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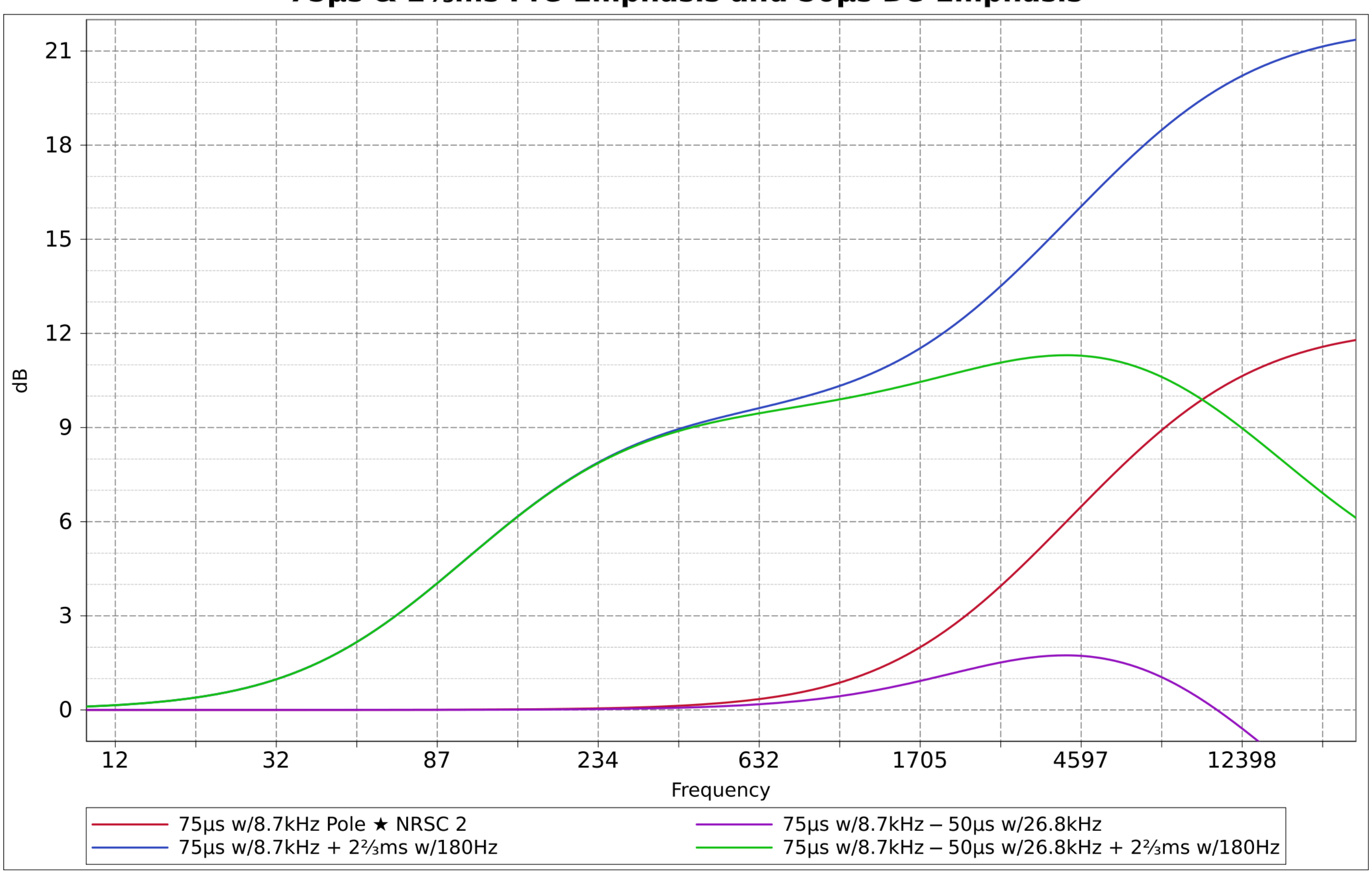
TX 100mW Mono 1610 AM

A Mono 100mW Transmitter on 1610kHz AM

This is presented to demonstrate what can be done to condition the modulating signal to maximize the signal quality during transmission and reception using a maximum of 100mW of input power into the final amplifier. (However, for live speech using an electronic voice compressor, or for music converting to mono, doing multi-band compression with a computer based sound program and storing it in a compressed audio file, and making minimal use of the compressor described here is the optimal arrangement.) A moderate and unique amount of pre-processing is used to take advantage of certain features and limitations of a short antenna (3 meters) and limited power (100mW). Some of it is conventional and found in use in commercial broadcasting, some is a modification of those features to optimize effectiveness for this application, and some are completely new. For **AM Stereo** implementing some of these features especially in non-linear systems like C-QUAM®, Kahn ISB®, or Magnavox PMX® require accurate math processing at the audio level for a clean signal and **DSP** is the only practical way to go. In Mono using a full analog path is fairly straightforward and relatively cheap using off the shelf components.

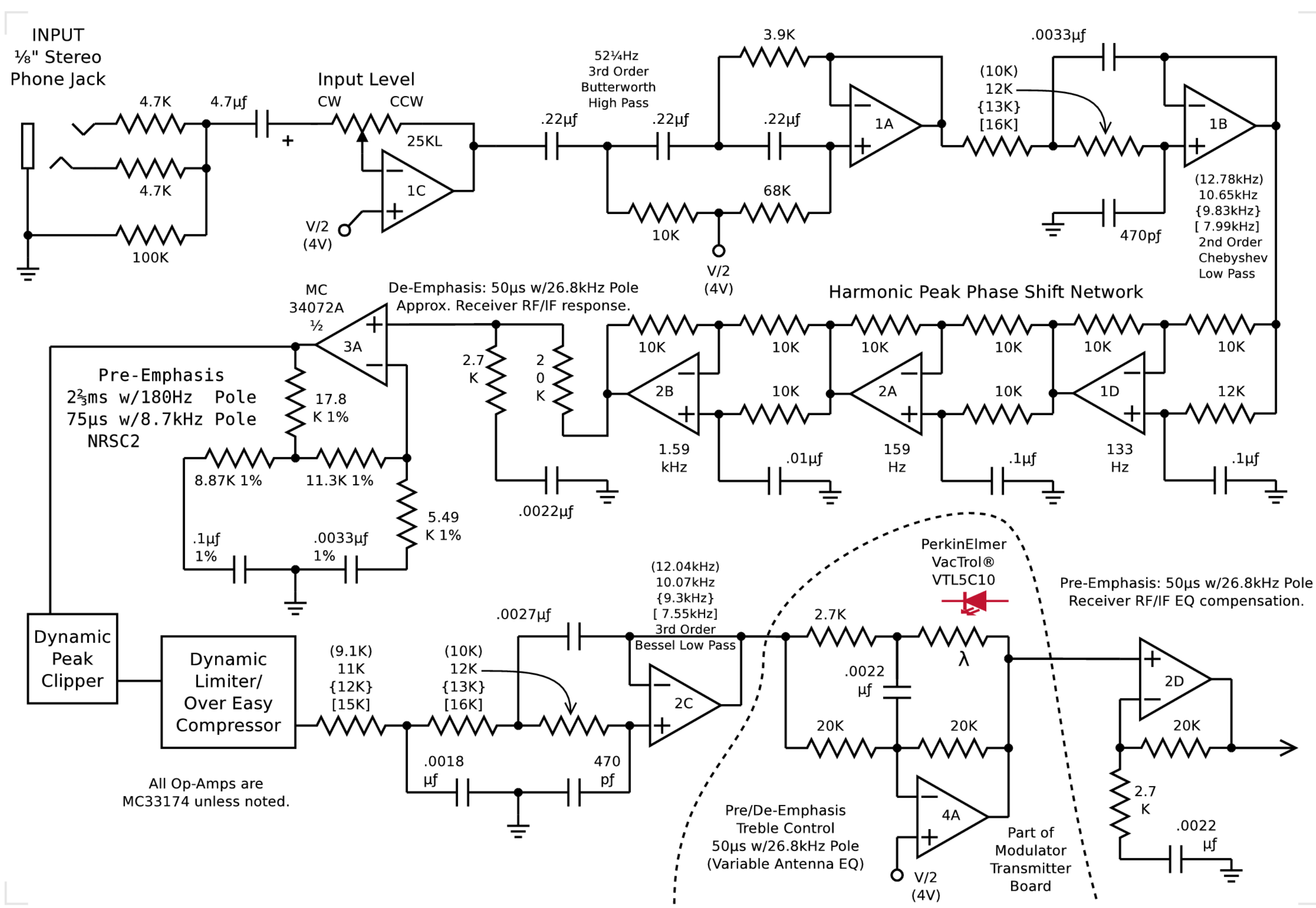
Pre-processing starts out with an adjustable input gain amp. Next is a 52¼Hz Butterworth 3rd Order High Pass Filter followed by a 10⅔kHz 2nd Order Chebyshev Low Pass Filter and then a Harmonic Peak **P**hase **S**hift **N**etwork which moves the harmonic peaks of of the peaks of their fundamentals reducing peak amplitude without clipping while maintaining loudness. Next is a 3.183kHz (50µs w/26.8kHz pole) De-Emphasis simulating the typical RF/IF receiver response prior to envelope detection. This will be restored after the Dynamic Peak Clipper and Limiter/Over easy Compressor has limited modulation to –100% for the frequencies below 3.183kHz (50µs) so the typical AM radio of limited bandwidth will remain fully envelope compatible. Part 15 Rules says nothing about exceeding –100% modulation that I know of and having the upper frequencies exceed this during

