Class AB Push-Pull Vacuum Tube Guitar Amplifier Analysis, Design, and Construction

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Abstract—**Analysis of a vintage Class AB Push-Pull audio amplifier is presented. Armed with the understanding gained from this analysis, techniques used by engineers of the past, and modern circuit analysis tools a redesign and improvement to this revered amplifier is produced. Using primitive electronic components such as vacuum tubes, magnetic transformers, and passive components, this improved design is realized and constructed.**

Index Terms—Class-AB, Audio, Electron Tube

I. Introduction

MANY MUSICIANS AGREE that the class AB vacuum tube electric guitar amplifier was perfected in the 1950's with the design of the Fender Bassman 5F6-A. Over the years since its inception, many manufacturers have attempted to improve this circuit, yet the basic layout has been largely unchanged. The purpose of this project is to understand the design of this classic amplifier, assess what differences I would like in a guitar amplifier, and using Spice simulation, attempt to adjust the circuit so as to achieve the desired sonic results, thereby developing a new design.

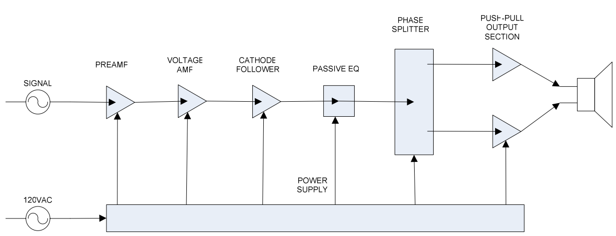


Fig 1. High level block diagram of the 5F6-A

Please Refer to Fig A1 (pg XX) for detailed schematic of the Fender Bassman 5F6-A.

II. Analyisis

A. Ear Analysis

Bud Purvine, a Local magnetics engineer and owner of the Onetics Transformer Company, was kind enough to allow me to come listen to his original Bassman. This amp immediately has a very familiar sound, and it is understandable why this is considered a favorite by many guitarists. It has a very clear and bright sound that ranges from clean to mildly distorted when turned up and played hard. It is labeled to put out 50 watts, has a very sensitive EQ control, and uses four 10" speakers in a closed wooden enclosure.

Based on my impression, I chose the following criteria in modifying the Bassman:

- More volume
- A stronger low end response
- More capability for distortion
- Less noise/ac hum
- More simplistic input network

In designing a modified amplifier, each stage of the circuit would need to be visited.

B. 12AY7 PreAmp

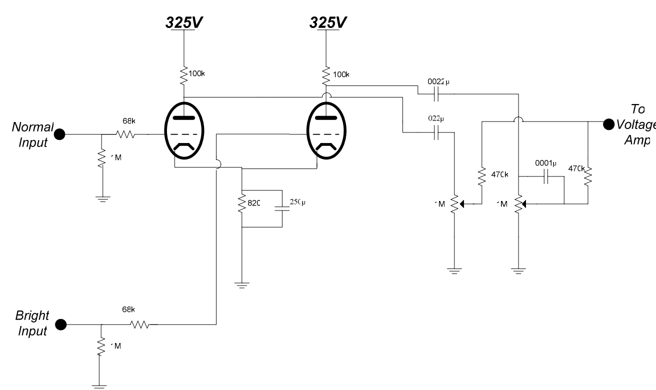


Fig 2. Preamp Circuitry

The Bassman preamp employs the use of two distinct preamp channels. Both inputs see a passive gamma network before interfacing a common cathode voltage amplifying stage using a 12AY7 medium μ triode. Each of these stages is loaded by a passive network which includes a potentiometer tapping a portion of the output voltage signal to ground.

Load line analysis of the plate current characteristic curves of the 12AY7 reveals the operating point and small signal characteristics of each stage. It's worth noting that since bias current from both parallel stages is shared in the cathode resistor(R_K), the *effective* cathode resistance($R_{K,eff}$) of each individual stage is seen as twice the value of R_K[1].

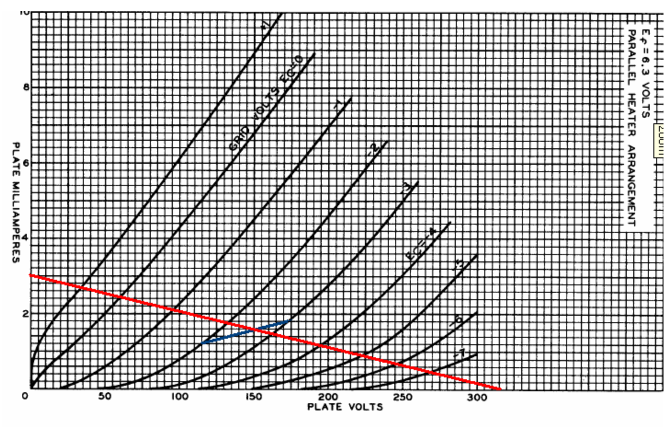


Fig 3. Load line associated with 12AY7 preamp stage

Fig 3 above displays the manufacturer's published typical anode characteristics of the 12AY7. The DC load line in red is composed of the resistive relationship between the HT supply of 325V and the load resistor value of 100kΩ. The blue grid line is composed of the relationship described in (1).

$$I_P = \frac{-V_{gk}}{R_{k,eff}} \tag{1}$$

The choice of the intersection of these lines as a bias point yields small signal parameters as shown in (2) – (4).

$$r_p = \frac{\Delta V_{pk}}{\Delta I_p} \tag{2}$$

$$gm = \frac{\Delta I_p}{\Delta V_{gk}} \tag{3}$$

$$\mu = \frac{\Delta V_{pk}}{\Delta V_{gk}} \tag{4}$$

$$\mu = gm \times r_p \tag{4.1}$$

Mid-band gain(G) is calculated from this circuit using (5) by noting that the triode behaves like a voltage controlled voltage source loaded by a voltage divider between the load resistor and plate resistor.

$$G = \mu \frac{R_L}{R_L + r_p} \tag{5}$$

High frequency response is dictated by the low-pass filter created by the Miller Capacitance (C_M) between the grid and ground, calculated by (6).

$$C_M = C_K + (1 - A)C_P \tag{6}$$

Note that C_K and C_P are the typical parasitic capacitances stated on the 12AY7's technical data sheet (both = 1.3pF). The HF -3dB cutoff point of this filter is found from (7).

$$f = \frac{1}{2\pi R_{GS} C_M} \tag{7}$$

Also note that R_{GS} is the 68kΩ "grid stopping" resistor. The input signal to the amplifier sees a voltage divider between R_{GS} and C_M.

