost should be under $30 if all parts are purchased new (not including shipping costs and taxes), and under $10 for a reasonably-stocked junkbox containing the common components such as R's and C's and scrap PC board material, etc; likely available to anyone who has done any modern homebrewing. The two most important components, the mosfets, are some of the least expensive items required. This is one thing that makes homebrewing today so great! So build, and enjoy. Then, build some more!

ACKNOWLEDGEMENT:

Thanks to my old friend and fellow homebrewer Bob Friess, N6CM, for alerting me to the Pentium heat sink/coolers available at low cost online, and for persuading me to enter the contest with my amplifier. The heat sink is the heart of this amplifier in that one needs plenty of cooling capability to enable use of the inexpensive TO-220 packaged devices at these power levels, especially with nearly unlimited key-down times. And, yes, the mosfets are beryllium-oxide free, as with all plastic-cased low-cost devices. Also, if it matters, they are available lead-free if you purchase the IRFZ24NPB part number.

Respectfully submitted,

D. W. Huff, W6JL 10 February, 2010

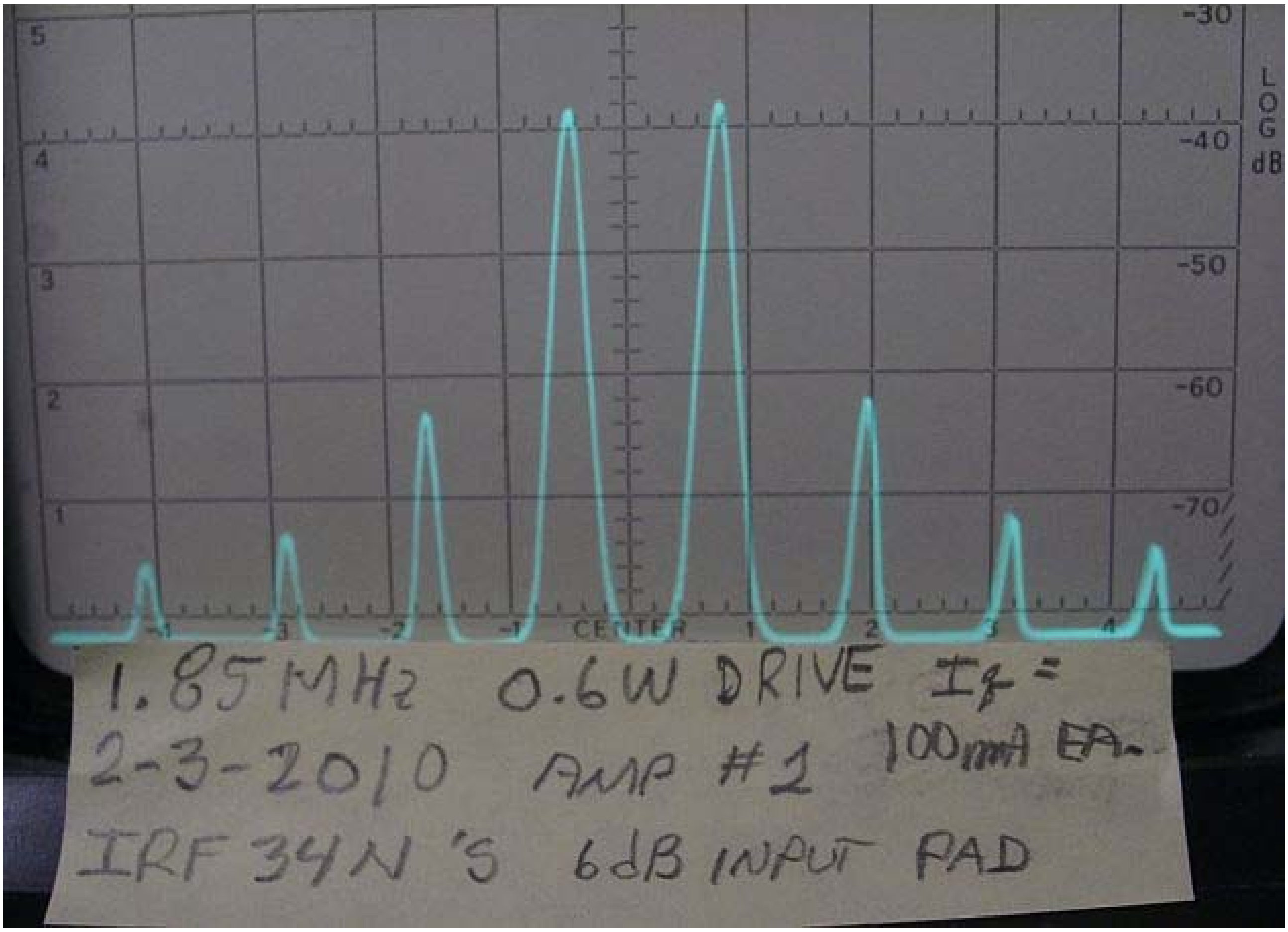
Notes:

1. See 2010 ARRL Radio Handbook, page 17.35. Also see sidebar for derivation.

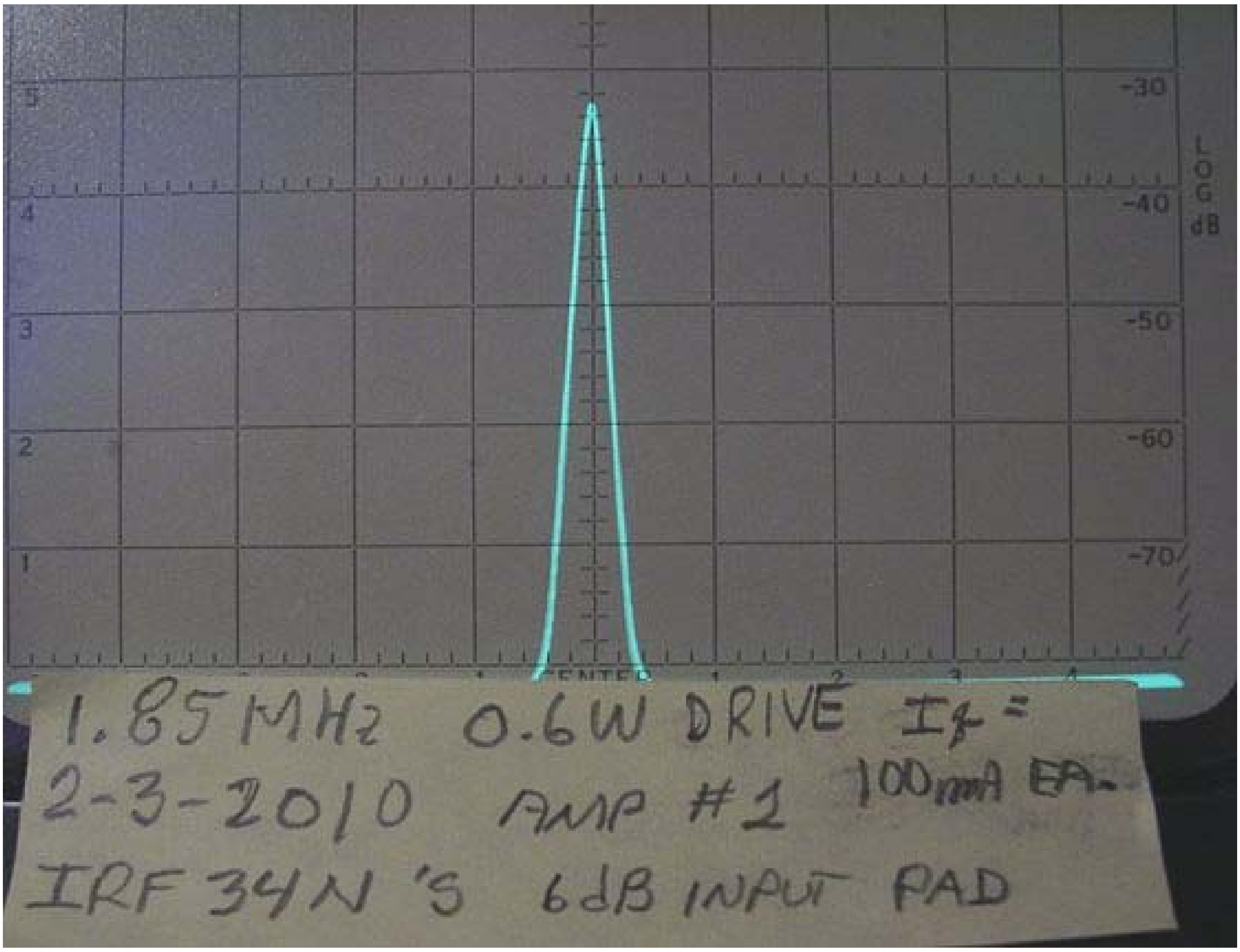
APPENDIX 1: Using other mosfets: the IRFZ34N or IRFZ34NPB

The author did some limited experimenting with a similar mosfet, the IRFZ34N. This is priced in the same range as the IRFZ24N, but is rated at 34 amps max, compared with 24 amps for the IRFZ24N. It has a lower ON resistance, but we don't get something for nothing, so it has about twice the capacitance of the IRFZ24N. This limits it further in high frequency performance but I was curious to see how it might work on the lower three HF bands, 40 through 160 meters. It does indeed appear to work somewhat better on these bands The data below shows a capability of at least 50W output on 40, 80 and 160 meters, with low drive power of 1W or less (about 0.5W ON 160m gives 50W of output when biased at 100 mA per device). This is a gain of 100 , or 20 dB, not bad for a device costing less than one dollar each. The IMD also met or exceeded the required -28 dBc on all these bands, at 12.5 watts per tone. So here is one way to improve performance on the lower HF bands.

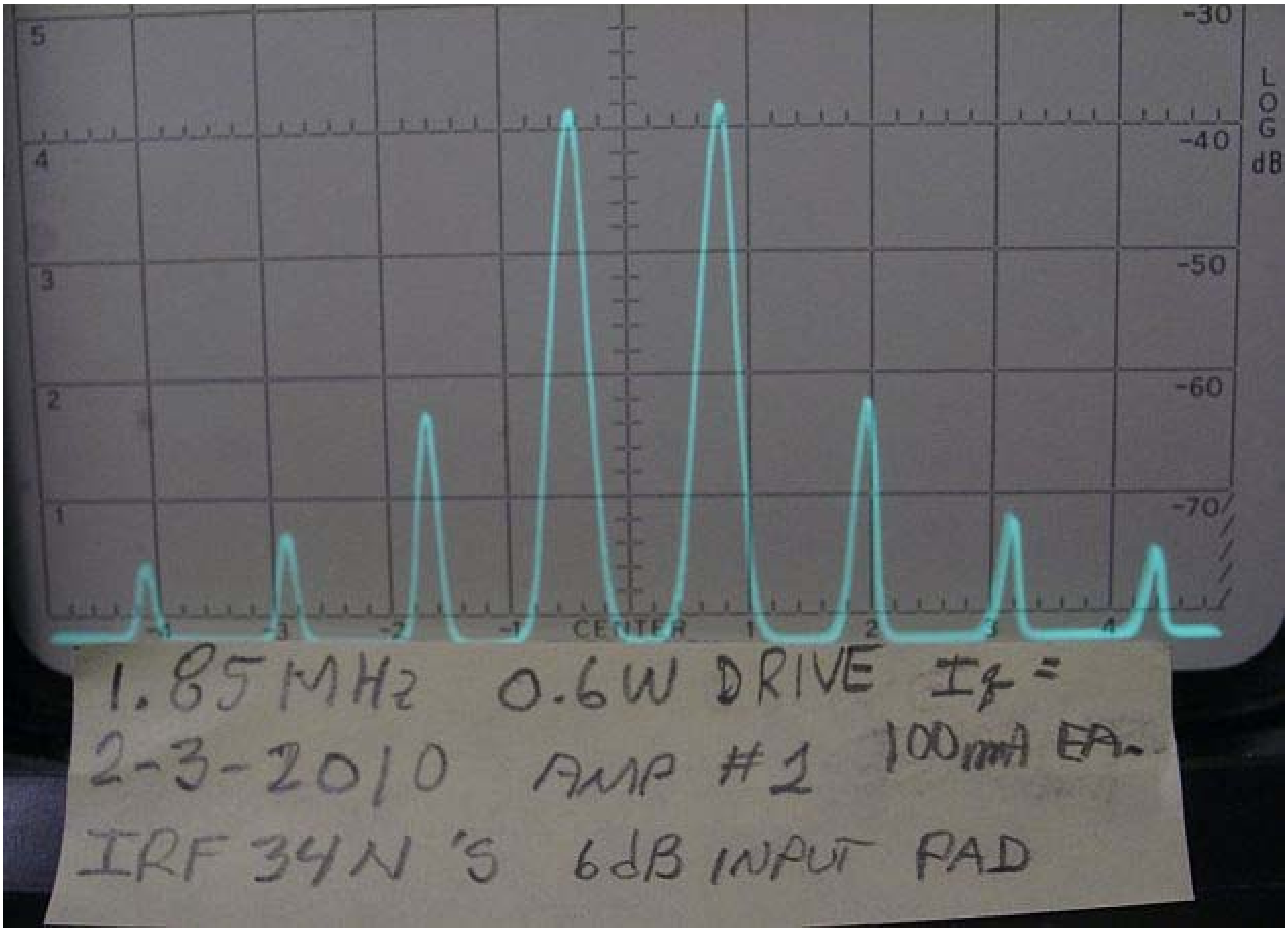
MEASUREMENT RESULTS: PHOTOS/GRAPHS



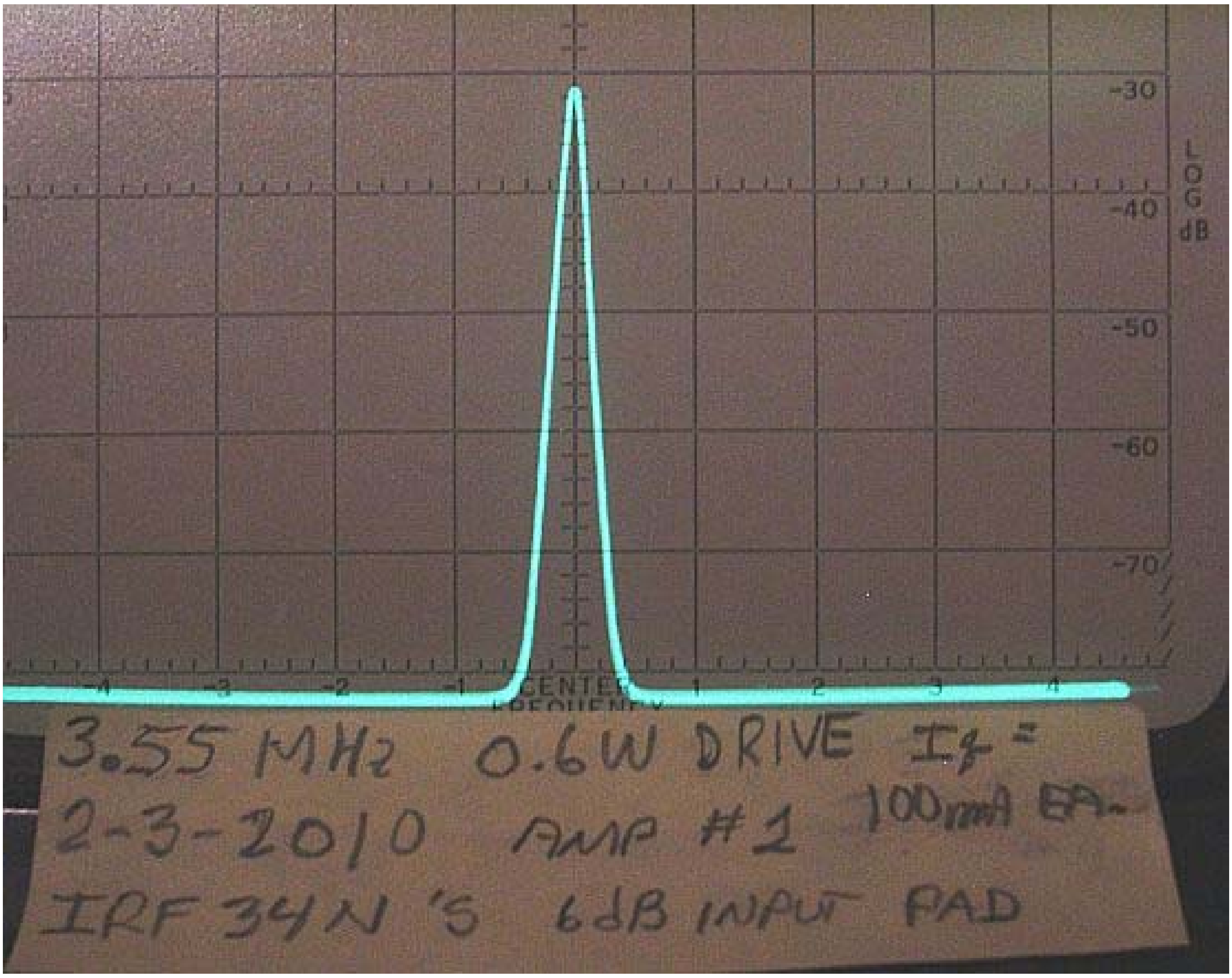
160M: About -31 dBc IMD at 12.5W per tone with less than 1W of drive.



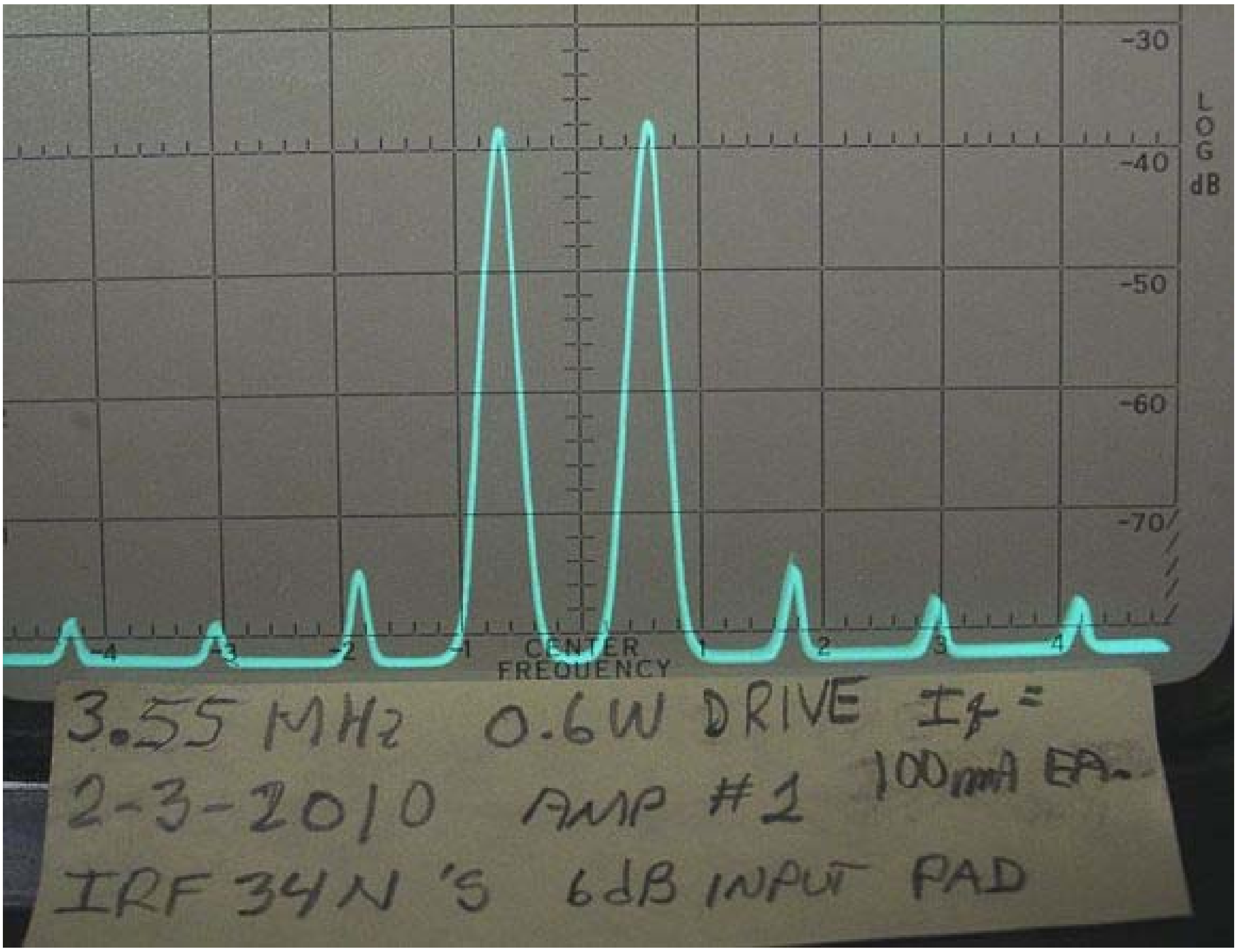
160M: 50W CW output with less than 1W of drive.