75µs & 2⅔ms Pre-Emphasis and 50µs De-Emphasis



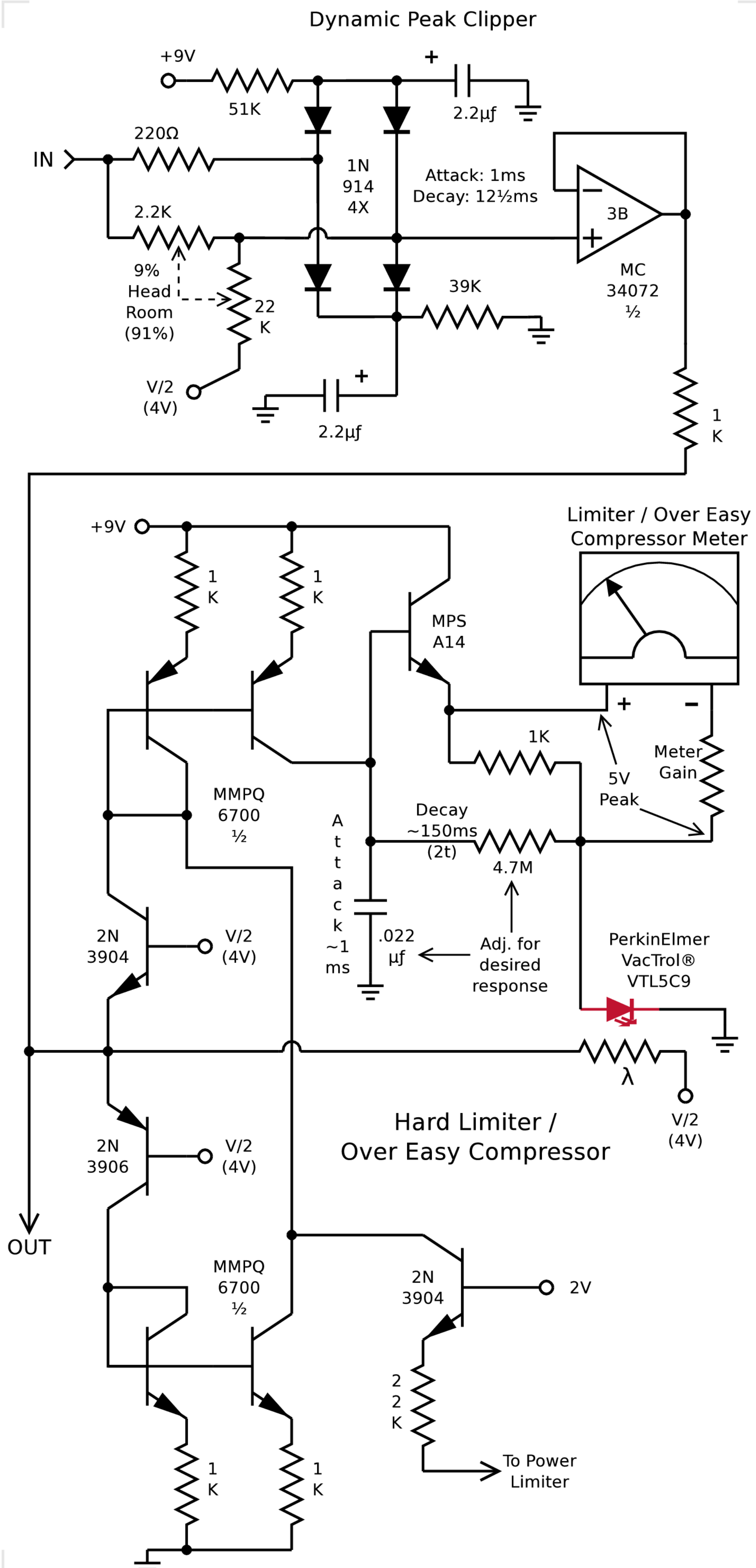
transmission and then attenuated by the receiver back down to ─100% or less will maintain compatibility while allowing a reduced carrier to be used placing more power in the sidebands. The graph above shows the standard 75µs w/8.7kHz Pole NRSC2 Pre-Emphasis (**Red** curve) along with a 2⅔ms w/180Hz (**Blue** curve with the **Red** curve combined) which reduces bass levels. In music bass program material is the strongest and this EQ reduces bass levels by ⅔ adding a average minimum headroom of 3dB for the higher frequencies with a potential up to 5½dB. This S/N loss for the bass frequencies is much less noticeable than the S/N loss in the higher frequencies and can be compensated for with the proper De-Emphasis EQ or turning up the bass control. The 75µs w/8.7kHz + 2⅔ms w/180Hz ─50µs w/26.8kHz (**Green** curve) is what the Clipper/Compressor will process and what the envelope detector in the receiver will mostly see. The **Purple** curve is the NRSC2 75µs w/8.7kHz and the RF/IF receiver response 50µs w/26.8kHz combined without the 2⅔ms w/180Hz bass reduction added. After Clipping and the Limiter/Compressor the signal peaks have been flattened generating harmonics that need to be removed prior to transmission. A 10kHz 3rd Order Bessel Low Pass Filter is used. The Bessel response has a maximally flat delay filtering out harmonics without producing overshoot. There are other resistor specifications for different cutoff frequencies for both the Chebyshev and Bessel filters and the combination of the two approximates a 5th Order Butterworth Low Pass response. The three other resistor choices provide high quality voice at 7¾kHz, a more compliant 9½kHz NRSC2 response, and Hi-Fi Music at 12⅖kHz. Depending on the majority of the type of program material transmitted this will dictate the choice of cutoff frequency. If mostly all voice use 7¾kHz or mostly all music use 12⅖kHz or a even combination of both use the 10⅓kHz response which is close to the NRSC2 10⅕kHz brick wall response. Next is the 3.183kHz (50µs w/26.8kHz) Pre-Emphasis response removing the RF/IF receiver filter De-Emphasis modeling.

Last is the dynamic antenna EQ treble control. In order to take advantage of the efficiency of a high Q antenna which are typical of vertical shorts a narrow bandwidth must be maintained but this



produces a low fidelity transmission. Reducing the Q to widen the bandwidth reduces transmitted signal strength so using a high Q narrow bandwidth antenna and EQing the modulating signal for a flat response is the way to go. In order for the transmitted signal to obtain the maximum –100% modulation response at the receivers envelope detector Pre-Emphasis for both antenna and RF/IF receiver response has been compensated for. This requires the signal modulator to exceed –100% modulation and a four quadrant modulator like a Gilbert Cell is needed. It is also beneficial to run the modulator in full suppressed carrier mode to obtain the best out of the modulator and use a two quadrant modulator to insert a carrier with the benefit of easy carrier level adjustment. Instead of using a pre made Gilbert Cell like the MC1496 one will be assembled from individual transistors for several reasons. The upper quad of the Gilbert Cell will be operated in linear mode not exceeding 27½mVp-p to eliminate harmonics so the same transistors need to be used for the carrier modulator. Using a balanced modulating signal input eliminates the need for the current sources in the 1496 that require additional voltage for proper operation and PNPs are used instead of NPNs since the output is connected to a center tapped choke near ground for biasing the driver transistors. NPNs could be used but an output transformer would be needed to isolate the supply and bias voltages and there will be core losses also.

To the upper right is the Dynamic Peak Clipper and Limiter/ Compressor. The Clipper is a compound one in that one resistor and two diodes independent of the

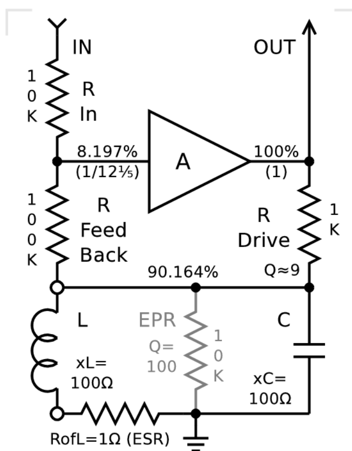


signal path are used to charge the capacitors defining the attack time and bleeders set the decay time. The signal path is a resistor divider which sets the adjustable headroom and uses the 2nd set of diodes to limit against the charged capacitors. The Limiter/Compressor input is a 1K resistor that clips the signal when it exceeds ~1⅛Vp-p by operation of the NPN & PNP emitters as their collectors transmit the clipping current through current mirrors to charge the .022µƒ capacitor in ~1ms while the 4.7MΩ resistor sets the 2t decay time of ~150ms. Selection of these two components allows the adjustment of attack and decay times. The capacitor sets the attack time and the capacitor and resistor combo sets the decay time. This control signal is buffered with a Darlington transistor that has a bias current similar to an Op-Amp so as not to affect the attack/decay times set by the capacitor/resistor combo too much. The emitter and its resistor controls the maximum current delivered to the LED within the Opto-Isolator containing the variable resistance CDS cell which shunts the signal reducing its amplitude until the signal is minimally clipped. The emitter resistor can be selected to adjust maximum attenuation levels. An equilibrium is obtained when the amount of clipping is enough to keep the capacitor charged and activate the CDS cell to attenuate the signal. The goal is to keep the signal ~1Vp-p maximum so the modulator will not exceed –100% for narrow band radios. The shortcomings of CDS cells are that their response times are less than stellar but they greatly simplify circuit design if they can be tuned well. The model chosen has the fastest attack and decay times and lowest on resistance. Setting the RC attack/decay times to somewhat slower will help to prevent overshoot. Finding the best RC combo with fast tracking with minimal overshoot and a decay time to minimize pumping effects will take some tuning of these components. A shorter decay time will produce a louder more compressed signal while a longer decay time will produce a quieter signal with a wider dynamic range.

The Oscillator (My Current Theory)

The best carrier oscillator is a quartz crystal. If you have one cut to the the desired frequency then use it. It's easy to build an oscillator with one that is reliable and very stable. That being said, and lacking one, needing to have a custom crystal cut, may not be cheap and turn around time may be slow. The other option is an LC oscillator but careful design is needed for good frequency stability that is low in amplitude and phase noise.

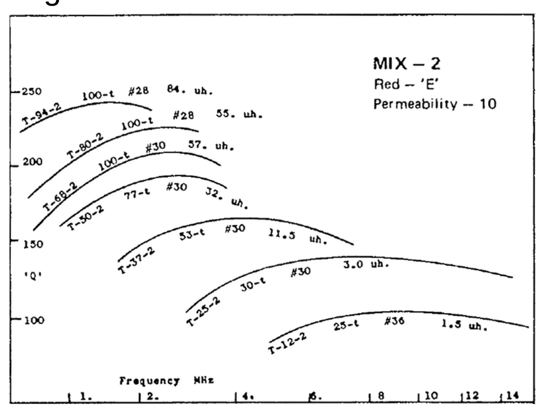
The Regenerative Receiver. It is hard to not talk about this when talking about oscillators because it is an oscillator when the regeneration gain is turned up too high. This type of receiver was popular before super-heterodyne was invented. Its operation was simple and positive feedback was used to increase the Q of the tank almost to the point of oscillation. Operating the regeneration control increased the positive feedback and the skill was to turn it up to to increase the Q and narrow the bandwidth to increase selectivity without going into oscillation. Turn it up too high and oscillation would occur. A blip was heard and the received signal was muted. To stop oscillation the control had to be turned down past the point where oscillation started for it to stop, having a hysteresis. It takes more feedback to start the oscillation that to maintain it. It was called regeneration because the signal passed through the band pass many times and each pass narrowed the response but another effect was also at play, negative resistance. In the diagram to the right the LC tank is assumed the be made with perfect theoretical components. An 1Ω resistor is placed in series with the inductor to represent the winding resistance. At resonance where xC=xL=100Ω the 1Ω ESR of the inductor converts to 10K EPR. This RLC tank has a static Q of 100 without the other components added. With A=0 making the output impedance of the amp 1K and the input grounded this represents the full static model. With all this positive resistance loading the LC tank it reduces the Q to ~9. As 'A' increases in value the 1K output impedance is being neutralized with negative resistance and at some point just before critical feedback it ceases to be a load on the tank and only the 10K EPR and the