The LF -3dB cutoff of this stage is dictated by the voltage division between the coupling capacitor connecting the12AY7 output voltage to the next stage and the 1M volume control potentiometer (R_V).

$$f = \frac{1}{2\pi R_V C_C} \tag{8}$$

If the bright channel is used, then the LF cutoff is higher in the frequency spectrum due to the smaller coupling capacitor used. In addition, there is a dependent relationship between the .0001μF and volume control that controls some brightness (see Fig 4.). If the volume is completely turned up, the capacitor is bypassed. Otherwise some high frequency signal bypasses the volume resistor through this capacitor.

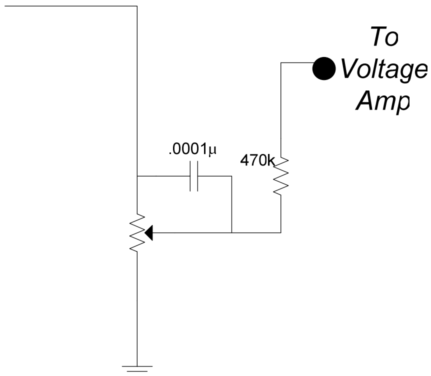


Fig 4. Bright capacitor

The final point of analysis for the preamp stage is headroom. Headroom can be defined as the input voltage amplitude threshold where the output signal becomes a non-linear representation of the input signal. Analyzing the load line in Fig 3, there are two conditions that establish headroom constraints. If the V_{gk} exceeds 0V, then grid current will flow, causing distortion. If V_{gk} become so negative that the non-linear cutoff region is reached, then distortion will also occur. It is worth noting that the cutoff threshold is vaguely defined, and grid current distortion poses a greater threat for non-linear distortion.

Because our quiescent V_{gk} = -2.7V, it is determined that a voltage swinging positively as high as 2.7V will drive the stage into grid current. Swinging the other direction, cutoff will be reached somewhere around -3V, so our headroom will be decided by V_{gk} not exceeding a peak of 2.7V. Therefore maximum input voltage may not exceed 5.4VPP.

Also worth noting is the fact that the AC behavior of the stage does not consider R_K, as the 25µF bypass capacitor shunts the all audio frequencies above the LF cutoff frequency of 4.24 Hz. The next stage does not include this feature.

The analysis of the preamp circuit is summarized in Table 1.

Spec	Value
V_{gk}	-2.7V
I_p	1.65mA
V_p	157V
r_p	29.9kΩ
gm	1.4mS
µ	41.9
G	-32.2
C_M	44.5pF
LF -3dB (norm)	7.23Hz
LF -3dB (bright)	72.34Hz
HF -3dB	52.6kHz
headroom	5.4VPP

Table 1. Preamp Analysis Results

C. 12AX7 Voltage Amp

Following the preamp stage, the amplified signal is fed to another common cathode voltage amplifier, this time using half (one triode) of a 12AX7 high µ tube. This stage uses a non-bypassed cathode resistor, employing negative feedback to the circuit.

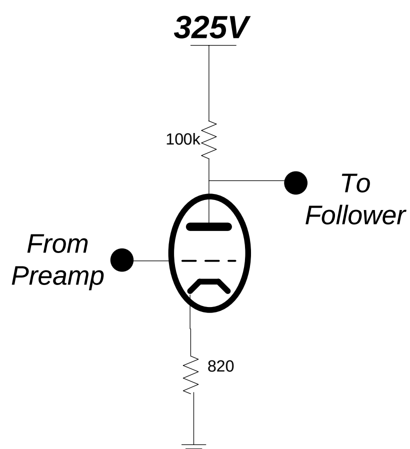


Fig 5. 12AX7 Voltage Amp

Noting the use of HT = 325V and R_L = 100kΩ, a load line can be transposed onto the 12AX7's anode characteristic graph. Using the same techniques as in section *B*, operating point and small signal parameters can be found (see Table 2).

Gain calculation is complicated some by the fact that R_K is not AC bypassed, and therefore becomes involved in the voltage division of equation (9)[1].

$$G = \frac{-\mu R_L}{R_L + r_p + (\mu + 1)R_K} \quad (9)$$

As input signal causes the plate circuit to draw current, a greater voltage is developed across R_K, causing v_{gk} to decrease. This constitutes negative feedback. The amount of negative feedback can be described by the feedback factor, β, which equals R_k/R_L in this circuit.

This circuit contains another HF filter between the 270k grid stopper resistor and the Miller capacitance of the stage, calculated as before.

Headroom for this stage is limited again by the threat of grid current when V_{gk} = 1.19V or greater. This is significant considering the fact that the signal has already been amplified by a gain of 32 by the previous stage. The ability to tap a portion of the pre-amplified signal to ground before the grid of this stage is crucial to the control of distortion as well as the overall volume of the amplifier.

Spec	Value
V_{gk}	-1.19V
I_p	1.43mA
V_p	181V
r_p	59kΩ
gm	1.7mS
μ	100
G	-41.4
C_M HF -3dB headroo m	72pF 8.5kHz 2.4VPP

Table 2. Voltage Amp Analysis Results

D. Cathode Follower and EQ

The next stage in the signal path of the bassman is a cathode follower circuit built around the second triode contained within the 12AX7 tube. This topology is designed to provide a low impedance source to the following equalizer section at a gain of approximately 1. The output impedance of the 12AX7 common cathode voltage amp is approximately r_p in parallel with R_L, equaling about 59kΩ. In order for the signal to react sensitively to the equalizer circuitry, this stage is a beneficial buffer.

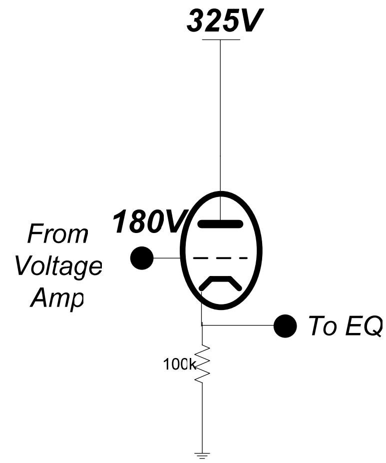


Fig 6. Cathode Follower

Although the load is seen at the cathode of the device, a DC load line can still be used for analysis just as above. Gain for the stage is approximated using (10).

$$G = \frac{R_K}{\frac{1}{gm} + R_K} \quad (10)$$

This gives a non-inverting gain slightly less than one.

Output impedance of the circuit is approximated by (11).

$$R_o = R_K \left\| \frac{1}{gm} \right. \quad (11)$$

This provides 531Ω source impedance to the equalizer circuit to follow.

The headroom of this stage is not an issue, as the large cathode resistance provides a substantial amount of negative feedback, keeping V_{gk} very close to its DC value[1].

