F5-HA Discrete All FET Class A Headphone Amplifier

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Background

One weekend, with poor weather outside, I dug up what I had in my drawers and build a Vero board prototype of the F5 Headphone Amplifier that I had drawn up and simulated a few days early, initially as a circuit design exercise [1]. But the simulation results were impressive enough that I was curious to find out how it would actually perform in reality. The results, both in measurement and subjective impressions, were so pleasant that I drew up a PCB layout subsequently and built one for a friend.

Those who know us know that we only do discrete circuits, mostly balanced, almost exclusive FET based, and with it all the complication of sourcing and matching active devices. Although many have built Nelson's F5 power amp successfully without matching, the best performance is still to be had through careful matching of active devices. At the very least, one wants to have both channels having identical sonic signature.

The F5-HA PCB was done primarily for our own use. In any case, this is definitely not a beginner's project. One needs to know what one is doing, and one needs to have more equipment than a simple DMM and a sound card.

Circuit Design & Device Choices

With almost all Toshiba audio FETs now obsolete, it is not easy to publish a circuit based on still-active parts. For the MOSFETs, the initial prototype was built using IRF510 / IRF9510. Later measurements revealed that the Fairchild FQP3N30 / FQP3P20 is a much better complementary pair in comparison (Fig. 12). One can of course also use other complementary pairs, such as 2SK2013/2SJ313. But for the input JFETs, the 2SK170/2SJ74 pair still has too much advantage over the rest, in terms of noise and transconductance. So the standard circuit is based on 2SK170BL/2SJ74BL and FQP3N30/FQP3P20.

The basic circuitry is not any different from the original F5, or its predecessor the Selectronics Profet. Because the headphone amplifier is supposed to drive a load of say 25 ohm to 300 ohm, using a low-impedance feedback network as in the original F5 will load the output too much, that distortions will become high. While this can be solved by increasing the feedback network resistor values by say a factor of 10, this in turn will degenerate the JFETs so much that the bias will be too low. Thus, it was necessary to revert to the "normal" type of feedback network, as found in the Profet on which the original F5 was based. In this case, the JFETs still run at their Idss, and one is pretty free to choose the feedback resistor values without affecting the rest of the circuit.

The circuit will also work with GR grades of 2SK170/2SJ74, which might be easier to obtain. Since these have lower Idss than the BL grade, the values of the JFET drain resistors have to be increased in order to provide the same biasing voltage for the second stage. In doing so, one has also doubled the open loop gain (OLG), and reduced the stability margin after feedback. To compensate for that, a simple approach is to double the value of the MOSFET gate stoppers resistors (to 2 ~3k ohms). A good starting point for the value of R3/R4 would be 7.5V / (Idss of JFETs). And the trim-pots should be a minimum of 8x that of R3.

Other alternative input pairs are 2SK246BL/2SJ103BL and the 2N5457/2N5460 family. Straightly speaking, these devices are also no longer in production, though one can still purchase the Fairchild devices in SOT-23 package and use an adaptor board to convert to TO92 pin layout. Apart from quite a bit higher noise, these devices also have much lower transconductance. One would therefore wish to use these with a low bias (say 3mA) so as to recover some OLG via the increased drain resistor values. This implies using 2SK246GR/2SJ103GR or 2N5457/2N5460, as always with matched Idss. If, however, you can only get 2SK246BL/2SJ103BL or 2N5458/2N5461, then you should revert to the original F5 feedback network using 1k/200R for a gain of 6. The shunt resistor of the feedback network will degenerate the BL grade Idss by 4~6mA, getting them closer to the desired bias current. All these, one has to bear in mind, are compromises. The different pin assignments to the 2SK170/2SJ74 also needs to be taken into account during soldering, without saying.

As in the original F5, a current limiting circuit has been included for each MOSFET. By changing values of the potential divider across the MOSFET source resistors, one can set the current limit at any value from 200mA upwards.

It has been reported by various F5 builders that the current limiter has a subjective negative impact on the sound quality. As in our own F5X, we do not use these at all. But we include them anyhow on the PCB as optional, as they do not take up extra space.

Power Supply

This is an amplifier with very high bandwidth (>1MHz at -3dB) when driven by a low impedance source, so a couple of power rail capacitors were placed right at the amplifier itself. As opposed to a fully-balanced, class-A amplifier such as the F5X, a single ended F5 still draws variable current under load, and will benefit from a regulated supply. The PCB layout was made with LM7815/LM7915 regulators pin assignments, but any pin-compatible devices that are stable with the amplifier circuit can be used. Again, for my own build, I chose to use LM317/337 pre-regulators followed by discrete low-noise Darlington cap multipliers. Each channel is further equipped with its own 25VA Talema toroidal PCB transformers with 2x15V dual secondaries, providing more than sufficient margins on current delivery. Each secondary has its own bridge rectifier (4x MSR 860), and these were only joined at the common flooding ground plane for each channel on the top layer of the PCB. Because the transformers come also with dual 2x115V primaries, simple jumpers will allow easy conversion between 115Vac and 230Vac mains.

The normal power consumption of 2 channels of F5-HA amounts to approximately 15W. A mini-fuse of 0.3A slow-blow should be sufficient for 230Vac mains. To prevent excessive inrush current, a CL90 NTC may be added between Mains Live input (after the power switch) and the mains connection on the PCB, though I found this unnecessary.

For those who want the ultimate noise performance and wishes to use batteries, provisions have been made to connect 15V battery packs to the PCB. One can simply saw off the transformer and rectifier sections of the PCB to make room for the batteries. Both LiPo and NiMH batteries of around 2000mAH should fit, and one may, though unnecessary, still use the regulators at a slightly reduced voltage (say 12V). One will need a 4TDP toggle switch with an inrush limiter to select between Power On and Charge. Battery charger is best placed outside the amplifier enclosure.

Cross-feed options

We have tried various cross feed buffers in our DAO headphone amplifier, and we definitely want to include those for our own F5-HA. As in the DAO, this will be on a separate daughter board, so that one may choice between Meier, CMoy-Linkwitz and Danyuk Cross Feed. A set of jumper will allow bypassing the cross feed circuit altogether. As they are not our own intellectual properties, we shall not be publishing schematics for them [2, 3, 4, 5, 6]. In Appendix 1 though, we have proposed a slight modification to the Danyuk Cross Feed to avoid the use of an additional follower within the filter section. This also allows the use of the same PCB as the CMoy-Linkwitz due to their similarities.

The power for the cross feed buffer is taken from the right channel power supply via 2x 240R 1W resistors. These isolate the buffer from the F5-HA, and forms a 6.6Hz low pass filter with the power supply capacitors on board the cross feed. It also reduces the rail voltages to about +/-10V for the buffer. I further added an additional LT1764 / LT1964 low noise LDO's in between to give 8.4V regulated rails. Bias current is about 20mA, depending on the Idss of the JFET Buffers. When not using the cross feed, one simply leaves out the 2 power resistors.

Note that all cross feed filter will attenuate the incoming signal by 6~10dB. For my own build, I prefer to increase the closed loop gain from 6x to 11x to (partly) compensate for that. This is simply done by halving the value of the feedback shunt resistor R10. Even with the higher gain, distortion is essentially the same for the same output voltage level.

Volume options

For a not-so-extravagant project like this, an immediate choice of volume control would be the ALPS RK27112 (Blue Velvet). This was used also as default for the PCB layout. The pin locations are such that with some leg bending, one can also make a TKD CP600 fit. On top of that, one can also obtain series-type switched resistor attenuators in ALPS and TKD compatible housings. One might prefer these to the carbon / conductive polymer tracks of the originals, as they might deteriorate over time by wear and tear. For our curiosity, we designed a double-sided PCB which will allow a ladder-type resistor attenuators using ALPS mechanical parts. But this is a rather complicated DIY solution, requiring substantial soldering skills and subsequent CNC machining. So we shall only publish pictures at a later date, just to demonstrate how it can be done.

For my own use however, I like placing the attenuator as close to the amplifier input as possible, and have therefore made further provisions on the PCB for a LDR-based volume. This uses a fixed resistor (10k bulk film) as series and an LDR as shunt. The LDR is driven by an adjustable constant current source based on an LED biased NPN-BJT. But one can just as well use the standard circuit in the TL431 datasheet. Similar to the cross feed buffer, power for the LDR CCS is taken from the left channel, which is further regulated down to 9V by a simple 78L09. The CCS current varies between 20µA and 20mA. The worst case dissipation of the 78L09 regulator is 120mW (at 11mA CCS current), well within the capability of the TO92 package.