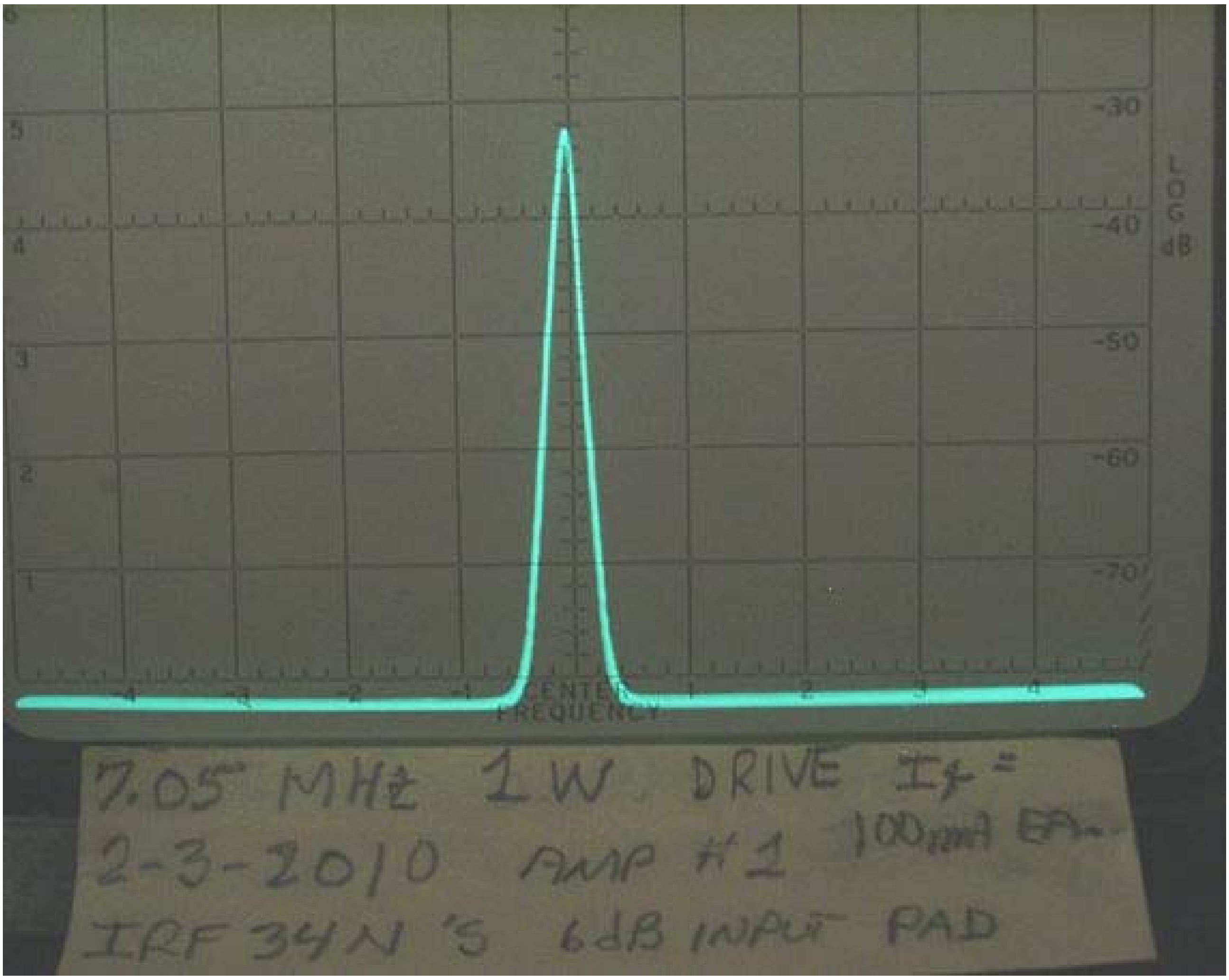
160M: About -31 dBc IMD at 12.5W per tone.



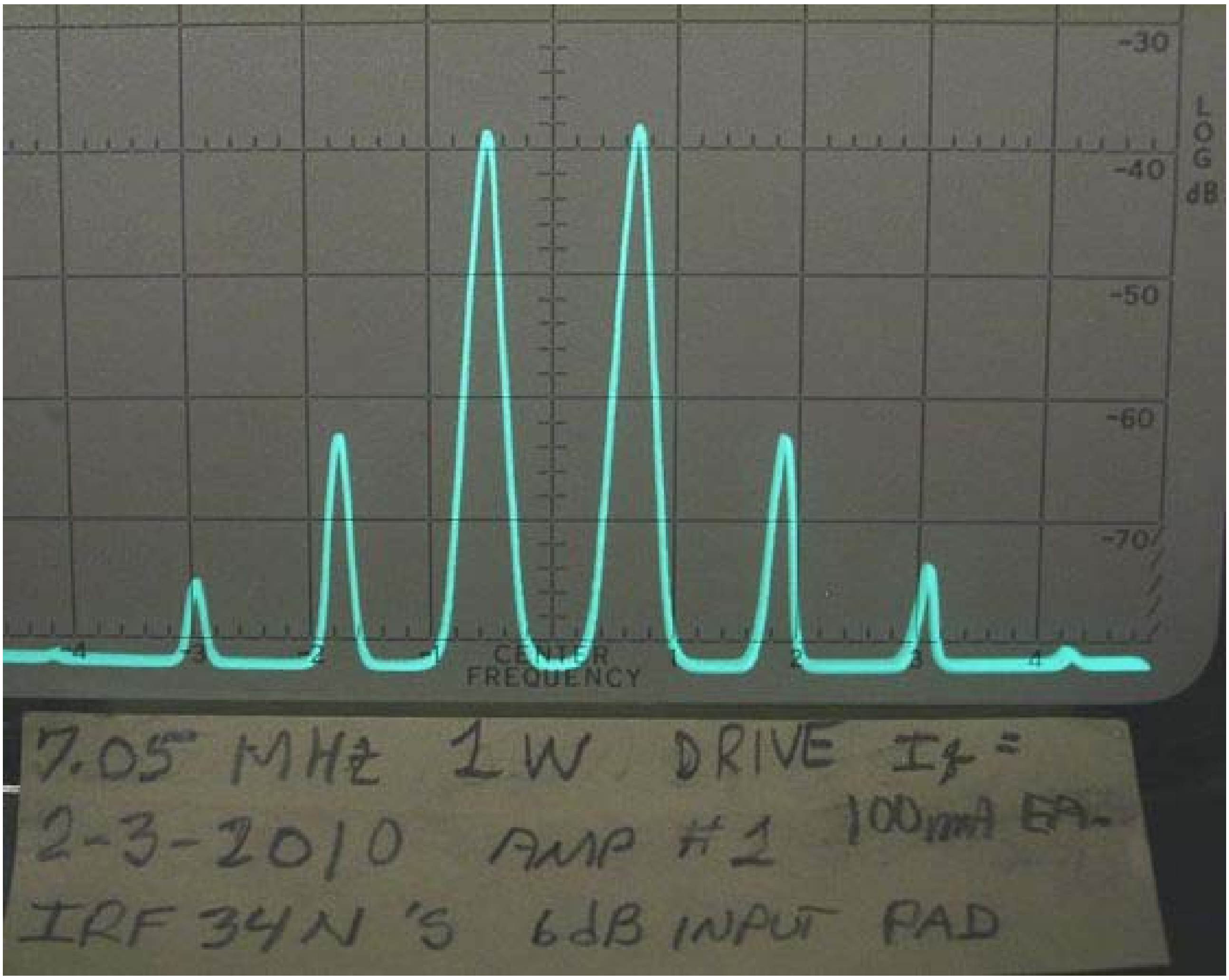
80M: About 63W CW output with less than 1W drive.



80M: Impressive IMD of -42 dBc with less than 1W drive. This approaches the IMD of the test two-tone source. IMD varies with temperature and biasing, especially when this clean. With circuit values unchanged from the IRFZ34N design, 80M seems to be the new "sweet spot" for low IMD when using the IRFZ34N. (The reference to a 6 dB input pad refers to a pi-attenuator made of three 2W resistors, to further reduce the output of the K3 2-tone source, which has a lower limit of resolution of a few tenths of a watt, so that a higher power setting could be used on the K3).



40M: 50W CW with about 1W of drive.



40M: IMD about -30 dBc, not as good as the IRFZ24N, but will be somewhat better with higher quiescent current. .

ADDENDUM:

The Homebrewing Experience

This project provides the amateur radio builder with an opportunity to experiment with a useful RF amplifier using very inexpensive modern devices. Building RF circuits on bare copper-clad is a joy. Progress is rapid, and soldering is easy with almost nothing required except the components themselves, soldered together in free-space style. A little forethought in layout will usually result in a circuit which “flows” naturally when reading the schematic and working on the actual circuit. It is my preferred method of constructing even QRO transmitters, and complete receivers.

This amplifier, with appropriate variation of bias point, supply voltage, and incorporating such features as RF negative feedback, one can achieve improved IMD performance and/or higher RF output, possibly even 100W, at a very low cost. It is also conceivable that four parallel devices could be used, at least at the lower end of the HF spectrum, for additional power output. Alternatively, two or more amplifiers can be combined with transformer combiners for even greater power output. There are a large number of possibilities waiting to be explored with such a basic circuit as a starting point. With devices costing under $1 each, experimentation need not be prohibitively expensive when sudden failures in devices occur. RF-rated devices rated at comparable power level cost 20 to 50 times as much per device as the IRFZ24N devices used in this amplifier, which are currently available online for $0.89 each, in single quantities (www.jameco.com.)

Expensive or elaborate test equipment is not required for this kind of project, even though the author has available a 35+ year old surplus HP spectrum analyzer. Assuming an adequately clean RF two-tone source as driver is available, even just an oscilloscope display can get you in the ballpark for IMD. (An Elecraft K3 transceiver became available for use in testing. Its design conveniently includes a clean built-in 2-tone RF generator with variable output power (down to 0.1W) on all HF bands. Before the K3 became available, a homebrew single mosfet Class A driver amplifier and simple homebrew hybrid combiner circuit was used, driven by two low level RF sources separated by a few KHz in frequency. This was adequate for IMD testing). An oscilloscope allowing observation of the two-tone RF output voltage envelope will enable a builder to get IMD performance at least equal to -28 dBc or better, just by carefully observing envelope waveform zero crossing and crest wave shapes. The author found this to be the case, and used his HP141T spectrum analyzer only for making more accurate measurements for documentation purposes. For CW ops like myself, even less equipment is needed, since a high degree of linearity is not a requirement. Incidentally, older analog test equipment, now available at low prices online, is more than adequate for ham radio homebrew projects such as this. Use of an inexpensive digital camera allows easy documentation of test results and waveforms. This author used his 10 year old Casio digital camera, hand held, for all screen shots in this project.

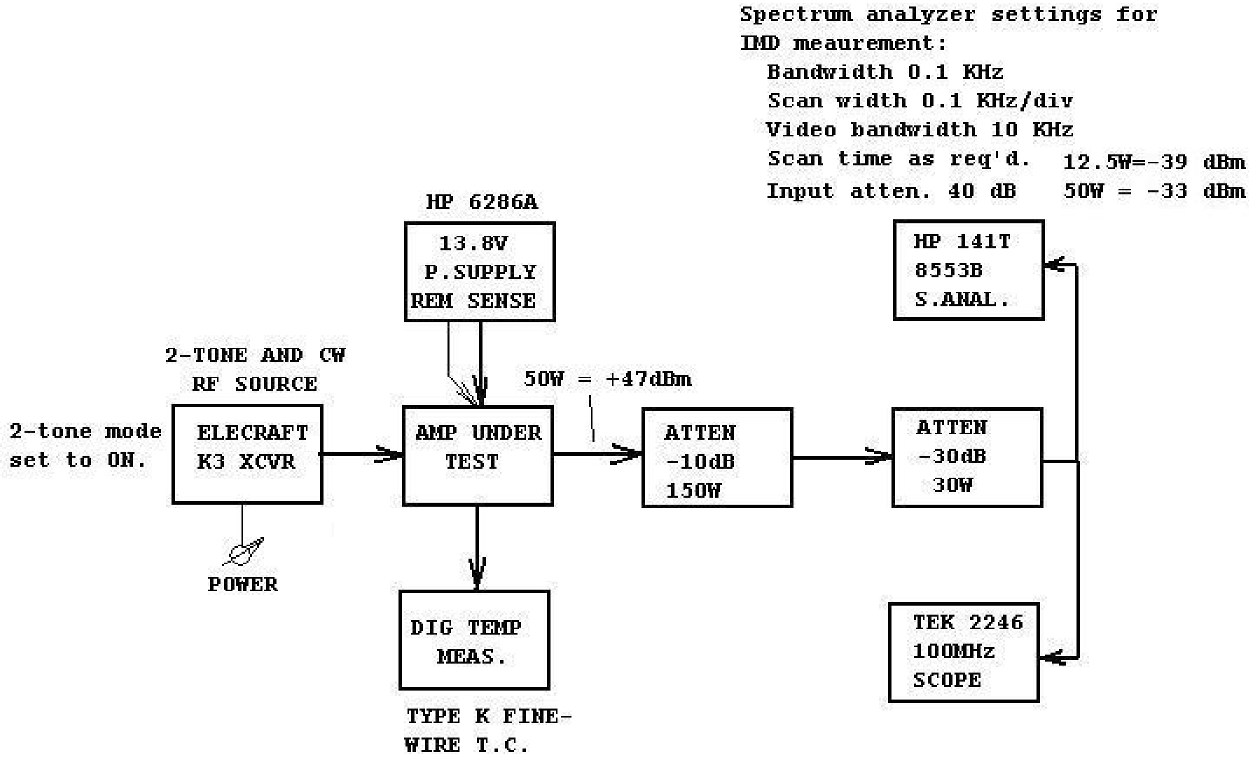
Homebrewing has never been easier, or less expensive, than it is today, for building excellent rigs! The author has operated all-homebrew equipment at QRO power levels for many years (decades), frequently updating as new design ideas, (plus free software circuit simulation tools), and devices become available. Parts are readily available and at low cost online from many sources today. See internet part sources list below, current as of this writing, at the end of this article. Excellent homebrew designs are also available by searching the internet. Many new and creative designs are coming from ham homebrewers in Europe also. So there is a lot to see, learn, (and copy!) if desired. All it takes is the interest and motivation to do so.

Full-featured free software simulation tools using PSPice are now available. One I recommend is Linear Technologies' LTSpiceIV, available for download from their website at: http://www.linear.com/designtools/software/#Spice. This program with built in schematic capture and parts libraries allows almost any builder simulate linear and nonlinear circuits and measure their performance, *before* you build them. And you can add models of other parts not already in the provided library, as well as custom components. This is a really fun program to use, and can be used as a learning tool to gain much insight into how circuits of all kinds work, without lifting a soldering iron.

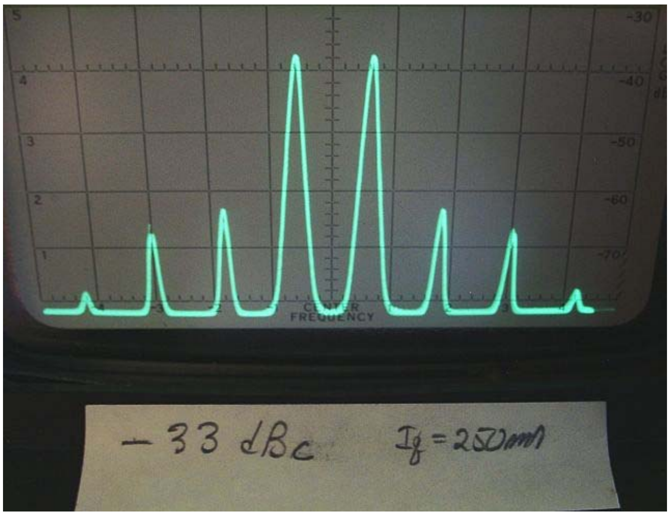
We are living now in a Golden Time for those of us interested in homebrewing. The author's 54 years of homebrewing experience supports this conclusion. This is just the opposite from what many hams I talk with on the air every day seem to believe. Many are convinced that parts are difficult to obtain and are more costly, but those who have shared this opinion with me have never actually tried obtaining modern parts online from the many sources available today. The mentality of many hams I talk with on the air who claim to have once been homebrewers but say they "can't homebrew any more" seems to be that of 30+ years ago, when electronic parts were purchased at local retail stores, which are now long gone. That era has (fortunately, in my view) long passed, with something much better taking its place with the much better situation we have now! So, what other excuses do YOU have for not enjoying homebrewing, if you are *really* interested?

APPENDIX:

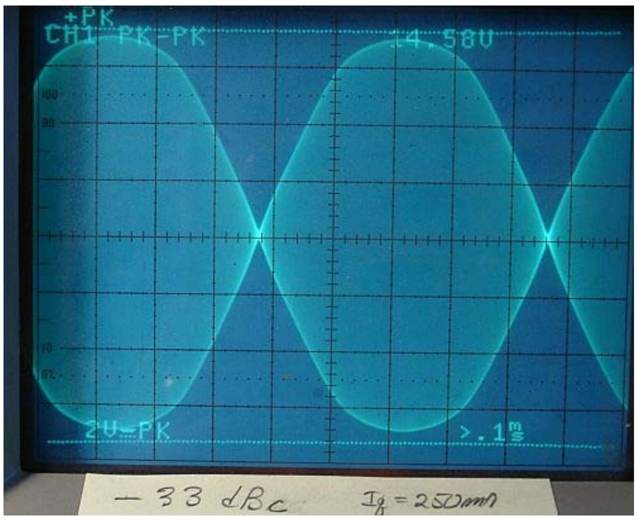
SUPPORTING DOCUMENTATION



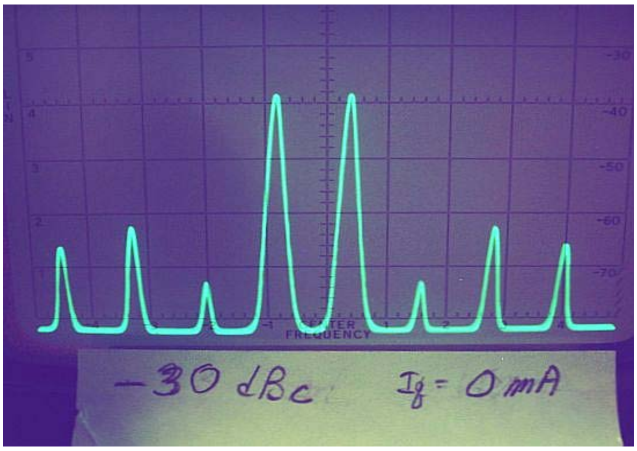
W6JL HOMEBREW CHALLENGE 50W AMP TEST SETUP
JANUARY 9, 2010



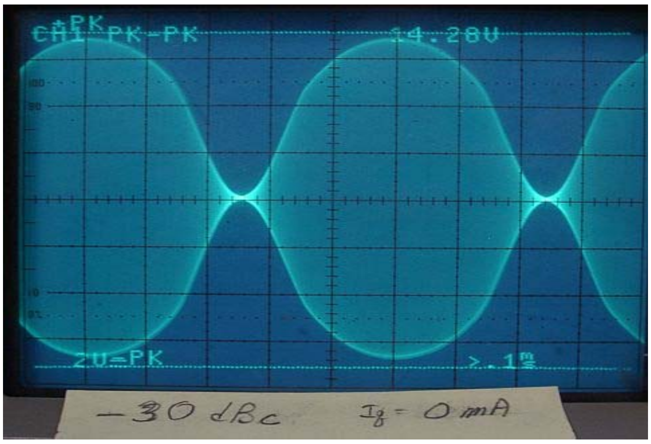
Example of reasonable IMD



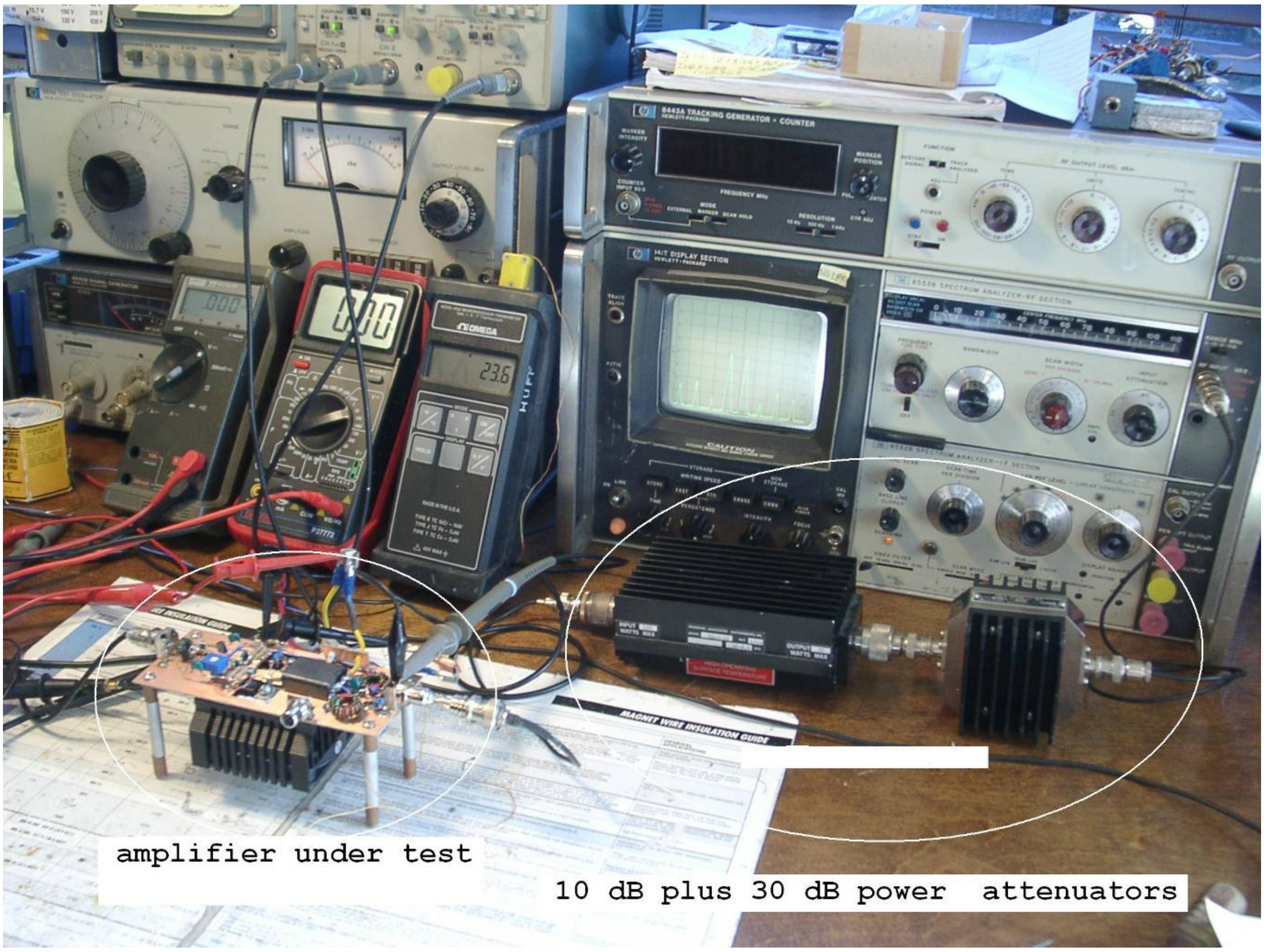
RF envelope of IMD plot at left. Some peak rounding.