Spec	Value
V_{gk}	-.59V
I_p	1.83mA
V_p	142V
r_p	50kΩ
gm	1.85mS
µ	93
G	.984
R_o	531

Table 3. Follower Analysis Results

Analysis of the frequency equalizer section of the amplifier is better left to computer aided analysis as hand calculations yield lengthy and complex expressions for frequency response. Please see appendix (A2-A7) for Spice analysis results.

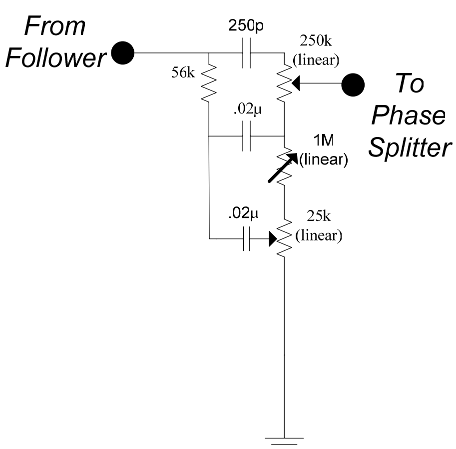


Fig 7. Three Band Equalizer

E. Long Tailed Pair Phase Splitter

In order to drive a push pull output stage of the amplifier, the pre-amplified signal must be split into two identical (more or less) signals 180˚out of phase from one another. The 5F6-A achieves this goal by employing a differential amplifier made up of two triode stages in a 12AX7 vacuum tube.

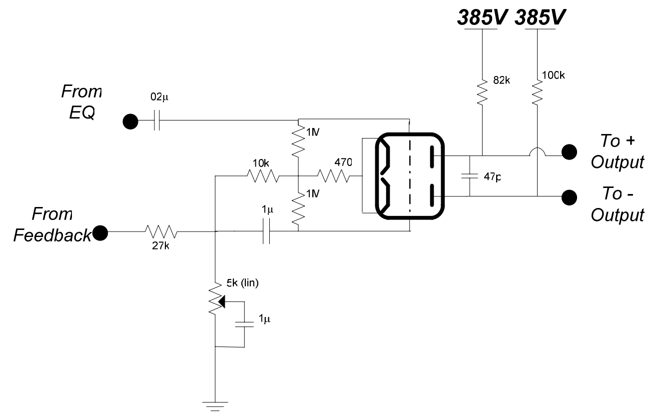


Fig 8. Differential Amplifier Phase Splitter, "Long-Tailed Pair"

In analyzing the DC behavior of this circuit, the following assumptions are made: All capacitors are open circuits, feedback voltage is zero, and the two load resistors are both equal to 100kΩ. Also, each plate circuit shares the cathode resistors, and so their value to each distinct triode is double.

A load line for each triode can be extracted between the supply voltage of 385V and equation (12) [1].

$$I' = \frac{V_p}{R_L + 2(R_K + R_{pot})} \tag{12}$$

$R_k = 470\Omega,\ R_{pot} = 5k\ \Omega$

An intersecting grid line can also be extracted from the relationship $V_{gk} = -2R_kI_p$. Bias point and small signal parameters can be extracted from this line as done above. To determine differential gain of the stage, we assume that the inputs to the different inputs are equal and opposite as in (13).

$$v_{g,left} = -v_{g,right} = \frac{v_{in}}{2} \tag{13}$$

Thus, partial differenial gains can be approximated by (14) and (15).

$$G_{left} = -\frac{gm}{2}(R_L \| r_p) \tag{14}$$

$$G_{right} = \frac{gm}{2}(R_L \| r_p) \tag{15}$$

Note that these gains are equal and opposite.

Common mode gain can be derived by considering the case where $v_{left}=v_{right}=v_{in}$. This gain is approximated in (16)[1].

$$G_{CM} = \frac{-\mu R_L}{R_L + r_p + 2(\mu+1)R_{pot}} \quad (16)$$

In the differential pair used in the 5F6-A, only one input is presented with the input signal as can be seen in Fig X. This means that $v_{in,\,left} = v_{in}$, and $v_{in,\,right} = 0$. Now, assembling the complete gain in terms of differential and common mode components:

$$v_{out,left} = G_{left}(v_{in,left} + v_{in,right}) + G_{CM}\left(\frac{v_{in,left} + v_{in,right}}{2}\right) \quad (16)$$

$$v_{out,right} = G_{right}(v_{in,left} + v_{in,right}) + G_{CM}\left(\frac{v_{in,left} + v_{in,right}}{2}\right) \quad (17)$$

Recalling that the two individual differential gains are opposite in sign, it can be observed that common mode gain contributes to the out-of-phase output, but subtracts from the magnitude of the in-phase output, thereby causing an imbalance in the amplitude of the phase split outputs. The 5F6-A accounts for this by using an 82kΩ load resistor in the left circuit.

Negative feedback is again introduced to the amplifier. This time, output signal from the secondary of the output transformer is voltage divided between a 27kΩ feedback resistor and the 5kΩ presence control potentiometer resistance. At its largest, the feedback factor, β, is the ratio of R_{pot}/R_{fb}, however the presence control plays into this. The voltage gain due to feedback can be found using KVL through the plate loops and feedback loop (see Table 4).

The .1μF capacitor between the finger of this potentiometer and ground controls the amount of mid to high frequency signal that is negatively fed back to the phase splitter tail. Frequencies affected by this control are above about 318Hz as calculated by (17).

$$f = \frac{1}{2\pi R_{pot} C_{shunt}} \quad (17)$$

The input impedance of this circuit is affected by the amount of negative feedback. The more high frequency content shunted from the feedback loop to ground by the .1μF cap, the lower the input impedance becomes. With zero high frequency shunting (minimum presence), the input impedance is at its maximum of about 2.3MΩ. At maximum presence, much of the high frequency feedback signal is shunted to ground, reducing the amount of linearization of these high frequencies as well as degrading the input impedance to about 1.9MΩ.

Using KVL techniques, it can be shown that the limiting factor for headroom in the phase splitter is the second triode reaching grid current, which corresponds to v_{in} = -2.61V.

Spec	**Value**
V_{gk}	-1.36V
I_p	1.44mA
V_p	199V
r_p	57.7kΩ
gm	1.75mS
μ	101
$G_{,left}$	-25
$G_{,right}$	26.6
$G_{,cm}$	-4.4
$G_{fb,\,left}$	3.9
$G_{fb,\,right}$	-4.14
β	.185
headroom	5.22VPP

Table 4. Phase Splitter Analysis Results

F. Push Pull Output Amplifier

The final chain in the signal path is the output amplifier. This stage is designed to use a pair of 5881 pentode in a push-pull topology. While one tube conducts, the other tube is in cutoff and visa versa, hence the moniker, “push-pull”. However, conduction periods of the two tubes overlap to a degree, thus operation is referred to as Class-AB. Class-A would be constant conduction by both tubes, whereas Class-B would show one tube “pushing” while the other is completely cutoff and visa versa. Class-AB yields some of the efficiency benefits of Class-B operation, while avoiding crossover distortion.