The frequency response measurements shown in Figs. 8 & 10 were performed using a low impedance signal source (< 100R). If the signal source has a high output impedance (>5k), the frequency response still measured flat, but at a lower bandwidth of around 200kHz. However, between 200R and 5k (as in the case of driving directly from a 10k volume pot), the source impedance interacts with the capacitances of the input JFETs, and the closed loop response showed a slight hump at the corner frequency (of 500kHz or above, depending on Zin) of up to 1.5dB. This is of course way above the audio band and is thus not really significant. But to flatten that as well, a 100p silver mica (Cin) was added between Vin (before R11, R12) and Gnd. Response is now flat all the way from 20R to 10k Zin. And even at 10k, the -3dB bandwidth is still 120kHz.

As capacitances vary between devices, it is best to actually measure the frequency response of the finished amplifier with various Zin values, and then choose a value for Cin to give a flat response for all Zin values.

If one however prefers to retain the high bandwidth of the amplifier itself, the easiest solution is to add a JFET follower buffer right after the potentiometer, using e.g. a matched pair of 2SK209s.

PCB Layout

The PCB layout is typical that of a headphone amplifier that is not overcrowded. At the rear are mains connections and RCA inputs; at the front volume control and output jack. It is thus logical to place the power transformers at the rear, and the amplifier at the front, as far away from the transformers as possible. Although the power MOSFETs are not placed near the middle for optimal heat removal, the dissipation per device is a mere 2W, such that one can place more priority on electrical layout rather than thermal management.

Since the build is essentially dual mono, the two channels can also be placed apart from each other to minimise any cross talk. The middle section is left for the input lines as well as for the cross feed, the latter by necessity.

To minimise any disturbances from the mains connections and the transformers, a ground shield is placed directly above the aforementioned. This is further connected via build-in tracks to the enclosure as well as to the mains earth. Furthermore, the two transformers are connected to the mains in anti-phase, so as to enable stray field cancellation between the two.

The RCA inputs are connected from rear to front, where the volume control potentiometer is placed, by copper tracks on the bottom side of the PCB, each surrounded by its own guard ring connected to its respective signal ground. To provide further shielding from the transformers, an additional flooding ground plane is placed on the top side of the PCB directly above the input signal tracks. This signal ground plane is separate from the mains shield, but can be connected to the latter via a power resistor.

Two additional ground planes are provided for each of the amplifier channels. These are connected to the star point of each amplifier close to the gates of the JFETs of the input stage These two ground planes are electrically separate from the central signal Gnd, with provisions to connect them together via 1206 SMD resistors. The exact connection scheme depends largely on the input and output configurations. However, by necessity, the two inputs have to share a common Gnd at the cross-feed PCB. And since the cross feed buffer is powered by the power supply of the right channel, this cross feed common ground is automatically connected to the latter. Thus, it makes sense to connect the input ground shield to the right channel ground plane with low impedance.

Most signal source normally has the signal ground connected to mains earth. So in theory no direct connection is required between the input ground and the mains shield. Even if that is not the case, a resistive connection between the two (e.g. 10 ohm power resistor) is more than sufficient. If the headphone has all four connections to the transducers accessible, then using a 4-wire connection to the amplifier is the much preferred solution. This allows the return current of the two channels to be routed separately instead of using a common conductor. For such an output connection, the ground planes of the two channels should not be connected at all.

In case a three-wire output connection (e.g. 1/4" stereo phone jack), the output grounds of the two channels are connected also at the output socket. No further direct connections of the left channel ground plane to the others are therefore necessary.

Thermal Management

The recommended bias per channel is 100~150mA. This will guarantee pure Class A output of at least 1.3Wrms for headphone impedance between 32 and 65 ohms for 15V rails, using the higher bias value for lower phone impedance. Adding to that the voltage drops across the transformer coils, the rectifiers and the regulators, the total dissipation for both channels amount to some 15W. As each of the 16 rectifier diodes dissipates a mere 0.15W, no additional heat sink is necessary for the TO220 devices. When using 7815/7915 type regulators, they have to get rid of some 0.6W each. It is good engineering practice to connect them thermally to the amplifier enclosure / heat sinks, using e.g. a simple 3~5mm thick aluminium angle. This becomes an absolute necessity when using shunt regulators, as these can dissipate up to 7.5W per channel.

Otherwise, the MOSFETs dissipate the most of the heat, each of them about 2W. This is of course no issue for such TO220 packages. But heat sinking is still needed, and thermally connecting the MOSFETs to the bottom plate of the enclosure (minimum 3mm aluminium) is a simple solution. The large sectional area of the bottom plate will spread the heat and conduct it to the rest of the enclosure, its entire surface area also acting as heat sink surfaces.

The F5, though complementary by design, is not 100% symmetrical and still has a small DC drift. Good heat sinking always means lower DC drift during warm up. It is therefore advantageous to use an enclosure equipped with full-length side heat sinks. These will help to keep any temperature rise, and thus DC drift, to a minimum.

Set-Up Procedure

The best approach to build the amplifier is to test each section sequentially, starting with the power supply and regulators. For convenience of testing, the regulator outputs are disconnected from the amplifier on the PCB, and needs to be bridged using SMD 1210 0R jumpers. One may of course use solid copper instead.

When using regulators such as 7815/7915, they are only stable with an output capacitor connected. With the 1210 jumpers removed, there is no output capacitor connected to the regulators, and they will oscillate. The simplest way to overcome this is to solder directly between the Output and Gnd pins of the regulators a WIMA MKS2 2.2µF film cap on the bottom side of the PCB. For the LM317/337 cap multiplier, this is not necessary. Other regulators have their own requirements to satisfy.

Once the regulator output voltages have been measured to be normal, the power supplies can be connected to the amplifiers using the said 1210 jumpers. Again, it is best to connect power to one amplifier at a time.

Personally I prefer to test the amplifiers on their own first using laboratory supplies. These not only allow easy adjustment of voltages and current limits, but are almost always equipped with voltmeters and ammeters to give you a good idea of what is happening. Using the on-board power supply would not give you such flexibilities. And you will certain appreciate them in case of any circuit malfunctions (e.g. oscillations).

If the Idss of the JFETs and the Vgs of the MOSFETs at bias current are known, one can calculate the exact value of drain resistor required. Assuming the standard circuit with 2SK170/2SJ74, the drain resistor value is given by :

R_drain = (MOSFET Vgs at bias + Bias current x 3.3R) / JFET Idss

The above only applies to the standard circuit where R10a, R10b = 0. In case they are not, the actual bias current of the JFETs after degeneration should be applied in the above equation.

Using a DMM, one can then pre-set the trim-pot to give that value on the PCB. This will ensure that you are quite close to the desired bias current at power up.

Once powered up, and with the input shorted to Gnd beforehand, measure the voltages across the two 3.3R power resistors consecutively. Adjust the drain resistor trim-pots, one at a time, until the voltage across the 3.3R's both reads 0.40V. The bias of the MOSFETs are now set at 121mA each. Now measure the output voltage with respect to Gnd, and adjust one of the trim-pots until any DC offset is halved. Then adjust the other trimpot until offset is zero. Recheck voltage across 3.3R and re-adjust bias as necessary. Re-check regulator output voltage to ensure all is functioning as intended.

The MOSFETs have rather high temperature coefficients, so that bias current will increase as temperature rises. Allow the amplifier to warm up for 30 minutes and readjust bias and DC offset as above. To attain 150mA bias, voltage across the 3.3R resistors should read 0.50V. For 100mA bias, it should be 0.33V.

Once this is set and allowed to settle thermally, one can apply an AC signal to the amplifier. The simplest test is a 2V pk-pk, 1kHz square wave at the input. The output without load should be around 12V pk-pk, without overshoot or ringing. If one does get ringing, increasing the MOSFET gate stopper resistors should cure that. Once the output waveform is clean, one might wish to check again by apply a dummy load, such as a 50ohm 5W wire-wound power resistor.

With more equipment, one could of course measure frequency response, distortion spectrum, etc.

And then it would be time for music. ☺

If you are using cross feed buffers, and / or LDR volume, it is best to test those off line with laboratory power supplies before installing them onto the main PCB. In both cases, it is worth checking the power supply voltages at both modules.

A Note on Current Drive

It appears that driving headphones with a transconductance amplifier (or voltage controlled current source VCCS) is coming into fashion, with commercial offerings from Bakoon and Erzetich. The idea is of course not new, at least for loudspeakers, but it has never really catch on commercially. Once can argue that it only really works well for single full range drivers, and the latter are also not main stream. Interesting enough, one of the most prominent promotors of current drive speakers, Esa Meriläinen [7], is less enthusiastic about driving headphones in current mode.