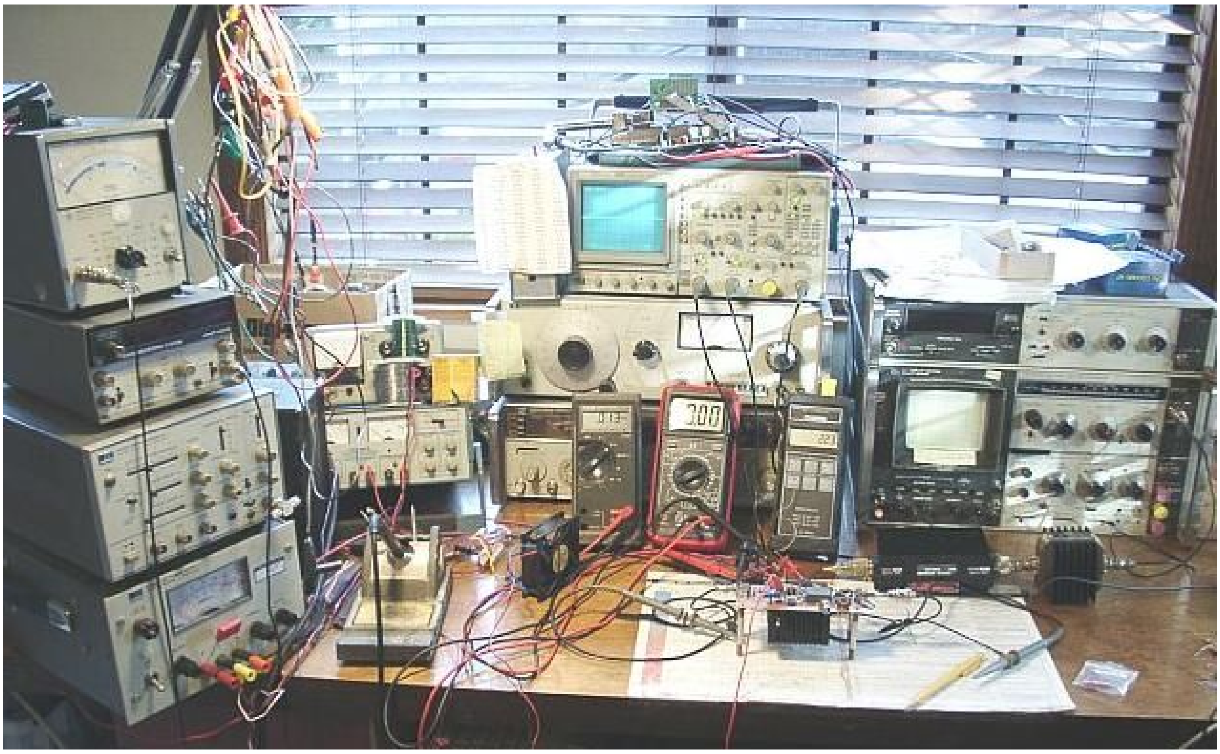
More IMD, higher order.



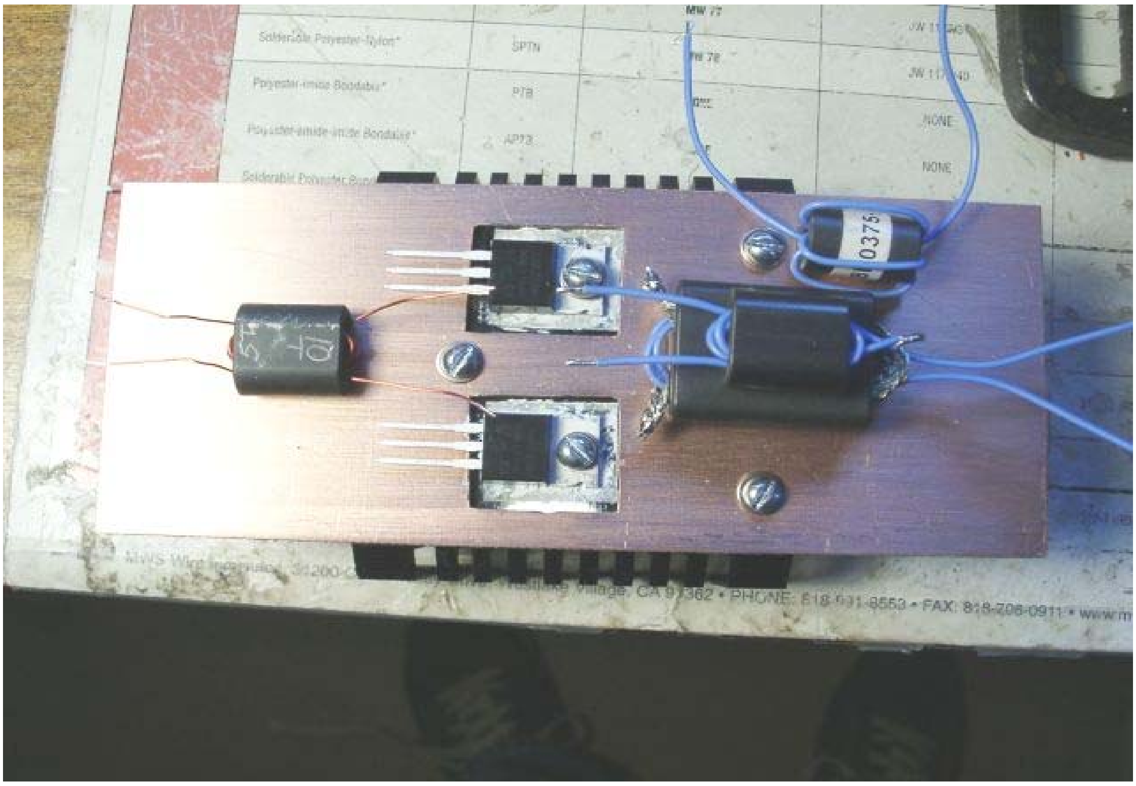
RF envelope of IMD plot at left. Distortion is obvious.



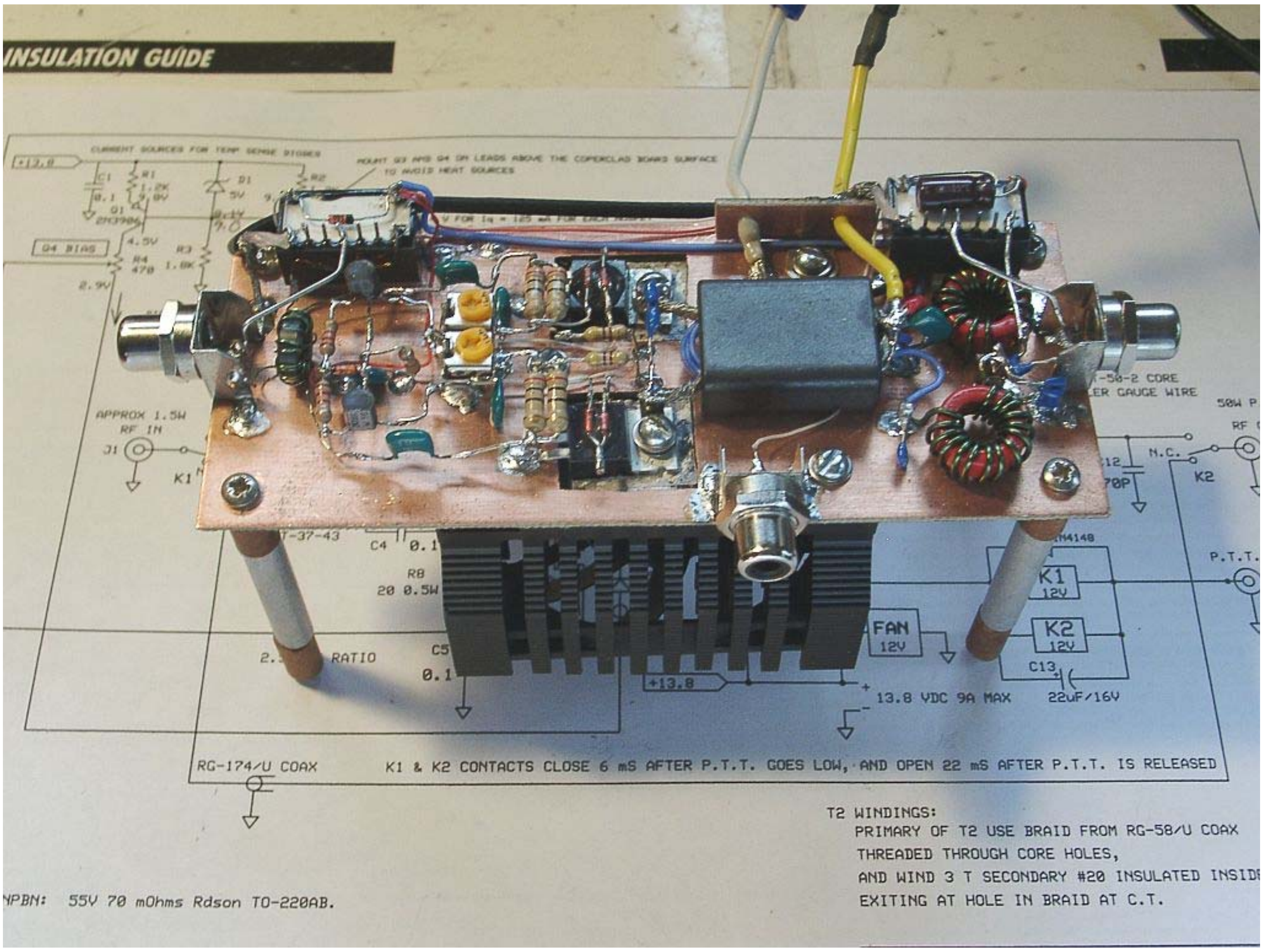
Test Bench showing Amp under test, attenuators, and spectrum analyzer. I found the power attenuators at bargain prices at the 2009 Dayton Hamvention flea market.



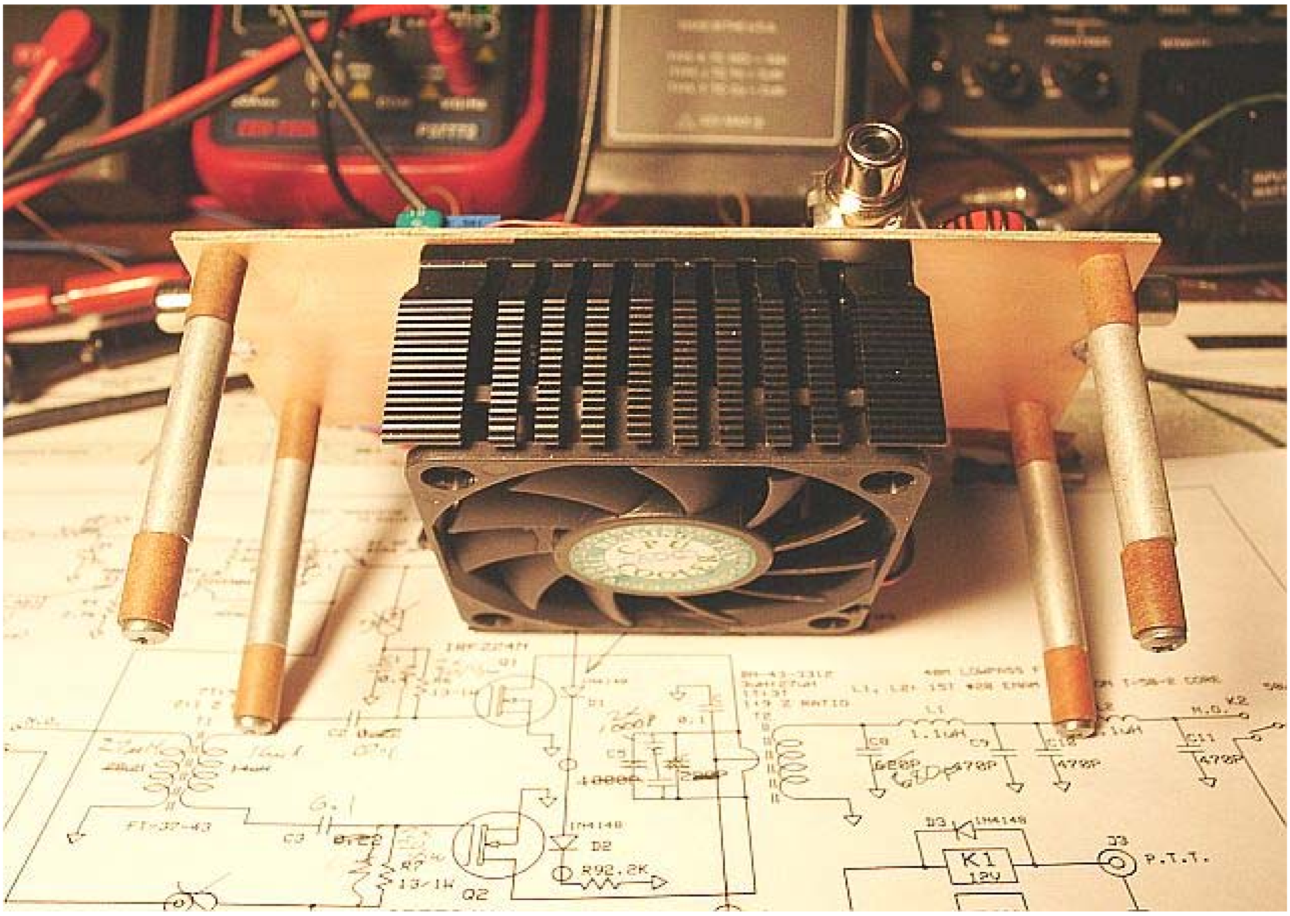
My “Homebrewing Bench”, wide view. This is adjacent to the main operating position. (I use a homebrew wireless paddle and wireless cans while ragchewing and working on the bench. That way, I can get a lot done while ragchewing). How much of this equipment was used for this project? Very little: power supply at lower left, scope at top center, one or more DVOMs, temperature meter, and, for IMD documentation, the older HP spectrum analyzer at lower right.



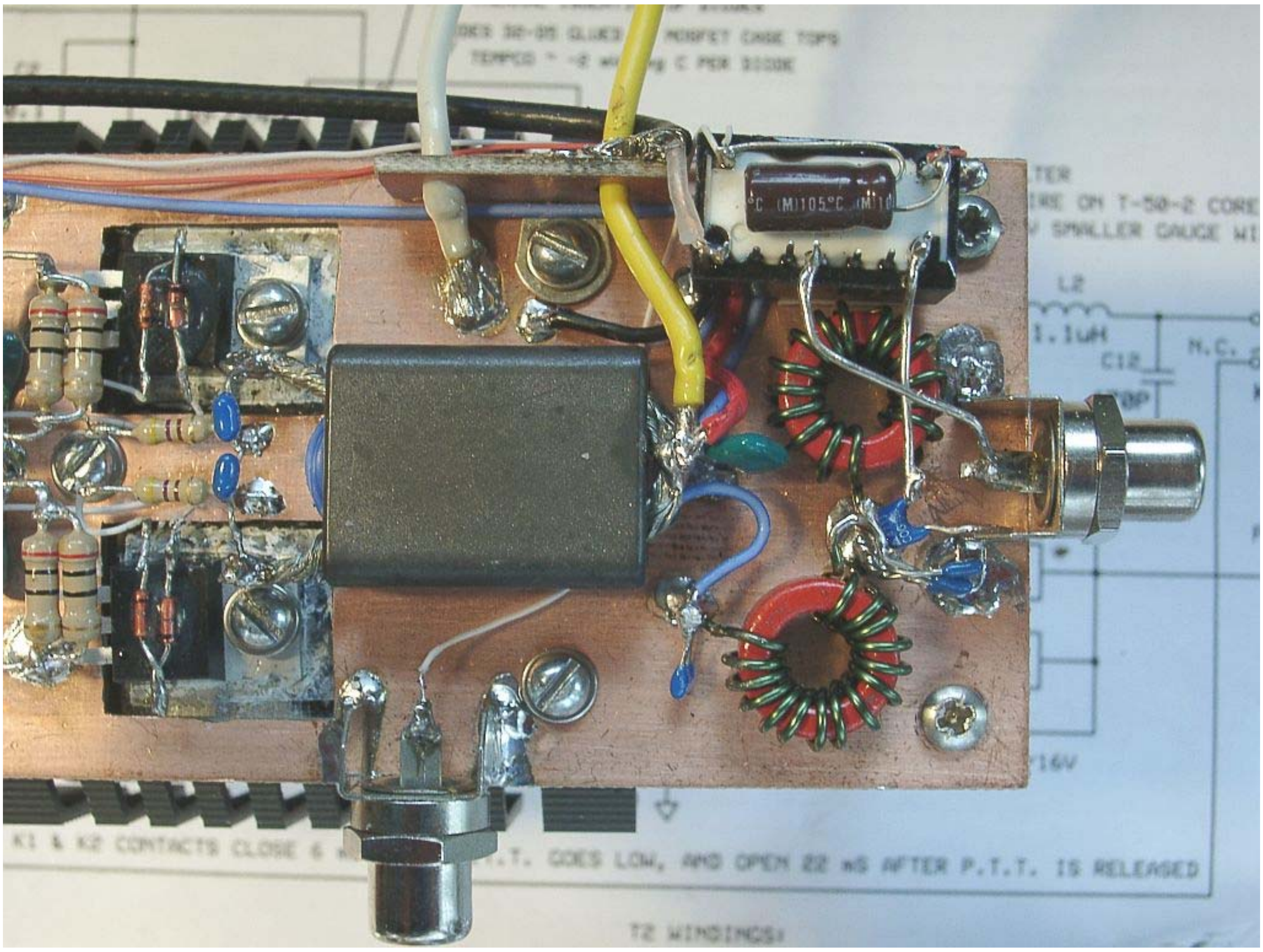
Beginning of amplifier assembly on bare copper-clad. Later the input binocular and DC feed binocular and bead chokes were found unnecessary and eliminated. Keep It Simple applies here.



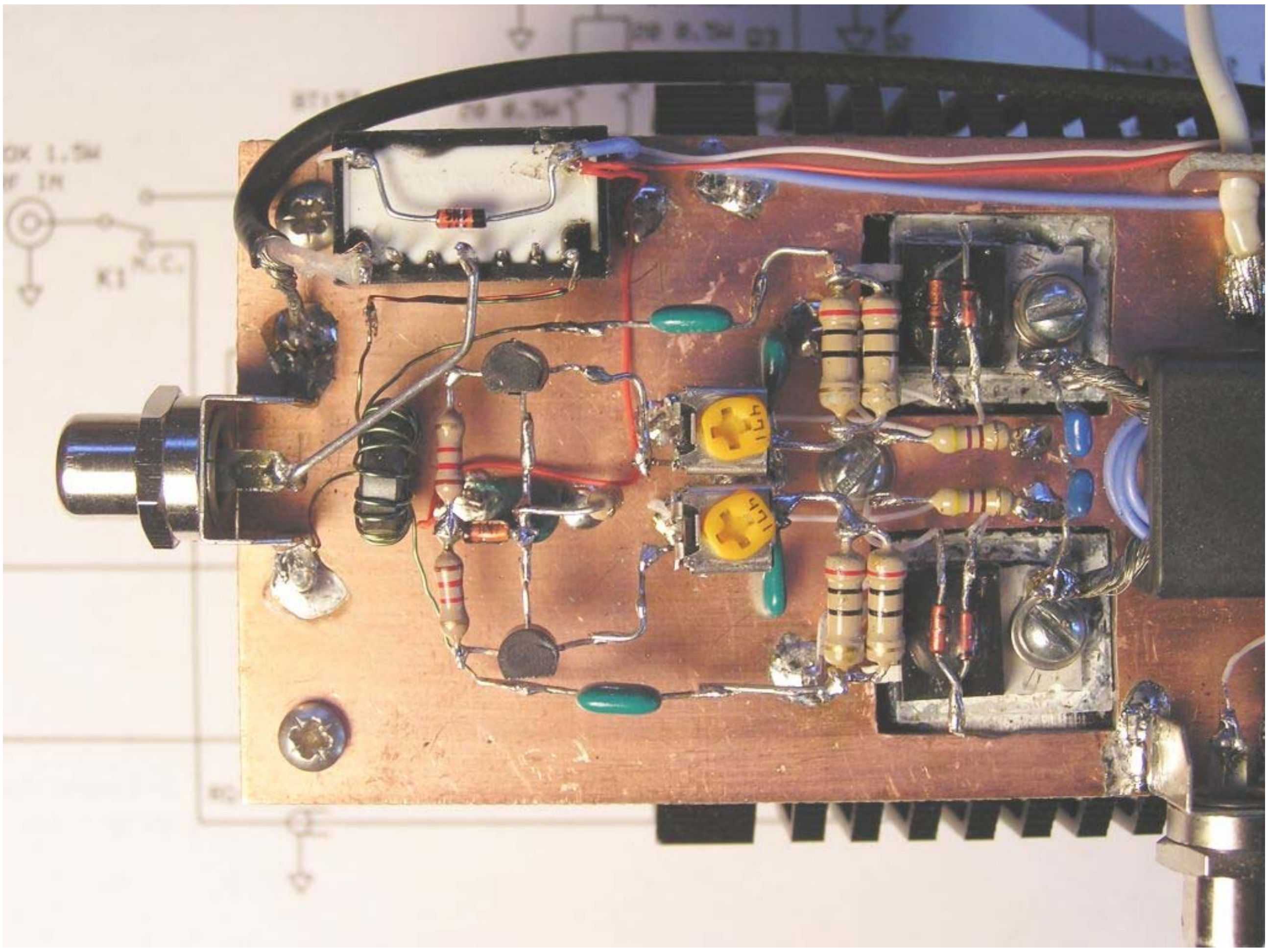
Overall view of bottom side of completed amplifier. Copper-clad is a beautiful material for building circuits, and is easy to work with.



Top three quarter view of amplifier showing temporary longer spacers for testing and adjustment access to bottom side while allowing sufficient clearance for the fan to operate.