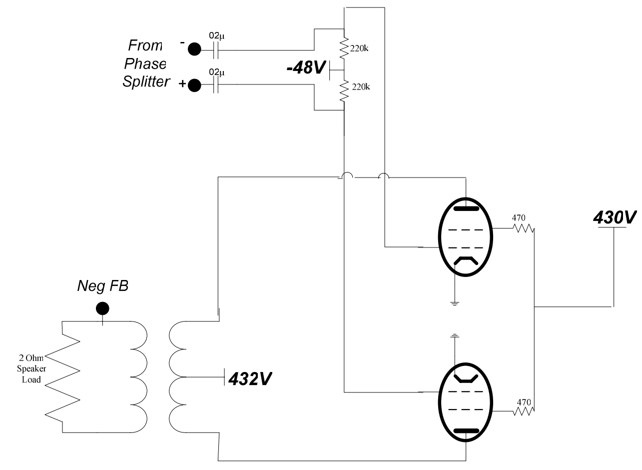


Fig 8. Push Pull Output Amplifier

Analysis of these tubes is complicated by their unique physical construction. This is specifically the addition of a fourth electrode, the screen grid. Whereas a triode's plate current is described by grid voltage and plate voltage as in (18),

$$I_P = K(V_G + \frac{V_P}{\mu})^{3/2} \tag{18}$$

the pentode's plate current is rather described as a function of grid voltage and *screen* voltage, V_S.

$$I_P + I_S = K(V_G + \frac{V_S}{\mu_s})^{3/2} \tag{19}$$

Note that K, μ_s, and μ are both factors describing the physical nature of the specific tube model, and I_P+I_S is the total space current through the pentode. Equation 19 alludes to the independence of current on plate voltage, and therefore a large plate resistance. This can be confirmed by the 5881 anode characteristics displayed in Fig 9.

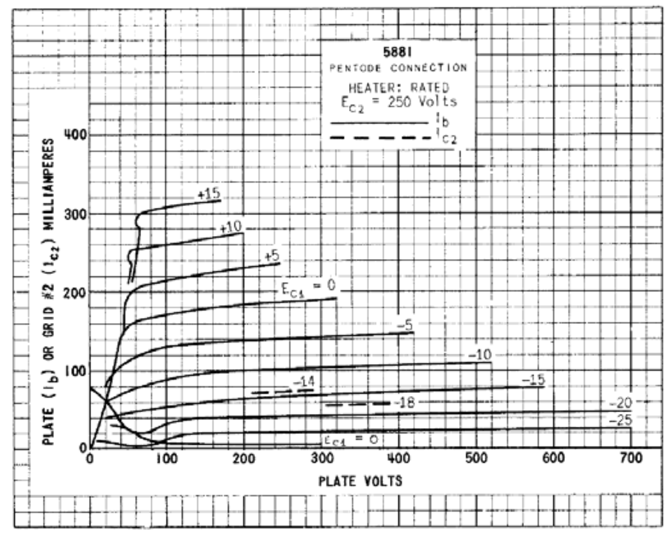


Fig 9. 5881 Anode Characteristics

Note that this figure shows that at V_{GK} less than -50V, the tube is in cutoff and almost no current flows. Also, note that this chart assumes V_S = 250V, as opposed to the 430V used in the 5F6-A, accounting for some variation. Noting the relationship in (19), idle plate current can be calculated (see Table 5). This amplifier is considered "fixed biased" as the V_{gk} of these tubes is set by a fixed supply voltage of -48V as opposed to a cathode biased configuration as we've seen previously.

The output transformer used in the 5F6-A uses a 4050Ω primary. Each tube is in parallel with half of the turns in the primary. Because the impedance varies with the square of the turns ratio, $R_p = 1/2^2(4050) = 1013\Omega$.

Traditional load line analysis isn't as revealing as in single ended triode stage analysis, the reason being that Fig 9 only describes one tube. A composite anode characteristic can be composed as in Fig 10, which shows the conduction of both tubes. Note that this graph theoretically describes net current from the perspective of one of the tubes, V_{gk} ranging from 0V to -96V.

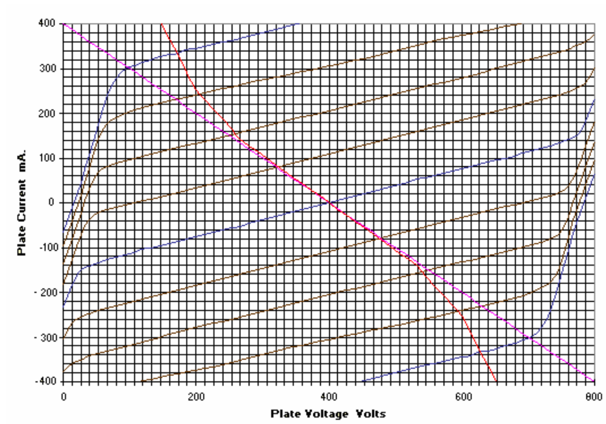


Fig 10. Composite Loadline

In the case of the 5F6-A, the grid line intersecting with the purple load-line would be the -48V grid voltage, as this is where both valves are at idle and zero current flows.

Plate resistance can be extracted from the characteristic curves to be about 8.2kΩ. Noting the parallel arrangement of the two tubes, it makes sense that the output transformer used is fixed with a primary impedance of 4.05kΩ. Output impedance of the amplifier is estimated by the output resistance reflected through the transformer as in (20).

$$R_o = r_p \frac{R_L}{R_p} \tag{20}$$

At maximum power, V_{gk}=2(48) =96VPP. The load line reveals that at this peak input, output voltage,$V_{o,max}$, is around 300V. Therefore average power is defined as in (21).

$$P_{avg} = \frac{(V_{o,\max} / \sqrt{2})^2}{R_P} \tag{21}$$

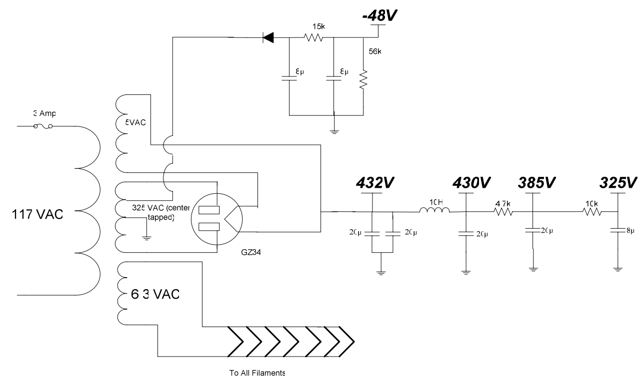
At maximum power, the pentodes contribute 3rd harmonic distortion. This is reined in using the negative feedback loop from the output transformer to the cathode circuit of the phase inverter.