His argument [8], which I agree, is that " in electrodynamic headphones, the achieved benefits of current drive are usually very minor compared with the improvement in loudspeaker operation. This is mainly because in the impedance of headphone transducers the relative proportion of the DC resistance is generally much higher than in speaker drivers, so the interfering current components produced by the electromotive forces are left rather small even on voltage drive. A greater problem is generally constituted by the unevenness of frequency reproduction and its dependence on the ear canal shape. As with loudspeakers, the frequency response of headphones also exhibits certain changes when moving to current-drive. These changes may, depending on the case, also result in undesirable impressions. "

He has a very informative website about current drive amplifier topologies for those interested [9] .

FETs, as we are using in the F5-HA, are of course transconductance devices by nature, i.e. they convert voltage input to current output. So converting the F5-HA to current drive is quite simple. As the F5-HA employs global negative feedback, all we need to do is to convert the feedback network (R9, R10) from voltage sensing to current sensing. The schematics can be seen in Fig. 11. One basically replaces the original R9 with the load (headphone), and R10 with a current sensing resistor. The current gain (transconductance) is approximately equal to 1 / R_{sense}. For +/-15V rails with 150mA bias, we suggest 3.3R, so that a 1V input signal will be sufficient for maximum Class A current output. If one is using a headphone with significantly higher impedance (e.g. 300R), one might wish to consider reducing the current gain. we suggest 1/5 to 1/10 of the phone impedance as a starting point. As in the voltage feedback network, R_{sense} should be low noise, high precision, and with sufficient wattage to avoid any thermal distortion.

Note also that the impedance of most headphones are not flat over frequency, and this is very likely to cause instability to the F5-HA due to additional open-loop phase shifts. For example, many headphones becomes increasingly inductive at around 10kHz. For such headphones to be driven in current mode, it will be necessary to use some form of Zoebel network to compensate for the impedance fluctuation. But then the attractiveness of current drive is lost in the process. An example is shown in [10] for the Sennheiser HD650. You can also find many more circuit ideas for balanced current drive amplifier topologies in [11]. The latest planar magnetic headphones seem to have quite constant impedance across the audio band. But most measurements published cease at 20kHz, so you might still need to use some form of R-C Zoebel at a higher frequency to ensure amplifier stability.

Or better still, use an open-loop VCCS circuit with sufficient voltage headroom, specially designed for that purpose. Something along the lines of a First Watt F1.

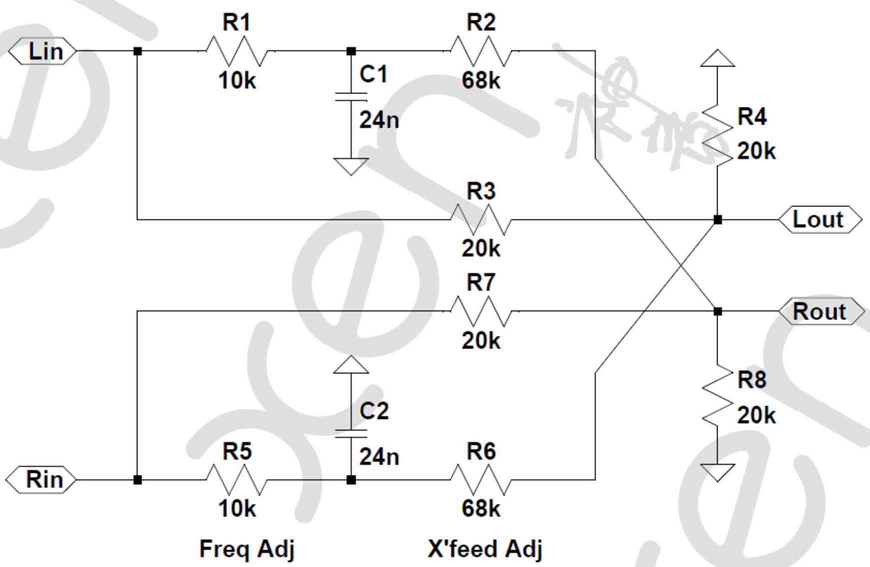
Appendix 1 Modified Danyuk Cross Feed

The Danyuk Cross Feed differs from the rest that it has no treble boost built-in. It also has the convenience that the level of cross feed as well as the cross-feed corner frequency can be adjusted independently by trim-pots. However, the penalty is a pair of additional active followers within the filter circuit [6]. As we wish to avoid additional active stages in the audio chain, we investigated how this additional follower stage can be avoided / eliminated.

Let's refer to the original schematics of Danyuk for analysis. For the left channel, as an example, the forward path of its own signal is formed by the potential divider R3 / TR3 with -6dB attenuation. The right channel signal first goes through a low-pass filter formed by (R5 + VR4) / C2, followed by a unity-gain buffer, a variable attenuator VR5, and is then summed to the left channel output via R6 connected to the junction of R3 / TR3.

Firstly, we are of the opinion that unless constant adjustment of the level and frequency of the cross feed is necessary, using fixed resistors to replace the trim-pots is a much preferred solution. Secondly, the attenuation of VR5 can also be achieved by combining it with the potential divider R6 / TR3. Under normal usage, the attenuation of VR5 is in the order of ¼ x to ½ x. Using an R6 with a value between 3x to 7x of VR5 would achieve the same effect. With such high values of R6, Once can directly connect R6 to the output of the LP filter without the need of the additional buffer.

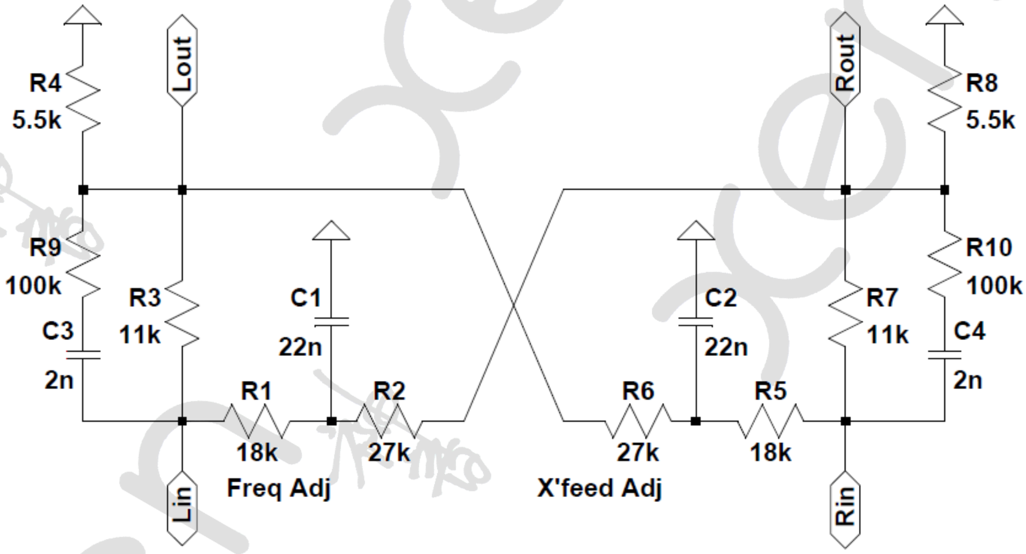
A possible alternative circuit is shown below :



It provides essentially the same functionalities as the original. The cross feed frequency can be adjusted by changing the values of either R1 / R5 or C1 / C2, although R1 / R5 should not be much lower than 5k to avoiding loading the signal source excessively. R6, on the other hand, will allow adjustment of the amount of cross feed. The only penalty is that the frequency and level adjustments are not entirely totally independent. So some iterations may be required for large changes, which we do not foresee.

This purely passive filter has an output impedance in the order of 10k, as shown. If one wish to follow this by a volume control attenuator, or drive the F5-HA directly to a high bandwidth, a follower stage (e.g. a pair of simple JFET source followers) can be used to provide the necessary low output impedance.

If we further compare the above passive network to that of the CMoy modified Linkwitz cross feed, we should notice that they have a lot of similarities, but for the treble boost section. One can therefore also adjust the values of a CMoy filter to achieve the same output characteristics as that proposed by Danyuk. The following example essentially does the same as the passive network shown above, but with an additional -4dB attenuation.



To allow initial adjustment of cross feed level and frequency, we recommend that R1 / R5 be first replace by a 50k stereo pot, and R2 / R6 by a 100k stereo pot. Once adjusted to your liking, you should then proceed to them by fixed resistors.

A private conversation with Dimitri Danyuk confirmed his approval of the adaption proposed.