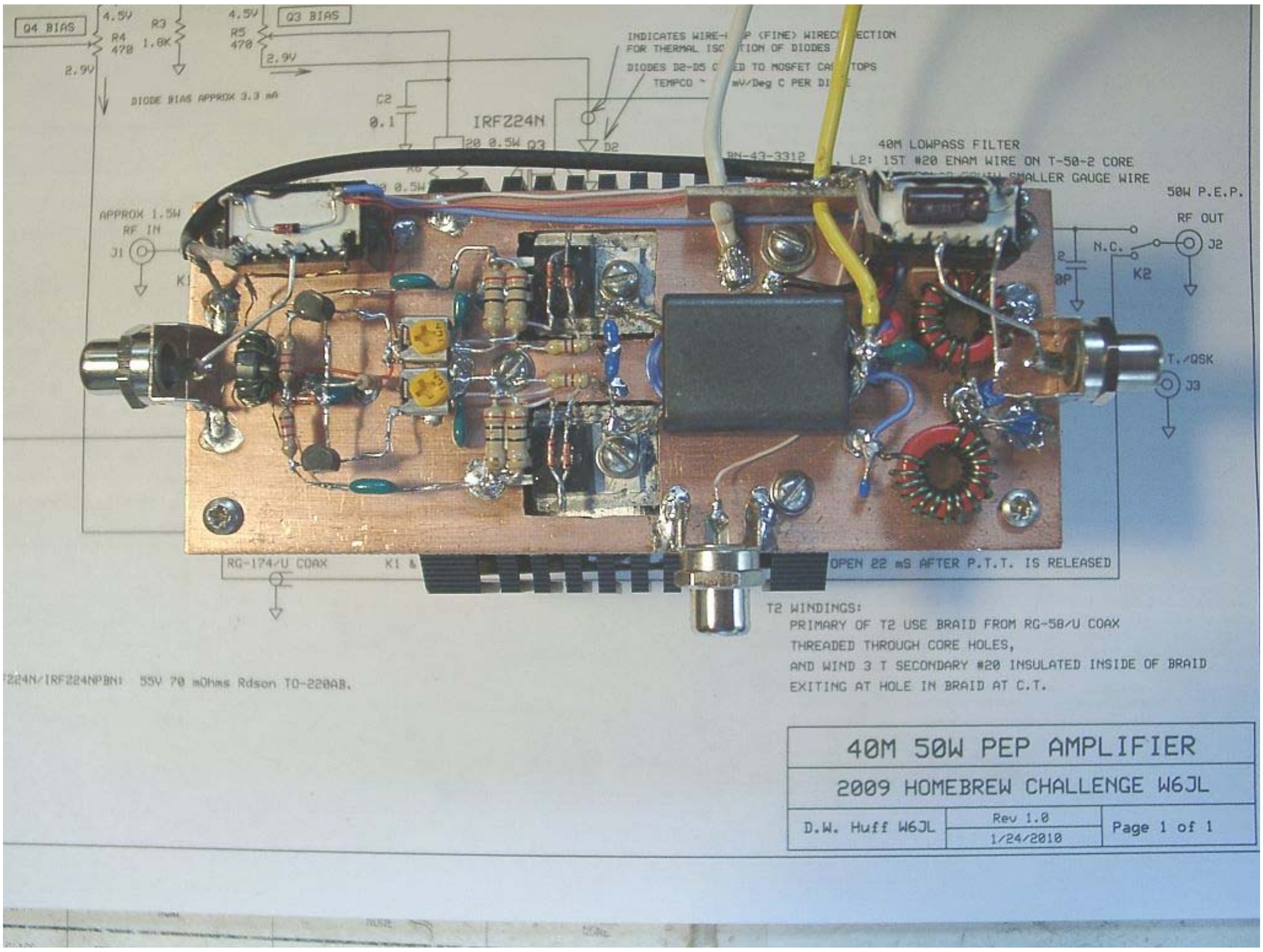


Close view of mosfets and output circuitry. Note 1N4148 diodes epoxied to mosfet case tops. These serve as temperature sensors for bias compensation. Wire-wrap wire serves as a minimal thermal-conductance electrical path for connections to diodes.

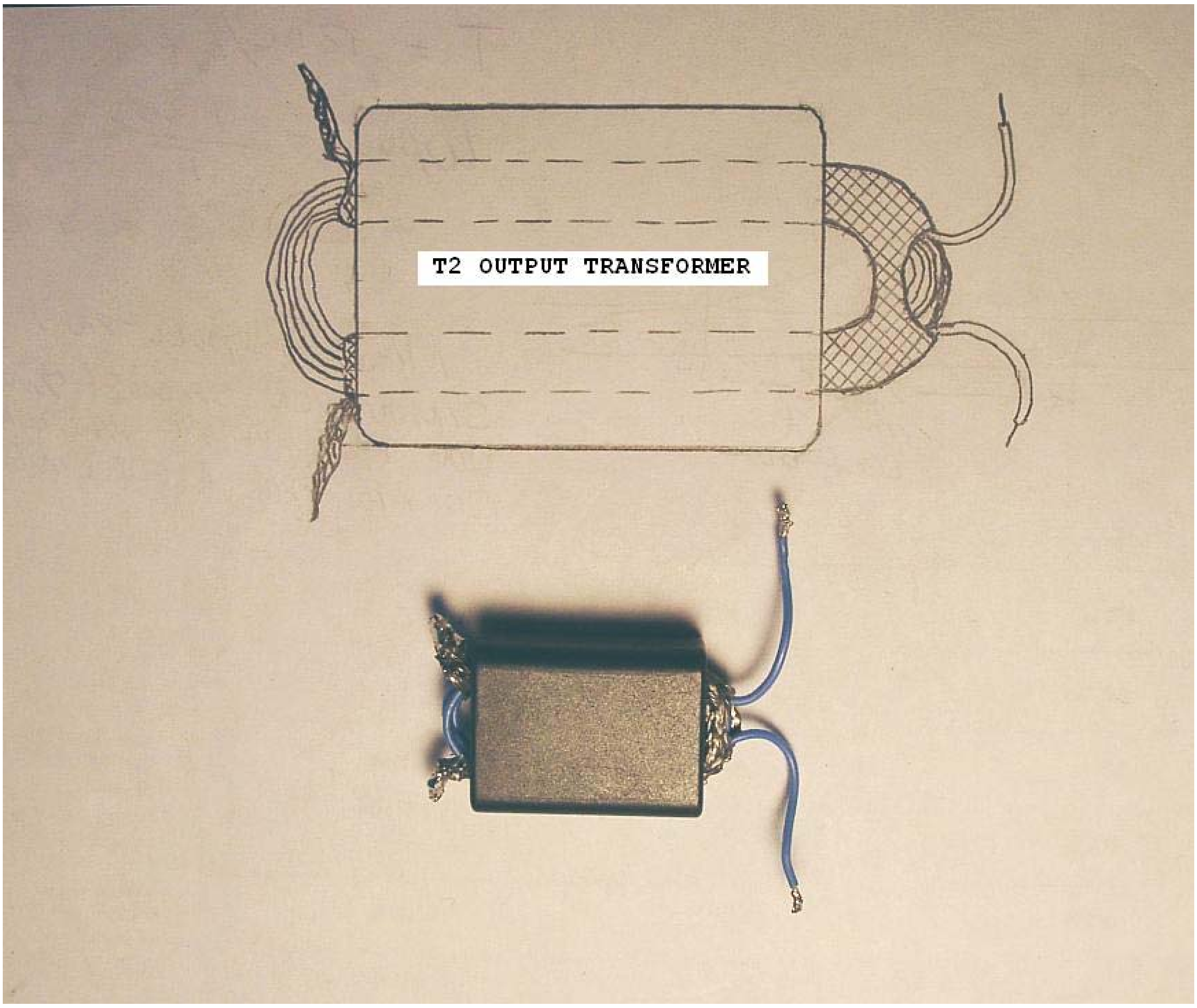


Input circuit of amplifier showing input coupling transformer and temperature compensated bias adjustment circuit.

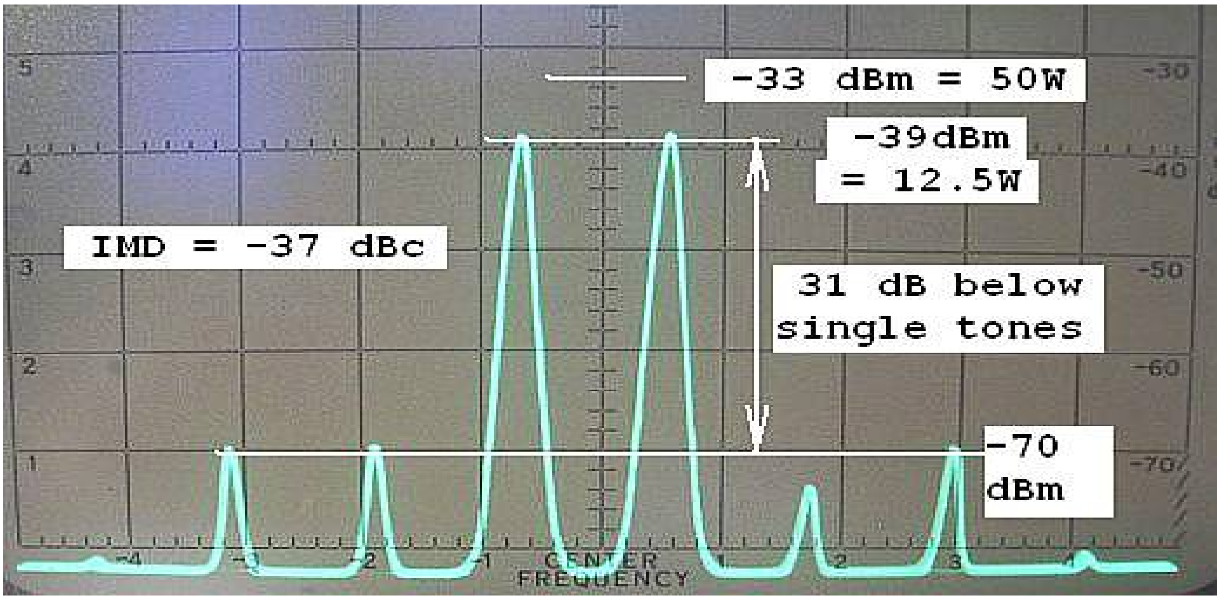
Top view, nearly completed amplifier mounted upright on its spacers, showing Pentium cooler fan/heat sink. Input/output and PTT utilize inexpensive solder-down RCA jacks. No labeling is shown; if you are the builder/user then no need to label anything; you know what each connection is for. Fancy labels are for "show & tell", mainly.



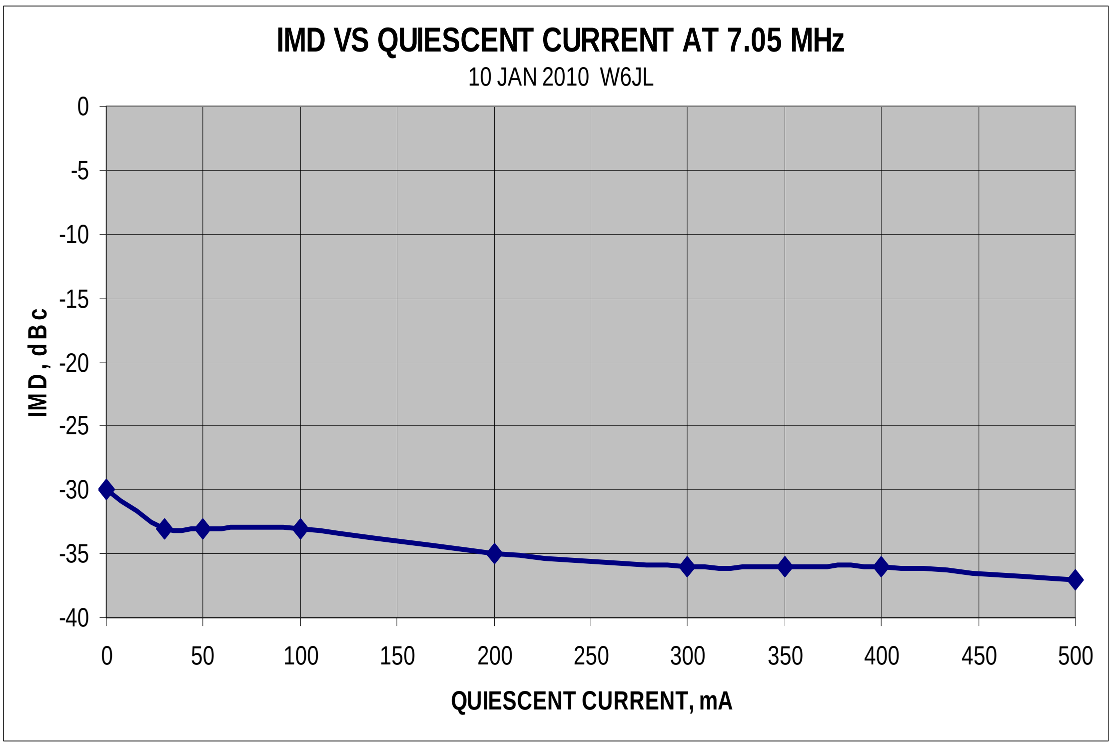
Amplifier bottom view.



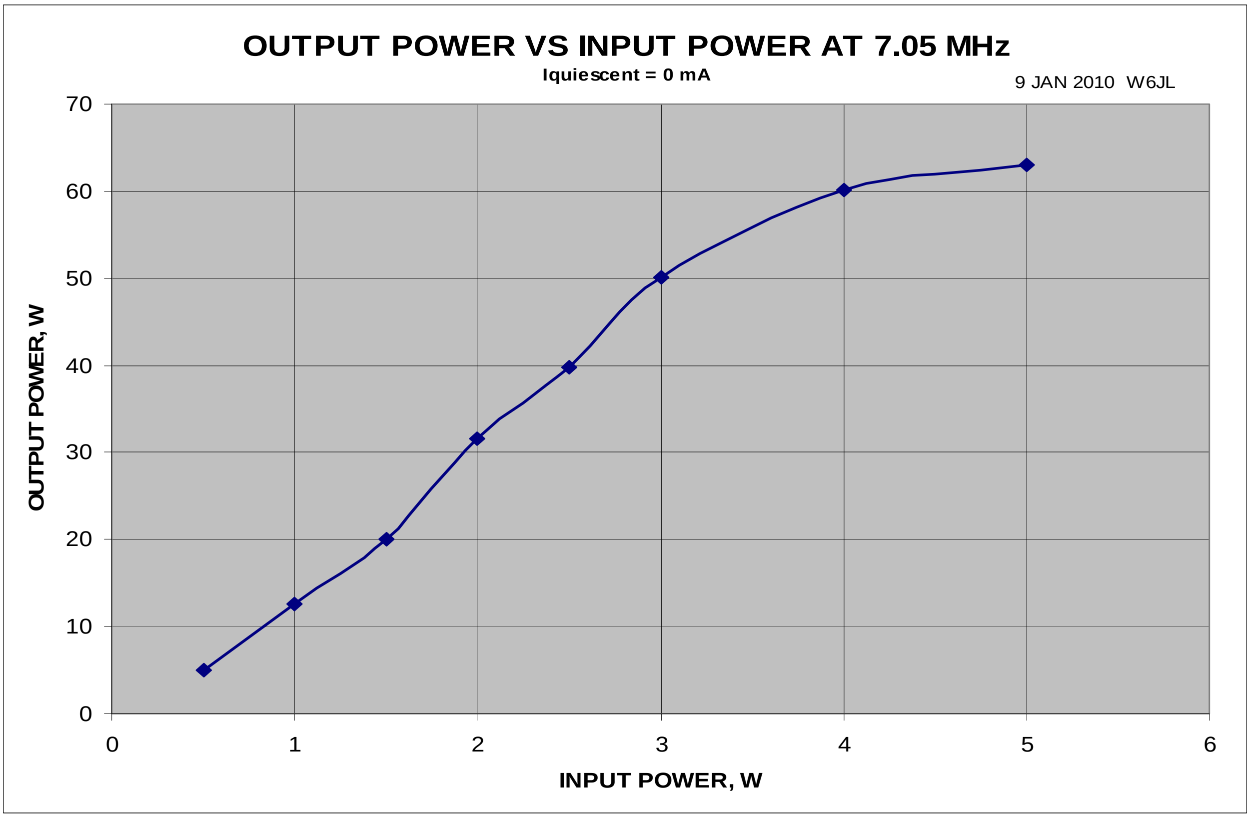
Output transformer construction details. Three turns of insulated wire make up the secondary, wound inside a section of RG-58/U or similar coaxial cable shield, which makes up the U-shaped single primary turn. The core is an Amidon BN-43-3312 ferrite binocular core. An equivalent construction is a stack of toroidal beads on each leg of the primary, at least equal in cross-sectional core area and total core volume as the BN-43-3312. Note that the two legs of the primary and the secondary are not coupled magnetically with each other; electrically the binocular cored transformer is equivalent to two transformers in series. This is why using stacked toroidal beads (or one equivalent larger cylindrical bead) for each half works the same as a binocular core.



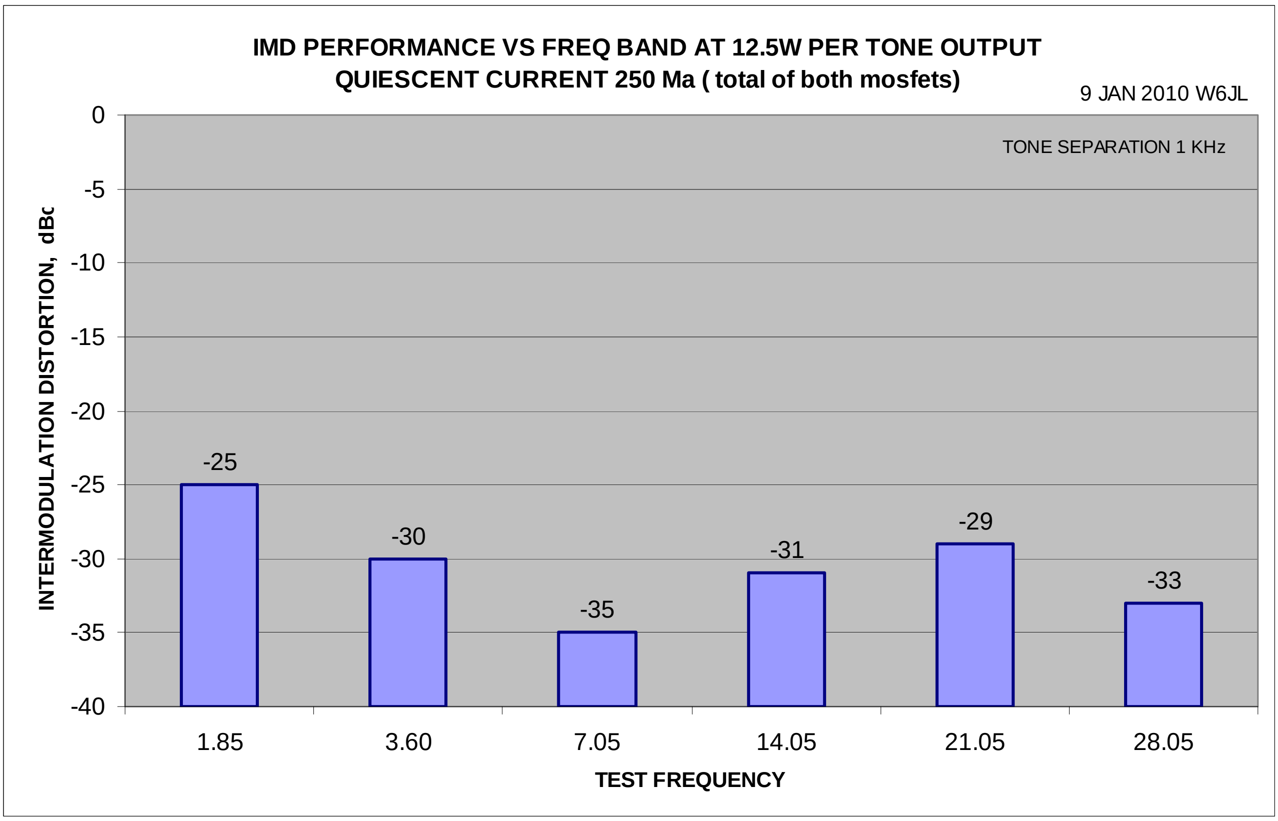
Example spectrum analyzer display of IMD on 7050 KHz, two-tone output at 12.5W per tone. (Quiescent bias is set at 250 mA drain current total, or 125 mA per device). Approximately -37 dBc IMD is shown (highest of the IMD products is about 31 dB below each of the two tones, which in turn are 6 dB below maximum output, adding up to 37 dB below carrier level, or dBc). To obtain actual power output in dBm from the traces, we just add 80 dB (sum of input attenuators used in front of the spectrum analyzer) to the levels shown. Thus, -39 dBm (top of the two tones shown), plus 80 dB gives +41 dBm, or 12.5 watts, per tone, created by the combining of equal level 7.000 KHz and 7.001 KHz input drive signals. The total power output of just the two tones is thus 25W, or 3 dB below 50W. In this particular case, the higher-order IMD products are about equal in level to the first-order ones. Which IMD products dominate depends on the nature of the distortion of the two-tone RF envelope shape.