Critical Feedback Model
Regeneration A<12⅕
Oscillator A≥12⅕

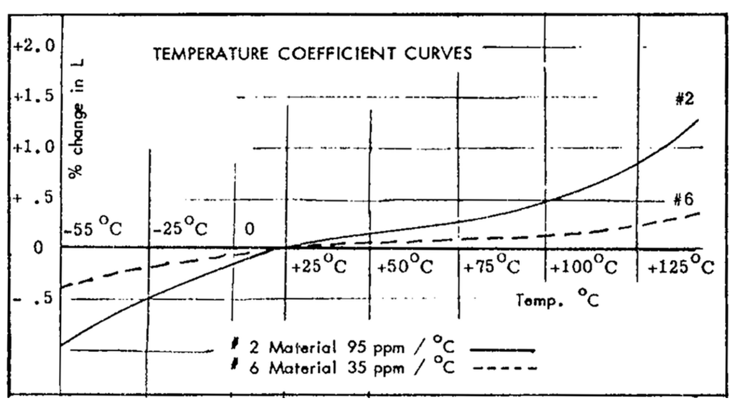
R In + R Feedback of 110K is keeping the Q below 100. As 'A' increases more towards critical feedback the 10K EPR and the 110K will also be neutralized with negative resistance and the Q will begin to rise above 100. In this theoretical model at the critical feedback point all load on the tank disappears, the Q of the LC tank is **∞** and so is its parallel resonance impedance. At the point that the 1K, EPR of 10K and 110K positive resistance loads are neutralized with negative resistance 'A' is inversely equal to the feedback reduction and is at unity gain. When 'A' rises just above unity gain oscillation starts and signal strength rises until the circuit limit is reached. At this point the oscillation can be sustained with less feedback which creates the hysteresis effect. With each pass the signal makes through the amp the signal grows in strength and eventually the output signal is larger than the amp can produce and the output is severely clipped and may resemble a square wave. Some crystal oscillators are made to run this way having the output level limited by the amp with little ill effect however this model does not optimize the benefits of maintaining a just over-unity feedback level for LC versions. Having too much negative resistance has the same detrimental effect on Q as positive resistance and reducing Q reduces the purity of oscillation. What is needed is the right amount of negative resistance over-unity feedback to maintain oscillation with the maximum possible Q. An **A**utomatic **L**evel **C**ontrol that maintains a just over unity feedback and maximized Q has to be designed into the oscillator.

Frequency stability is the other main factor in a good LC oscillator. Choosing high quality resonant components with high Q and combining them in proper balance to neutralize temperature coefficients is a must. The Q limiting factor of many LC oscillators is the inductor as the capacitors used usually have Qs in the 1000s. The kind of oscillator is also important for good stability, one with low phase and amplitude noise so the Clapp oscillator will be used. From some preliminary calculations the size of inductor needed is ~35µh. Fortunately Amidon has some example windings on their cores with the resultant inductance and Q. Using Mix #2 and in the chart to the right on a size 50 core there is a 32µh example with 77 turns of #30. At the desired operating frequency its approximate Q is 180. Adding 4 more turns should get it up to the desired inductance but the next smaller wire #31 is needed and will probably lower the Q by 10 to 170. The capacitors having a Q in the 1000s in combination with the inductor should have a estimated combined Q of ~150. In the analysis this value will be used to calculate the ESR of the inductor and the capacitors pure to simplify analysis.

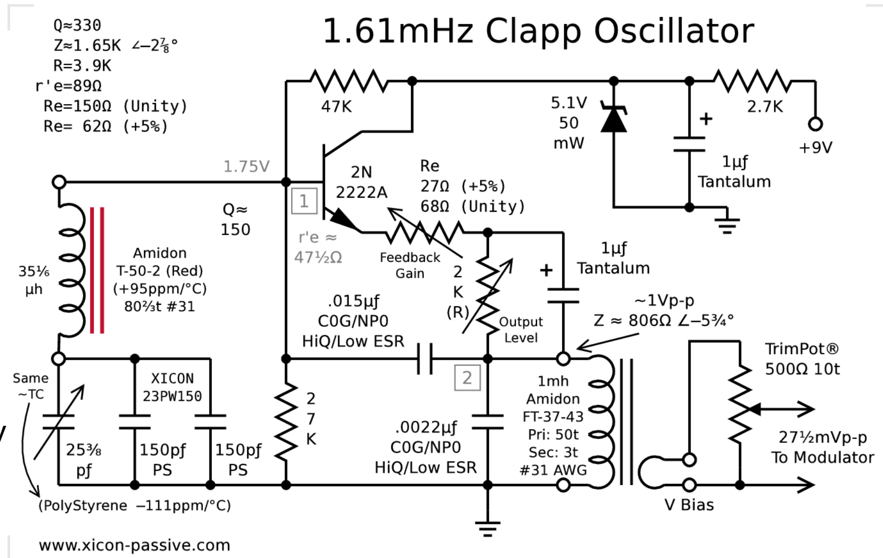


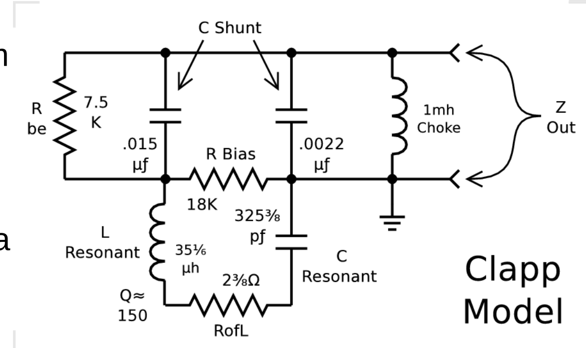
Matching the temperature coefficients to cancel out the drift requires that the total resonant capacitance has an equal but opposite coefficient of the inductor.

To the right the graph specifies Mix #2 having a +95ppm/°C. The curve is not straight and from 35°C to 75°C this straight line area extrapolates to +55ppm/°C. The slope increases around 25°C so the +95ppm/°C spec. is assumed to be for this temperature. At –25°C it is +125ppm/°C. For practical use a 20°C to 30°C operating range using +95ppm/°C is realistic. Finding a type of capacitor with with the exact same but opposite coefficient is slim. Usually a combination of capacitors with different



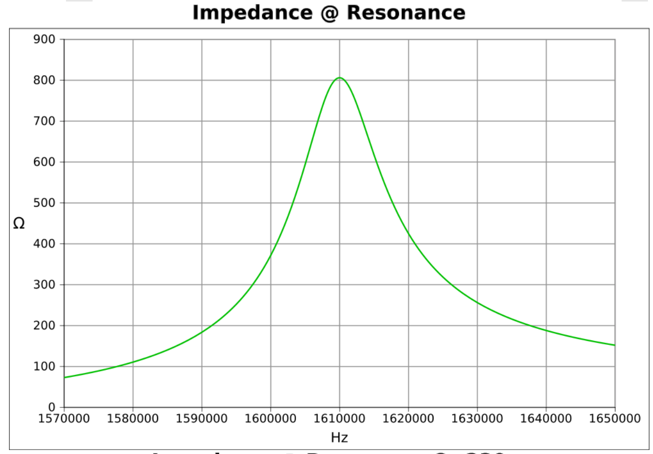
coefficients are combined in the proper ratio to match the inductor's coefficients. In this Clapp oscillator the resonant capacitance will be polystyrene. One source for for them is XICON and the temperature coefficient graph in the datasheet is a fairly straight line with a slight curve. There is a ½ % change over 45°C equating to $-111.\overline{1}$ppm/°C. To make this easy using C0G/NP0 (zero drift) types for the shunt capacitance and adjusting the ratio between the two types the polystyrene's coefficient can be reduced and balanced with the inductor's coefficient. 95/111.$\bar{1}$=85½, the polystyrene's reactance needs to be 85½% leaving 14½% for the shunt capacitance. 85½÷14½≈5.9:1 . The shunt capacitors of .015µ*f* & .0022µ*f* in series is 1918⅗p*f* and the needed resonant capacitance is 325⅜p*f*. All of these in series is 278⅕p*f* and the inductance needed for resonance is 35⅙µh. If the coefficient cancellation is kept to <5ppm/°C then the frequency drift should be kept to within ±20Hz over a ±5°C (68°F to 86°F) range for this frequency. If a wider temperature range is desired then the temperature coefficient of Mix #15 is +190ppm/°C and a straight line from 0°C to 125°C. Using a similar resonant to shunt capacitor ratio the resonant ones would need to be −225ppm/°C with the shunts being C0G/NP0 zero drift types.



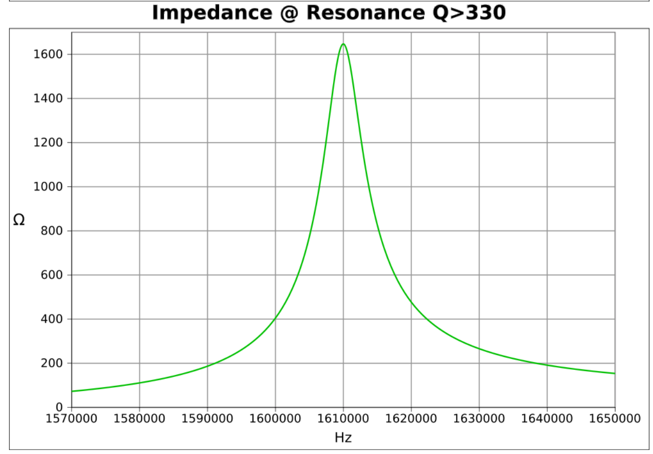


In the top right is the schematic for a Clapp oscillator using a bipolar transistor. If the parts arrangement looks somewhat familiar it is from the ARRL Handbook, referred to as a Colpitts, with some variations. This 1st version was the starting base to develop an analysis model and does not have an **ALC** so it is hard to predict exactly what the real output level will be. In order to determine the necessary amount of feedback the output impedance needs to be calculated and the analysis model to the right is used. To reduce the math complexity only the most critical components are part of this model. Using complex numbers in a spreadsheet and assigning the resistance and reactance to real and imaginary, (r,i) respectively, inductance +i and capacitance -i, and plugging the complex numbers into series and parallel resistance formulas the output impedance was determined over a frequency range and the static response is plotted in a graph to the upper right. It peaks at 806Ω. The shunt capacitance voltage divider between points 1 & 2 is 87⅕%. To have just over-unity gain feedback with the 806Ω output impedance Re + r'e will need to be <118Ω for oscillation to start.