Spec	Value
V_{gk}	-48V
I_p	33mA
I_s	0.6mA
V_p	432V
V_s	430
r_p	8.2kΩ
$R_{primary,per\ tube}$	1013Ω
R_L	2Ω
R_o	17.3Ω
P_{avg}	44watts,RMS
Headroom	96VPP

Table 5. Power Amplifier Analysis Results

G. *Power Supply*

The 5F6-A power supply topology is shown in Fig 11. 117VAC is translated to two AC lines of 325V sharing a common ground. These voltages use the GZ34 tube dual diode as a full wave rectifier entering an input capacitance followed by several low pass filter stages.



An important sonic characteristic of the power supply is the amount of voltage “sag” experienced by the power tube screens upon maximum load conditions due to a large signal. Sag is the result of the tube diode’s internal resistance as is displayed by Fig 12. Sag amount and the time it takes to achieve this sag translate to how “compressed” the sound of the amplifier is.

The Phillips datasheet for the GZ34 rectifier displays the amount of sag to be expected in power supply voltages

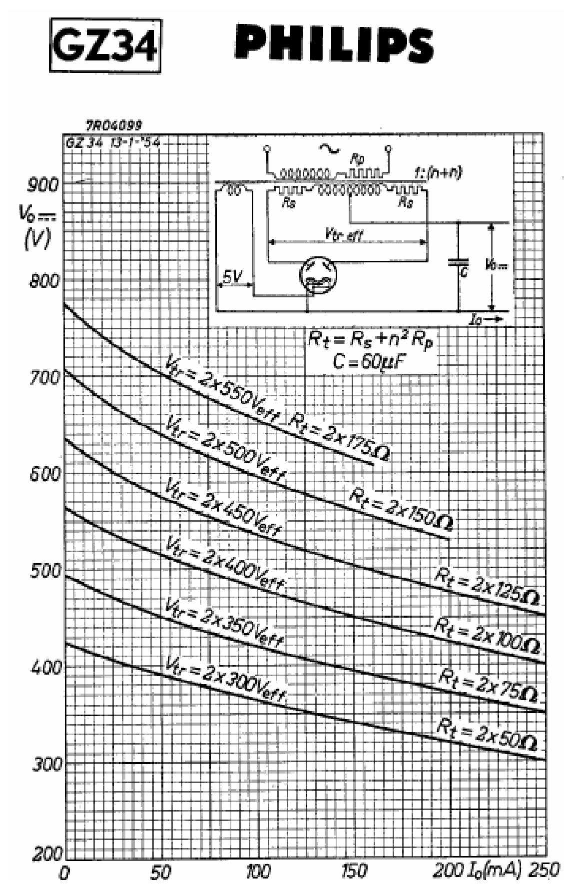


Fig 12. GZ34 Supply Voltage vs. Load Current

The amount of Total load currents of the amplifier at idle are equated to 77mA as in (21).

$$I_{total} = I_{plate} + I_{screen} + I_{splitter} + I_{12AX7} + I_{12AY7}$$
(21)

Using an interpolated curve between 2 x 350V and 2 x 300V, a load current of 77mA is shown to drop peak voltage 50V from 482V to 432V, which is the case for the 5F6-A.

Maximum input signal can be considered a 96VPP signal at the grid of the pentodes. Using the pentode current equations averaged over 360˚ of phase as in (23) and (24)˚,

$$I_{\max} = I_{plate,\max} + I_{screen,\max} + I_{triodes} \quad (22)$$

$$I_{plates,\max} = \frac{1}{360}\sum_{\theta=1}^{360}\left[I_s(\theta) + I_s(\theta+180)\right] \quad (23)$$

$$I_{screens,\max} = \frac{1}{360}\sum_{\theta=1}^{360}\left[I_s(\theta) + I_s(\theta+180)\right] \quad (24)$$

These formulas give a maximum signal average plate current of 160mA and a maximum signal average screen current of 17mA. The additional currents through the triodes are dwarfed by the pentode currents, and so are maintained at their bias sum of about 10mA, leading to a total load current sum of 187mA.

Using Fig 12, total voltage sag is found to be an additional 55V. If bias current is used as a zero reference, the additional current load from maximum power signal is 187 – 77 = 110mA. If the triodes are excluded, the entire supply rectification characteristics are approximated by R_o, and the choke is considered a short, an estimation of sag delay can be found using the model in Fig 13[1][6].

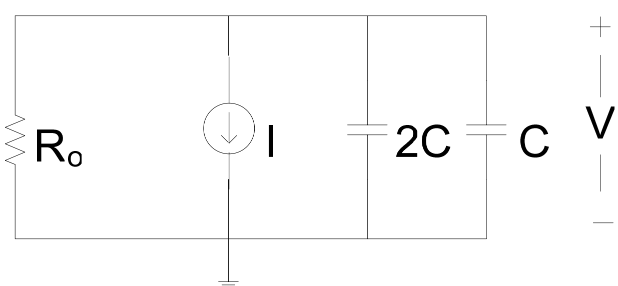


Fig 13. Approximated Model of Voltage Sag

In Fig 13, C is the 20μF capacitance, I is the additional current load of 110mA, and R_o is the rectifier output impedance of 500Ω found as in (25)[6].

$$R_o = \frac{V_{sag}}{I_{\max Load}} \quad (25)$$

Fig 13 yields the relationship (26), which can be graphed as in Fig 14, displaying about a millisecond of delay before the power supply has reacted to the additional load of a max power signal.

$$v(t) = -IR_o\left[1 - e^{\left(\frac{-t}{3R_oC}\right)}\right] \quad (26)$$

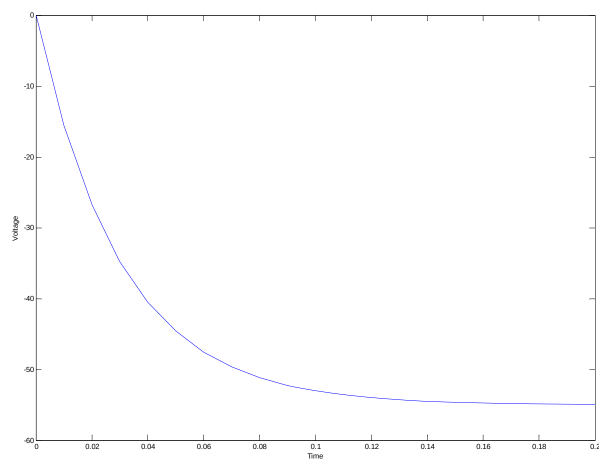


Fig 14. Approximated Voltage Sag vs. Time

Another important characteristic of the power supply is its ability to filter voltage ripple. The first filter applied to the plate supply is largely unimportant as far as ripple is concerned because the push pull behavior of the stage cancels any common ripple voltage between the two tubes.