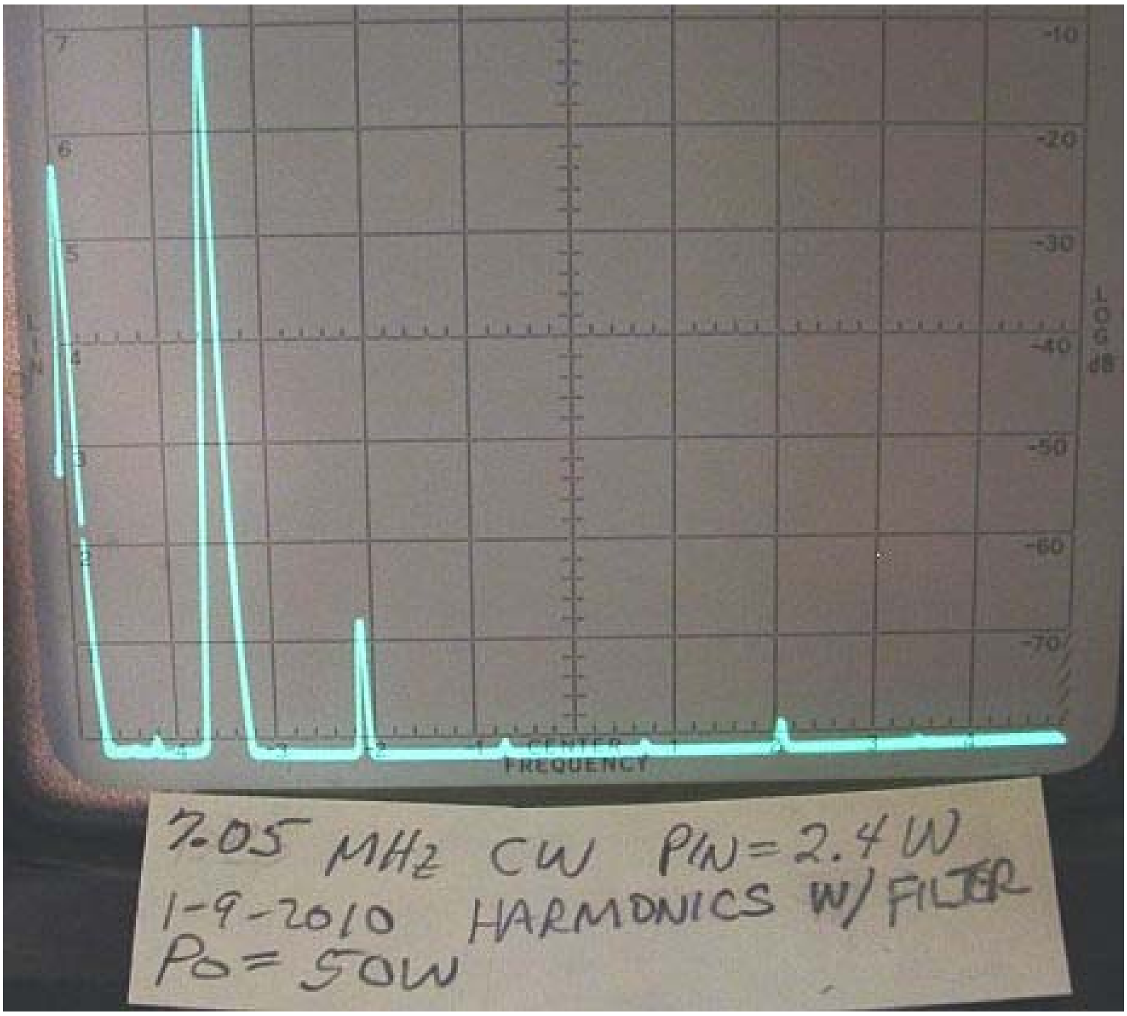


IMD improves somewhat with increasing quiescent drain current, as expected.
This reduces crossover distortion in the RF envelope waveform and also increases gain.

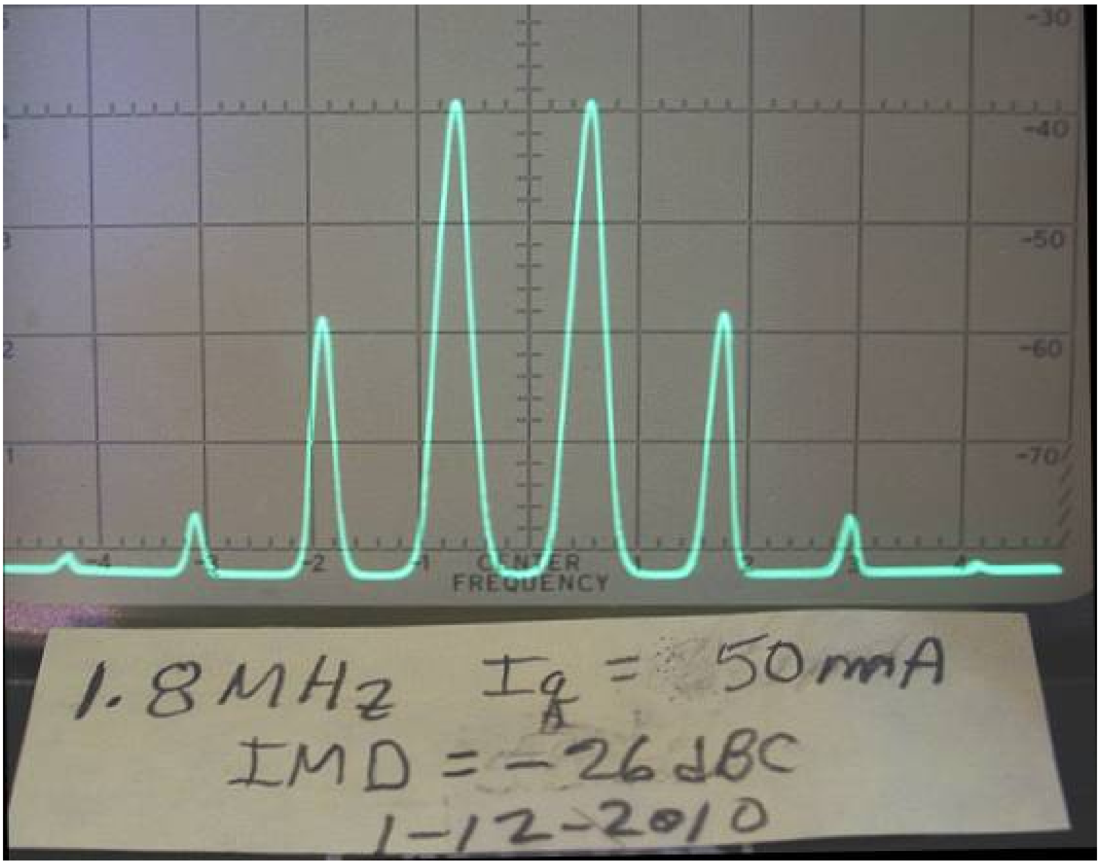


Up to 63W CW output with 5W drive, class B.





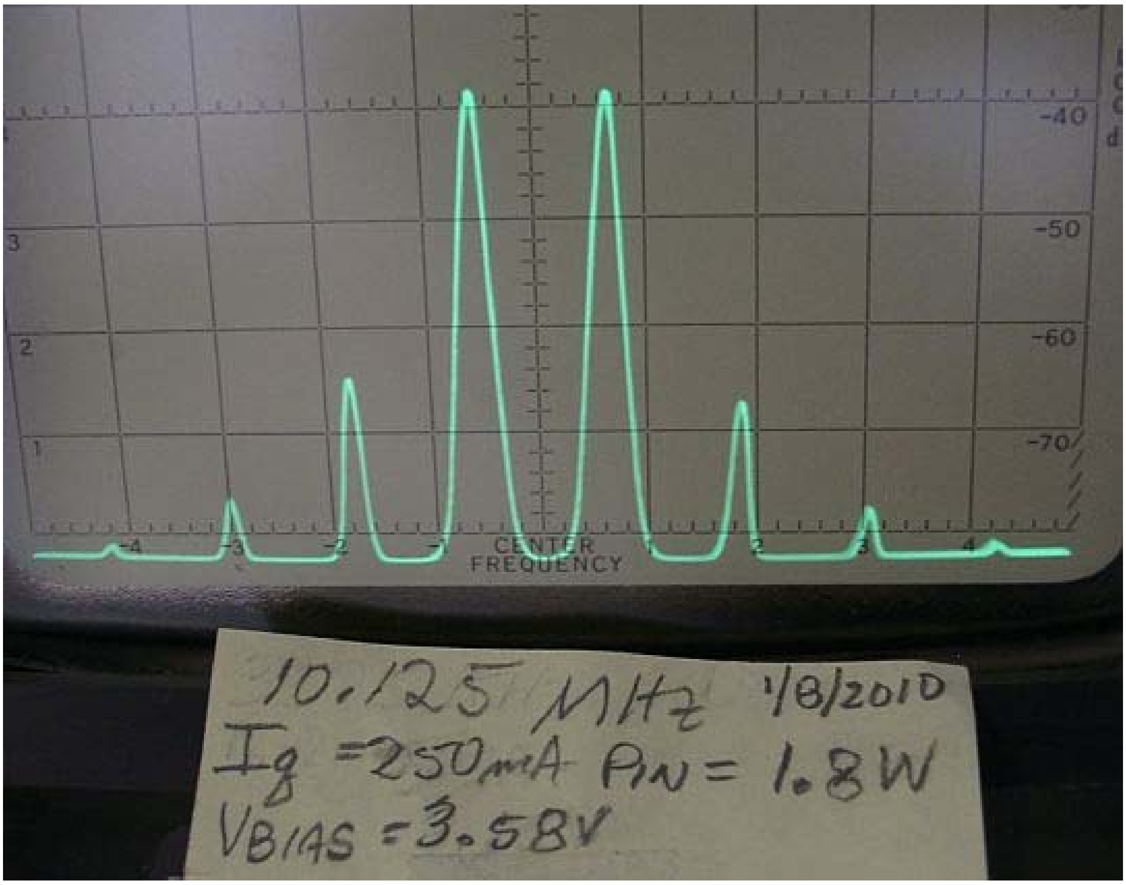
40 meter harmonic output at 50W. Highest is the second harmonic at 58 dB below 50W. Display shows fundamental output at 7.05 MHz, and has been adjusted to be at a level of -10 dBm (second spike from left). The 3rd spike from left is the second harmonic, at 14.1 MHz, which is 58 dB below the fundamental output. The next spike, which is the third harmonic, is greater than 70 dB down. The spike at far left is the "zero frequency" reference of the spectrum analyzer.



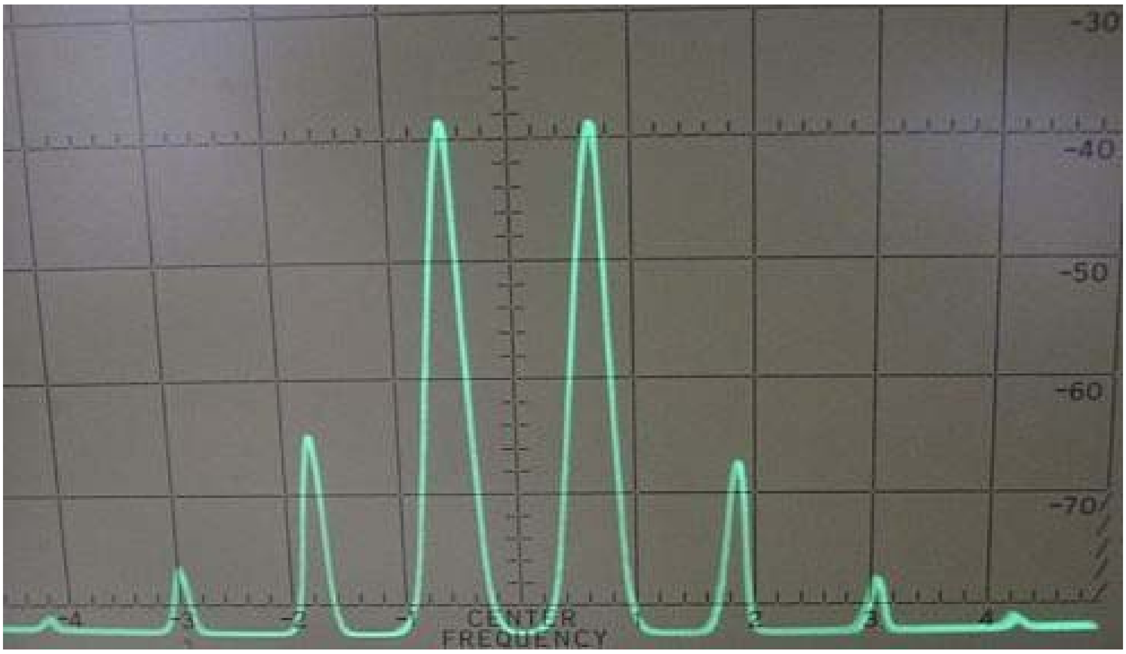
Best obtainable IMD on 1.85 MHz without modifying parts values is only -26 dBc, and this occurs at a low quiescent current of 50 mA.

3600 KHz, IMD -31 dBc at 12.5W per tone

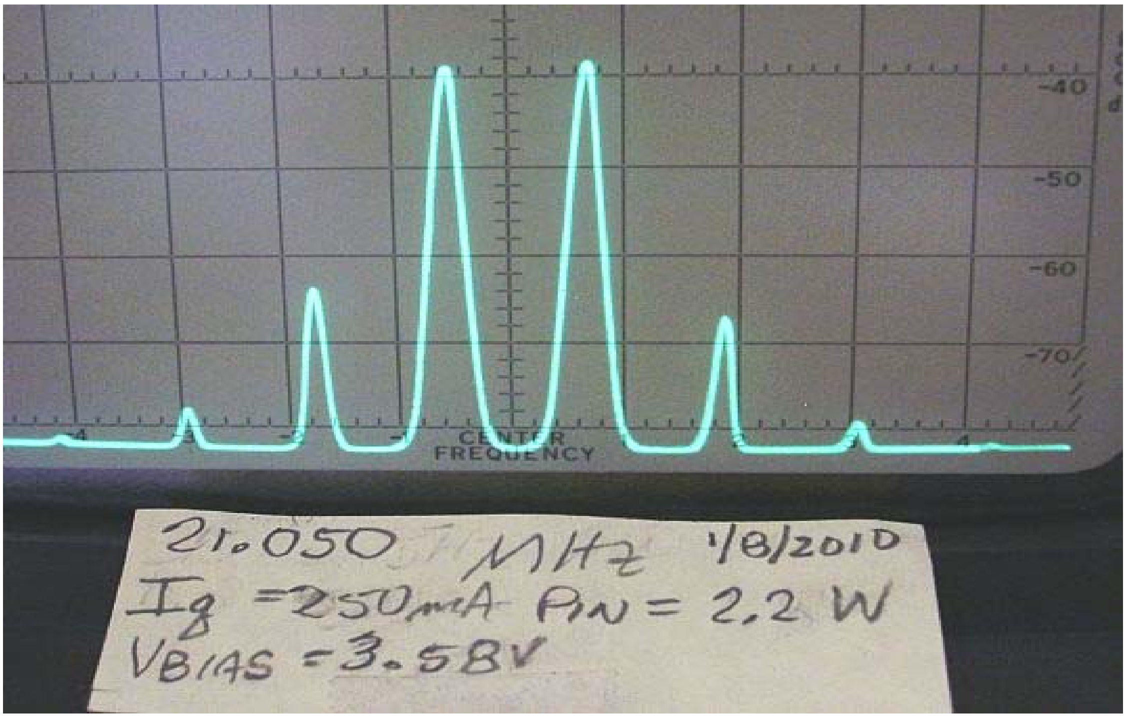
7050 KHz IMD -37dBc. Iq = 250 mA total for both devices.



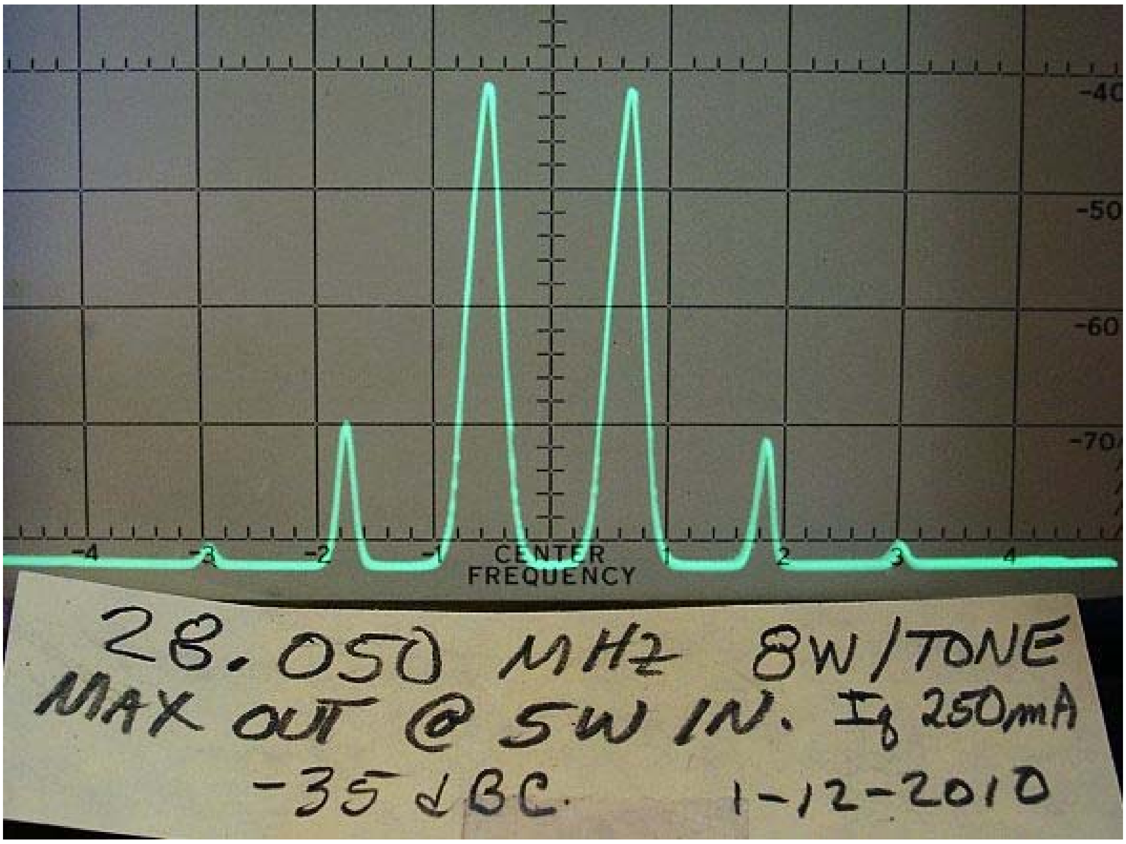
10.125 MHz IMD -32 dBc at 12.5W per tone.



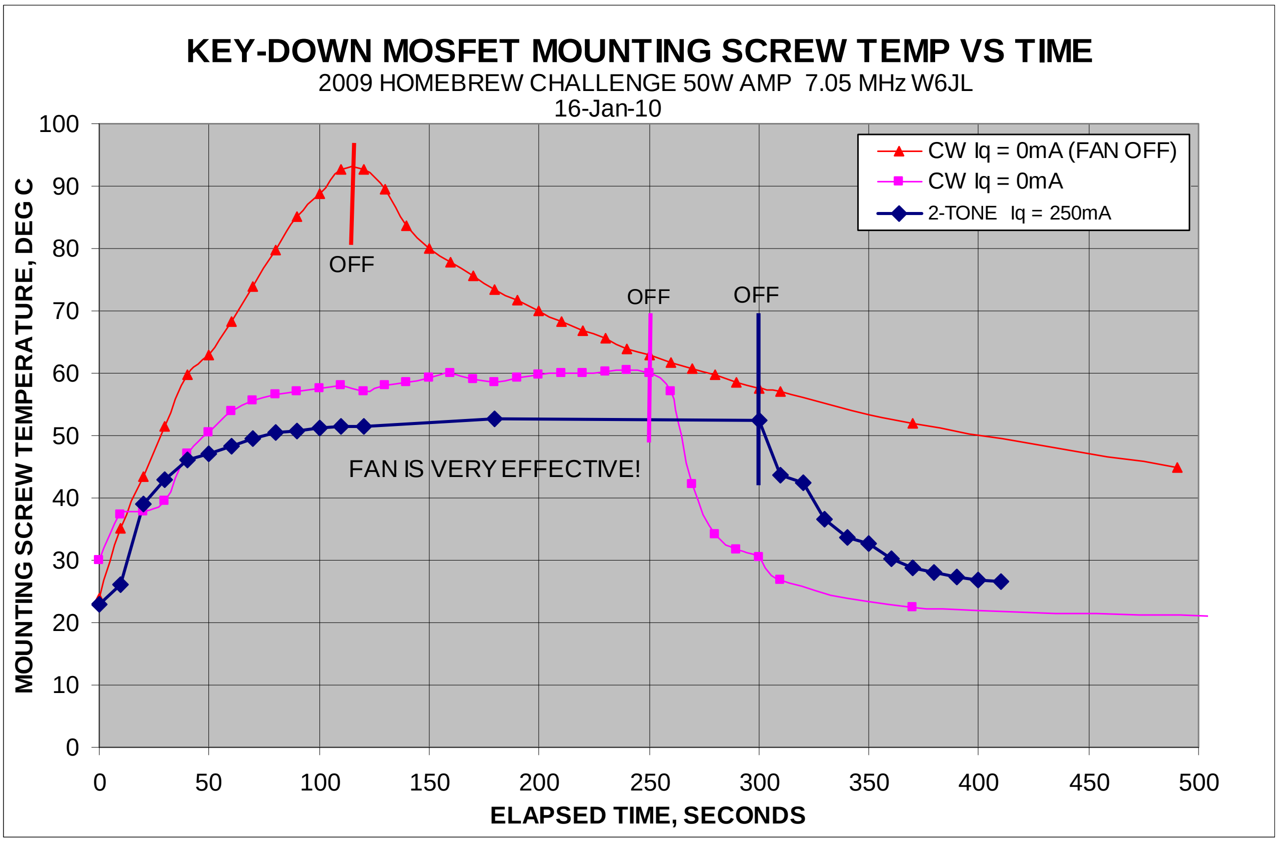
14.050 MHz IMD, no output filter, = -32 dBc at 12.5W per tone