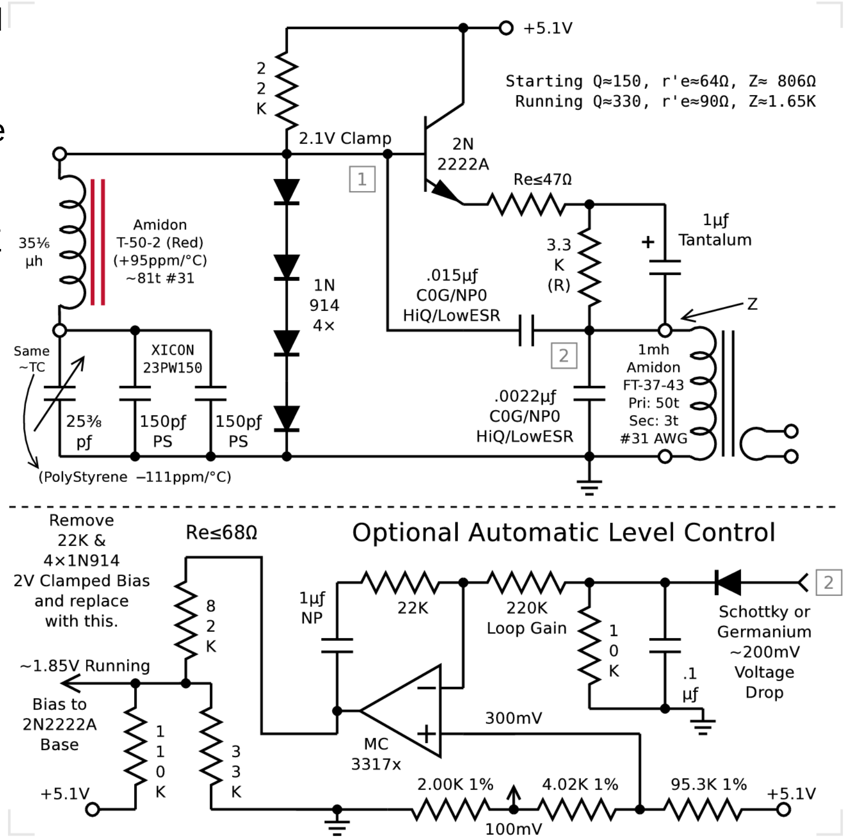
Impedance @ Resonance



Impedance @ Resonance Q>330

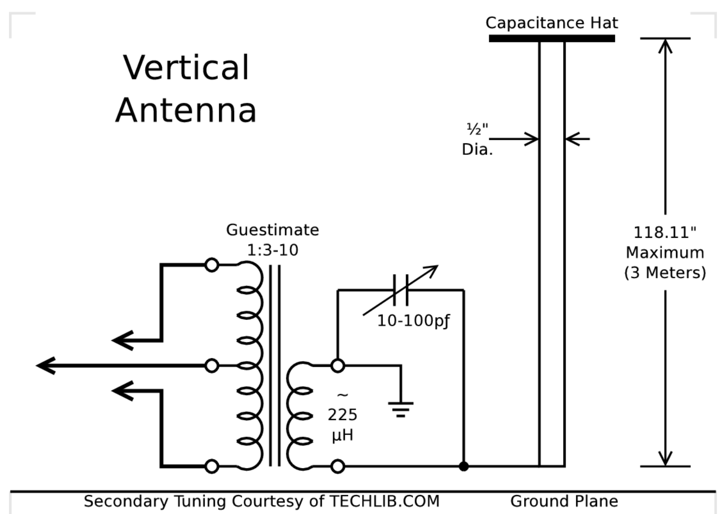


Once oscillation starts the feedback will produce negative resistance which will neutralize the positive resistance. In practical applications it is unrealistic to expect all positive resistance to be neutralized and obtain a Q of ∞ and Z of ∞. The Q limiting factor is the inductor. The minimal contributions to positive resistance are base loading of the transistor, biasing and ESR of the capacitors, and all will be mostly neutralized. The inductor winding resistance accounts for <25% of the Q limitations and will be mostly neutralized also. This leaves the core losses. To be conservative if ⅓ of the core losses can be neutralized and adding in the other neutralizing factors the Q would be above 330 with a Z of 1.65K. With this value of Z then for just over-unity gain Re + r'e only needs to be <241Ω but this does not account for the negative resistance feedback needed. This is not enough to start oscillation, hence the hysteresis effect.



Automatic **L**evel **C**ontrol. This is needed to maintain a constant output level and to keep the oscillator operating in an optimal range with just the right amount of negative resistance over-unity feedback to maximize Q. In the drawing to the right the bias is clamped. This is adapted from the J-FET version of this oscillator described in the ARRL Handbook. Given the differences in biasing of J-FET and bipolar transistors the layout is slightly different but the effect is the same. As the amplitude increases the bias level decreases because the clamped signal reduces the bias voltage. For both types of transistors the reduction in bias voltage reduces the idle current which in turn reduces transistor gain. Lower current also means lower output level for a given output impedance. For the J-FET it operates lower on the trans-conductance curve and for the bipolar r'e increases. For both types of transistors it is hard to determine what the output level will be but on start up the gain is high with enough feedback to start oscillation and as the output level increases the gain reduces to a point of equilibrium where there is just enough negative resistance over-unity feedback to sustain oscillation and operate at a near maximum Q. If tight control of signal output level is desired then in the 2nd ½ of this drawing a PLL style loop amp controlled by the rectified signal level is used to control bias voltage level which effects feedback level. This may not operate at the equilibrium point so at the controlled output level the feedback level may not be optimized. Once output level is set adjustment of Re can be used to obtain the just over-unity feedback to maximize Q.

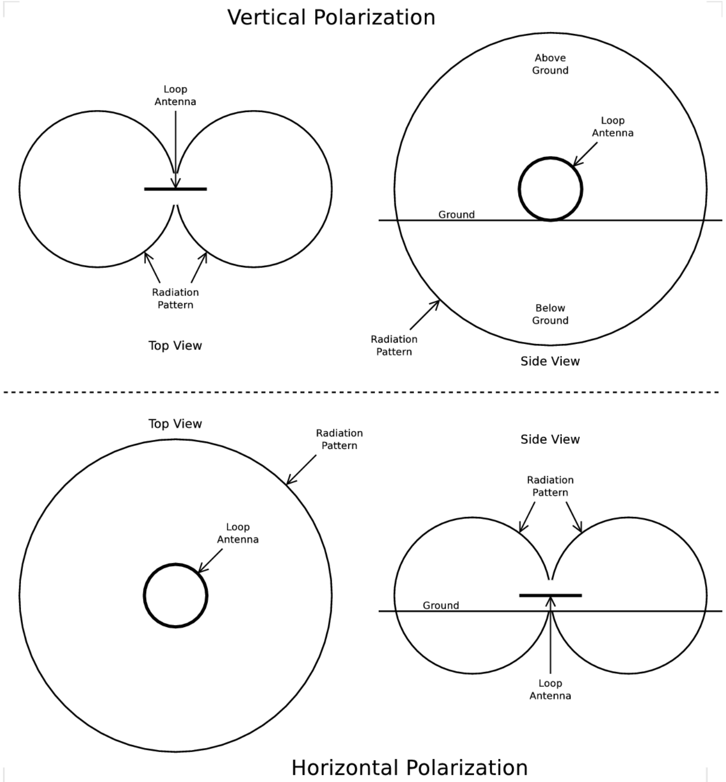
The Antenna



The typical antenna used for 100mW transmission is a short vertical with a loading coil no longer than 3 meters. A good ground plane is needed to maximize load on the amplifier and reach the 100mW limit of input power. The antenna needs the ground plane for reflection to reduce its load impedance. This involves planting enough radials in the ground especially if soil conductivity is poor. If you are lucky to live in a salt marsh and few radials produce excellent ground conductivity then this is about the best situation you can have. Or if there is a big box store that has a corrugated metal support system underneath the roofing material where you may set your transmitter in the middle of it then that would be the 'Bee's Knees'. Under situations like these it is not uncommon to have a usable signal travel 1¼ miles. That being said these situations are hard to obtain and maintain. A change in weather could affect ground conductivity and thus radiation efficiency if not enough radials are in place. A change from wet to dry when the antenna was tuned for proper loading after a rain and the dry season lowered the radiation efficiency then the final amp would not have enough loading and the signal could be clipped. Or vice-versa where the antenna was tuned for proper loading during the dry season, when the wet season came around the radiation efficiency would increase and overload the final amp and may cause it to draw more than the permitted 100mW of input power. - - - - - For safety it is a good idea to ground the transmitter through a 100K resistor or a 10mh choke to bleed off any static electricity that might accumulate from the high voltage at the top of the antenna. You don't want a Van de Graaff generator that could shock someone or damage equipment.

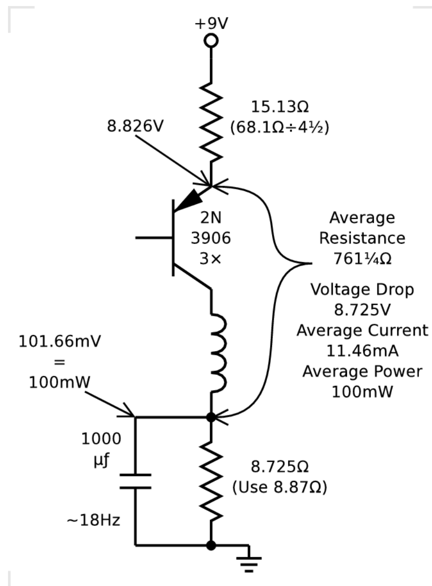
In balancing power output with the antenna's load resistance in the image above the proper amount of inductance in the transformer's secondary is needed for the loading coil. Using the output transformer's secondary to double as the loading coil is more efficient and transfers power better. To cover the whole AM band 2 to 3 different inductance taps in the secondary along with the 10-100pf trimmer should allow the antenna to be tuned to resonance. After establishing the ground plane and determining the load resistance the proper primary to secondary impedance match must be made which also includes the supply voltage to the finals in the calculations to meet the 100mW power input limit. The class AB amp is the most efficient for linear amplification and the primary of the output transformer is center-tapped for this type. Although an auto-transformer is more efficient it would be a difficult setup for this type of class AB amp.