Figures

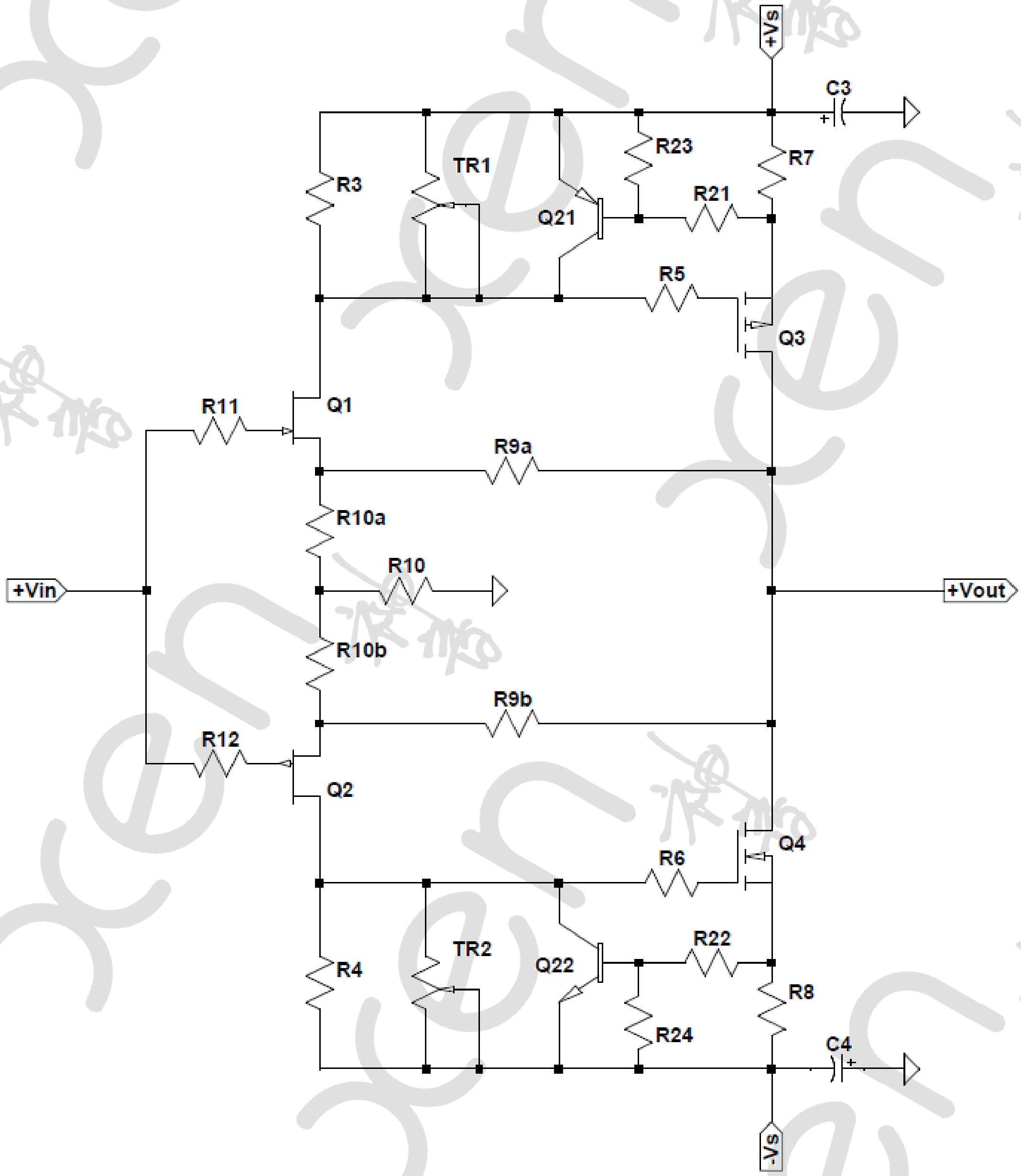


Fig. 1 **F5-HA Amplifier General Schematics**

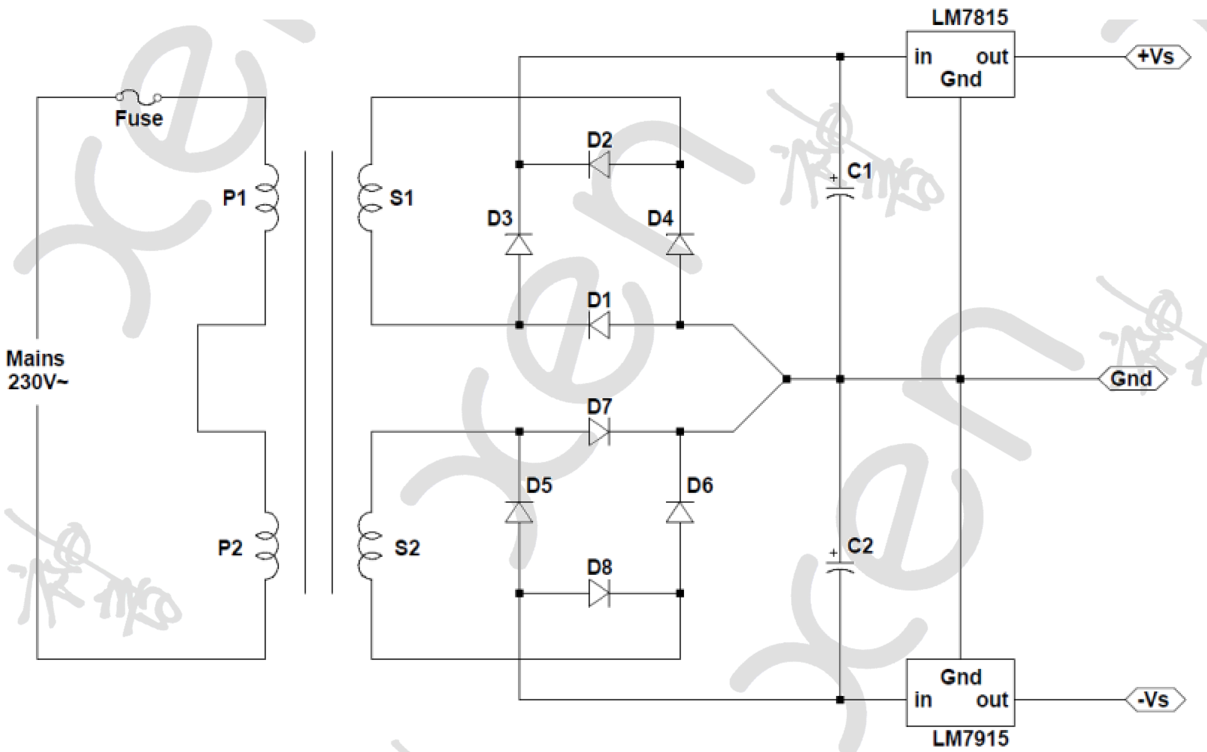


Fig. 2 **Typical Example of Power Supply for 230Vac Mains**
* when not connected to C3, C4, LM7815/7915 require 2.2 µF minimum between Outputs & Gnd for stability