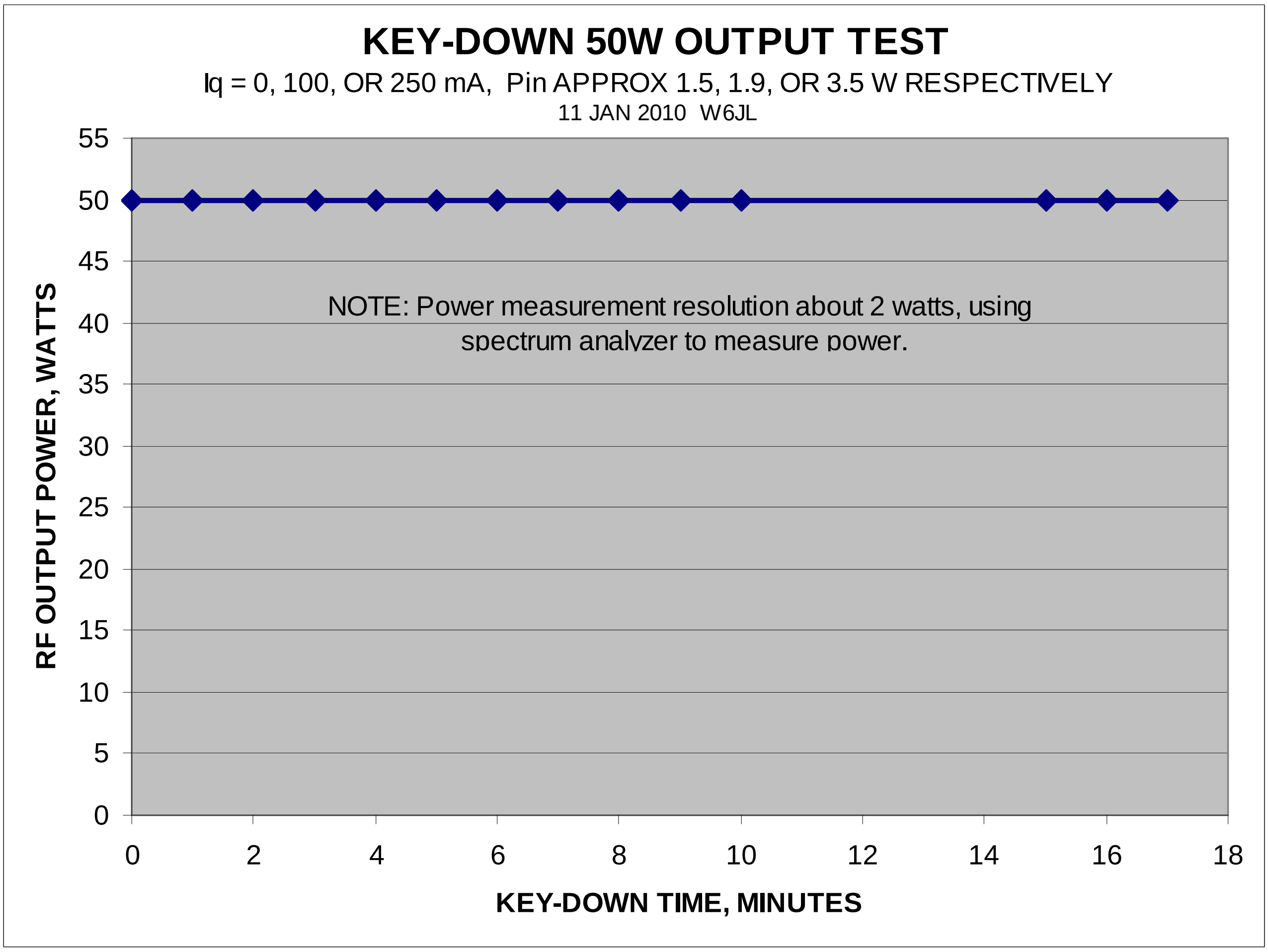
21.050 MHz IMD, no output filter = -30dBc at 12.5W per tone



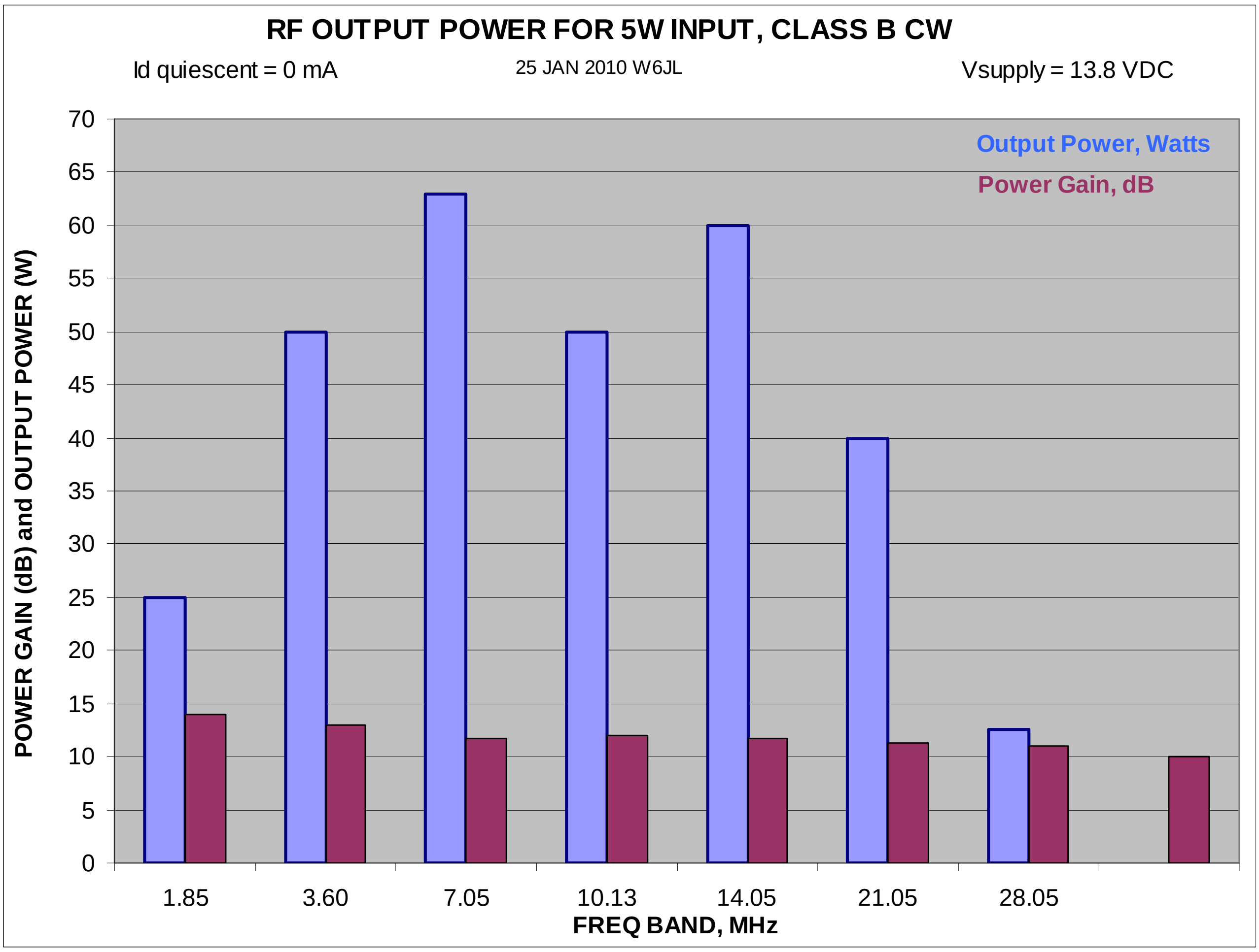
IMD at max two-tone output available on 28.05 MHz, at 5W input. This is equivalent to only about 32 W PEP, but IMD is good at -35 dBc.

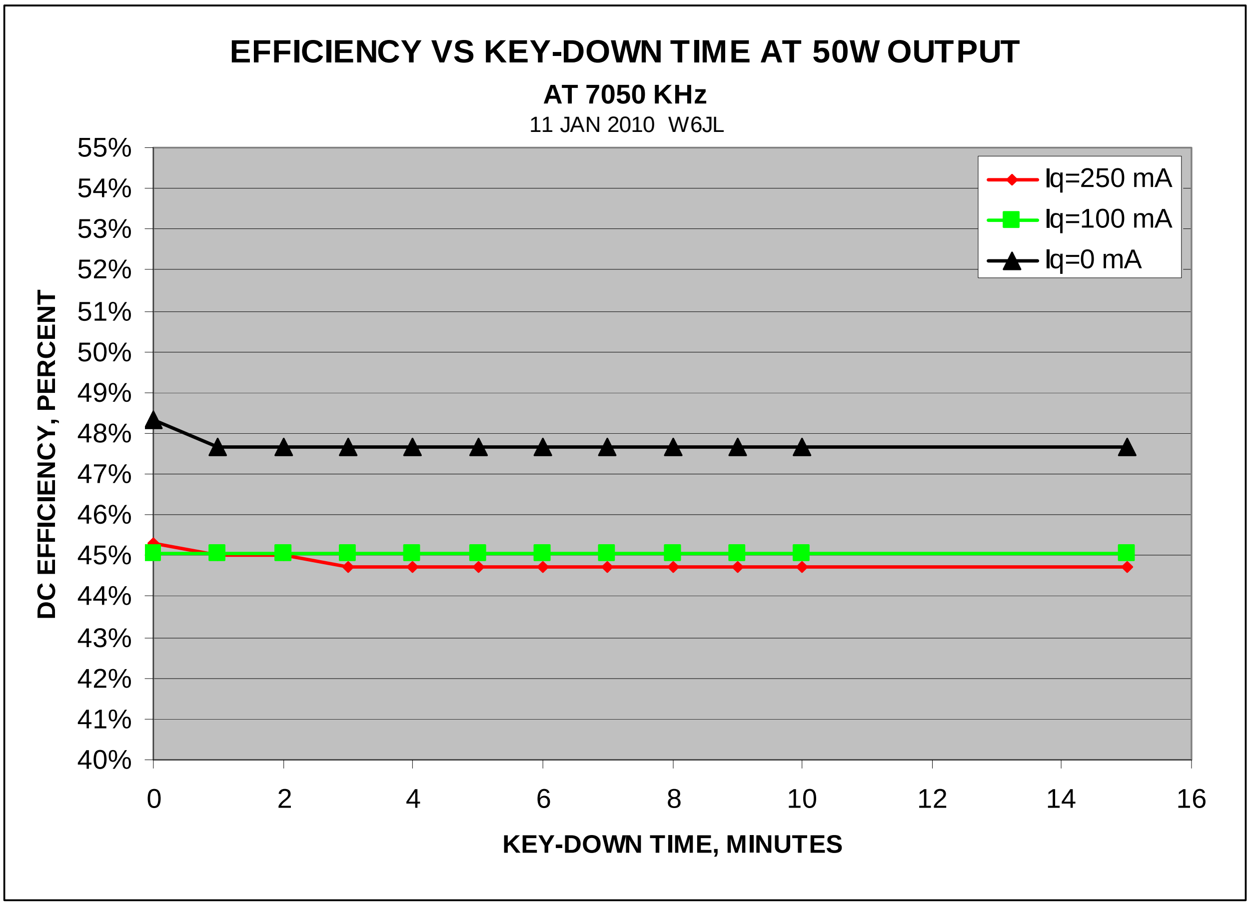


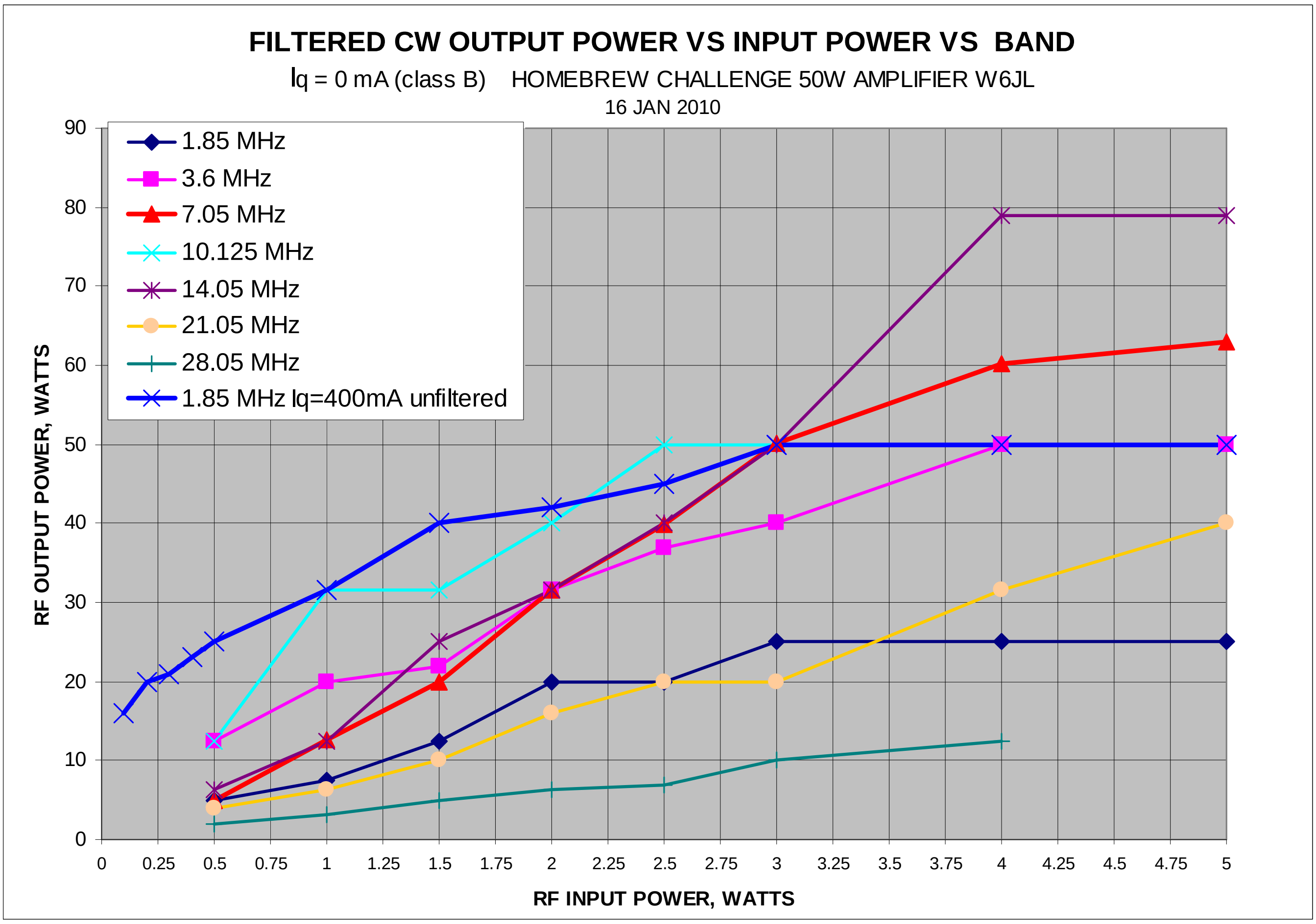
Key-down thermal test at 50W output, 7.05 MHz. Junction temperatures are likely 25 deg C or more above mounting screw temperature.



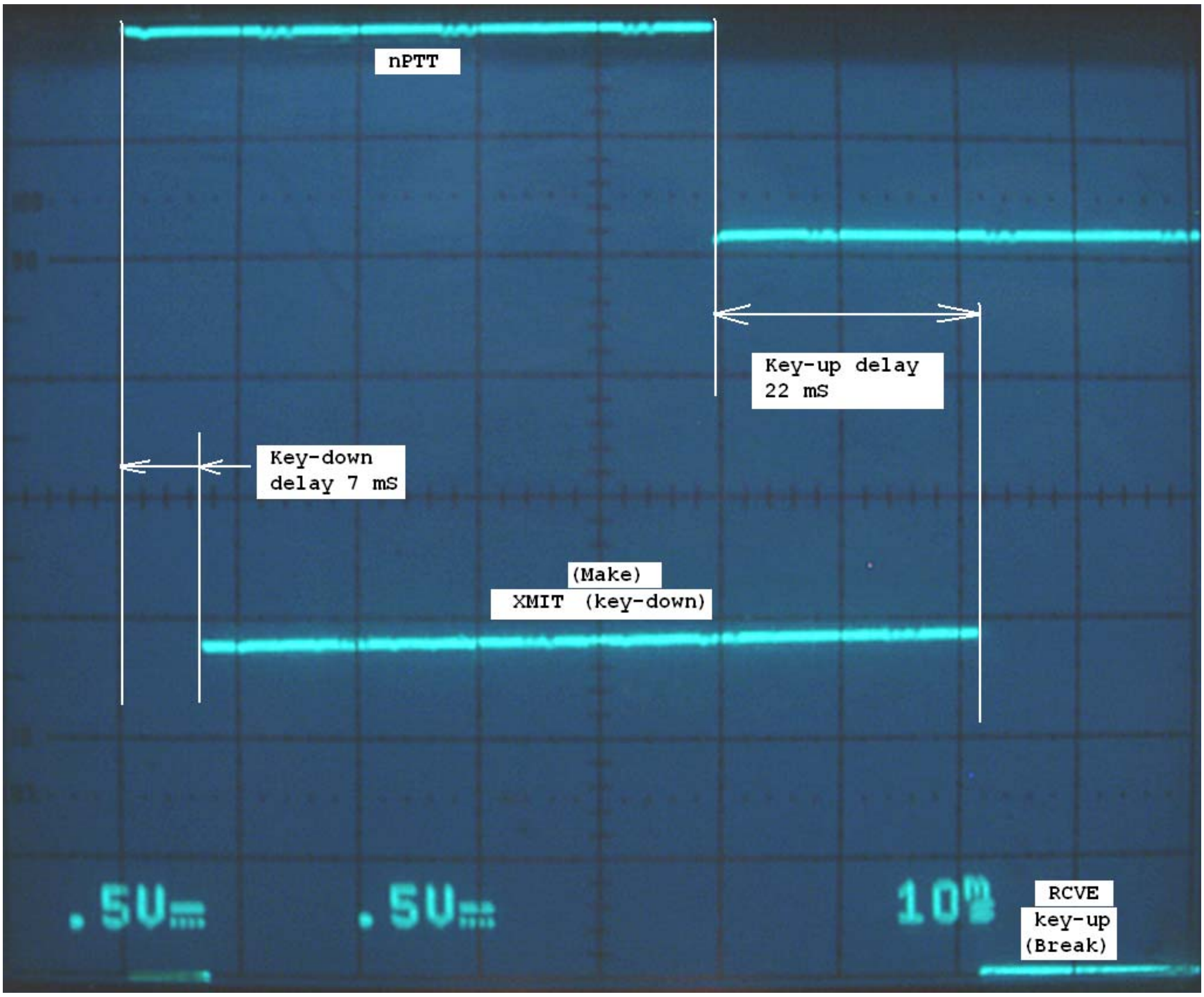
Key-down CW output, 7.05 MHz.


RF OUTPUT POWER FOR 5W INPUT, CLASS B CW
Id quiescent = 0 mA
25 JAN 2010 W6JL
Vsupply = 13.8 VDC
Output Power, Watts
Power Gain, dB
POWER GAIN (dB) and OUTPUT POWER (W)
70
65
60
55
50
45
40
35
30
25
20
15
10
5
0
1.85
3.60
7.05
10.13
14.05
21.05
28.05
FREQ BAND, MHz


EFFICIENCY VS KEY-DOWN TIME AT 50W OUTPUT
AT 7050 KHz
11 JAN 2010 W6JL
DC EFFICIENCY, PERCENT
55%
54%
53%
52%
51%
50%
49%
48%
47%
46%
45%
44%
43%
42%
41%
40%
0
2
4
6
8
10
12
14
16
KEY-DOWN TIME, MINUTES
Iq=250 mA
Iq=100 mA
Iq=0 mA