Analyzing the ripple attenuation applied to the screen supply, consider the voltage division caused by the 20μF capacitor and the 10H inductor at the 120Hz rectified supply.

$$\frac{v_{screen}}{v_{plate}} = \frac{\left(\frac{1}{Cs}\right)}{Ls + \left(\frac{1}{Cs}\right)} = -41dB$$
(27)

At the phase splitter, voltage division between the 4.7kΩ resistor and another capacitor gives an additional -37dB of ripple attenuation as shown in (28).

$$\frac{v_{ps}}{v_{screen}} = \frac{\left(\frac{1}{Cs}\right)}{R + \left(\frac{1}{Cs}\right)} \tag{28}$$

In the same way, ripple is attenuated an additional -36dB at the first two triodes. This amounts to a total of over -114dB of ripple attenuation before reaching the sensitive 12AY7 circuit.

H. Amplifier Headroom

Distortion being such an important part of the 5F6-A's sound, it's important to understand what stage(s) of the amplifier distort first, and how the controls play into this balance.

An input signal on the brink of distorting the 12AY7 at 5.4VPP would be amplified by the gain of 32, attenuated slightly by the voltage division between preamp output impedance and 12AX7 input impedance, and would certainly distort the 12AX7 input.

If the volume control were used to back off the input to the 12AX7 voltage amp to the brink of distortion, this stage would amplify by a gain of 41, then the signal would buffer through the cathode follower and assuming no attenuation at the equalizer, the phase splitter's 5.44VPP headroom would be far breached.

If the equalizers controls were backed off to the point where the phase splitter was linear, then 5.44VPP would be amplified by a gain of about 25, producing a 125V signal at the output. This signal would breach the 96VPP headroom threshold of the pentodes.

This interplay displays how much of the 5F6-A's distortion characteristic comes from the pentodes. If these pentode's are generally driven into maximum power at V_{gk} > 48V, then power supply sag is an issue that is a part of this amplifier's normal operation.