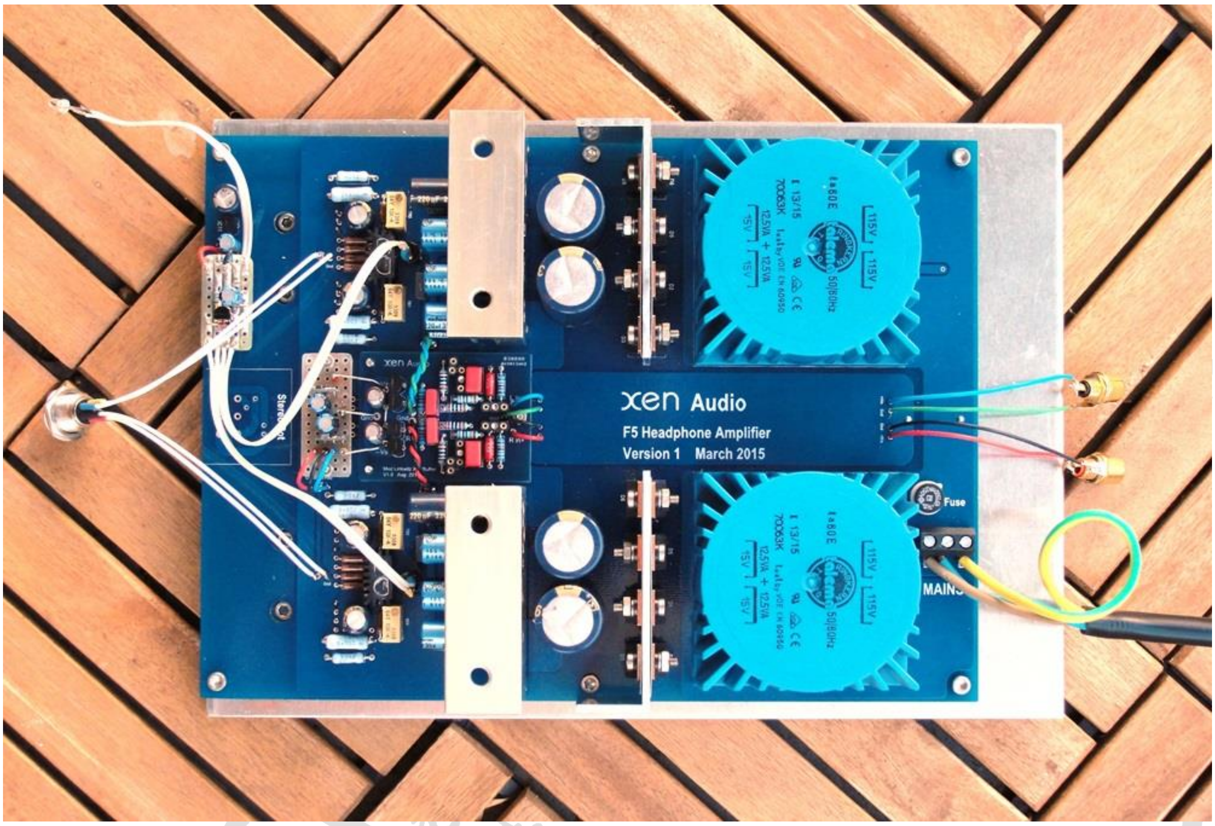


Fig. 3 Fully finished PCB with LDR volume and cross feed buffer, without heat sink enclosure, Top View