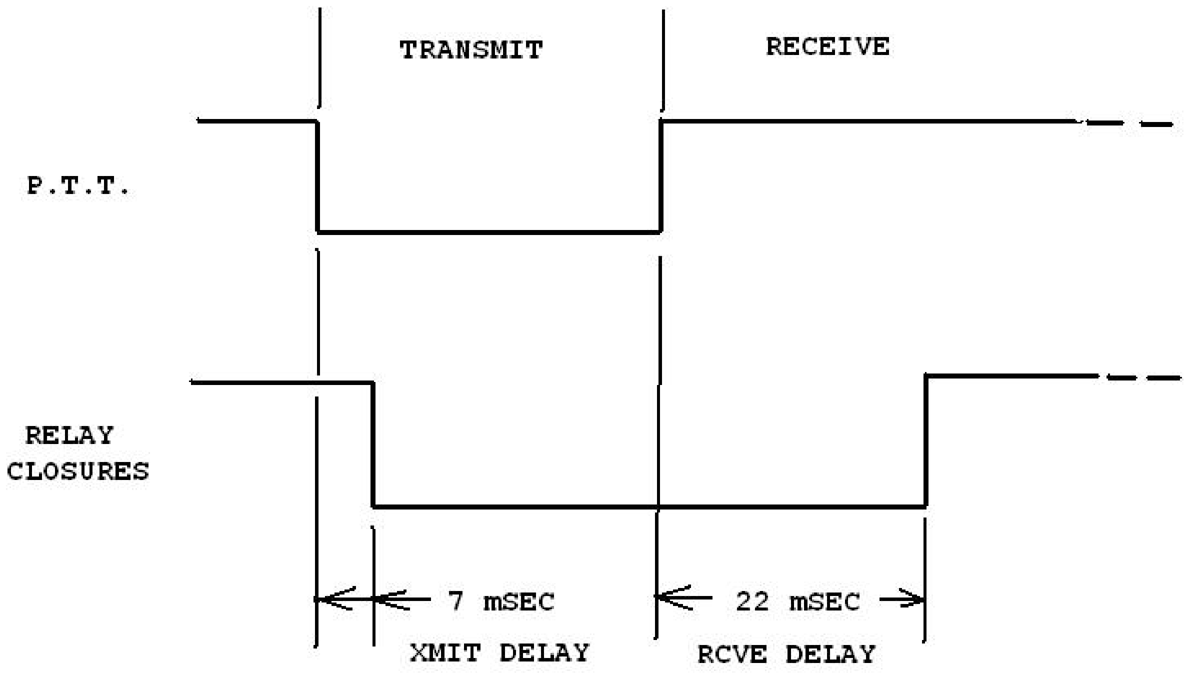


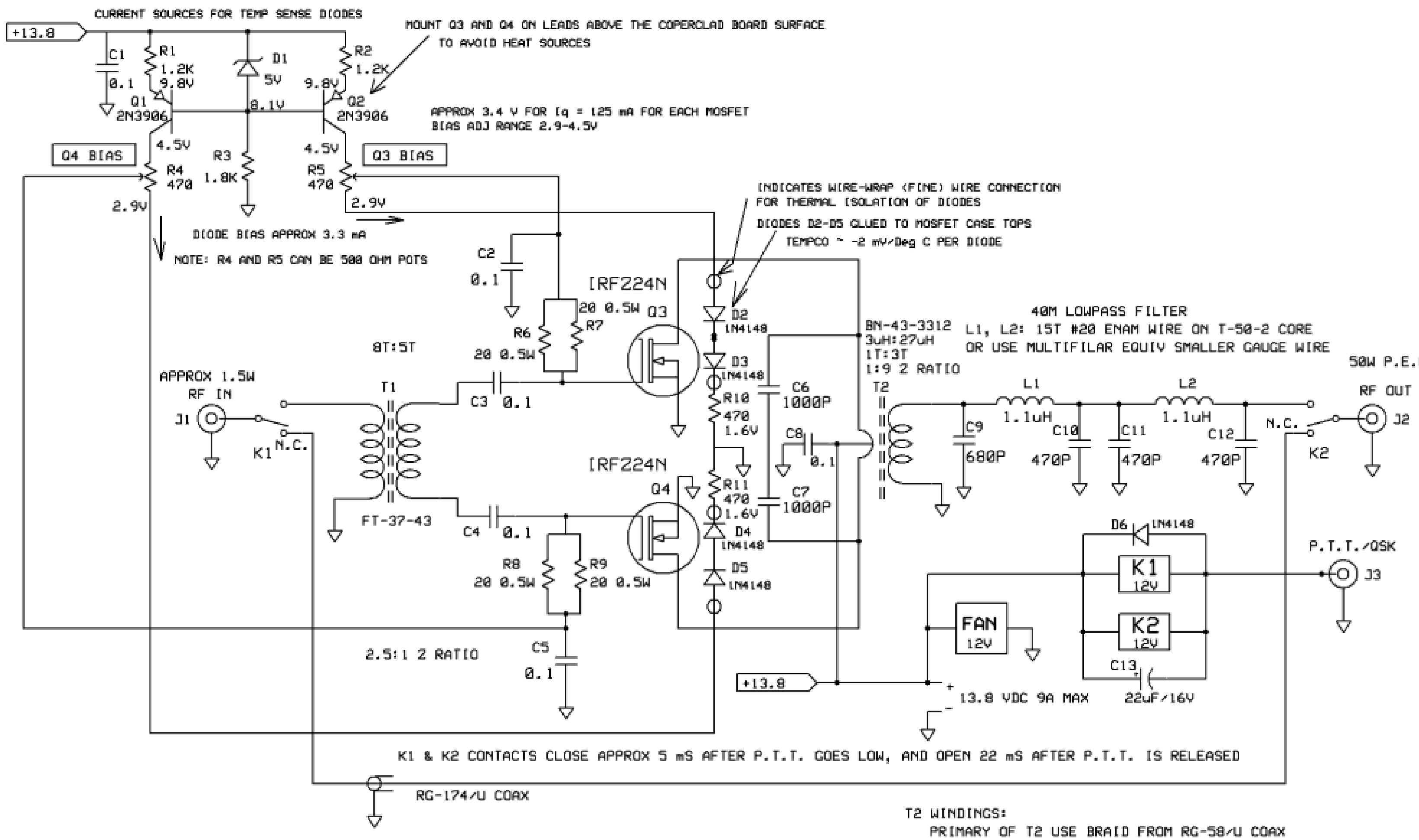
Multiband power output vs input, CW Class B.



PTT/QSK T/R switching timing. With C12 = 22uF and using Omron G5Y-1-H relays, both relays close (key-down, when nPTT is high) 7 mS after PTT is pulled low (which is key-down) and open 22 mS after PTT goes high (which is key-up). This should be sufficient for QSK at keying speeds up to about 30 wpm if the exciter's RF output does not rise before 7 mS after key-down, and stays on no longer than 22 mS after key-up. This can usually be easily arranged (especially if the exciter is also homebrew!). It is assumed that the PTT line is controlled by the CW keying line of the exciter. NOTE: The sense of the PTT and resultant relay closures are reversed in these scope traces because a pulse generator was used to drive a mosfet which pulled the relay coils to ground to close the relays. The "nPTT" trace is the positive-going gate drive signal to this mosfet. The closure of the relays was sensed by probing the voltage across a 1K resistor connected between the relay contacts and ground, (with the other set of relay contacts connected to +13.8 VDC), thus this trace also goes high when the relays are closed. Below is a clearer graphic equivalent of the above scope trace photo.

OMRON G5Y-1-H RELAY PULL-IN AND DROP-OUT TIME
IN-CIRCUIT WITH 22uF SHUNT CAPACITOR





T2 WINDINGS:
PRIMARY OF T2 USE BRAID FROM RG-58/U COAX
THREADED THROUGH CORE HOLES,
AND WIND 3 T SECONDARY #20 INSULATED INSIDE OF BRAID
EXITING AT HOLE IN BRAID AT C.T.

IRF224N/IRF224NPBN: 55V 70 mOhms Rdson TO-220AB.

IRFL224N/IRL2NPBF: SAME DEVICE BUT LOWER THRESHOLD VOLTAGE 1-2V. CHANGE R10/R11 TO 0 OHMS AND R1/R2 TO 1K OHMS CURRENT SOURCES NOW BECOME 4.0 mA AND BIAS POT RANGE BECOMES ~ 1.2-2.0 VDC.

IRF234N/IRF234NPBF: WILL WORK BUT SLIGHTLY POORER IMD AND POORER HIGHER BAND PERFORMANCE DUE TO HIGHER CAPACITANCE AND LONGER SWITCHING TIMES.

40M 50W PEP AMPLIFIER		
2009 HOMEBREW CHALLENGE W6JL		
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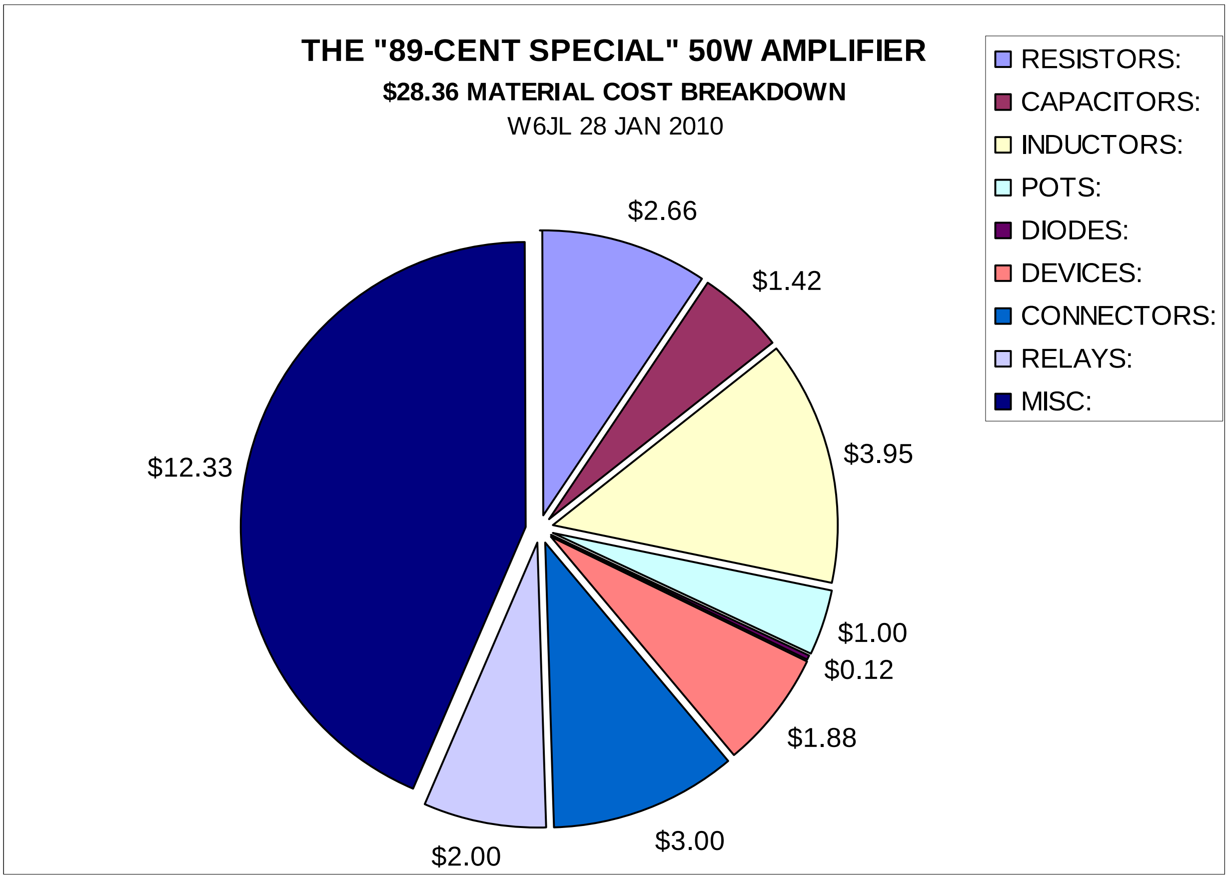
HOMEBREW CHALLENGE 2009 50W AMP. W6JL 26 JAN 2010

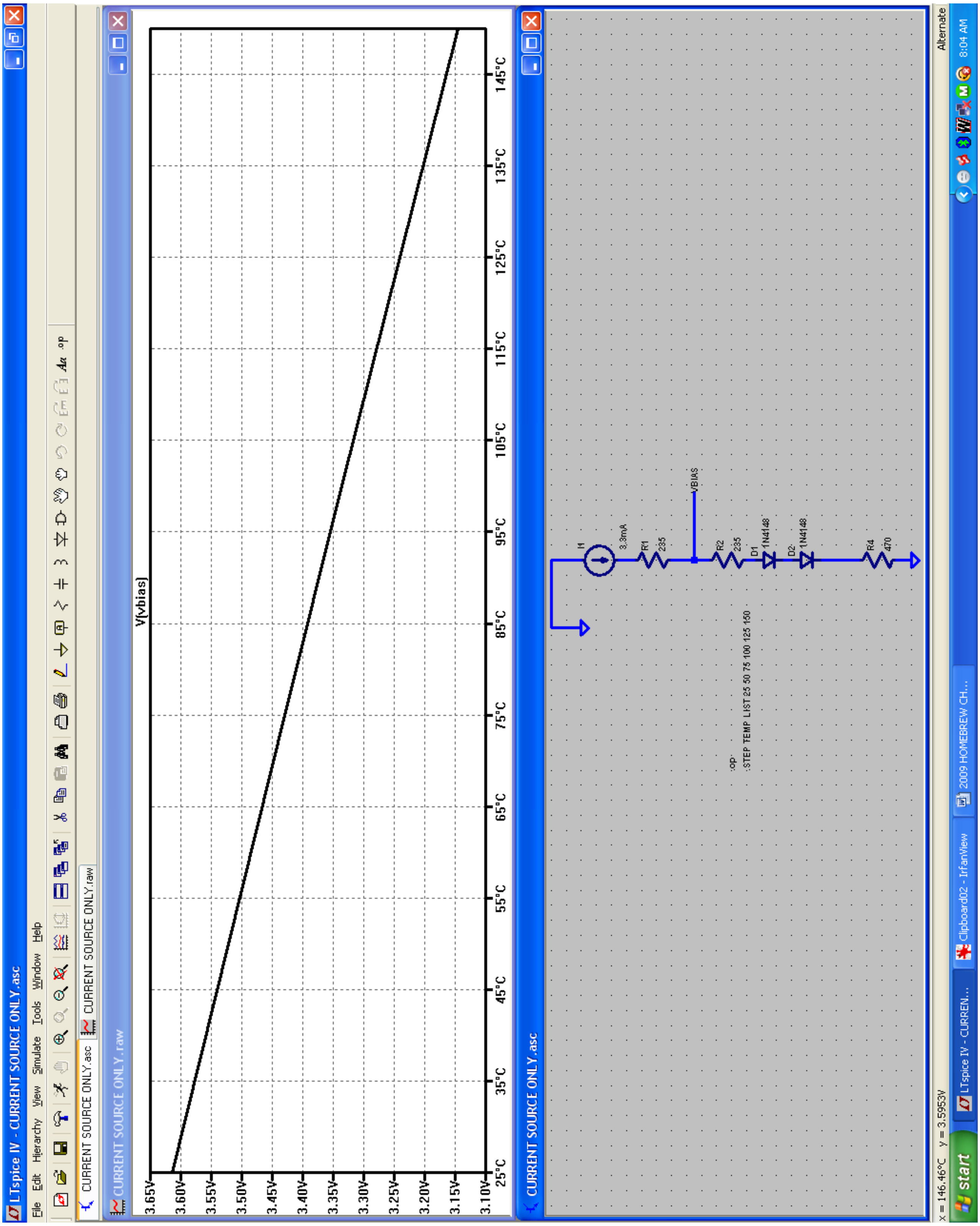
MATERIAL LIST/COSTING:

DESCRIPTION	QTY	COST	EXTENDED	SOURCE	P/N	REF DES.	NOTES
RESISTORS:							
20 OHMS 1/2W CARBON	4	$0.23	$0.92	WWW.MOUSER.COM	30BJ500-20	R6-R9	
3.6K OHMS 1/4W CARBON	1	$0.29	$0.29	WWW.MOUSER.COM	791-RC1/4-362JB	R4	
820 OHM 1/4W CARBON	1	$0.29	$0.29	WWW.MOUSER.COM	791-RC1/4-821JB	R5	
470 OHM 1/4W CARBON	2	$0.29	$0.58	WWW.MOUSER.COM	791-RC1/4-272JB	R1, R2	
2.2K OHMS 1/4 W CARBON	1	$0.29	$0.29	WWW.MOUSER.COM	791-RC1/4-222JB	R10	
100 OHMS 1/4W CARBON	1	$0.29	$0.29	WWW.MOUSER.COM	791-RC1/4-101JB	R11	
		TOTAL	$2.66				
CAPACITORS:							
0.1uF/50V MONOLITHIC	6	$0.10	$0.60	ELEXP.COM	14KK050.1U	C1-C6	
680PF 500V DISC	1	$0.12	$0.12	ELEXP.COM	14DK500680P	C9	
470PF 500V DISC	3	$0.12	$0.36	ELEXP.COM	14DK500470P	C10-C12	
1000PF 100V MONOLITHIC	2	$0.10	$0.20	ELEXP.COM	14MN050.001U	C6, C7	
22UF 16V ELECTROLYTIC	1	$0.14	$0.14	DIGIKEY.COM	P808-ND	C13	
		TOTAL	$1.42				
INDUCTORS:							
T-50-2 CORE (FOR 1.1uH) 15T 4-FILAR #28	2	$0.65	$1.30	CWS BYTEMARK.COM	T-50-2	L1, L2	ALT: DANSSMALLPARTSANDKITS.NET: 3 FOR $1.45
BN43-3312 BINOC CORE OUTPUT XFMR	1	$2.00	$2.00	CWS BYTEMARK.COM	BN-43-3312	T2	
FT-37-43 TOROID CORE INPUT XFMR	1	$0.65	$0.65	CWS BYTEMARK.COM	FT-37-43	T1	ALT: DANSSMALLPARTSANDKITS.NET: 3 FOR $1.60
		TOTAL	$3.95				
POTS:							
470 OHM TRIMPOT	1	$1.00	$1.00	GOLDMINE-ELEC-PRODUCTS.COM	G9226 (FOR 3 PCS)	R3	
		TOTAL	$1.00				
DIODES:							
1N4148	3	$0.02	$0.06	MOUSER.COM	512-1N4148	D1-D3	
5.2V ZENER	1	$0.06	$0.06	MOUSER.COM	512-1N5231C	D4	
		TOTAL	$0.12				
DEVICES:							
IRFZ24ZN MOSFET 55V 0.07 OHMS	2	**$0.89**	$1.78	JAMECO.COM	669871	Q3, Q4	ALT: DIGIKEY IRFZ24NPBF-ND @ $1.62 1's OR $1.00 10's
2N3906 NPN TRANSISTOR	2	**$0.05**	$0.10	JAMECO.COM	178597	Q1, Q2	ALT: JAMECO IRLZ24N P/N 669901 @ $0.69
		TOTAL	$1.88				ALT: JAMECO IRFZ34N P/N 669935 @ $0.85
							ALT: JAMECO IRLZ34N P/N 668843 @ $0.79
CONNECTORS:							
RCA JACK SOLDER-DOWN	3	$1.00	$3.00	GOLDMINE-ELEC-PRODUCTS.COM	G13173	J1-J3	PER 8 PCS
		TOTAL	$3.00				
RELAYS:							
RELAY SPDT 12V COIL	2	$1.00	$2.00	GOLDMINE-ELEC-PRODUCTS.COM	G13442	K1, K2	ALT: (9V COIL)MOUSER P/N 653-G5V-1-DC9 @ $1.52
		TOTAL	$2.00				ALT: (9V COIL) NEWARK P/N 89C6069 @ $1.71
MISC:							ALT: (5V COIL) DIGI-KEY P/N Z773-ND @ $1.90
4-40 THREADED ALUM SPACER 3/4"	4	$0.29	$1.16	http://www.oselectronics.com			ALT: (5V COIL) ARROW P/N G5V1DC9 @ $1.32
PC COPPER CLAD 3X6 IN 0.032	1	$1.72	$1.72	GOLDMINE-ELEC-PRODUCTS.COM			ALT: (12V COIL) ALLILED P/N 821-1042 @ $1.62
CPUT FAN+HEATSINK	1	$3.75	$3.75	JAMECO ELECTRONICS	EC-K6-6B		
TO-220 INSULATOR KIT MICA	2	$1.85	$3.70	http://www.oselectronics.com			
WIRE 15 FT #22 L1 AND L2	1	$1.00	$1.00	DANSSMALLPARTSANDKITS.NET: 15FT FOR $1.00			
WIRE 25 FT #28 T1 AND T2	1	$1.00	$1.00	DANSSMALLPARTSANDKITS.NET: 25FT FOR $1.00			
		TOTAL	$12.33				
	GRAND TOTAL		**$28.36**				

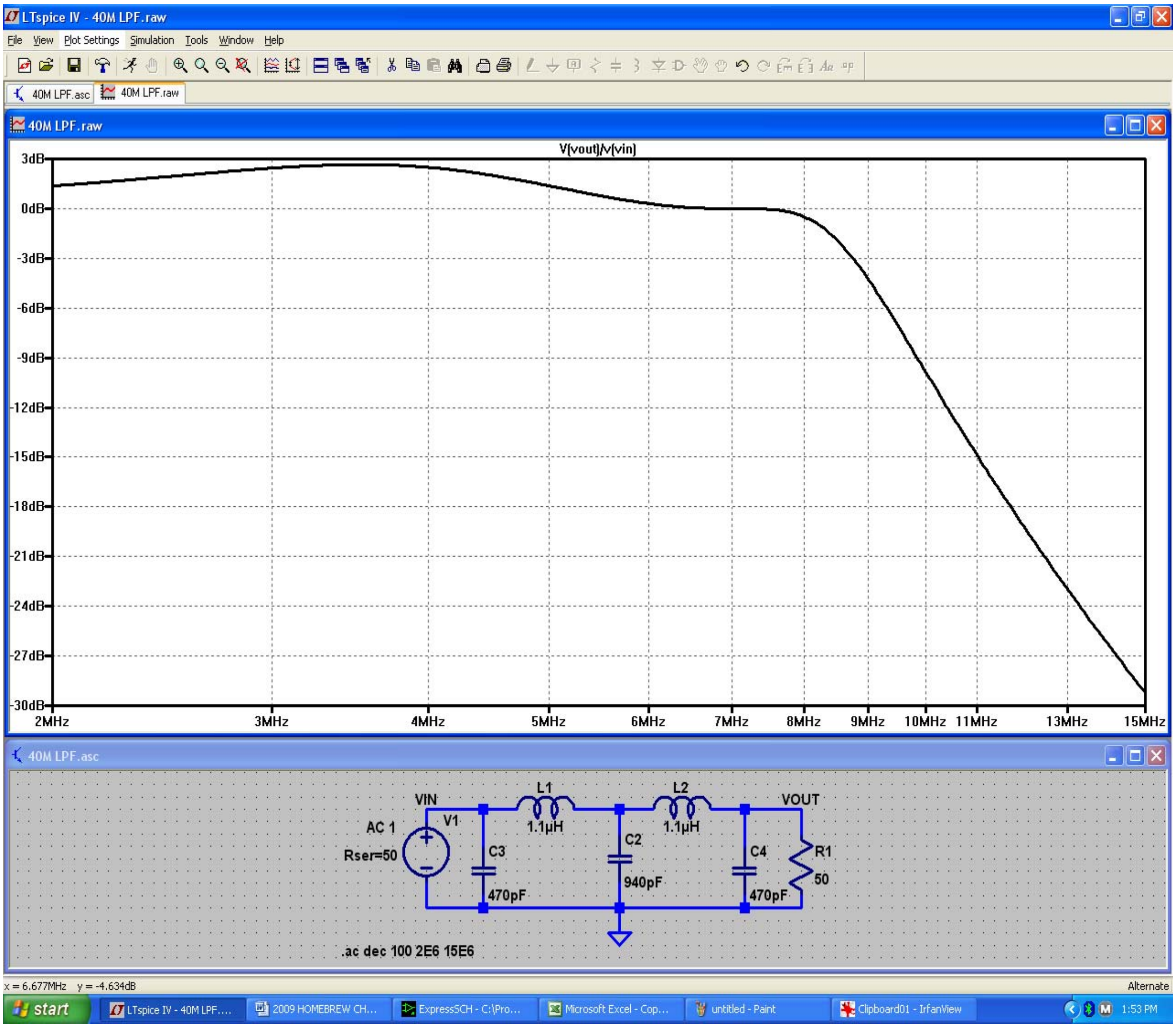


Material cost breakdown for the amplifier. In many cases, the “Miscellaneous” segment (as well as several other segments) of the pie chart can be made $0, if your junk-box is reasonably well-stocked.