APPENDIX

A. Schematics

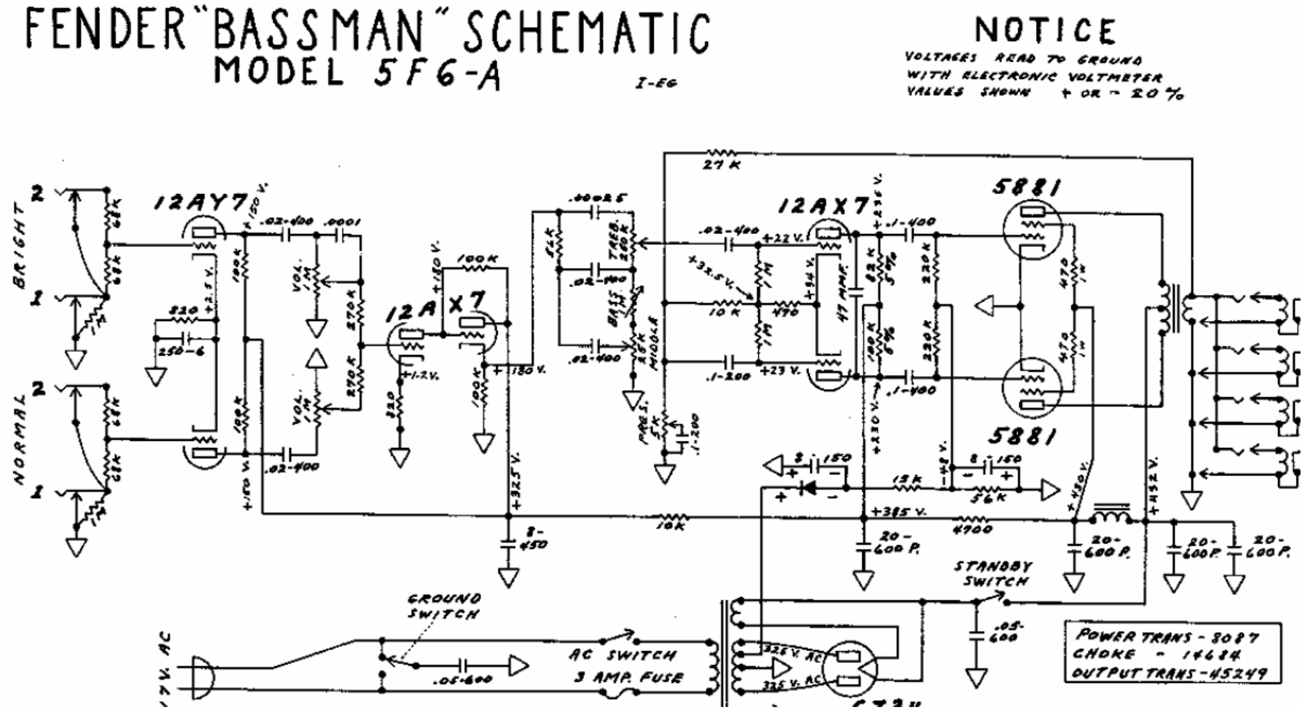


Fig A1. Schematic for Bassman 5F6-A

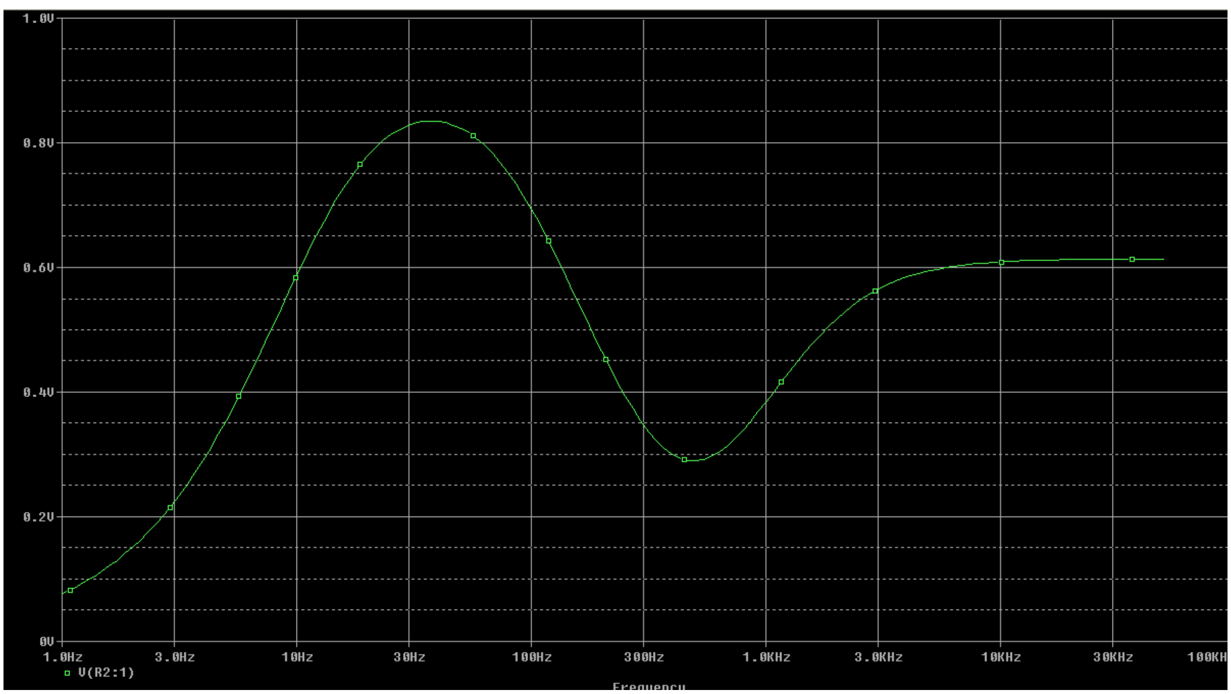


Fig A2. Flat EQ AC Response: Treble = 5, Bass = 5, Mid = 5

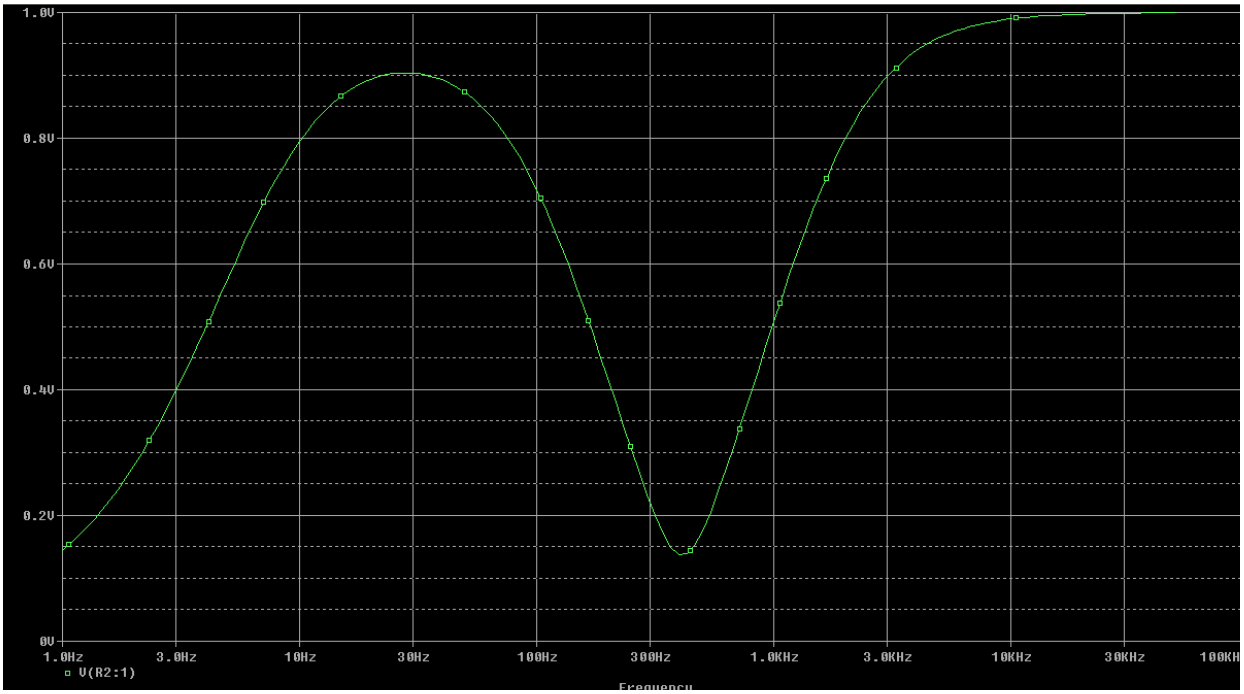


Fig A3. "Scooped"EQ AC Response: Treble = 10, Bass = 10, Mid = 0

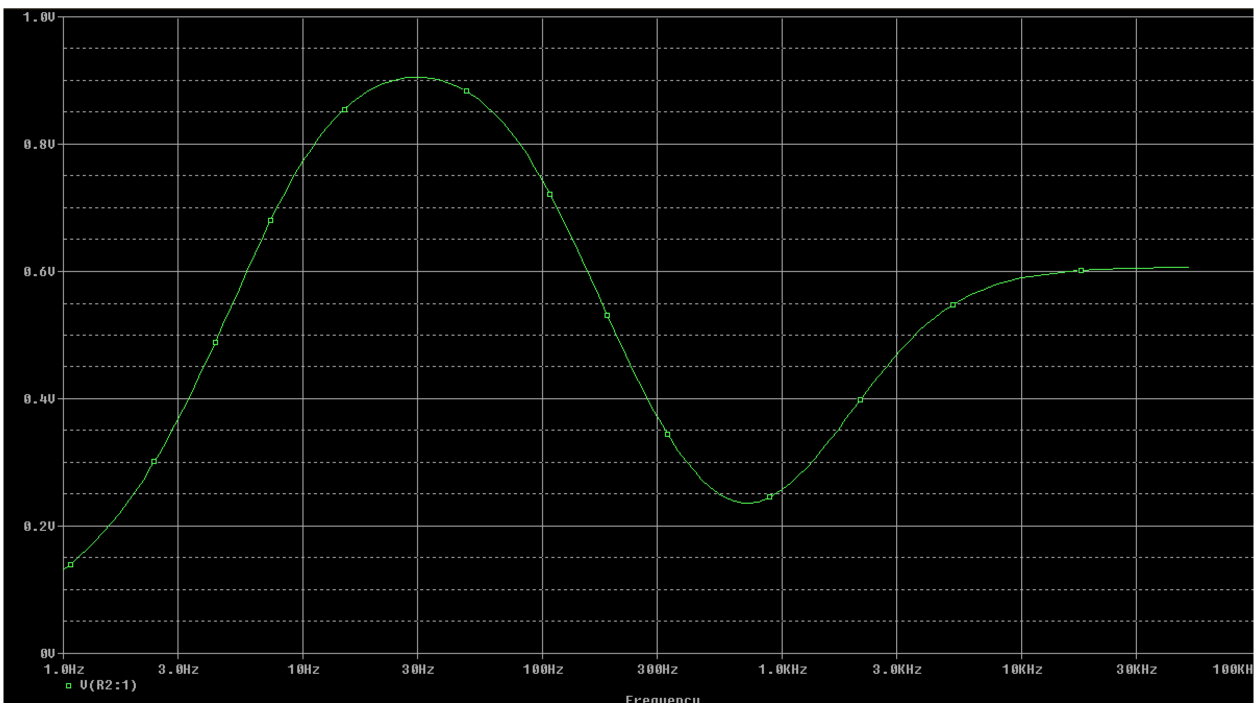