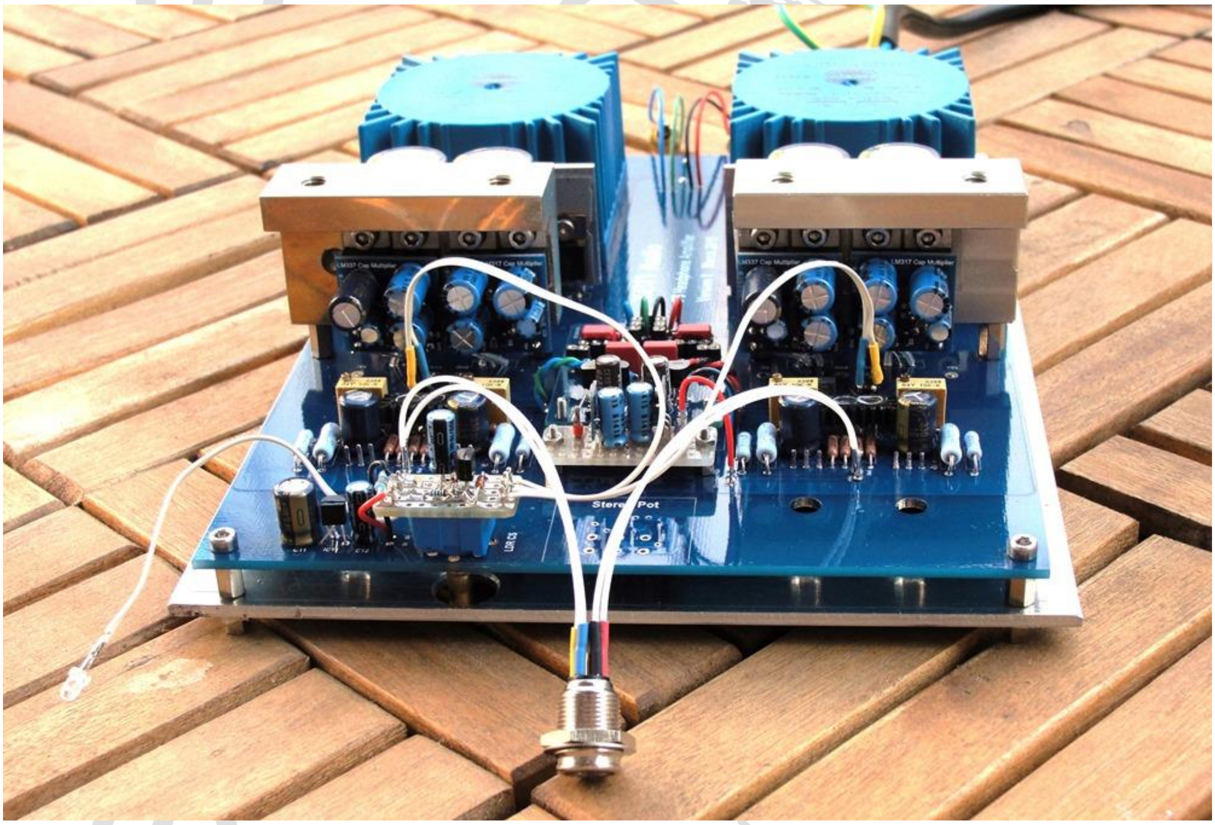


Fig. 4 Front View

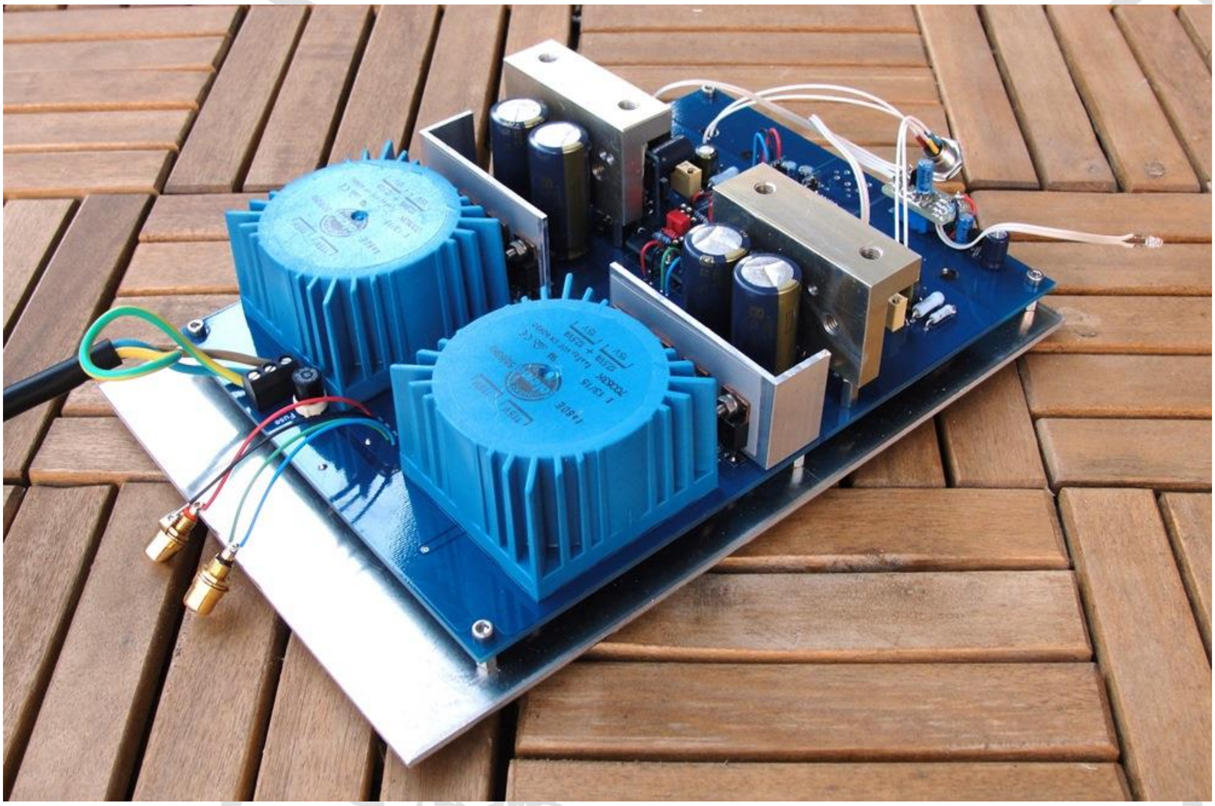


Fig. 5 **Rear View**

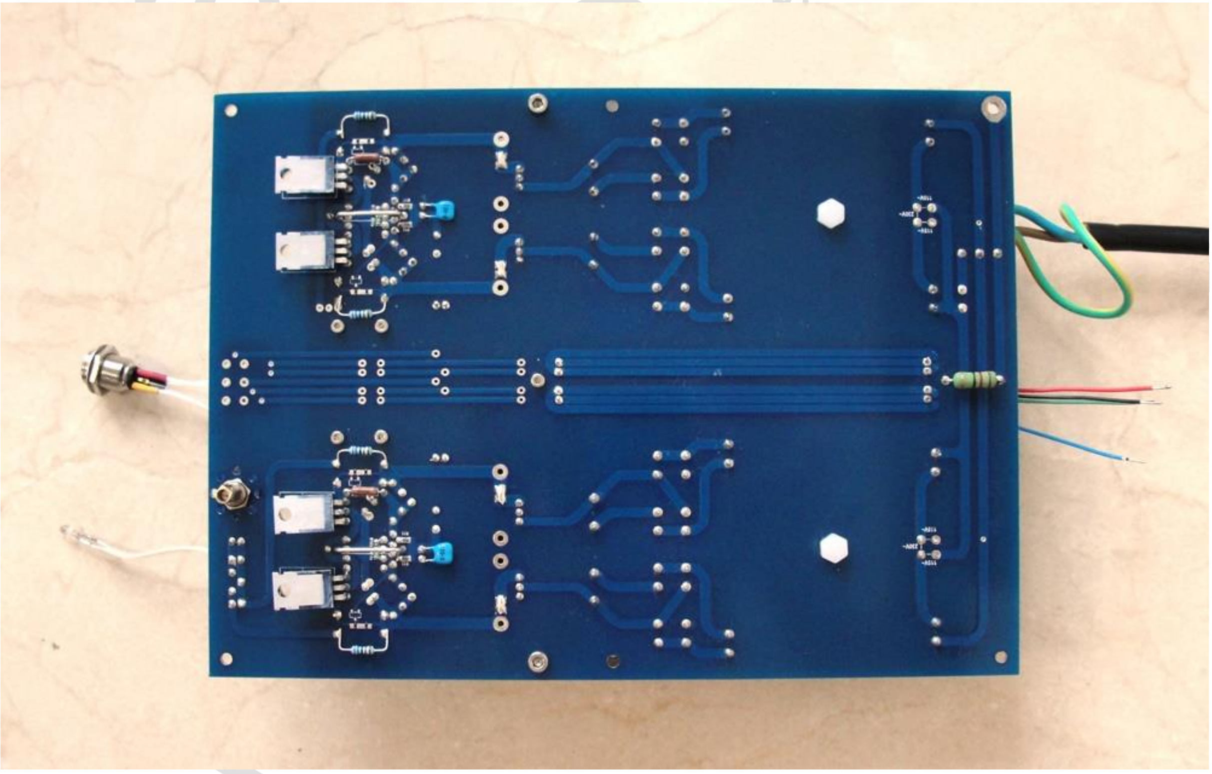


Fig. 6 **Bottom View**

Measurements of First Prototype

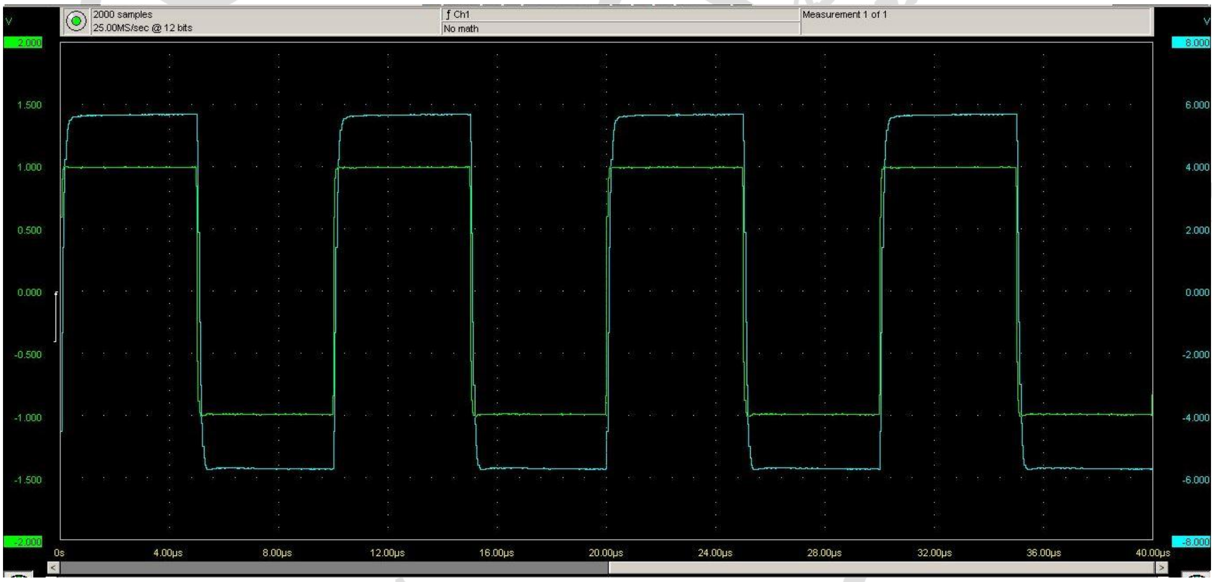


Fig. 7 100kHz Square Wave Response, No Load

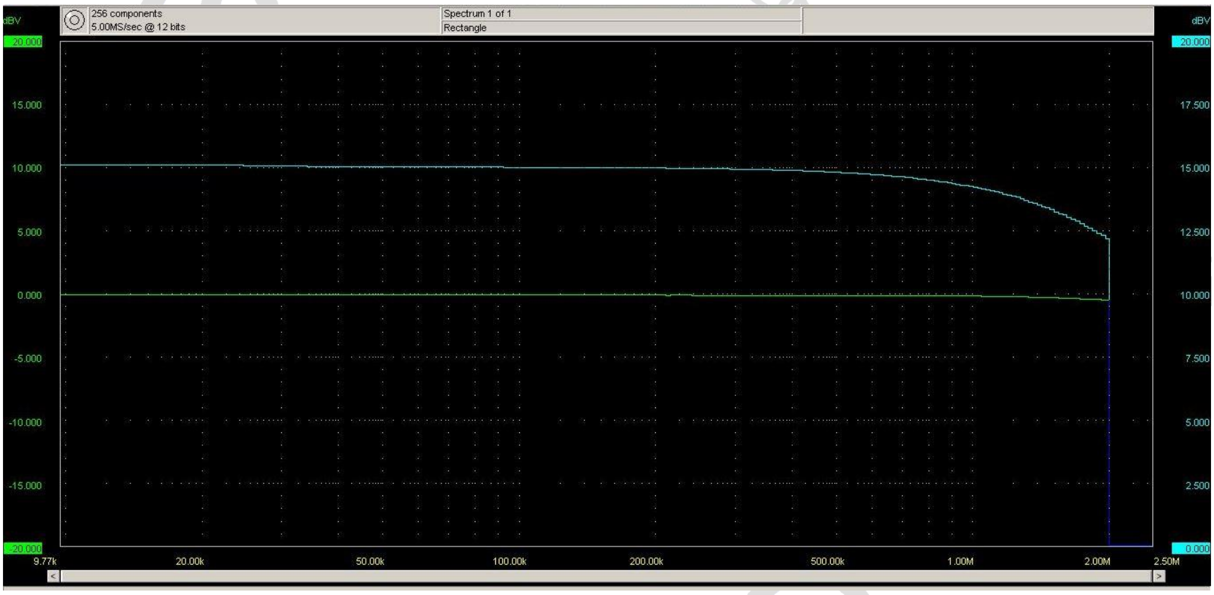