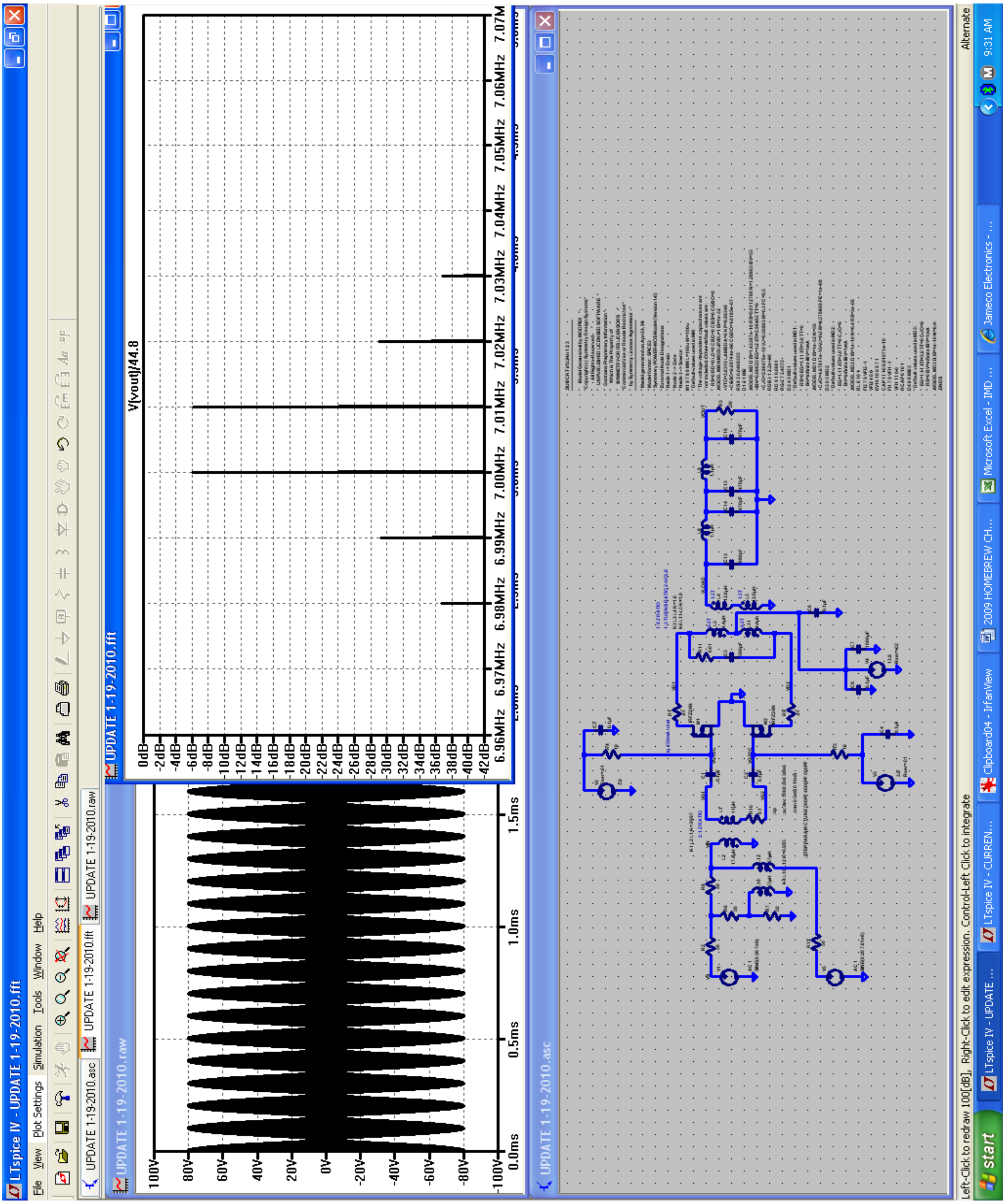
AMPLIFIER SIMULATIONS USING LTSPICE:



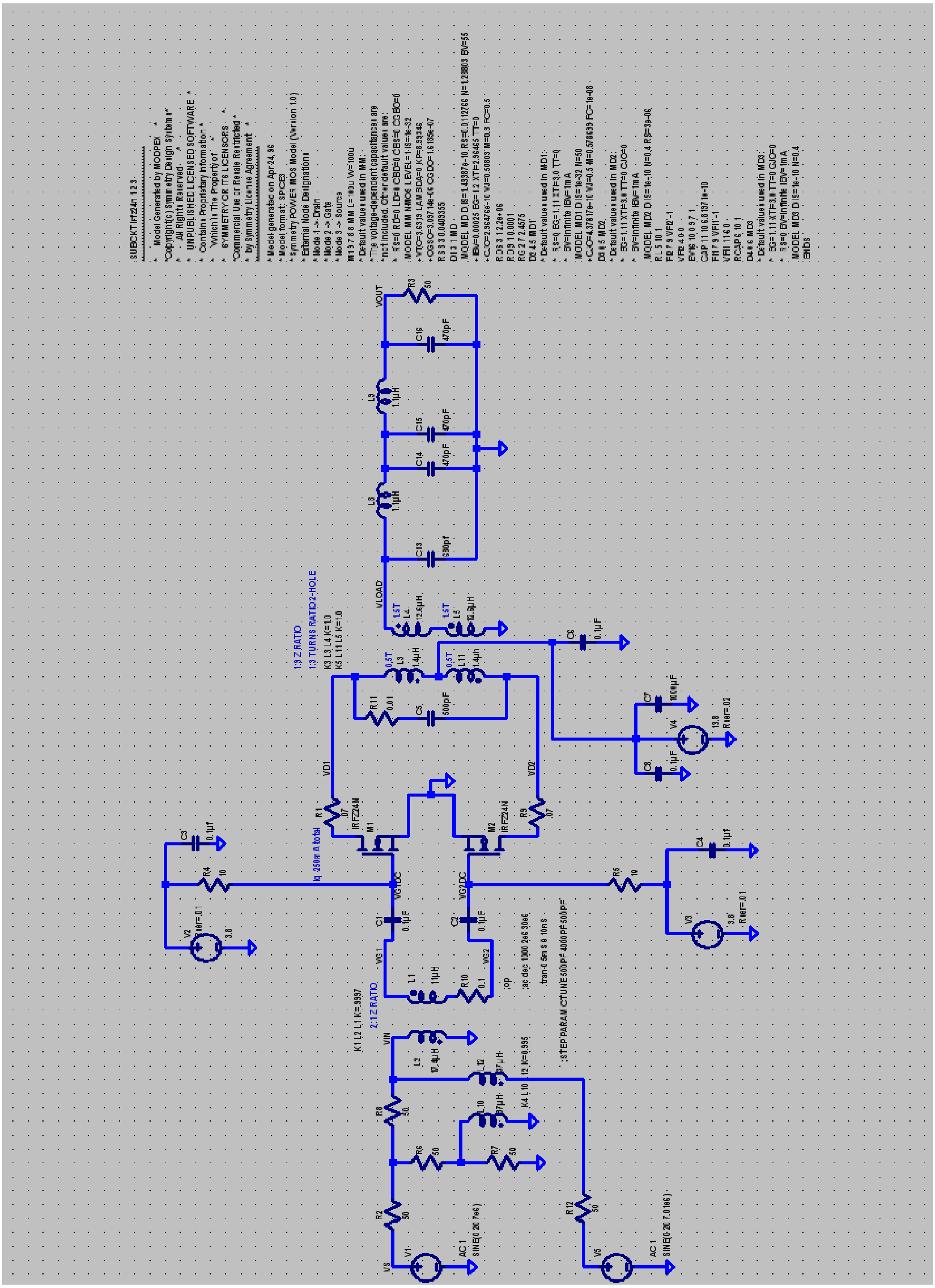
Biasing circuit temperature dependence simulation run using LTSpice, showing linear decrease of bias voltage with temperature. .



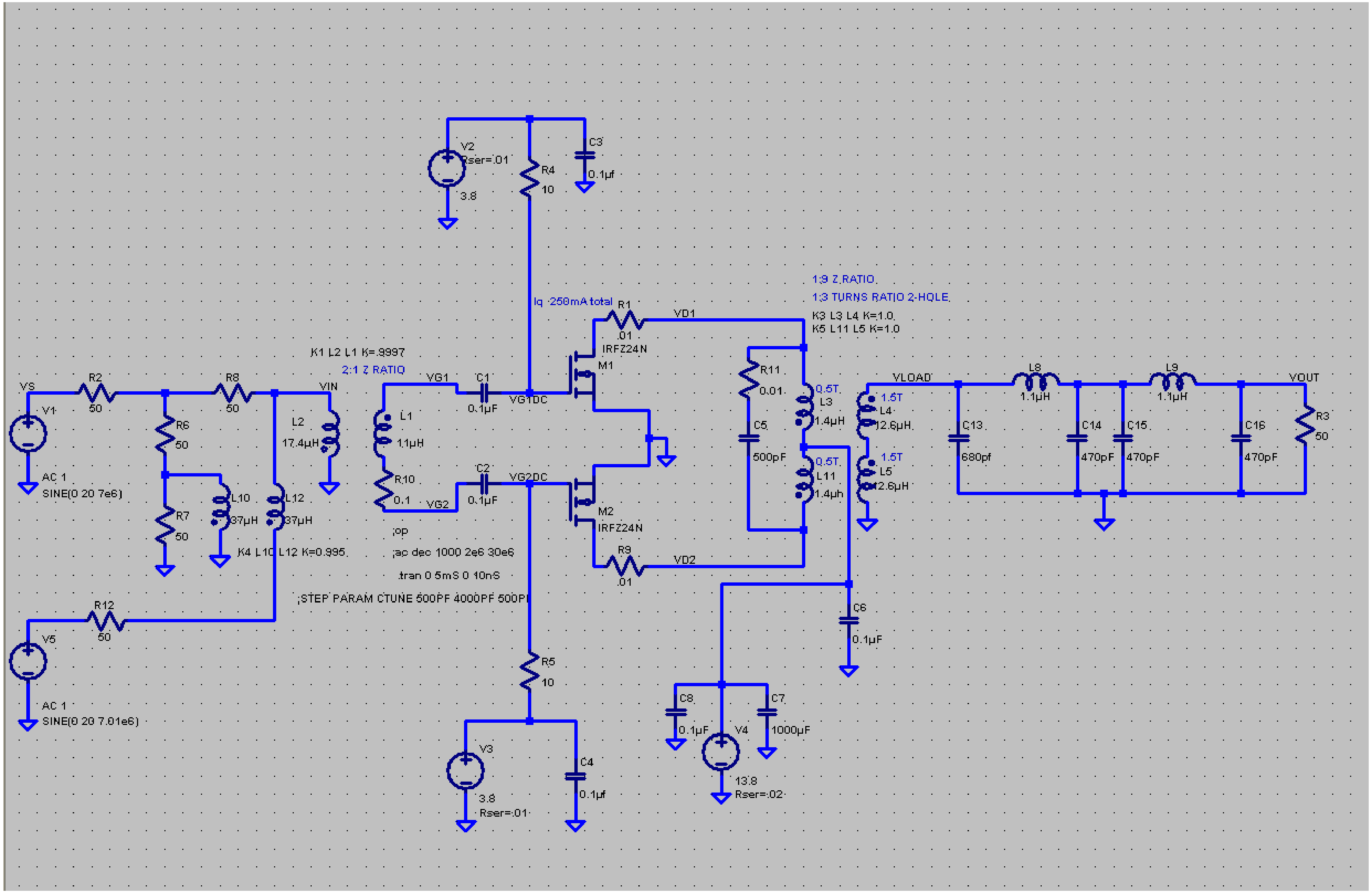
40M low-pass filter simulation run using LTSpice



Screen shot of entire amplifier simulation for IMD using LTSpice simulator. Simulation shows about minus 29 dBc IMD for this case. A fast fourier transform (FFT) function is available in LTSpice, making it easy to plot IMD directly. The two-tone time domain RF envelope is also plotted. Note the output transformer is modeled as two transformers in series, which it is, electrically. The Fast Fourier Transform (FFT) function in LTSpice can be used to determine the resulting frequency spectrum of any voltage or current in the circuit. In this case the FFT is of the output RF waveform and the plot has been scaled to show directly the resulting IMD products which are tightly clustered around the center frequency of the output RF envelope.



LTSpice schematic for amplifier simulation. Quiescent operating point, gain, frequency response, IMD, and temperature dependence are easily simulated using the free LTSpice simulation software downloadable (and updatable) from Linear Technologies' web site (www.linear.com) . Us homebrewers could not imagine having these tools, especially freely available, when I started homebrewing back in the mid 50's! The spice model text for the IRFZ24N mosfet is put directly onto the schematic as a text file and the mosfet symbols are associated with it. The mosfet spice model is available on the International Rectifier web site (www.irf.com) The simulation schematic includes at the input a homebrew discrete hybrid combiner circuit (see ARRL's "Experimental Methods in RF Design" book). It enables combining the two closely-spaced RF sources needed to generate the beat note pattern for IMD testing, without each source being affected (modulated) by the other.


VS
R2
50
V1
AC 1
SINE(0 20 7e6)
R6
50
R7
50
R8
50
L10
37μH
L12
37μH
K4 L10 L12 K=0.995
R12
50
V5
AC 1
SINE(0 20 7.01e6)
VIN
L2
17.4μH
K1 L2 L1 K=.9997
2:1 Z RATIO
L1
1.1μH
R10
0.1
VG1
VG2
C1
0.1μF
C2
0.1μF
VG1DC
VG2DC
;op
;ac dec 1000 2e6 30e6
.tran 0 5mS 0 10nS
;STEP PARAM CTUNE 500PF 4000PF 500PF
V2
Rser=.01
3.8
R4
10
C3
0.1μf
Iq 250mA total
R1
.01
IRFZ24N
M1
M2
IRFZ24N
R9
.01
VD1
VD2
R5
10
V3
3.8
Rser=.01
C4
0.1μf
1:9 Z RATIO
1:3 TURNS RATIO 2-HOLE
K3 L3 L4 K=1.0
K5 L11 L5 K=1.0
R11
0.01
C5
500pF
0.5T
L3
1.4μH
0.5T
L11
1.4μh
1.5T
L4
12.6μH
1.5T
L5
12.6μH
C6
0.1μF
C8
0.1μF
V4
13.8
Rser=.02
C7
1000μF
VLOAD
C13
680pf
L8
1.1μH
C14
470pF
C15
470pF
L9
1.1μH
C16
470pF
VOUT
R3
50

SOME USEFUL INTERNET PARTS SOURCES FOR HOMEBREWERS:

www.mouser.com
www.digikey.com
http://www.jameco.com/
http://kitsandparts.com/
http://www.kangaus.com/
http://cbjohn.com/aa0zz/index.html
http://www.farcircuits.net/
http://www.danssmallpartsandkits.net/
http://comtrolauto.com/
http://www.tubesandmore.com/
http://www.elexp.com/
http://www.cwsbytemark.com/mainpage.php
http://futurlec.com/index.shtml
http://www.mfjenterprises.com/Search.php
http://www.radiodaze.com/
http://www.goldmine-elec-products.com/departments.asp
http://www.cwsbytemark.com/mainpage.php
http://www.communication-concepts.com/
http://www.bgmicro.com/
http://www.aade.com/
http://www.allelectronics.com/
http://www.angelfire.com/electronic2/index1/
http://www.rfparts.com/
http://www.qrpkits.com/
http://www.oldradioparts.com/
http://www.rentron.com/
http://www.mtechnologies.com/index.html
http://www.sdr-kits.net/
http://www.oselectronics.com

SIDEBAR

Where do design equations come from? One should always try to understand where "formulas" come from, instead of just accepting the author's use of them. All formulas are based on assumptions and simplifications, which may or may not apply to the case at hand. We can easily derive the equation for calculating required drain-to-drain load resistance from supply voltage and desired output power of Class B or AB mosfet RF amp:

From our knowledge that for a push-pull transformer-coupled class B or AB amplifier, we know that for each mosfet the drain voltage excursion will be about twice the supply voltage (2Vcc). This is because each half of the primary swings from near zero (mosfet ON, opposite mosfet OFF), to near twice Vcc (mosfet OFF, opposite mosfet ON), centered around Vcc. And, both halves of the primary are in series and are 180 degrees out of phase (hence the term "push-pull"). Each mosfet drain sees a total swing of 2Vcc pk-pk, thus the total primary voltage swing is the *difference* between the two drain voltages, allowing for the fact that the lower drain voltage is always 180 degrees out of phase with the upper one (ie, it is always the *negative* value of the upper voltage). Now, with any sine wave, we know that we just divide its peak-to-peak value by 2.828, which is twice the square root of 2, to obtain its RMS value. We now have all the information we need to derive an equation relating the power output and the drain-to-drain primary load impedance.

Reiterating from above,

$$Vprimary_{rms} = \frac{Vprimarypk - pk}{2\sqrt{2}} = \frac{Vdrain1 - Vdrain1}{\sqrt{2}} = \frac{(2Vcc - (-2Vcc))}{\sqrt{2}} = \frac{(4Vcc)}{2\sqrt{2}} = \sqrt{2}Vcc$$

And, remembering from Ohm's Law how to calculate power in a resistor:

$$Pout = \frac{(Vout_{rns})^2}{Rout}$$

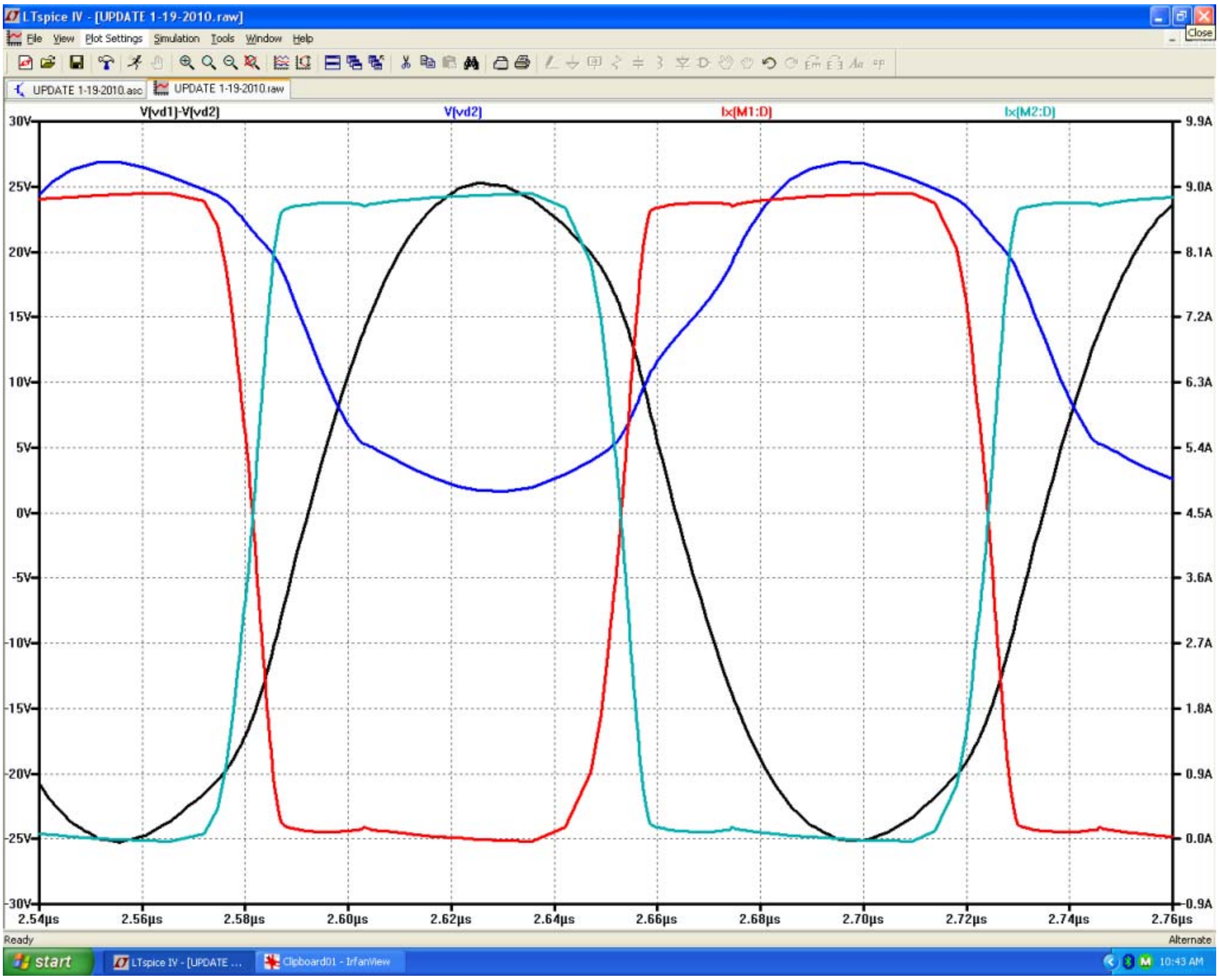
Now, substituting from above we get, $Pout = \frac{(\sqrt{2}Vcc)^2}{Rout} = \frac{(2)(Vcc)^2}{Rout}$

Starting to look familiar? Now, rearranging the equation to solve for Rout:

$Rout = \frac{(2)(Vcc)^2}{Pout}$ Voila! Now, that was not so hard, was it?

And now we know where that equation from the 2010 ARRL Handbook, page 17.35, comes from and we can use it with more confidence.

A really instructive way to verify the above and get a feel for how things work, is to use a circuit simulator, such as LTSpiceIV. Let's do it. Below is a transient response run of the entire amplifier circit. From it we can plot any voltage or current in the circuit, vs time:



The four waveforms plotted are the drain voltage and current for each mosfet, plus the total primary voltage end to end, which is V(Vd1)-V(Vd2) (black trace). We can see that the peak to peak total primary voltage is near four times Vcc, as we expected. Note that the primary voltage is approximately a sine wave, whereas the primary current is nearly a square wave. Circuit simulators are great design tools, but equally useful as learning tools. And since this one is FREE (and even has free online support via Linear Technologies' LTSpice User's Forum), why not use it? These waveforms are somewhat idealized however, as the simulation model of the mosfet does not take into account the voltage-dependent capacitances in the mosfets. Any simulation's accuracy is only as good as the models of the circuits and devices used in the simulation. These waveforms do not look quite so clean on the oscilloscope (reality). However, even a limited simulation is a very valuable design, verification, and learning tool, one which any homebrewer today should have in his toolbox.