These issues with a vertical short antenna make it difficult to adapt to the goals of this document. Over on TECHLIB.COM's Personal Radio Station page the use of a loop antenna is described. As the article mentioned its radiation is magnetic and unlike a vertical it cuts through solid objects pretty well, up to the outer limits of the coverage area and nothing much affects its resonance or radiation pattern, well, unless you stick something magnetic in the middle of it like a large piece of ferrite ;-). It does not require a ground plane and small loops are usually considered poor radiators but in this application it is well suited. In the picture the antenna is positioned for vertical polarization and its radiation pattern is similar to a dipole but perpendicular with the two lobes aligned with the plane of the antenna. Perpendicular to the plane of the loop are the null points. The radiation pattern resembles a Static Warp Bubble with the radiation extending up into space as far as it does on each of the lobes. It also extends below ground but depending on soil composition the signal may extend further below ground or be absorbed. If you're not concerned with the ground wave using horizontal polarization may be a better option. Consider this, in space (not time) your signal will be in quadrature, and your signal will be the strongest at all other signal's null point that are vertically polarized. The radiation pattern will also be omni-directional. Your signal's null point will also be vertically polarized greatly reducing the potential for interference to licensed broadcasters. It's like your signal is radiated on a whole other plane of existence. You control the horizontal, they



control the vertical ;-). The only requirement for reception is that the receiver's ferrite antenna inside the radio must be positioned vertically instead of the normal horizontal position. This means placing the radio on on its side in most cases. This also nulls out any signals received that are vertically polarized. The loop antenna pictured and described on the page is made out of ½" copper pipe, is a square 30"×30" O.D. and the inductance he has estimated to be is 2µh but **EEWeb** says 2¼µh, although using 0.889 premeability will produce 2µh. It is usually assumed that for air core inductors the permeability is 1 but under certain circumstances it may be less. With the circle being the shape of the most efficient loop antenna, maximum area for minimum circumference, this is the shape to be used. For bendability using ⅜" copper pipe makes it easy to form. After using the online calculator on **EEWeb** to determine the inductance of a round loop made with a ⅜" copper pipe with a 3 meter circumference the result is 2.64µh, or 2.81µh for ¼" pipe, so this is the value that will be used although it may actually be a little different but it is close to inductance of the square version so some adjustments can be made to accommodate resonance. Using a permeability of 0.889 then the inductance for ⅜" and ¼" pipe would be 2.35µh and 2½µh respectively. These inductance values are ballpark figures so actual inductance may not faithfully reflect the formula calculations.

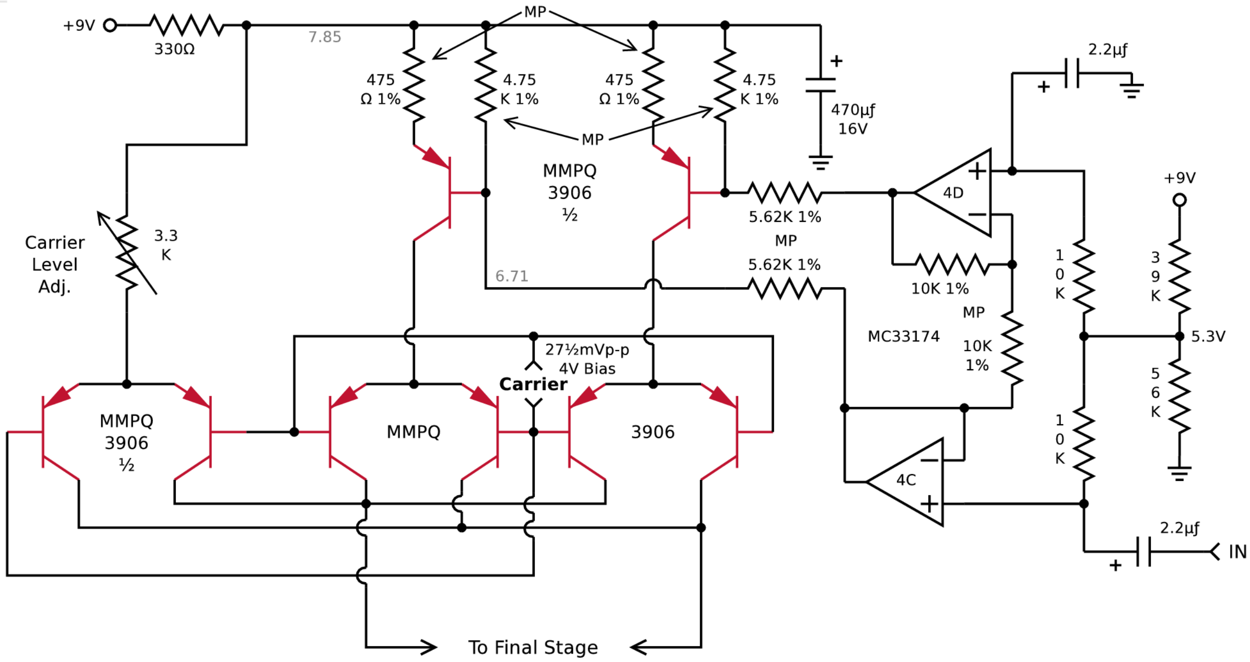
The 100mW Final Input Power Limit

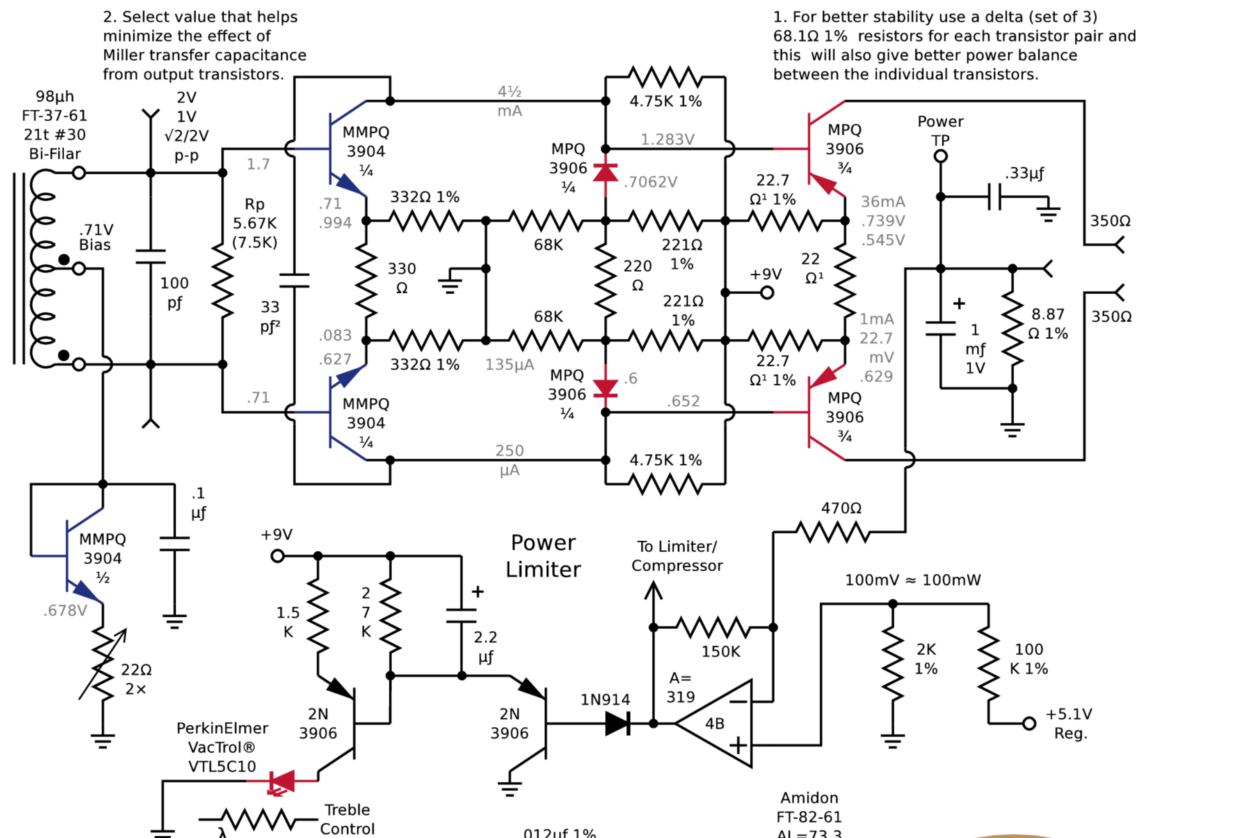


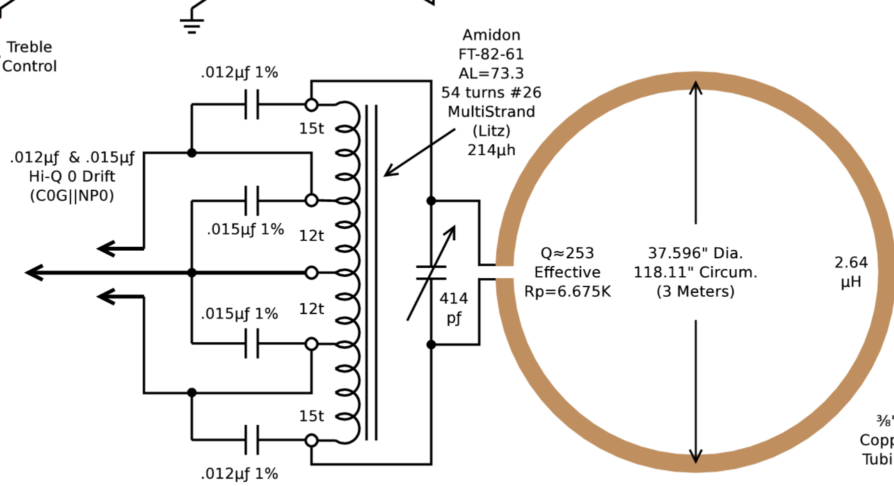
The Final will operate from a 9V supply. Minus the voltage dropped across the resistors the input power is calculated from the voltage drop across the antenna load and output transistors. To calculate the amount of load needed to reach the 100mW limit the power divided by the voltage will define the average current allowed, 100mW÷8.725V≈ 11.46mA, making average load resistance 8.725V÷11.46mA≈761¼Ω. A Sine wave's average voltage is 2/π or 0.6366 so to get the same amount of current to flow the resistive load would need to be 2×761¼Ω÷π≈484⅝Ω. An unmodulated carrier is ½ the amplitude of the carrier during +100% modulation so the load resistance would need to be ½ of this, 484⅝Ω÷2≈242⅓Ω. For reasons described and clarified a reduced carrier will be used. The current level of the unmodulated carrier will be reduced by 3dB and will use 50mW of input power. This requires the antenna load resistance to be increased by √2, √2×242⅓Ω≈**342⅔Ω**. The loop antenna has a very high Q and at 1.61mHz and a ±3.183kHz (±50µs) bandwidth this makes the Q 252.9 . The inductance of the loop is 2.64µh, the step up output auto-transformer is 213¾µh and in parallel is 2.608µh. The **E**ffective **P**arallel **R**esistance for the desired bandwidth is 6675Ω. If the Q is greater then De-Q it with a resistor. If it's less and can't be increased then determine to Q and calculate bandwidth/2 for the audio corner frequency for the Variable Antenna EQ Treble Control. If this Q is a dramatic reduction from the desired Q then the finals have the potential to draw more power than and permitted. For maximum efficiency the proper tap ratio will need to be determined for best input power to output radiated energy and components properly selected. The voltage step up from the final outputs to the loop antenna is 1:4½. It is defined by the .015µf & .012µf capacitive divider and the auto-transformer is tapped at the same ratio. Just the capacitive divider could have been used for the step up alone but a center-tapped choke is still needed to supply the power to the transistors so why not also use it to do some of the lifting as an auto-transformer. With the 1:4½ voltage step up the impedance step down to the output transistors is 20¼:1 making the load on the transistors **329⅗Ω**, close enough.