Fig. 8 100kHz Frequency Response, No Load

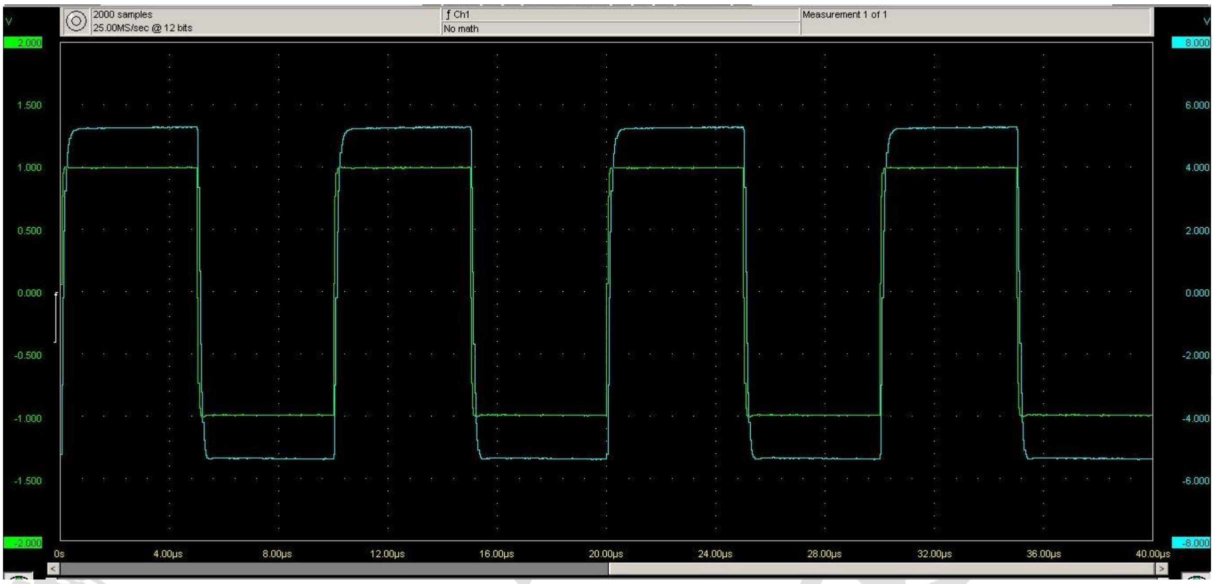


Fig. 9 100kHz Square Wave Response, 68R Resistive Load

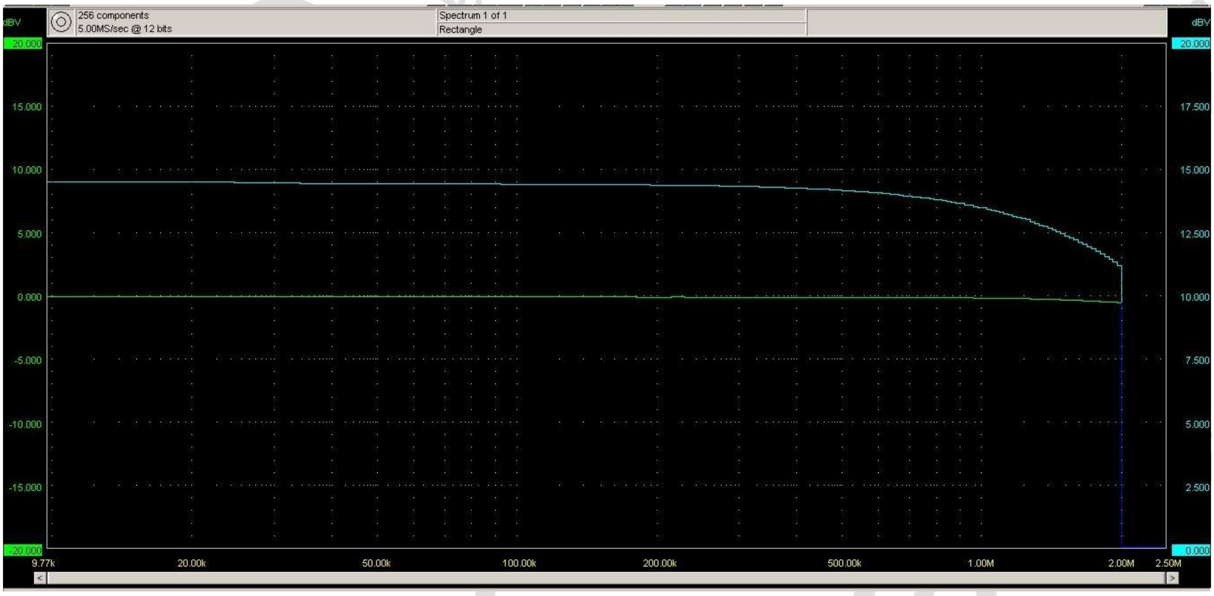


Fig. 10 100kHz Frequency Response, 68R Resistive Load

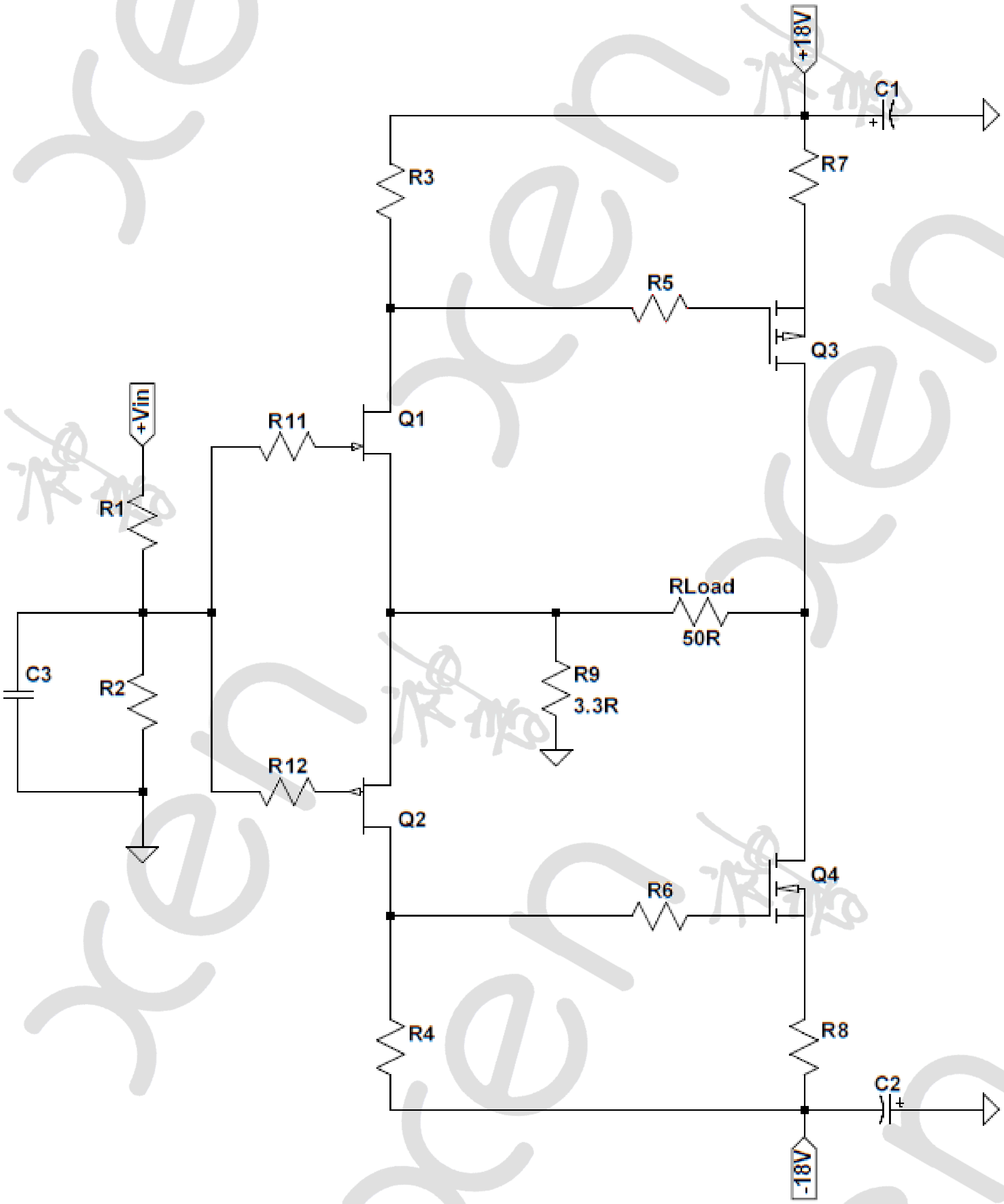


Fig. 11 **F5-HA Amplifier in Current Mode**
(RL = headphone, assumed 50R resistive)