Fig A4. Bass Boosted EQ AC Response: Treble = 5, Bass = 10, Mid = 5

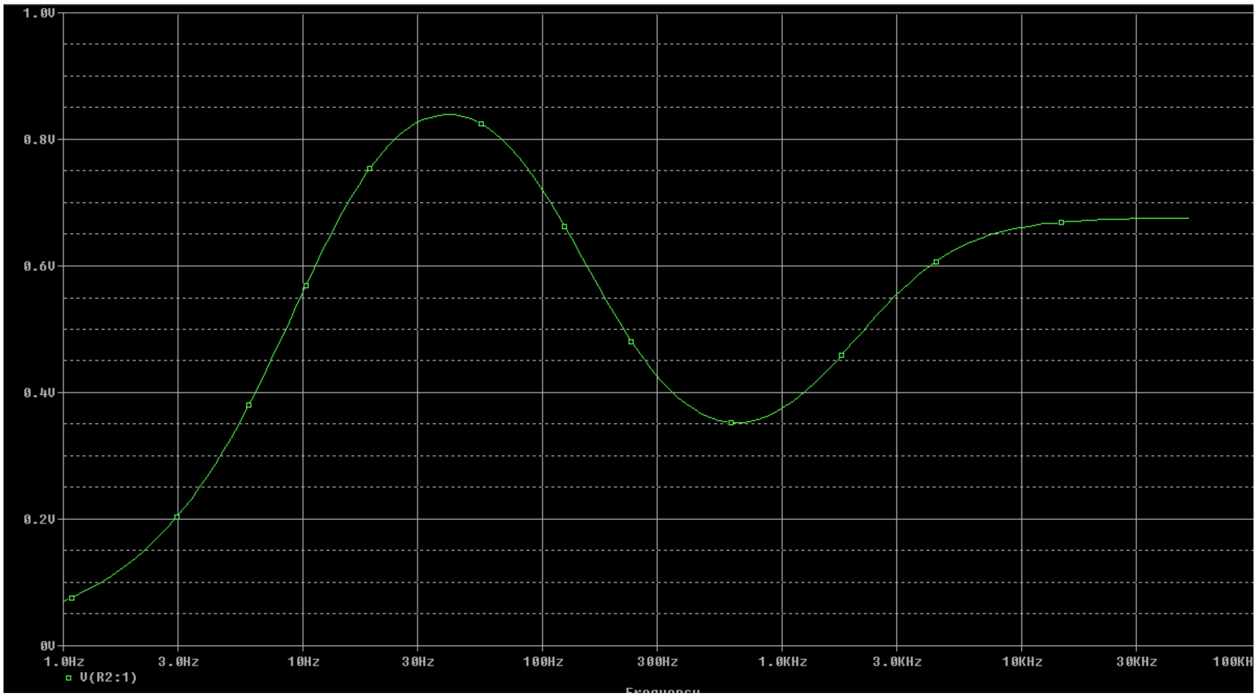


Fig A5. Mid Boosted EQ AC Response: Treble = 5, Bass = 5, Mid = 10

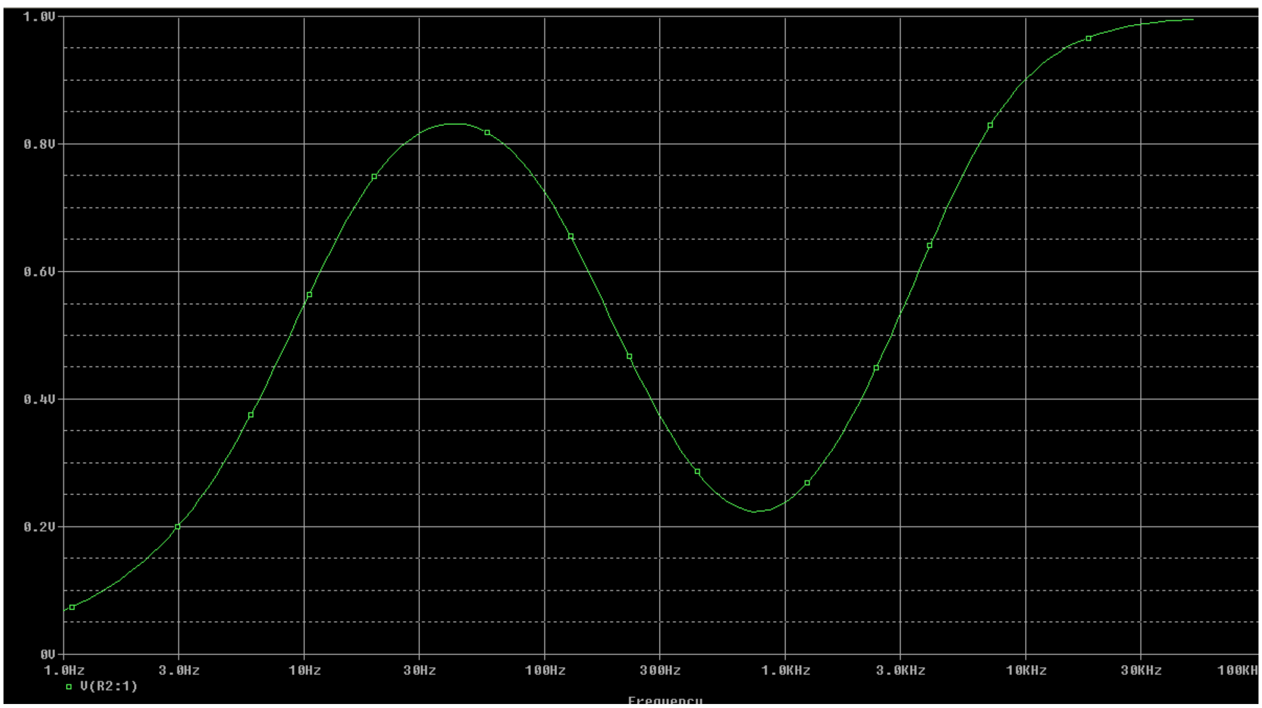


Fig A6. Treble Boosted EQ AC Response: Treble = 10, Bass = 5, Mid = 5

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