The Modulator & Final Output

The modulator is a Gilbert Cell running in suppressed carrier mode. There is a 2^{nd} modulator to insert the carrier. The signal modulator is operated to ±50% of clipping level for 100% modulation for frequencies below 3.183kHz (50µs) leaving a 6dB headroom. Calibration: Place a center tapped dummy load of ~100µh transformer primary on the final output with a 1.4K connected p-p (that's c-c in







1:4.4	
.020µf	12t
.024µf	10t
.024µf	10t
.020µf	12t
5455pf	142µh

1:4.$\bar{4}$	
.018µf	11t
.022µf	9t
.022µf	9t
.018µf	11t
4950pf	117µh

1:4½	
.016µf	10t
.020µf	8t
.020µf	8t
.016µf	10t
4444pf	95µh

1:4.4	
.015µf	18t
.018µf	15t
.018µf	15t
.015µf	18t
4091pf	319µh

1:4.46	
.013µf	16t
.016µf	13t
.016µf	13t
.013µf	16t
3586pf	247µh

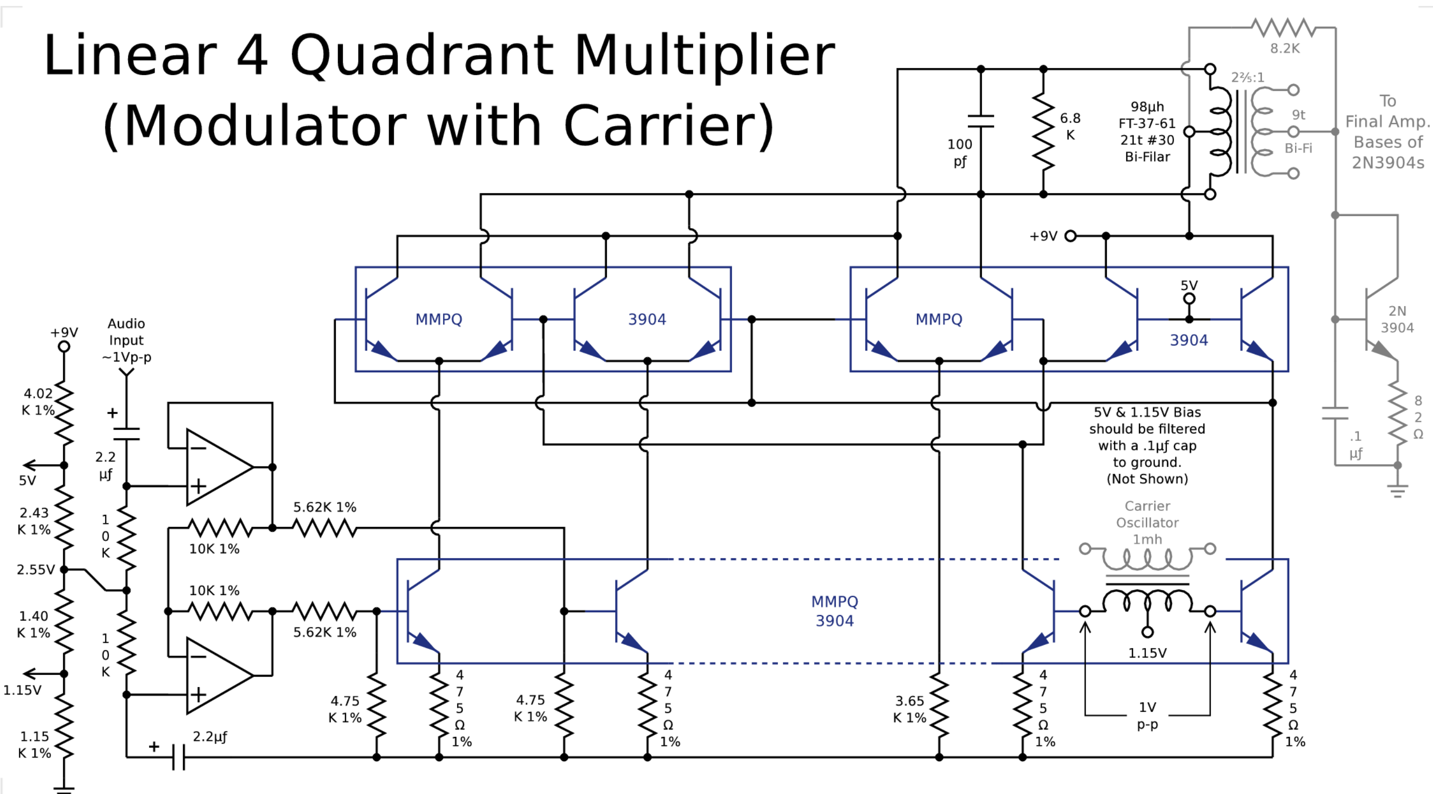
1:4.36	
.011µf	13t
.013µf	11t
.013µf	11t
.011µf	13t
2979pf	169µh

transistor speak). Add a few turns to the secondary for a scope signal. Insert a 500hz tone into the input at a level where the compressor is engaged and limiting the amplitude. Vary the input level to see if the compressed signal varies. Turn up the input to the point of clipping by the voltage swing limitations of the prior input stages and then back off some till no clipping occurs. This is probably the maximum level the compressor will see. Observe the output of the modulator to see if modulation is at ─100%. If stronger than ─100% then decrease the 3.3K tail resistor in value on the carrier modulator until ─100% is achieved. If weaker than ─100% increase the 3.3K in value. Remove input signal for just the unmodulated carrier. Place a 1K or less resistor across the Rp 5.67K(7.5K) resistor to reduce input signal level to a minimum. Adjust for minimum bias level until almost all of the crossover distortion disappears on the scope via the two 22Ω bias control resistors. The less bias the more power available for output signal. Once set remove the 1K from Rp 5.67K(7.5K). Adjust the unmodulated carrier level until 77mV is obtained on the 8.87Ω power level sensing resistor via Rp 5.67K(7.5K) value. (If there is any clipping at ±100% modulation for frequencies below 3.183kHz (50µs) the antenna load may not be enough and re-adjusting the level from 70mV down to 65mV may be necessary.) There is 23mV(40mW) of headroom remaining for frequencies above 3.183kHz (50µs). In this 23mV range is where the treble control operates. The 319 gain on the power sensing amp will cover the range from no cut to full cut on the treble control. In the last ¼ of its range a current signal is also sent to the compressor to ensure that 100mV(100mW) is not exceeded. With most music material the high frequencies are of minimal amplitude compared to the rest of the signal so the treble control may see only moderate action. The wider audio bandwidth of 12⅖kHz will produce more treble control action while the 7¾kHz bandwidth will produce less. A signal source with strong treble content could see a lot of cut and produce a low fidelity sound but that is the limitation of keeping the power input under 100mW.

The final output transistors are quad 2N3906 (Ceramic) DIP packages, 1 chip for each ½ of the swing of the signal. One transistor in each chip is connected as a super diode while the other 3 handle ⅓ of the current per ½ of the swing. The super diode along with the other 3 sets up a current mirror as a 1:10 multiplier. Distributing the power between the 3 transistors allows the operation of them in their highly linear range. Transistor gain is a fairly straight line with a slight increase in gain as current increases for the range used but for current levels above this range the gain starts to drop off rapidly and the output admittance changes drastically at Vce<400mV. The emitter resistors are setup in a delta configuration. This allows a larger amount of resistance at zero crossing for better bias control but on each ½ if the swing the resistance drops to ⅔ providing less voltage drop and a little more headroom. It also smooths the zero crossing transition some allowing for a lower bias current. While the delta arrangement using 22.7Ω resistors in the schematic is for all 3 push-pull transistor pairs in application these are 68.1Ω resistors in a delta for each push-pull transistor pair, 3 deltas for 3 sets of push-pull transistors for a total of nine 68.1Ω resistors. While this final output design work done has been to optimize it for a highly linear output current to drive the antenna for a low distortion Hi-Fi sound not much work has been done on stabilization except for the suggested 33pf capacitor to reduce miller feedback issues. Additional capacitors are probably needed to eliminate other parasitic issues. As it is it may preform OK but there will probably be performance issues that will need to be tuned out.