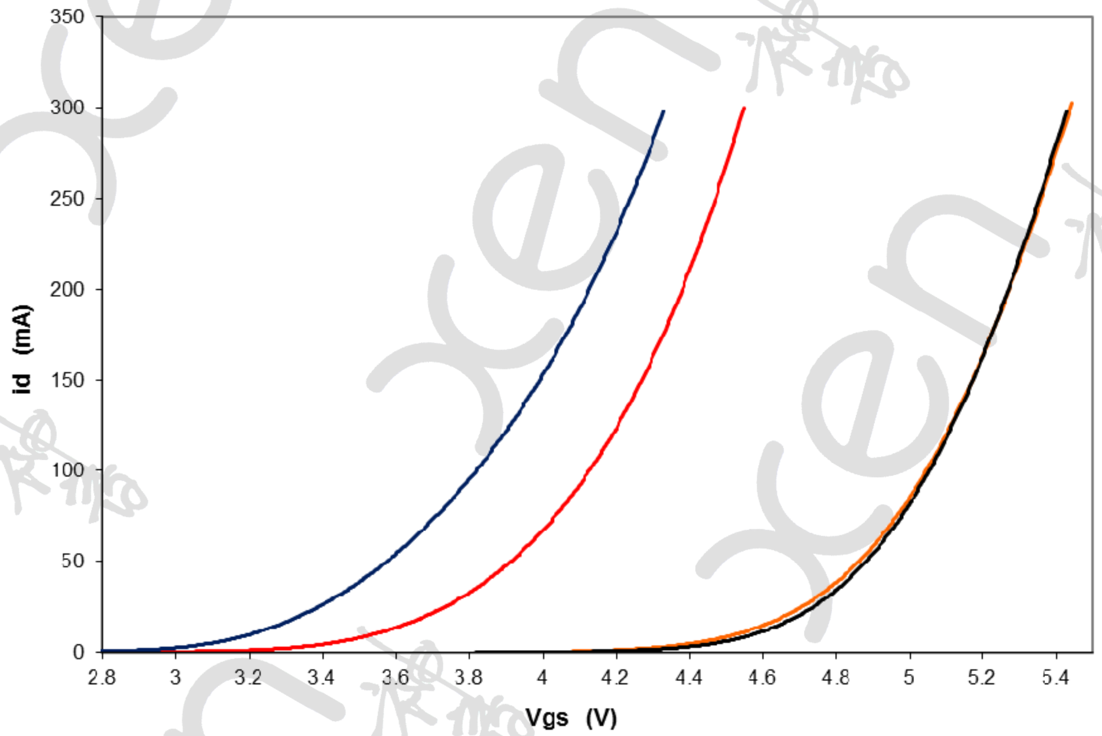


Fig. 12 **MOSFET Comparison**
(Red = IRF510, Blue = IRF9510, Orange = FQP3N30, Black = FQP3P20)

Bill of Materials
(+/- 15V rails, 150mA bias, Gain = 6)

F5-HA Amplifier	**(one channel)**	
Designation	**Value**	**Type**
Q1	2SK170BL	
Q2	2SJ74BL	Matched Idss to Q1
Q3	FQP3P20	
Q4	FQP3N30	Matched Vgs to Q3 at 150mA
TR1, TR2	10k	Bourns 3296Y ~ 10 x R3
R3, R4	= 7.5V / Idss Q1	Vishay Dale RN55, see Description
R5, R6	2k	Vishay Dale RN55
R7, R8	3.3R	Panasonic 1W, matched to <1%
R9a, R9b	1k	Vishay Dale RN55
R10	100R	Vishay Dale RN55
R10a, R10b	0R	Panasonic 0805
R11, R12	100R	Susumu 0805
R41, R42	240R	Panasonic 1W, for LDR CCS
R43, R44	240R	Panasonic 1W, for Cross Feed Buffer
Ci	100p Silver Mica	* Optional when using 10k input pot
C3, C4	330µ 25V	Nichicon KA

Optional Current Limiter @ 400mA

Q21	BC807-40	
Q22	BC817-40	
R21, R22	1k	Susumu 0805
R23, R24	1.1k	Susumu 0805

PSU	**(one channel)**	
Tx	2x115V; 2x15V	Talema 70063K
D1~D8	MSR860	
IC1	LM7815	
IC2	LM7915	
C1, C2	4700µ 35V	Nichicon KA
Fuse	0.5A	Little Fuse TR5 374 Series
Fuse Holder		Little Fuse 560 Series
RGnd	10R	Panasonic 2W
Mains Connector	3 pole, 5mm pitch	Phoenix 1729021

Volume Control LDR (stereo)

Designation	Value	Type
Stereo Pot	--	Not populated
LDR	NSL32SR3 LDR	Matched
IC11	78L09 TO92	OnSemi MC78L09ABPRAG
C11, C12	47µF 35V	Nichicon KA
J1 to J12, J4 to J13		Jumper Wires
J2 to J11, J3 to J14		Jumper Wires
J21	10k	Caddock MK132

LDR CCS

Q1	BC337-40	Fairchild TO92
IC1	TL431	
D1	Blue LED 3mm	
R1	120R	Dale RN50 1%
R2	2k	Dale RN50 1%
R3	2k	Dale RN55 1%
TR1	100k Log Pot	Bourns 51CAD-D16-D20L

Volume Control ALPS (for stereo)

Stereo Pot	10k Log	ALPS RK27112A00A5
LDR	--	Not populated
IC11	--	Not populated
C11, C12	--	Not populated
J1~8, J11~14	0R	Apply jumper
J21 (2x)	0R	Apply jumper

When using Cross Feed Buffer + LDR

J1	To X'fd Left channel +Vin
J2	To X'fd Left channel Gnd
J3	To X'fd Right channel Gnd
J4	To X'fd Right channel +Vin
J11	To X'fd Left channel Gnd
J12	To X'fd Left channel +Vout
J13	To X'fd Right channel +Vout
J14	To X'fd Right channel Gnd

CMoy Linkwitz Cross Feed Buffer

Qty.	Designation	Value	Type
2	R1a	11k	Caddock MK132 / Dale RN55
2	R1b	33k	Caddock MK132 / Dale RN55
2	R2	5.49k	Caddock MK132 / Dale RN55
2	R3	50k	Caddock MK132 / Dale RN55
2	R4	18k	Caddock MK132 / Dale RN55
2	R5	18k	Caddock MK132 / Dale RN55
2	R11	220R	Susumu 0805 0.5%
2	C1	3.3n	WIMA MKP2 1%
2	C1a	680p	Panasonic PPS 16V
2	C2	22n	WIMA MKP2 1%
2	C3, C4	100µF 16V	Nichicon KA
4	Q1 ~ Q4	2SK170BL	Idss matched
2	J1		Perspective Jumper
2	J2		Cross feed On
2	J3		Cross feed Bypass

Danyuk Cross Feed Buffer
(uses Linkwitz PCB)

Qty.	Designation	Value	Type
2	R1a	11k	Caddock MK132 / Dale RN55
2	R2	5.49k	Caddock MK132 / Dale RN55
2	R3	100k	Caddock MK132 / Dale RN55
2	R4	50k log pot	* Adjust for cross feed frequency
2	R5	100k log pot	* Adjust for cross feed amount
2	R11	220R	Susumu 0805 0.5%
2	C1	1.5n	WIMA MKP2 1%
2	C1a	680p	Panasonic PPS 16V
2	C2	22n	WIMA MKP2 1%
2	C3, C4	100µF 16V	Nichicon KA
4	Q1 ~ Q4	2SK170BL	Idss matched

2	J1	Open
2	J2	Closed
2	J3	Open

Meier Cross Feed Buffer

Qty.	Value	Type
5	10k	Caddock MK132 or PRP 1/4W
2	22k	Caddock MK132 or PRP 1/4W
2	10n	WIMA MKP2 1%
2	22n	WIMA MKP2 1%
2	68n	WIMA MKP2 1%
4	2SK170BL	Idss matched
2	100µF 16V	Nichicon KA
2	220R	Susumu 0805 0.5%
2	J0	Zero Cross Feed
2	J1	Medium Cross Feed
2	J1 + J2	Low Cross Feed
0	No Jumper	Standard Cross Feed

LM317 Cap Multiplier
(Replaces 1x 7815)

Designation	Value	Type
IC1	LM317	Onsemi Adj Regulator
Q1	BC817-40	NXP NPN SOT23
Q2	2SC4793	Toshiba NPN TO220
D1, D2	MSE1PJ	Vishay Low Leakage SMD Diode
R1	5.6k // 5.6k	Beyschlag MELF 0204
R2	240R	Susumu 0805 0.5%
R3	1k	Susumu 0805 0.5%
R4	1k	Susumu 0805 0.5%
Rb	100R	Susumu 0805 0.5%
C1	47µF 35V	Nichicon KA
C2	330µ 25V	Nichicon KA
C3	330µ 25V	Nichicon KA
C4	330µ 25V	Nichicon KA

LM337 Cap Multiplier
(Replaces 1x 7915)

IC1	LM337	Onsemi Adj Regulator
Q1	BC807-40	NXP PNP SOT23
Q2	2SA1837	Toshiba PNP TO220
D1, D2	MSE1PJ	Vishay Low Leakage SMD Diode
R1	5.6k // 5.6k	Beyschlag MELF 0204
R2	240R	Susumu 0805 0.5%
R3	1k	Susumu 0805 0.5%
R4	1k	Susumu 0805 0.5%
Rb	100R	Susumu 0805 0.5%
C1	47µF 35V	Nichicon KA
C2	330µ 25V	Nichicon KA
C3	330µ 25V	Nichicon KA
C4	330µ 25V	Nichicon KA

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