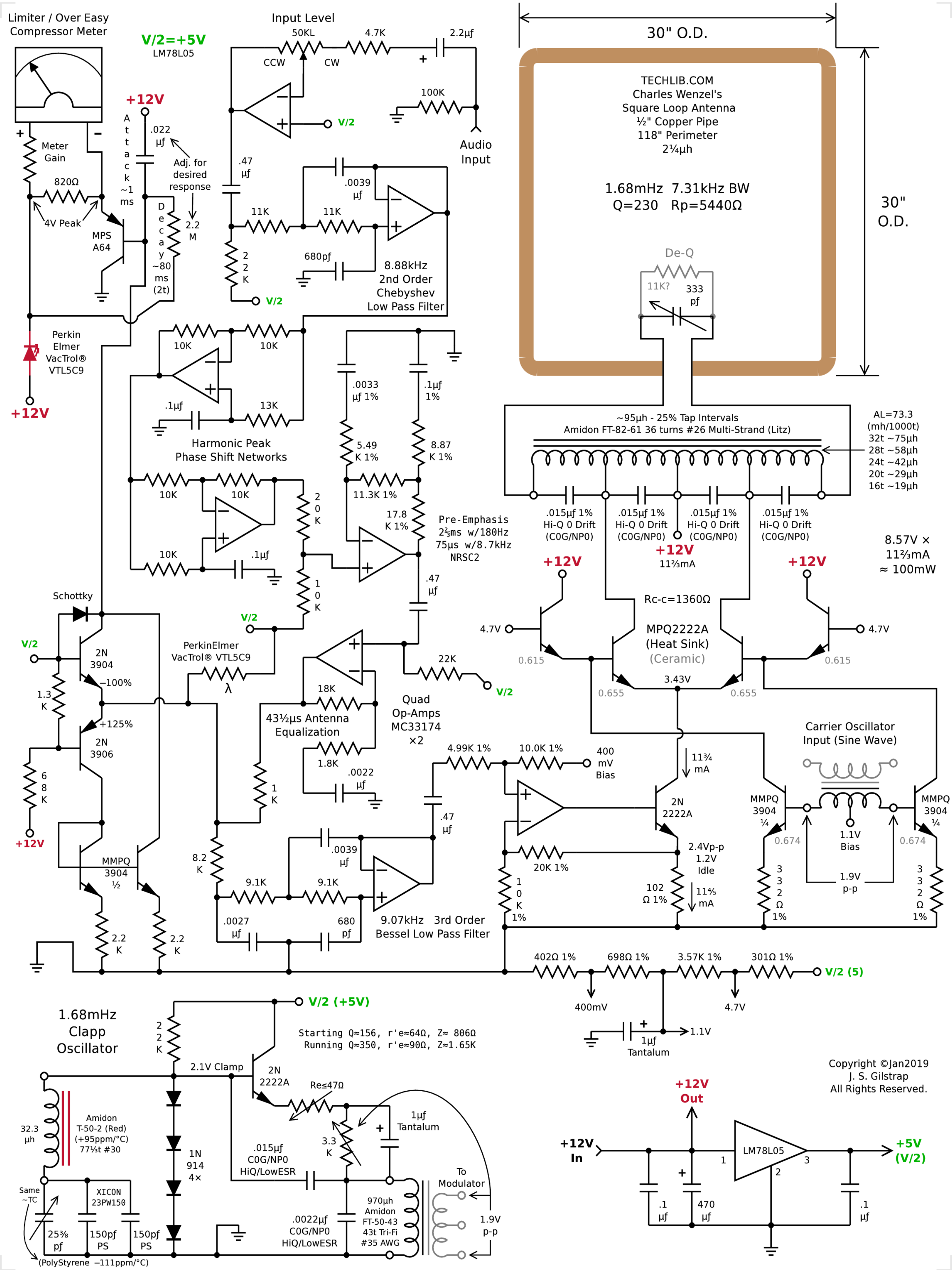
NPN Modulator

On the next page is an NPN version of the PNP 4 Quadrant Multiplier Reduced Carrier Modulator at the top of page 12. This NPN version uses 2 of the transistors as linearizing diodes to bias the other 6 transistors in the upper half of the Gilbert Cell. They are supplied with current from 2 transistors in the bottom half identical to the setup for the audio signal input and driven by the oscillator signal. This makes it a true Linear 4 Quadrant Multiplier (±200% modulation peak with the carrier at 100%) allowing for a much stronger oscillator signal input than 27½mVp-p. A different oscillator output arrangement will be needed, one like the setup for the Square Loop 'Lite' version at the end of this document. The oscillator input is driven to 50% of current peak providing a stronger output allowing a lower output impedance defined by the 6.8K on the transformer's primary which is stepped down 2⅖:1 to drive the NPNs input to the final amp. If this version is used then omit all parts on page 12 connected to the bases of the 2N3904s and use the suggested layout in the NPN version. Bias of the idle current for the MPQ3906 outputs is adjusted via the ~82Ω resistor. If the gain into the 2N3904 drivers is too strong then placing a resistor across the transformer's primary or secondary can reduce it. If too weak reducing the delta resistor values connected to the 2N3904 emitters will increase gain. With these two adjustments this should be capable of setting the unmodulated carrier level to draw 70% of maximum current, ~70mV measured across the 8.87Ω resistor. While this NPN version is more robust and provides a stronger quality signal and better control over the 2N3904 driver transistors it does draw 4½mA more than its PNP version. If powered by a regulated 9V power supply or a larger battery like a 1½V×6AA battery caddie then this really isn't an issue. If powered from a single 9V battery then the PNP version may be better.



Square Loop 'Lite'

This version on the next page has most of the pre-processing but doesn't exceed –100% modulation with its 2 quadrant modulator using a Class 'A' Push-Pull setup. A 5^{th} Order Low Pass Filter, –3dB @ 8¼kHz. Good for HQ Voice and/or Medium Low Fidelity Music. +12V operation, 1½V×8AA battery caddie. Loop is at +12V potential so supplying power from a car's 12V system is not advised unless the loop is well insulated. Having the un-insulated loop touch the body of the car may cause the transmitter to release blue smoke and noxious odors. Once released the transmitter is Fried. If supplying power from a car it is highly recommended that an isolated and regulated +12V supply be used.



Limiter / Over Easy Compressor Meter
V/2=+5V
LM78L05
Input Level
50KL
4.7K
2.2µf
CCW
CW
100K
Audio Input
V/2
Meter Gain
820Ω
4V Peak
MPS A64
+12V
Attack ~1 ms
.022 µf
Adj. for desired response
Decay ~80 ms (2t)
2.2 M
.47 µf
11K
11K
.0039 µf
680pf
2 2 K
8.88kHz 2nd Order Chebyshev Low Pass Filter
Perkin Elmer VacTrol® VTL5C9
10K
10K
13K
.1µf
Harmonic Peak Phase Shift Networks
.0033 µf 1%
.1µf 1%
5.49 K 1%
8.87 K 1%
11.3K 1%
17.8 K 1%
Pre-Emphasis
2⅔ms w/180Hz
75µs w/8.7kHz
NRSC2
20K
10K
.47 µf
Schottky
2N 3904
−100%
+125%
2N 3906
1.3 K
68K
MMPQ 3904 ½
2.2 K
2.2 K
PerkinElmer VacTrol® VTL5C9
λ
18K
43½µs Antenna Equalization
1.8K
.0022 µf
22K
Quad Op-Amps MC33174 ×2
1 K
8.2 K
9.1K
9.1K
.0039 µf
.0027 µf
680 pf
9.07kHz 3rd Order Bessel Low Pass Filter
30" O.D.
TECHLIB.COM
Charles Wenzel's
Square Loop Antenna
½" Copper Pipe
118" Perimeter
2¼µh
1.68mHz 7.31kHz BW
Q=230 Rp=5440Ω
De-Q
11K?
333 pf
~95µh - 25% Tap Intervals
Amidon FT-82-61 36 turns #26 Multi-Strand (Litz)
AL=73.3 (mh/1000t)
32t ~75µh
28t ~58µh
24t ~42µh
20t ~29µh
16t ~19µh
.015µf 1% Hi-Q 0 Drift (C0G/NP0)
11⅔mA
8.57V × 11⅔mA ≈ 100mW
Rc-c=1360Ω
MPQ2222A (Heat Sink) (Ceramic)
4.7V
0.615
0.655
3.43V
Carrier Oscillator Input (Sine Wave)
4.99K 1%
10.0K 1%
400 mV Bias
11¾ mA
2N 2222A
MMPQ 3904 ¼
1.1V Bias
1.9V p-p
0.674
2.4Vp-p 1.2V Idle
20K 1%
10K 1%
102 Ω 1%
11⅘ mA
332 Ω 1%
402Ω 1%
698Ω 1%
3.57K 1%
301Ω 1%
V/2 (5)
400mV
4.7V
1.1V
1µf Tantalum
1.68mHz Clapp Oscillator
V/2 (+5V)
Starting Q≈156, r'e≈64Ω, Z≈ 806Ω
Running Q≈350, r'e≈90Ω, Z≈1.65K
2.1V Clamp
2N 2222A
Re≤47Ω
1µf Tantalum
3.3 K
32.3 µh
Amidon T-50-2 (Red) (+95ppm/°C) 77⅓t #30
1N 914 4×
.015µf C0G/NP0 HiQ/LowESR
.0022µf C0G/NP0 HiQ/LowESR
970µh Amidon FT-50-43 43t Tri-Fi #35 AWG
To Modulator
1.9V p-p
Same ~TC
XICON 23PW150
25⅜ pf
150pf PS
(PolyStyrene −111ppm/°C)
+12V Out
+12V In
LM78L05
+5V (V/2)
.1 µf
470 µf
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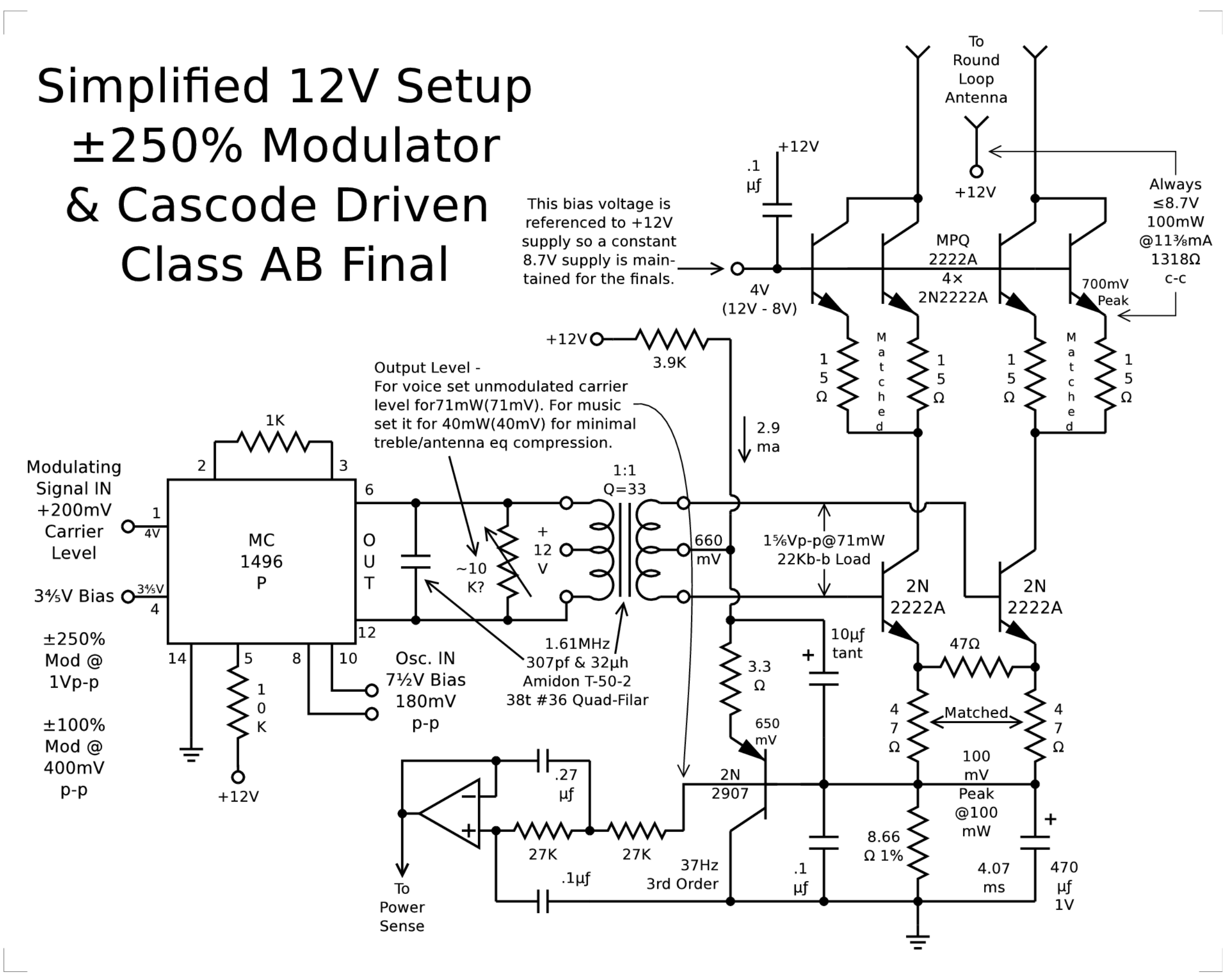
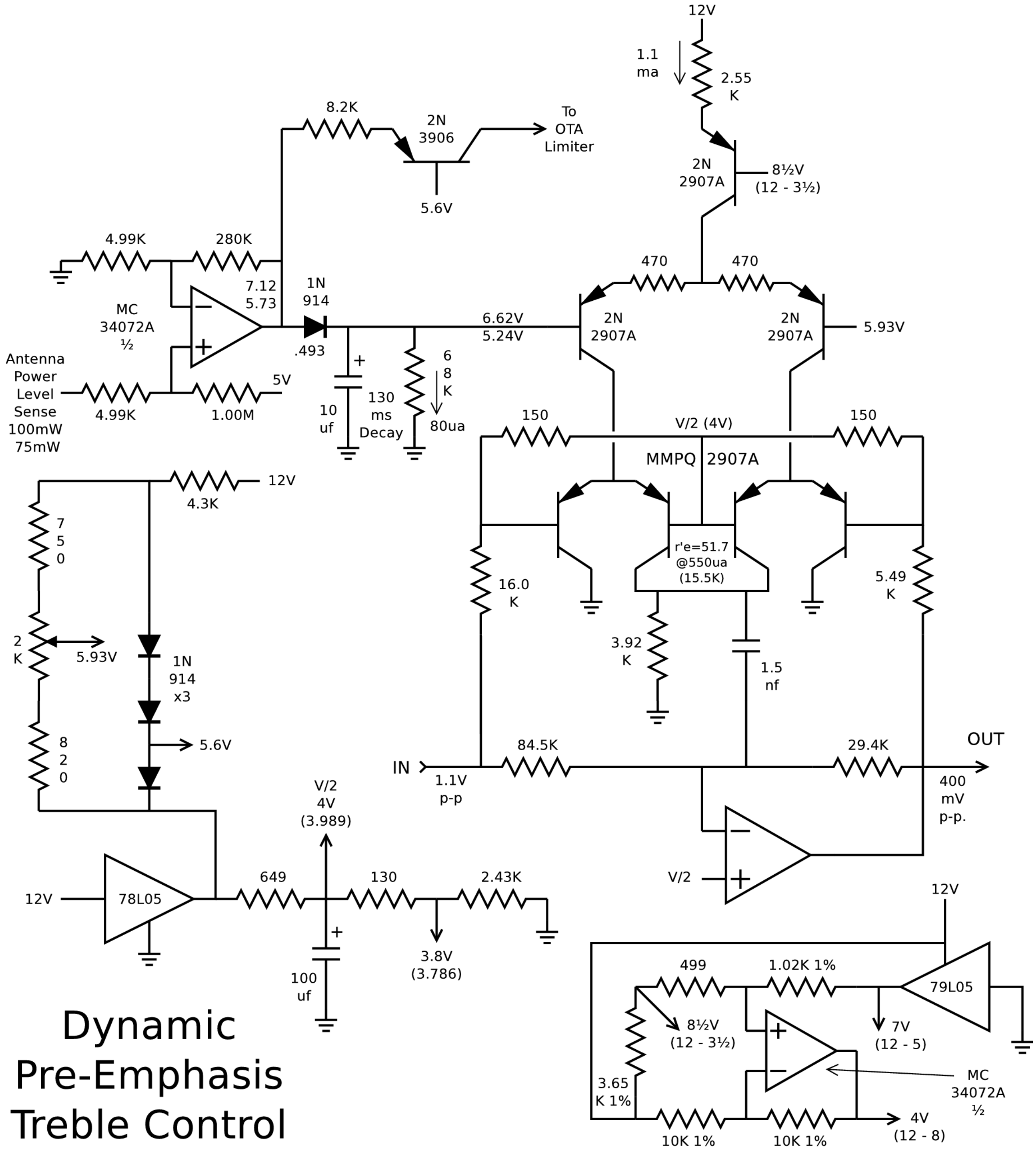
Legal

As of this writing the U. S. Federal Government is shut down and access to Part 15 Rules on the FCC website is unavailable. A quick rundown of a Cliff Notes version of the most important issues are: The input power into the final amp must not exceed a continuous average of 100mW, The antenna must be ≤3 meters in length including the coax lead and/or grounding, Part 15 transmissions must not cause harmful interference to any radio frequency transmission and/or device that operates under rules other than Part 15 that has a protected emission mask under those other rules, Emissions outside the AM band must be suppressed to ≥20dB below the unmodulated carrier. There are other less critical requirements not listed here to be met. It is the responsibility of the user that [he|she|it] comply with the law regarding Part 15 transmission particularly § 15.219 . Obtaining a full copy of the **Part 15 Rules** and carefully reviewing them is highly recommended. At some point in time in the future when the rules become available, again, a full copy may be appended to the end of this document. This document does not intentionally condone or encourage the operation of a Part 15 transmitter outside these rules. It is written for educational purposes and care has been taken to try to present information in a way that complies with the rules.

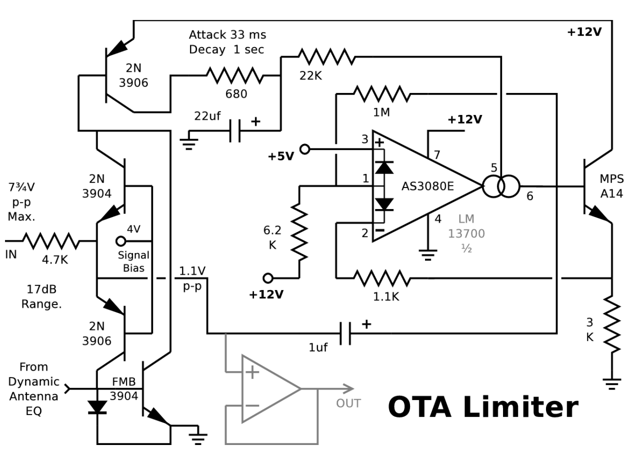
Use this greatly simplified modulator and finals in place of the one on page 12 with the described round antenna. Run the pre-processing on pages 4 and 5 on 12V.



In lieu of using a Vactrol based dynamic pre-emphasis control in the case one can't be obtained here is a fully electronic version that offers tighter controlled attack/decay timing performance of ~9ms/~130ms respectively.



Dynamic Pre-Emphasis Treble Control



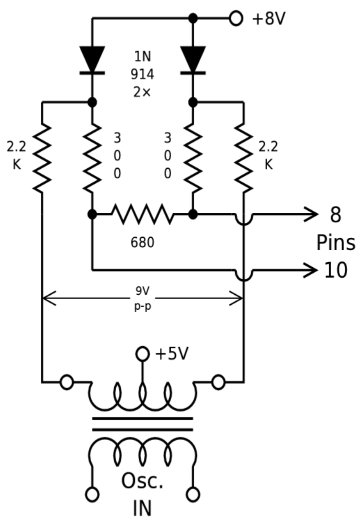
OTA Limiter

OTA Based Limiter

(Borrowed from a popular AM Stereo processor)

In the drawing to the left is a Limiter using an **O**perational **T**ransconductance **A**mplifier. The original was designed to operate on a ±15V supply. It has been modified to operate on a +12V supply to replace the Vactrol based version if one can't be obtained.

For RF modulation the MC1496 modulator's upper quad transistors are usually driven way past the point of linearity in order to get maximum signal output from it. As a result the output is rich in harmonics which would need to be filtered. Its predecessor, the MC1495 using linearization diodes, was a four quadrant multiplier capable of linear multiplication producing only the desired modulated sidebands. Why they didn't just leave the diodes in place without the differential pair driving them I don't know. Unless this is already an undocumented feature of the 1496 the future manufacture of them should contain these tied to pin 9 as was done in the AS3080E upgrade of the CA3080A. If not biased their presence would not affect the performance in older circuits designed without them. In the goal of producing a clean output linearizing the input to the 1496 is needed.



Since the antenna has such high Q this would eliminate the need for a tuned output transformer on the 1496 opting for a broadband one instead; FT-50-43, 116t #34 CT for both primary and secondary in 2 layers, or as many turns (~58) of a twisted quad-filar winding that would fit if the self resonance is well above the operating frequency. In the drawing to the left is a linearization diode circuit for the 1496 oscillator input. For the same %Δi the ΔV drop across the 1N914 diode is about twice that of the bases of the transistors in the 1496. In order to match the log response of the 1N914 diodes to the 1496 transistors a resistive divider network is used. To provide the needed Δi across the diodes to produce the ΔV needed across the transistors' bases and still provide a linear i input 2.2K resistors are used to feed the signal to the diodes, however this requires a 9Vp-p input @2¼mW. To get this power from the oscillator a buffer amp should be used to take the load off the oscillator.